metal enclosure. The regulator IC is mounted directly on the pc board, and it does not require a heat sink.

After the pilot lamp, the power switch, and the binding posts are installed on the front panel, T1 and T2 can be bolted in place near the front of the box. The transformer primaries can be tied in parallel for operation from 117 volts ac, or in series for 235-volt ac operation. The T1 and T2 secondaries must be connected in series and in proper phase for the power supply to operate correctly. If the unloaded ac output voltage as measured with a VOM is in the neighborhood of 20 volts, the windings are connected properly. If, however, the VOM reads approximately 6 volts, the secondaries are out of phase and the leads from one of the transformers must be reversed. If the primary leads are brought out to four separate terminal posts, changing from 117-volt to 235-volt operation will be a simple matter of changing appropriate jumpers. Alternatively, a 117/235 switch may be installed easily on the rear panel if frequent line voltage changes are anticipated. In either case attention should be paid to the matter of proper phasing of the windings. The use of a 3-wire ac cord installed in a properly grounded outlet is intelligent practice for this and any line-operated power supply. If a transformer with a secondary rating of approximately 18 volts at 3 amperes is available, it may be used in place of T1 and T2. Details for the modification of a 24-volt secondary transformer are given in *QST* for January, 1975.

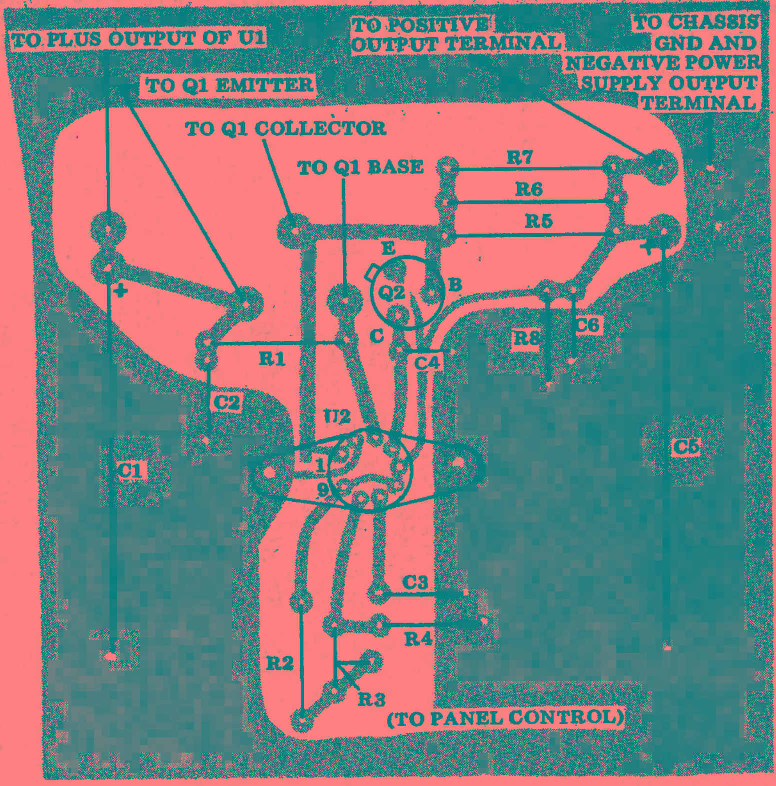


Fig. 2 — Foil pattern end parts layout for the regulated power supply.

A UNIVERSAL POWER SUPPLY FOR THE AMATEUR STATION

Presented here is a general-purpose unit with provisions for 117-220-volt operation, and it is adapted easily for use with most commercially available gear by constructing appropriate power cords. The supply delivers 800 V at 300 mA dc, 300 V at 175 mA dc, and 0 to -130 V at 25 mA. In addition the supply provides ac filament potentials of 6.3 V at 11 A or 12.6 V at 5.5 A.

Often the station power supply is a heavy black box that is tucked away in a corner and just sits there. A large cable interconnects this device with the station transmitter or transceiver and the amateur never comes directly in contact with it; all of the supply functions are remotely controlled

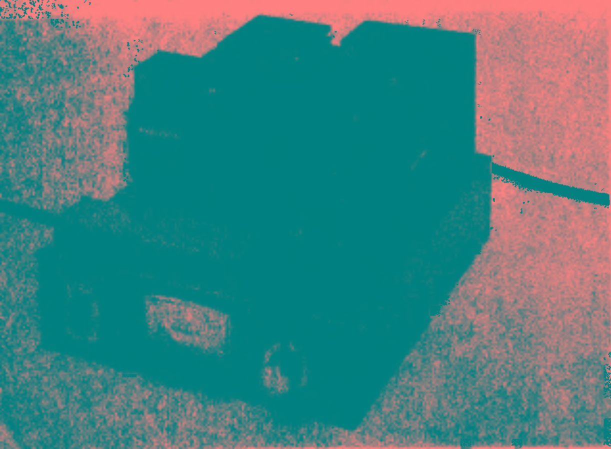


Fig. 1 — The Universal Power Supply is constructed on e standard-size aluminum chassis. Back-to-back plugs with appropriate jumper wires make changing from 117-V to 220-V input operation or from 6.3-V to 12.6-V filament operation a simple metter of reversing e plug.

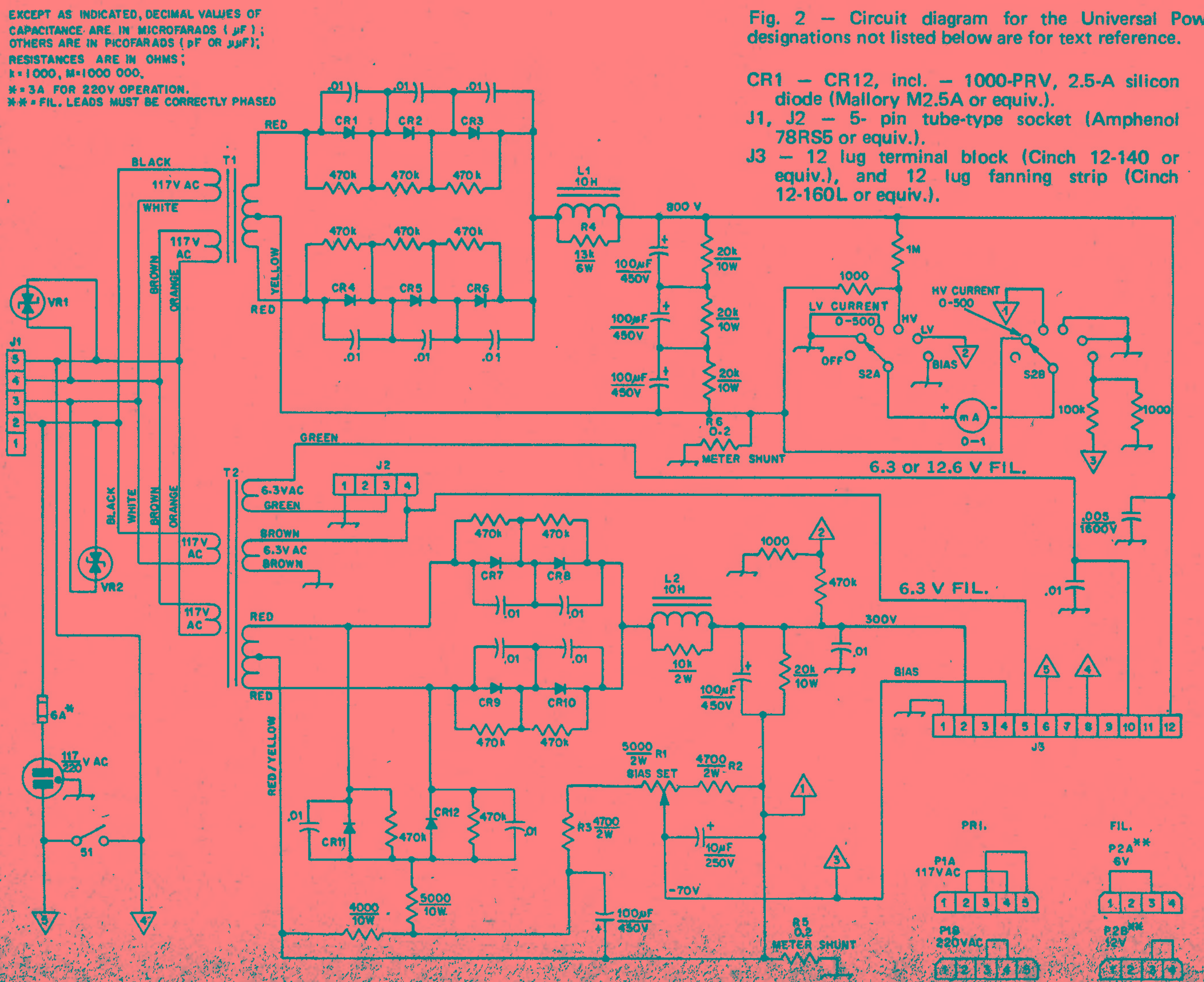


Fig. 2 — Circuit diagram for the Universal Power Supply. Component designations not listed below are for text reference.

CR1 — CR12, incl. — 1000-PRV, 2.5-A silicon diode (Mallory M2.5A or equiv.).
J1, J2 — 5- pin tube-type socket (Amphenol 78RS5 or equiv.).
J3 — 12 lug terminal block (Cinch 12-140 or equiv.), and 12 lug fanning strip (Cinch 12-160L or equiv.).
L1 — 10 H, 200 mA (Hammond 193J).
L2 — 10 H, 300 mA (Hammond 193M).
P1, P2 — 5-pin plugs to mate J1 and J2, 4 req'd (Amphenol 86-PM5 or equiv.).
R1 — 5-watt linear-taper control.
R2, R3 — For text reference.
R4 — Three 39,000-ohm 2-watt resistors connected in parallel.
R5, R6 — See text.
S1 — Spst toggle rated at 6 A or greater.
S2 — 2-pole 6-position rotary, nonshorting (Centralab 1411 or equiv.).
T1 — Dual primary, 117 or 220 V ac; secondary 890 volts each side of center tap at 300 mA (Hammond type 101059).
T2 — Dual primary, 117 or 220 V ac; secondary 350 volts each side of center tap at 175 mA, 6.3 volts ac at 6 A, 6.3 volts ac at 5 A (Hammond spacial 273BX).
VR1, VR2 — Thyrector assembly (G E 6RS20SP4B4).

Fig. 3 — Bottom view of the Universal Power Supply.

from the panel of the station transmitting gear. But what happens if an instance arises where a particular voltage (or combination of voltages) is needed for an experimental project? Can that "black box" in the corner be pressed readily into service? And what about the amateur who buys two power supplies for his station because his mobile transceiver cannot be plugged directly into his home-station transmitter power supply? This supply is designed to fill all these needs.

Many of today's commercially available ac supplies are not equipped for 220-volt operation. If the station includes a two-kilowatt amplifier, a separate 220-volt line should be available in the shack. Blinking house lights are not always a result of running a high-powered amplifier. It could be caused by the intermittent 400- or 500-watt load presented by an exciter power supply to the 117-volt source. Connecting the exciter supply to a 220-volt outlet (providing a dual-primary transformer is used) can be helpful in this regard.

Circuit Details

The supply is shown in Figs. 1 through 3. Primary power may be applied to the supply in two ways. First, terminals 6 and 8 of J3 may be shorted together; this is normally the function of the station transmitting equipment on-off switch (see Fig. 2). On the other hand, S1 may be actuated when the supply is used independently. Transient voltages on the ac line are eliminated by Thyrector assemblies VR1 and VR2.

Full-wave rectification is employed in the secondary circuit of each power transformer to develop the three dc operating voltages. Choke-input filtering provides adequate regulation of both the 300- and 800-volt outputs. Both L1 and L2 are shunted with suitable resistors to reduce the possibility of diode damage when primary power to the supply is removed.

The bias voltage is adjustable and may be set to any value between −40 and −80. Should a range between −80 and −130 volts be required, R1 may be interchanged with R3. Likewise, if a range from 0 to −40 volts is needed, R1 may be swapped with R2.

Metering

A six-position switch and a 0-1-mA meter allows monitoring of high and low voltages, the current for each of these, and the bias voltage. The sixth position permits the meter to be disabled. The meter shunts for both current positions of S2 are homemade and provide a full-scale reading of 500 mA on each range. The proper resistance for the shunts is determined by dividing the meter internal resistance (approximately 100 ohms in this case) by 500, and is equal to 0.2 ohm. No. 30 enameled copper wire has a resistance of 105 ohms per 1000 feet, or 0.105 ohm per foot. Extending the division another step, one inch of wire has a resistance of .008 ohm. Approximately 23 inches of wire provided the correct value for the shunts. Each 23-inch length of wire is wound on a 100,000-ohm, two-watt composition resistor which serves as a form.

Construction

The supply is built on a 10 × 8 × 3-inch aluminum chassis. The spot welds at the four corners are reinforced with No. 6 hardware since the transformers are quite heavy. The total weight of the completed supply is slightly more than 40 pounds. Several one-inch-diameter holes are cut in the chassis bottom plate to allow adequate air circulation.

All of the power-supply output voltages are present on a 12-connection terminal block. The end of the cable used to interconnect the supply to the station transceiver is equipped with a 12-lug fanning strip, providing a convenient means to disconnect it.

One special wiring precaution is necessary; the bleeder resistors for both the high and low-voltage circuits should be mounted in the clear to allow plenty of air circulation around them. Perforated aluminum stock is placed over a 1 × 3-inch cut in the chassis which is directly above the mounting position for the 800-volt bleeder network.

Operation

Two jumper plugs are mounted "back-to-back", making the change from 117-volt operation to 220 volts a simple matter of reversing P1. P2 performs an identical function to select 6 or 12 volts for the filament line.

The cost for this project should be under $100, even if all of the parts are purchased new. The price of the two power transformers and two filter chokes comprises approximately 60 percent of the total cost.[1]

[1] A package including the two power transformers and the two filter chokes is available from Hammond Manufacturing Company, Inc., 1051 Clinton Street, Buffalo, NY 14240, for approximately $60. In Canada, the address is Hammond Mfg. Co., Ltd., 394 Edinburgh Rd., North Guelph, Ontario. Catalog available.

A 3000-VOLT POWER SUPPLY

This high-voltage power supply may be used with linear amplifiers that are capable of operating at maximum legal input power levels. It was designed for use with a one-kilowatt 3-500Z amplifier, but with minor modifications to the control circuitry to suit individual circumstances it can be used with amplifiers having a pair of 3-500Z tubes, a single 3-1000Z, 4-1000A, or any tube or tubes calling for 2500 to 3000 volts at up to 700 mA. Examples of such amplifiers may be found in Chapter 6.

Fig. 1 — Top chassis view of the 3000-volt power supply as constructed by WA1JZC. The circuit board in the foreground holds the bleeder resistors, which are spaced apart and supported a short distance above the board for proper cooling. The large transformer is for the high-voltage supply, and the small transformer provides filament power for the amplifier.

The Circuit

A voltage-doubler circuit connected to the secondary of T1 provides approximately 3000 volts dc. See Fig. 3. The primary of T1 can be operated from either a 117-volt line or a 220-volt source; the latter voltage is preferred. VR1 and VR2 are suppressors included to prevent transients from damaging the high-voltage capacitor bank or the rectifier diodes. Since T1 has two 117-volt primary windings, a suppressor is connected across each. The windings and suppressors are connected in parallel for 117-volt operation, and they are series connected for a 220-volt line.

A relay (K1) is necessary to switch the high-current inrush when the supply is activated. Ordinary toggle switches cannot be used to activate the power supply directly. Surge protection is accomplished by placing R1 in series with one lead of the ac line. K2B shorts out this resistor a few seconds after the main power switch (S1, located on the amplifier front panel) is actuated. A separate line cord for the power supply allows this section to be operated on 220 volts while permitting other circuits in the amplifier to operate on 117 volts. The 120 volts needed to energize the coil of K2 are taken from a half-wave rectified dc supply located on the amplifier chassis. Note that the B-minus terminal is held a few volts above ground by the 15-ohm, 2-watt resistor, for metering purposes in the companion amplifier.

Construction

The power supply is built on a standard 10 × 12 × 3-inch aluminum chassis. Construction is straightforward, as can be seen from Figs. 1 and 2. The front and rear panels are made from 9 × 10-inch pieces of 1/16-inch thick aluminum, and the bottom plate and the U-shaped top cover are made out of perforated aluminum stock.

The primary and control-circuit components, as well as the rectifier board and capacitor bank, are mounted underneath the chassis. Reasonable care must be taken to prevent any part of the primary or control wiring from coming into contact with the high-voltage components. Each of the 100-μF capacitors in the capacitor bank is shunted by a 25,000-ohm, 20-watt wirewound resistor. These resistors equalize the voltage drops across the series-connected capacitors, and also serve as the bleeder resistance. Since these resistors get quite hot during normal operation, they are mounted away from the electrolytic capacitors on a separate circuit board above the chassis, to allow for adequate ventilation. The other large heat-generating components are the power and filament

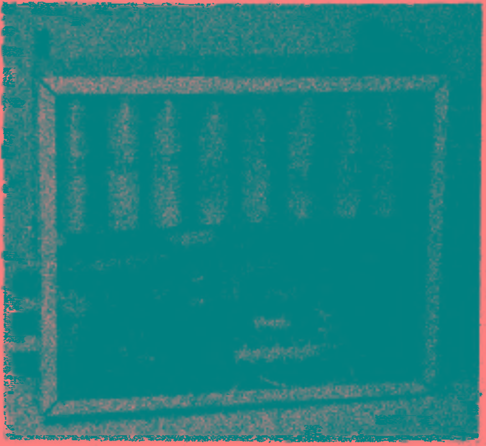


Fig. 2 — The primary and control-circuit components are grouped at the bottom, with the high-voltage capacitor bank and rectifier board occupying the upper portion of this bottom chassis view of the power supply. R1 is visible in the lower right-hand corner.

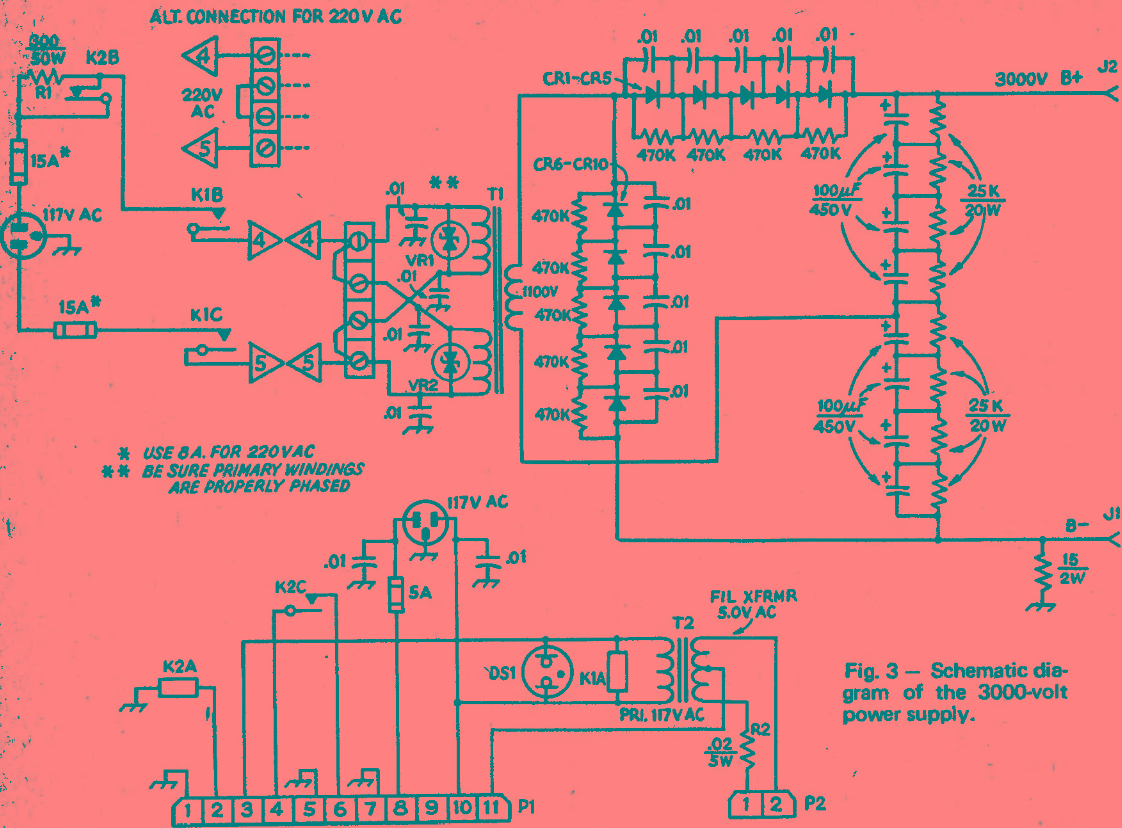


Fig. 3 — Schematic diagram of the 3000-volt power supply.

CR1 — CR10, inc. — 1000-PRV, 2.5-A (Mallory M2.5A or equiv.).
DS1 — 117-volt ac neon pilot lamp assembly.
J1, J2 — High-voltage chassis connector (Millen 37001). K1 — Power relay, dpdt, 117-volt coil (Potter and Brumfeld PR-11AY or equiv.).
K2 — Dpdt 10 A contacts, 120-V dc coil (Potter and Brumfeld KA11DG or equiv.).
P1 — Cable-mounted 11-pin power connector.
P2 — Cable-mounted 2-pin power connector.
R2 — 8 feet No. 14 enam. wire wound on 3-inch long, 3/4-inch die Plexiglas rod.
T1 — Dual 117-volt primary, 1100-V secondary, 600 VA (Berkshire 6181 or equiv.).
T2 — 117-volt primary; secondary 5.0 volts at 15 A (Stancor P6433 or equiv.).
VR1, VR2 — Transient-voltage suppressor, 120-volt (General Electric 6RS20SP4B4 or equiv.).

transformers (T1 and T2), which are also mounted above chassis.

A small etched circuit board supports CR1 through CR10 and their associated equalizing resistors and transient-suppressing disk capacitors. In actual operation, the filament voltage measured at the amplifier tube socket exceeded the maximum voltage recommended by the tube manufacturer slightly, so R2 was included to reduce the voltage to a suitable value. To avoid excessive voltage drop in the cable connecting T2 with the amplifier, it is recommended that the cable be made of No. 10 wire or larger (in many cases, R2 will not be necessary).

NICKEL-CADMIUM BATTERY CHARGER

Any advantage that a NiCad (nickel-cadmium) battery may have over other types can be lost through improper charging. This information concerning NiCad charging techniques was contributed by WAØUZO. NiCads can even be ruined on the *first* recharging cycle. If connected to a constant-voltage source, initial current may be quite high. Normally, no damage would result unless the battery voltage is low (fully discharged). Using a constant current for battery charging is permissible at the start of the charging cycle, however, as the battery reaches full charge, the voltage may rise to an excessive value.

The correct solution is a combination of the two methods. Any circuit used for charging NiCads should limit both the current and voltage, such as the one described here.

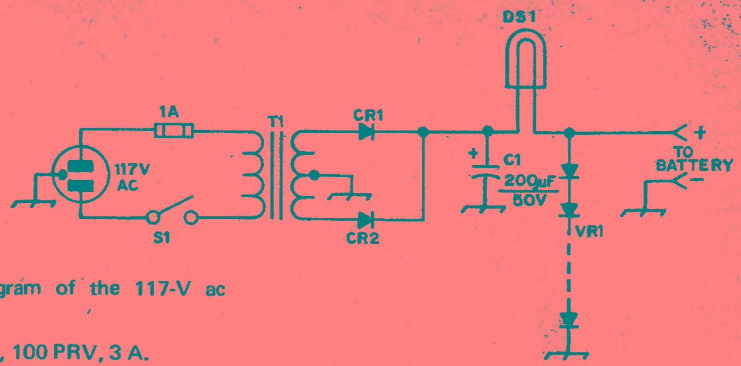


Fig. 1 — Schematic diagram of the 117-V ac charger.
C1 — Electrolytic.
CR1, CR2 — Silicon diodes, 100 PRV, 3 A.
DS1 — See text.
T1 — Primary 117 V ac, secondary 25.6 V at 500 mA. Calactro D1-752 (or equiv.).
VR1 — See text.

Some other precautions which should be observed while charging NiCads are:

1) Battery temperature should be between 40°F and 80°F. It should never exceed 100°F.

2) Two or more batteries with the same voltage rating may be charged in parallel, but be sure that the charger has sufficient current capability.

3) Check the manufacturer's data sheet for the maximum allowable charging rate. A typical figure would be ten percent of the ampere-hour rating (a 10-ampere-hour battery would require a current of 1A).

4) Do not attempt to charge two batteries in series with a constant current unless the batteries are of the same type and capacity, and are in the same state of charge (voltage on one may be excessive).

5) To determine the approximate charging time, divide the ampere-hour rating by the charging current used, and multiply the resulting time by 1.25.

Suitable Charging Circuits

Figs. 1 and 2 show two versions of the same basic charging circuit. The circuit shown in Fig. 1 is used with 117 V ac, and the one in Fig. 2 can be used with the car battery. The latter circuit could be connected to the cigarette lighter, and is suitable for battery packs of up to 14 volts.

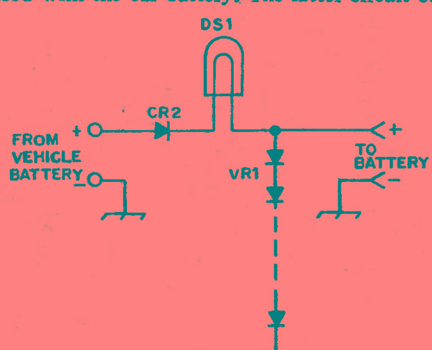


Fig. 2 — Schematic diagram of NiCad battary charger suitable for mobile use. See text for explanation of DS1 and VR1. CR2 protects the components in tha event of accidental reversal of input laads. See Fig. 1 for CR2.

The dial lamp (DS1) is used to limit the current. One with a rating of 100 to 150 mA should work fine with most batteries. The voltage rating should be approximately that of the charging source (for example, two 12-V bulbs in series may be necessary if a 26-V supply is used).

The voltage regulator shown in Fig. 3 is based on the fact that a forward-biased diode will not conduct until approximately 0.75 V dc is applied. By adding a suitable number of diodes in series as shown, a voltage regulator for the maximum battery voltage can be built easily. The circuit shown in Fig. 3 can be used in either Fig. 1 or 2, for VR1. It will draw little current until the battery voltage reaches a permissible value during charge. Once the voltage reaches a preset level, the diodes start to conduct and limit any further increases.

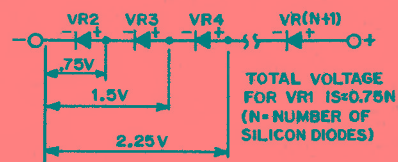


Fig. 3 — Schematic diagram of the voltage regulator (VR1, Figs. 1 and 2).

Initial Testing

After the circuit is wired and checked, apply power (without a battery connected for charging). The bulb should light to less than full brilliance. Measure the voltage across the regulator. It should be 3 to 8 percent above the rated voltage of the batteries to be charged. Adding or removing some diodes in VR1 may be necessary. Connect the discharged batteries and measure the charging current (either a built-in meter could be used, or a temporary one could be connected in series with the battery). The current should be typically 100 mA with partially discharged batteries. The current will decrease as the charging time increases, and a value of 5 mA indicates a fully charged condition. No damage will result if the batteries are left on charge continuously.

HF Transmitting

Regardless of the transmission mode – code, a-m, fm, single sideband, radioteletype, amateur TV – vacuum tubes and semiconductors are common elements in all transmitters. They are used as oscillators, amplifiers, frequency multipliers and frequency converters. These four building blocks, plus suitable power supplies, are basically all that is required to make any of the popular transmission systems.

The simplest code transmitter is a keyed oscillator working directly into the antenna; a more elaborate (and practical) code transmitter, the type popular with many beginners, will include one or more frequency-multiplication stages and one or more power-amplifier stages. Any code transmitter will obviously require a means for keying it. The bare skeleton is shown in Figs. 6-2A and B. The rf generating and amplifying sections of a double-sideband phone transmitter (a-m or fm) are similar to those of a code transmitter.

The overall design depends primarily upon the bands in which operation is desired and the power output. A simple oscillator with satisfactory frequency stability may be used as a transmitter at the lower frequencies, but the power output obtainable is small. As a general rule, the output of the oscillator is fed into one or more amplifiers to bring the power fed to the antenna up to the desired level.

An amplifier whose output frequency is the same as the input frequency is called a **straight amplifier**. A **buffer amplifier** is the term sometimes applied to an amplifier stage to indicate that its primary purpose is one of isolation, rather than power gain.

Because it becomes increasingly difficult to maintain oscillator frequency stability as the frequency is increased, it is most usual practice in working at the higher frequencies to operate the oscillator at a low frequency and follow it with one or more frequency multipliers as required to arrive at the desired output frequency. A frequency multiplier is an amplifier that delivers output at a multiple of the exciting frequency. A **doubler** is a multiplier that gives output at twice the exciting frequency; a **tripler** multiplies the exciting frequency by three, etc. From the viewpoint of any particular stage in a transmitter, the preceding stage is its **driver**.

As a general rule, frequency multipliers should not be used to feed the antenna system directly, but should feed a straight amplifier which, in turn, feeds the antenna system.

Good frequency stability is most easily obtained through the use of a **crystal-controlled oscillator**, although a different crystal is needed for each frequency desired (or multiples of that frequency). A **self-controlled oscillator** or **VFO** (variable-frequency oscillator) may be tuned to any frequency with a dial in the manner of a receiver, but requires great care in design and construction if its stability is to compare with that of a crystal oscillator.

Many transmitters use tubes, but for low-power hf and channelized vhf fm transmitters, transistors are dominant. New solid-state devices are being developed which allow dc inputs of 100 watts or more with a low-level of IM distortion products. As the cost of these transistors is reduced it can be assumed that at some point in the future tubes will be used only for high-power amplification.

The best stage or stages to key in a code transmitter is a matter which is discussed in a later chapter. The oscillator/multiplier/amplifier type of transmitter (Fig. 6-2B) has long been popular. However, the excellent frequency stability and the advantages of grid-block keying (which are explained in the Code Transmission chapter) have

Fig. 6-1 – An amateur's transmitter is his on-the-air voice. He is judged by the quality of that "voice," whatever the mode that he chooses to operate.

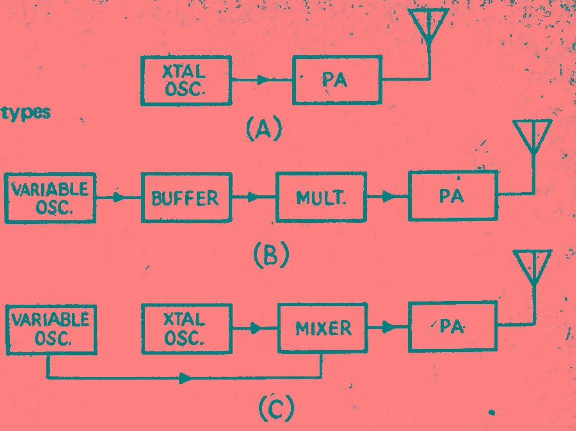


Fig. 6-2 — Block diagrams of the three basic types of transmitters.

made the heterodyne exciter of Fig. 6-2C increasingly popular, in spite of the slightly more complex circuitry required.

An fm transmitter can only be modulated in or following the oscillator stage. An a-m phone transmitter can only be modulated in the output stage, unless the modulated stage is followed by a linear amplifier. However, following an amplitude-modulated stage by a linear amplifier is an inefficient process, convenient as an expedient, but not recommended for best efficiency.

Following the generation of a single-sideband phone signal, its frequency can be changed only by frequency conversion (not multiplication), in exactly the same manner that signals in a receiver are heterodyned to a different frequency. Complete details of ssb transmitter design and construction are given in Chapter 13.

CRYSTAL OSCILLATORS

The frequency of a crystal-controlled oscillator is held constant to a high degree of accuracy by the use of a quartz crystal. The frequency depends almost entirely on the dimensions of the crystal (essentially its thickness); other circuit values have comparatively negligible effect. However, the power obtainable is limited by the heat the crystal will stand without fracturing. The amount of heating is dependent upon the rf cyrstal current which, in turn, is a function of the amount of feedback required to provide proper excitation. Crystal heating short of the danger point results in frequency drift to an extent depending upon the way the crystal is cut. Excitation should always be adjusted to the minimum necessary for proper operation.

The most stable type of crystal oscillator is that which provides only a small voltage output (lightly loaded), and which operates the crystal at a low drive level. Such oscillators are widely used in receivers and heterodyne transmitters. The oscillator/multiplier/amplifier type of transmitter usually requires some power from the oscillator stage. For either type of crystal oscillator, the active element may be a tube or a transistor.

Oscillator Circuits

The simplest crystal-oscillator circuit is shown in Fig. 6-3A. Feedback in this circuit is provided by the gate-source and drain-source capacitance. The circuit shown at B is the equivalent of the tuned-grid, tuned-plate circuit discussed in the chapter on vacuum-tube principles, using the crystal to replace the tuned grid circuit. Although JFETs are shown in the sample circuits at A and B, MOSFETs or triodes may also be employed, using the connections shown in 6-3C through F.

For applications where some power is required from the crystal oscillator, the circuits shown in

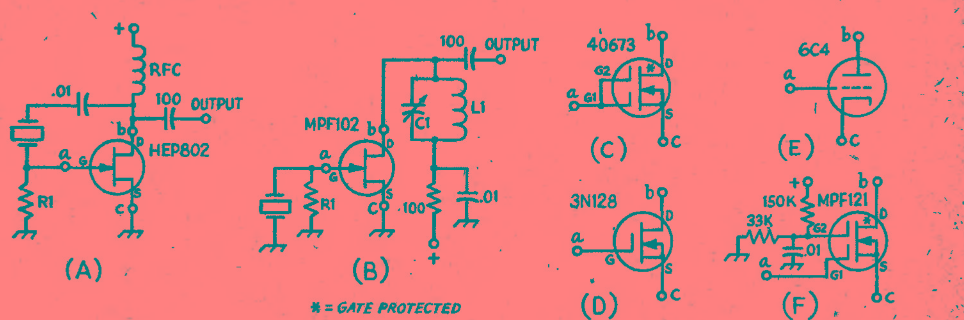


Fig. 6-3 — Simple crystal oscillator circuits. (A) Pierce, (B) FET, (C-F) other devices that can also be used in the circuits of A and B with appropriate changes in supply voltage.

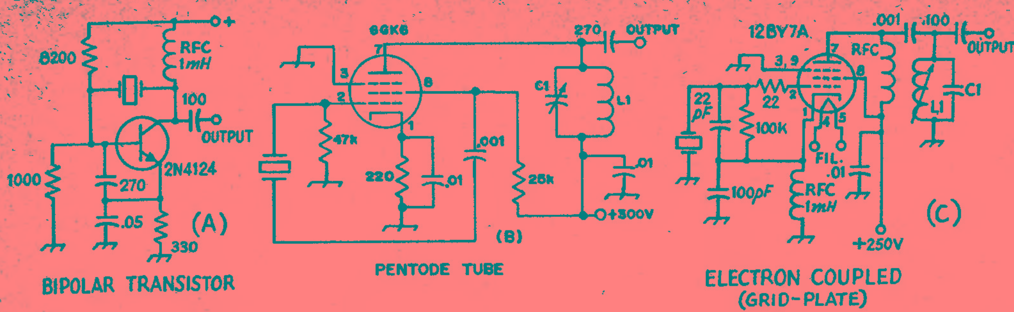


Fig. 6-4 — Crystal-oscillator circuits that are designed to deliver power. L1/C1 resonata at the crystal frequency, or a multiple thereof if the second, third, or fourth harmonic is the desired output frequency.

Fig. 6-4 may be employed. At A, a bipolar transistor is used, while the tube circuits (B, C) are somewhat more complicated. They combine the functions of oscillator and amplifier or frequency multiplier in a single tube. In these circuits, the screen of a tetrode or pentode is used as the plate in a triode oscillator. Power output is taken from a separate tuned tank circuit in the actual plate circuit. Although the oscillator itself is not entirely independent of adjustments made in the plate tank circuit when the latter is tuned near the fundamental frequency of the crystal, the effects can be satisfactorily minimized by proper choice of the oscillator tube.

The oscillators of Fig. 6-4B and 6-4C are a modification of the grid-plate circuit of Fig. 6-3B. In Fig. 6-4C the ground point has been moved from the cathode to the plate of the oscillator (in other words, to the screen of the tube). Excitation is adjusted by proper proportioning of 22- and 100-pF feedback capacitors.

When some types of tubes are used in the circuits of Fig. 6-4B, oscillation will stop when the output plate circuit is tuned to the crystal frequency, and it is necessary to operate with the plate tank circuit critically detuned for maximum output with stability. However, when the 6GK6, 12BY7A, 5763, or the lower-power 6AH6 is used with proper adjustment of excitation, it is possible to tune to the crystal frequency without stopping oscillation. These tubes also operate with less crystal current than most other types for a given power output, and less frequency change occurs when the plate circuit is tuned through the cyrstal frequency (less than 25 Hertz at 3.5 MHz).

Crystal current may be estimated by observing relative brilliance of a 60-mA dial lamp connected in series with the crystal. Current should be held to the minimum for satisfactory output by careful adjustment of excitation. With the operating voltages shown, satisfactory output should be obtained with crystal currents of 40 mA or less.

In these tube circuits, output may be obtained at multiples of the crystal frequency by tuning the plate tank circuit to the desired harmonic, the output dropping off, of course, at the higher harmonics. Especially for harmonic operation, a low-*C* plate tank circuit is desirable.

Practical Considerations

The operation of a crystal oscillator is often hampered because vhf parasitic oscillations also occur in the circuit. An effective way of killing parasitics is the use of a low-value composition resistor or ferrite bead, as shown in Fig. 6-5. The parasitic stopper can be located on the gate (grid or base) lead, and it should be placed as close as possible to the transistor. The circuit at A may be used for low-power applications. If a crystal above 1 MHz is to be used it may be advisable to include a trimmer capacitor across the crystal to allow the crystal frequency to be set exactly.

It is often desirable in fm and ssb gear to use several crystals, switch-selected in a single oscilla-

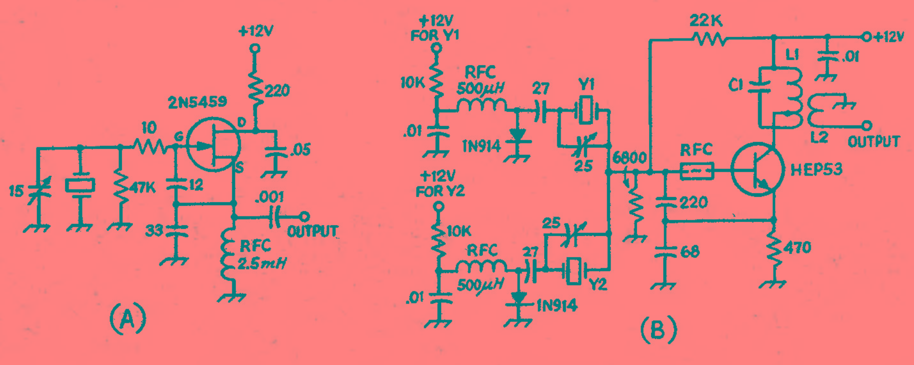


Fig. 6-5 — Two practical crystal-oscillator designs. (A) For low-power output applications such as a conversion oscillator or BFO, (B) an example of diode switching of crystals. The rf choke on the base lead of the transistor is a ferrite bead which prevents vhf parasitic oscillation.

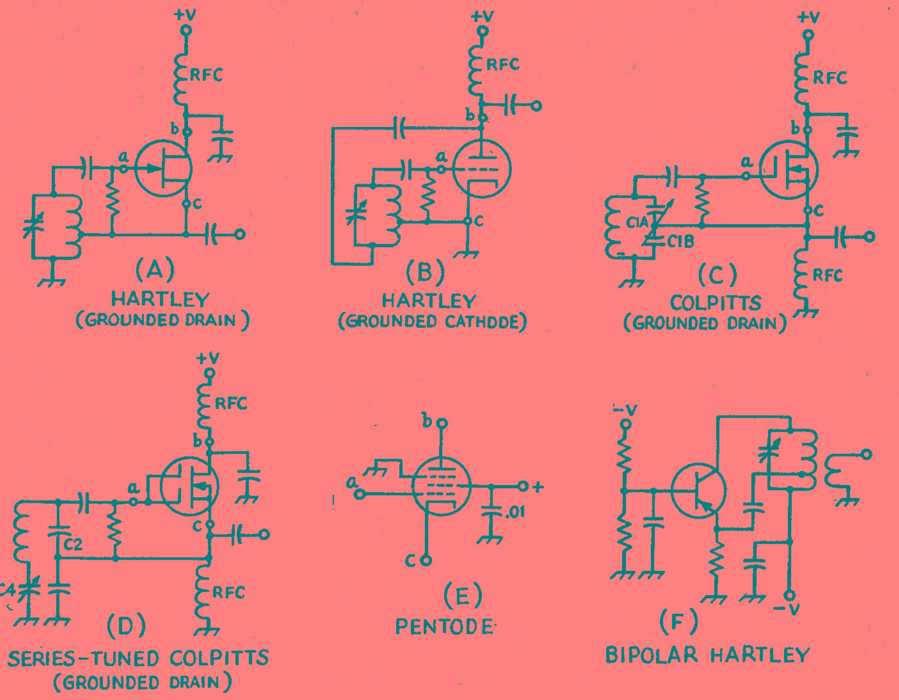


Fig. 6-6 — VFO circuits. The devices shown in Fig. 6-2C through F may also be employed as the active component.

tor. If manual switching is used, the leads to the switch may introduce sufficient additional capacitance to upset the operation of the circuit. Therefore, the use of diode switching, such as shown in Fig. 6-5B, is now popular. Any high-speed switching diode may be employed. The use of diode switching for low-level tank circuits, especially in receivers, has gained wide acceptance. A special diode known as the PIN has been developed for this purpose. In any diode-switching circuit it is important to insure that the switching bias is many times larger than the peak rf voltage present.

VARIABLE-FREQUENCY OSCILLATORS

The frequency of a VFO depends entirely on the values of inductance and capacitance in the circuit. Therefore, it is necessary to take careful steps to minimize changes in these values not under the control of the operator. As examples, even the minute changes of dimensions with temperature, particularly those of the coil, may result in a slow but noticeable change in frequency called drift. The effective input capacitance of the oscillator tube, which must be connected across the circuit, changes with variations in electrode voltages. This, in turn, causes a change in the frequency of the oscillator. To make use of the power from the oscillator, a load, usually in the form of an amplifier, must be coupled to the oscillator, and variations in the load may reflect on the frequency. Very slight mechanical movement of the components may result in a shift in frequency, and vibration can cause modulation.

In the past different techniques have been used to design the VFOs for transmitters and receivers. However, today the same circuits may be used for either application. In receivers the VFO is usually called an HFO.

VFO Circuits

Fig. 6-6 shows the most commonly used circuits. They are all designed to minimize the effects mentioned above. The oscillating circuits in Figs. 6-6A and B are the Hartley type; those in C and D are Colpitts circuits. (See chapter on vacuum-tube principles.) In the circuits of A, B and C, all of the above-mentioned effects, except changes in inductance, are minimized by the use of a high-*Q* tank circuit obtained through the use of large tank capacitances. Any uncontrolled changes in capacitance thus become a very small percentage of the total circuit capacitance.

In the series-tuned Colpitts circuit of Fig. 6-6D (sometimes called the Clapp circuit), a high-*Q* circuit is obtained in a different manner. The tube is tapped across only a small portion of the oscillating tank circuit, resulting in very loose coupling between tube and circuit. The taps are provided by a series of three capacitors across the coil. In addition, the tube capacitances are shunted by large capacitors, so the effects of the tube — changes in electrode voltages and loading — are still

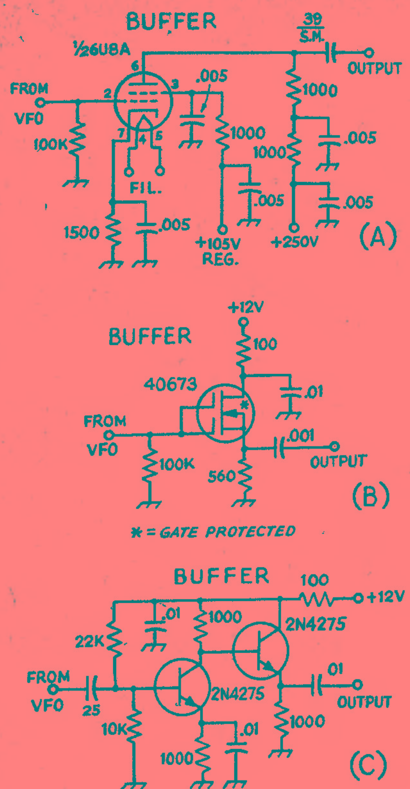


Fig. 6-7 — Isolating stages to be used between a VFO and the following amplifier or mixer stage.

further reduced. In contrast to the preceding circuits, the resulting tank circuit has a high L/C ratio and therefore the tank current is much lower than in the circuits using high-C tanks. As a result, it will usually be found that, other things being equal, drift will be less with the low-C circuit.

For best stability, the ratio of C2 to C4 should be as high as possible without stopping oscillation. The permissible ratio will be higher the higher the Q of the coil and the mutual conductance of the tube. If the circuit does not oscillate over the desired range, a coil of higher Q must be used or the capacitance of C2 and C3 reduced.

The pentode tube of 6-6E or any of the active devices shown in Fig. 6-3 may be used in either the Hartley or Colpitts circuits. Good results can be obtained with both tubes and transistors, so the choice of the active device is often a matter of personal preference.

Load Isolation

In spite of the precautions already discussed, the tuning of later stages in the transmitter may cause a noticeable change in frequency. This effect can be reduced considerably by designing a pentode oscillator for half the desired frequency and doubling frequency in the output circuit.

It is desirable, although not a strict necessity if detuning is recognized and taken into account, to approach as closely as possible the condition where the adjustment of tuning controls in the transmitter, beyond the VFO frequency control, will have negligible effect on the frequency. This can be done by adding isolating stage or stages whose tuning is fixed between the oscillator and the first tunable amplifier stage in the transmitter.

Fig. 6-7A shows such an arrangement that gives good isolation. A pentode tube is operated with a low-impedance resistive load, and regulated screen voltage. At B a simple follower circuit is used. The disadvantage of this circuit is that the level of the output will be quite low, usually less than one volt. Bipolar transistors are used in a direct-coupled follower arrangement in Fig. 6-7C, providing a higher level of output (above 3 V) than was possible with the design shown at B. The ability of a buffer stage to isolate the VFO from the load can be tested simply. Use a receiver to monitor the VFO, and listen as the buffer output is first left open and then shorted. A good buffer will hold the frequency change to less than 100 Hz. Often the frequency change may be in the order of several kHz when this test is made, an indication that the buffer is not doing its job.

Chirp, Pulling and Drift

Any oscillator will change frequency with an extreme change in plate screen voltages, and the use of stabilized sources for both is good practice. But steady source voltages cannot alter the fact of the extreme voltage changes that take place when an oscillator is keyed or heavily amplitude-modulated. Consequently some chirp or fm is the inescapable result of oscillator keying or heavy amplitude modulation.

A keyed or amplitude-modulated amplifier presents a variable load to the driving stage. If the driving stage is an oscillator, the keyed or modulated stage (the variable load) may "pull" the oscillator frequency during keying or modulation. This may cause a "chirp" on cw or incidental fm on a-m phone. In either case the cure is to provide one or more "buffer" or isolating stages between the oscillator stage and the varying load. If this is not done, the keying or modulation may be little better than when the oscillator itself is keyed or modulated.

Frequency drift is minimized by limiting the temperature excursions of the frequency-determining components to a minimum. This calls for good ventilation and a minimum of heat-generating components.

Variable capacitors should have ceramic insulation, good bearing contacts and should preferably be of the double bearing type. Fixed capacitors should have zero-temperature coefficients. The tube socket should have ceramic insulation.

Temperature Compensation

If, despite the observance of good oscillator construction practice, the warm-up drift of an oscillator is too high, it is caused by high-tempera-

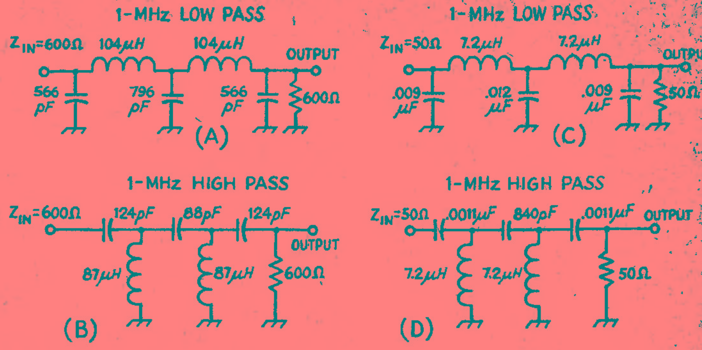


Fig. 6-8 — Universal high- end low-pass filter designs. The values given are for 1 MHz; they may be divided by the desired cutoff frequency (in MHz) to determine the value of the components needed. For example, if the 600-ohm high-pass design at A was to be used at 10 MHz, the values of *C* and *L* shown are divided by 10. The input and output impedance remains at 600 ohms.

ture operation of the oscillator. If the ventilation cannot be improved (to reduce the ultimate temperature), the frequency drift of the oscillator can be reduced by the addition of a "temperature-coefficient capacitor." These are available in negative and positive coefficients, in contrast to the zero-coefficient "NP0" types.

Most uncorrected oscillators will drift to a lower frequency as the temperature rises. Such an oscillator can be corrected (at a frequency f) by adding an N750-type capacitor (-750 parts per million per ^{0}C) of a value determined by making two sets of measurements. Measure the drift f_1 from cold to stability (e.g., 1 1/2 hours). To the cold (cooled-off) oscillator, add a *trial* N750 capacitor (e.g., 50 pF) and retune the cold oscillator to frequency f_1 (by retuning a padder capacitor or the tuning capacitor). Measure the new warm-up drift f_2 over the same period (e.g., 1 1/2 hours). The required corrective N750 capacitor is then

$$\text{Corrective } C = C_{\text{trial}} \frac{f_1}{f_1 - f_2}$$

If the trial capacitor results in a drift to a higher frequency, the denominator becomes $f_1 + f_2$.

Oscillator Coils

The Q of the tank coil used in the oscillating portion of any of the circuits under discussion should be as high as circumstances (usually space) permit, since the losses, and therefore the heating, will be less. With recommended care in regard to other factors mentioned previously, most of the drift will originate in the coil. The coil should be well spaced from shielding and other large metal surfaces, and be of a type that radiates heat well, such as a commercial air-wound type, or should be would tightly on a threaded ceramic form so that the dimensions will not change readily with temperature. The wire with which the coil is wound should be as large as practicable, especially in the high-C circuits.

Mechanical Vibration

To eliminate mechanical vibration, components should be mounted securely. The capacitor should preferably have small, thick plates and the coil braced, if necessary, to prevent the slightest mechanical movement. Wire connections between tank-circuit components should be as short as possible and flexible wire will have less tendency to vibrate than solid wire. It is advisable to cushion the entire oscillator unit by mounting on sponge rubber or other shock mounting.

Output Filtering

The output of an oscillator contains a good deal of harmonic energy in addition to the desired frequency. Often these harmonics can cause the generation of spurious products in heterodyne transmitters which result in signals being radiated outside the amateur bands. In receivers, the oscillator harmonics cause "birdies" and sprious responses in the tuning range. In general, transistor circuits generate a higher level of harmonic energy than do tube designs. Thus, the output of a tube VFO often can be used without filtering, while most solid-state VFOs require an output low-pass filter.

W7ZOI has provided general-purpose filter designs shown in Fig. 6-8. These designs have been developed using computer-aided design (CAD), where a digital computer models or "synthesizes" the circuit. Not only filters, but models of tubes, transistors and complete circuits may be developed in this way, allowing a designer to optimize a circuit without taking days of cut and try. However, the CADs are just models, and often once a circuit is built additional refinement is still required.

The filter shown in Figs. 6-8A and B are designed for 600-ohm input and output impedances, while C and D are for use in 50-ohm lines. The values shown are for 1 MHz. A design for higher frequencies may be obtained by dividing the values shown by the desired frequency. For example, if a VFO were to be used in tuning 5 to 5.5 MHz, the output filter of Fig. 6-8A might be employed with a cut-off frequency of 6 MHz. The *LC* values shown would be divided by 6. The result will usually be an odd number, so the closest standard value may be used.

A PRACTICAL VFO CIRCUIT

The circuit shown in Fig. 6-9 is for a solid-state VFO covering 3.5 to 4 MHz. A number of measures have been taken to prevent harmonic and spurious outputs that so often plague transistor designs. Examination of Fig. 6-9 will show that a diode, CR2, is connected between the signal gate

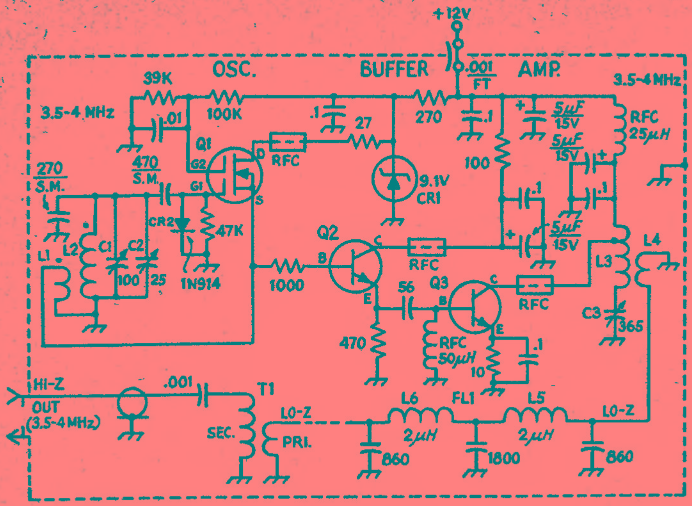


Fig. 6-9 — A typical VFO design showing extensive use of buffering and filtering to achieve a highly stable output with low spurious-frequency content.

of Q1 and ground. This diode should be designed for high-speed switching – a 1N914 is suitable – and should be connected with its anode toward gate 1. It clamps on the positive-going half of the cycle to prevent Q1 from reaching high peak transconductance, the time period when the output from the oscillator is rich in harmonic energy. This technique should be applied to any JFET or MOSFET oscillator, but does not work with bipolar-transistor oscillators. CR2 does not impair the performance of the VFO. Additional harmonics can be generated at Q2 and Q3, so attention must be given to that part of the circuit as well. Note that the collector of Q3 is tapped well down on L3. The tap provides an impedance match for the circuit, but still represents a high impedance at the harmonic frequencies, if not located too near the cold end of L3, thus contributing to a cleaner output signal. However, even though these precautions are taken, it is not uncommon to find that the second and third harmonics from a transistor output stage are only down some 10 to 15 decibels in level from the fundamental signal. By taking the VFO output at low impedance, L4, a low-pass, double-section filter can be used to diminish the harmonic to a level that is some 30 decibels or more below that of the desired output signal. FL1 is designed for 3.5 to 4-MHz use, and assures a clean output signal from the VFO.

VFO Output Level and Impedance

One of the things that perplexes many first-time users of transistorized VFOs is the matter of sufficient signal output to properly excite a transmitter input stage, or to supply adequate injection voltage to a receiver or transmitter mixer. The rms output of a solid-state VFO is limited by its low-impedance output port. In the circuits of Fig. 6-9 the output would usually be taken across the emitter resistor of Q2, the buffer. Typically, the rms output voltage at that point in the circuit will be on the order of 0.5 to 2 volts. Tube mixers can require up to several volts of oscillator signal in order to function properly. Most solid-state transmitters need from 3 to 10 volts of drive on the base of the first power stage, and a reasonable amount of driving power is needed to satisfy this requirement. Driving power is generally required by the grid of the first stage of a tube transmitter. The VFO should, therefore, be capable of supplying from 0.5 to 1 watt of power output. The Class-C amplifier, Q3, provides the needed power output. Should the driven stage present a low-impedance to the VFO, output can be taken directly from the side of FL1 opposite Q3. If, however, the driven stage of the transmitter or receiver has a high input impedance, some method must be used to provide the required impedance transformation, low to high. A broad-band toroidal step-up transformer, T1, is used for this purpose in Fig. 6-9. The secondary of the transformer is resonant somewhere in the operating range of the VFO, and takes advantage of the stray circuit capacitance, normally around 10 pF, to establish resonance. The impedance-transformation ratio is set by adjusting the number of turns on the primary winding. Alternatively, T1 can be replaced by a tuned circuit of conventional design. It can be equipped with a fixed-value capacitor and a slug-tuned inductor, or a fixed-value inductor can be used with a variable capacitor to permit peaking the output at the operating frequency. The use of a tuned circuit will assure somewhat better efficiency than will the broadband transformer, T1. Thus, it can be seen that the circuit must be tailored to the need.

Checking VFO Stability

A VFO should be checked thoroughly before it is placed in regular operation on the air. Since succeeding amplifier stages may affect the signal characteristics, final tests should be made with the complete transmitter in operation. Almost any VFO will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, well-shielded monitoring arrangements is a receiver combined with a harmonic from a frequency standard. (See the Measurements chapter for suitable circuits.) The receiver BFO is turned off and the VFO signal is tuned to beat with the signal from the crystal oscillator instead. In this way any receiver instability caused by overloading the input circuits, which may result in "pulling" of the hf oscillator in the receiver, or by a change in line voltage to the receiver when the transmitter is keyed, will not affect the reliability of the check. Most crystals have a sufficiently low temperature coefficient to give a check on drift as well as on chirp and signal quality if they are not overloaded.

Harmonics of the crystal may be used to beat with the transmitter signal when monitoring at the higher frequencies. Since any chirp at the lower frequencies will be magnified at the higher frequencies, accurate checking can best be done by monitoring at a harmonic.

The distance between the crystal oscillator and receiver should be adjusted to give a good beat between the crystal oscillator and the transmitter signal. When using harmonics of the crystal oscillator, it may be necessary to attach a piece of wire to the oscillator as an antenna to give sufficient signal in the receiver. Checks may show that the stability is sufficiently good to permit oscillator keying at the lower frequencies, where break-in operation is of greater value, but that chirp becomes objectionable at the higher frequencies. If further improvement does not seem possible, it would be logical in this case to use oscillator keying at the lower frequencies and amplifier keying at the higher frequencies.

Premixing

It is difficult to build a variable-frequency oscillator for operation above 10 MHz with drift of only a few Hz. A scheme called premixing, shown in Fig. 6-10A, may be used to obtain VFO output in the 10- to 50-MHz range. The output of a highly stable VFO is mixed with energy from a crystal-controlled oscillator. The frequencies of the two oscillators are chosen so that spurious outputs generated during the mixing process do not fall within the desired output range. A bandpass filter at the mixer output attenuates any out-of-band spurious energy. The charts given in Chapter 8 can be used to choose oscillator combinations which will have a minimum of spurious outputs. Also, Chapter 8 contains a discussion of mixer-circuit design.

PLL

Receivers and transmitters of advanced design are now using phase-locked loops (PLLs) to generate highly stable local oscillator energy up into the microwave region. The PLL has the advantage that no mixing stage is used in conjunction with the output oscillator, so the output energy is quite "clean." The Galaxy R-530, the Collins 651S-1, and the National HRO-600 currently use PLL high-frequency oscillator systems.

The basic diagram of a PLL is shown in Fig. 6-10B. Output from a voltage-controlled oscillator

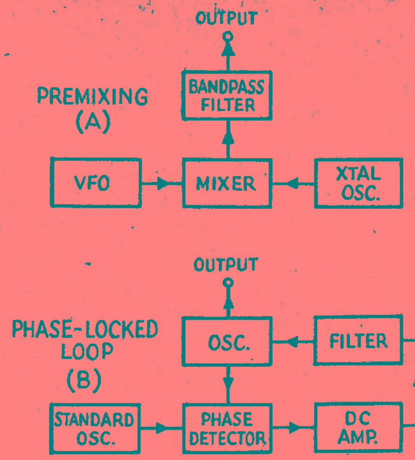


Fig. 6-10 — Block diagrams of the (A) premixing and (B) phase-lock-loop schemes.

(VCO) and a frequency standard are fed to a phase detector which produces an output voltage equal to the difference in frequency between the two signals. The *error* voltage is amplified, filtered, and applied to the VCO. The error voltage changes the frequency of the VCO until it is locked to the standard. The bandwidth of the error-voltage filter determines the frequency range over which the system will remain in phase lock.

Three types of phase-locked loops are now in use. The simplest type uses harmonics of a crystal standard to phase-lock an HFO, providing the injection for the first mixer in a double-conversion receiver. A typical circuit is given in Fig. 6-11. Complete construction details on this PLL were given in *QST* for January, 1972. A second type of phase-locked loop uses a stable mf VFO as the standard which stabilizes the frequency of an hf or vhf VCO. This approach is used in the receiver described by Fischer in *QST*, March, 1970.

The other PLL system also uses a crystal-controlled standard, but with programmable frequency dividers included so that the VCO output is always locked to a crystal reference. The frequency is changed by modifying the instructions to the dividers; steps of 100 Hz are usually employed for hf receivers while 10-kHz increments are popular in vhf gear. The use of a PLL for fm demodulation is covered separately in Chapter 14.

VFO DIALS

One of the tasks facing an amateur builder is the difficulty of finding a suitable dial and drive assembly for a VFO. A dial should provide a sufficiently slow rate of tuning — 10- to 25-kHz per knob revolution is considered optimum — without backlash. Planetary drives are popular because of their low cost; however, they often develop objectional backlash after a short period of use. Several types of two-speed drives are available. They are well suited to homemade amateur

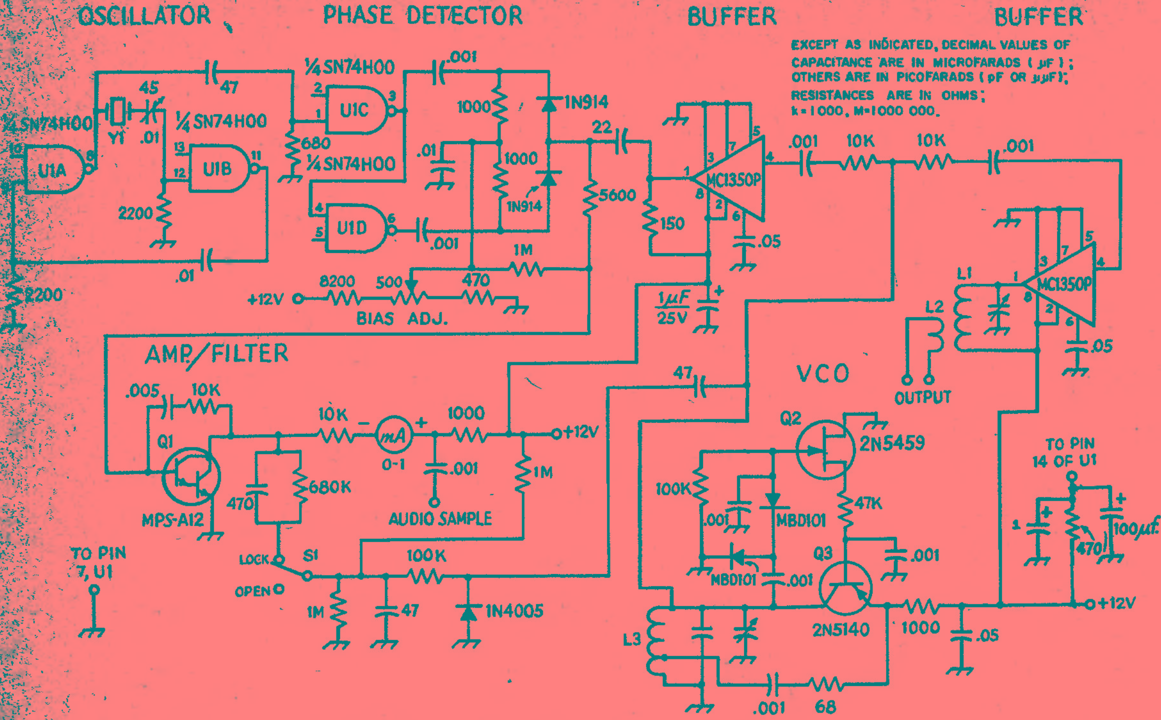


Fig. 6-11 — A practical phase-locked oscillator intended for application as tha crystal-controlled HFO in a transmitter or receiver. The crystal frequency should be chosen so that the harmonic content of the standard is sufficient at the desired output frequency. A 200-kHz crystal is good to 40 MHz, 500-kHz crystal to 60 MHz, and 1-MHz crystal to 80 MHz. L1 and L3 are chosen to resonate at tha desired output frequency.

equipment. Several of the construction projects described elsewhere in this book employ this type of dial. The Eddystone 898 precision dial has long been a favorite with amateurs, although the need to elevate the VFO far above the chassis introduces some mechanical-stability problems. If a permeability tuned oscillator (PTO) is used, one of the many types of turn counters made for vacuum variable capacitors or rotary inductors may be employed.

Linear Readout

If linear-frequency readout is desired on the dial, the variable capacitor must be only a small portion of the total capacitance in the oscillator tank. Capacitors tend to be very nonlinear near the ends of rotation. A gear drive providing a 1.5:1 reduction should be employed so that only the center of the capacitor range is used. Then, as a

Fig. 6-12 — A 5-digit readout using light-emitting diodes.

final adjustment, the plates of the capacitor must be filed until linear readout is achieved. In a PTO, the pitch of the oscillator coil winding may be varied so that linear frequency change results from the travel of the tuning slug. Such a VFO was described in *QST* for July, 1964. A different approach was employed by Lee (*QST*, November, 1970), using a variable-capacitance diode (Varicap) as the VFO tuning element. A meter which reads the voltage applied to the Varicap was calibrated to indicate the VFO frequency.

Electronic Dials

An electronic dial consists of a simplified frequency counter which reads either the VFO or operating frequency of a transmitter or receiver. The advantage of an electronic dial is the excellent accuracy (to one Hertz, if desired) and the fact that VFO tuning does not have to be linear. The readout section of the dial may use neon-glow tubes called Nixies (a trade name of the Burroughs Corp.), or a seven-segment display using incandescent lamps, filament wires in a vacuum tube, or LEDs (light-emitting diodes). A typical LED display is shown in Fig. 6-12. The use of MSI and LSI circuits, some containing as many as 200 transistors on a single chip, reduces the size required for an electronic dial to a few square inches of circuit-board space.

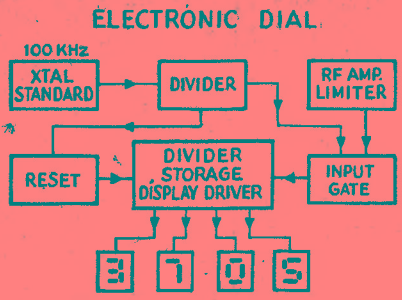


Fig. 6-13 — Block diagram of a frequency counter.

A typical counter circuit is given in Fig. 6-13. The accuracy of the counter is determined by a crystal standard which is often referred to as a clock. The output from a 100-kHz calibration oscillator, the type often used in receivers and transceivers, may be employed if accuracy of 100 Hz is sufficient. For readout down to 1 Hz, a 1- to 10-MHz AT-cut crystal should be chosen, because this type of high-accuracy crystal exhibits the best temperature stability. The clock output energy is divided in decade-counter ICs to provide the pulse which opens the input gate of the counter for a preset time. The number of rf cycles which pass through the gate while it is open are counted and stored. Storage is used so that the readout does not blink. At the end of each counting cycle the information that has been stored activates the display LEDs, which present the numbers counted until another count cycle is complete. A complete electronic dial arranged to be combined with an existing transmitter or receiver was described in *QST* for October 1970. Also, Macleish *et al* reported an adapter which allows a commercially made frequency counter to be mated with ham gear so that the counter performs as an electronic dial (*QST*, May, 1971).

FREQUENCY MULTIPLIERS

Single-Tube Multiplier

Output at a multiple of the frequency at which it is being driven may be obtained from an amplifier stage if the output circuit is tuned to a harmonic of the exciting frequency instead of to the fundamental. Thus, when the frequency at the grid is 3.5 MHz, output at 7 MHz, 10.5 MHz, 14 MHz, etc., may be obtained by tuning the plate tank circuit to one of these frequencies. The circuit otherwise remains the same as that for a straight amplifier, although some of the values and operating conditions may require change for maximum multiplier efficiency.

A practical limit to efficiency and output within normal tube ratings is reached when the multiplier is operated at maximum permissible plate voltage and maximum permissible grid current. The plate current should be reduced as necessary to limit the dissipation to the rated value by increasing the bias and decreasing the loading.

Multiplications of four or five sometimes are used to reach the bands above 28 MHz from a lower-frequency crystal, but in the majority of lower-frequency transmitters, multiplication in a single stage is limited to a factor of two or three. Screen-grid tubes make the best multipliers because their high power-sensitivity makes them easier to drive properly than triodes.

Since the input and output circuits are not tuned close to the same frequency, neutralization usually will not be required. Instances may be encountered with tubes of high transconductance, however, when a doubler will oscillate in t.g.t.p. fashion.

Frequency multipliers using tubes are operated Class C, with the bias and drive levels adjusted for plate-current conduction of less than 180 degrees.

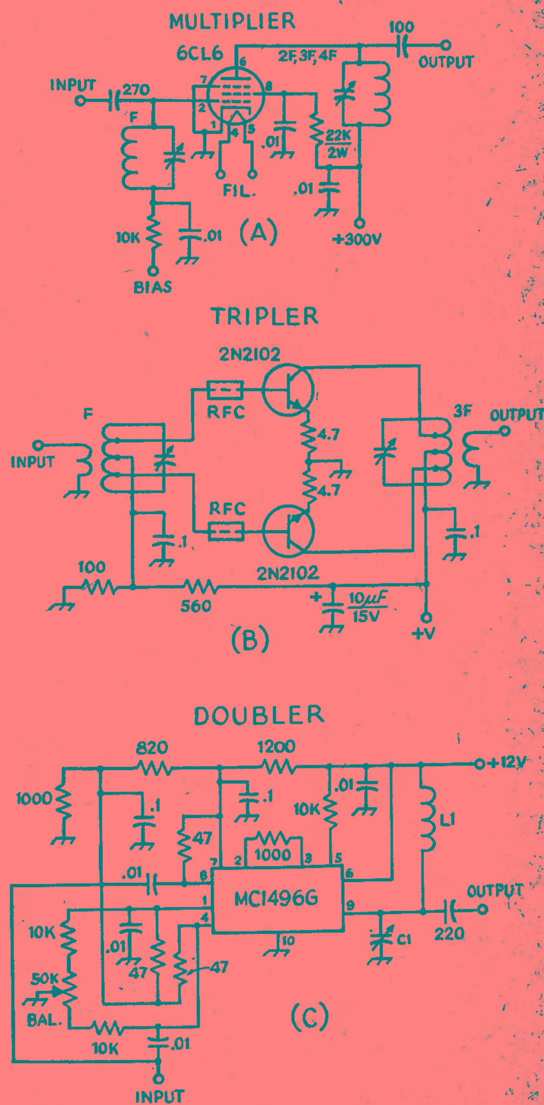


Fig. 6-14 — Frequency-multiplier circuits.

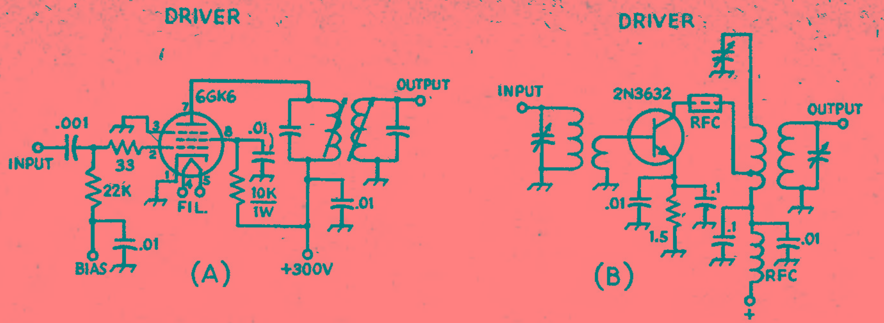


Fig. 6-15 — Driver stages using (A) a pentode tube and (B) a bipolar power transistor.

For maximum efficiency, a doubler requires a plate-conduction angle of about 110 degrees, while a tripler needs 100 degrees, a quadrupler 80 degrees, and a quintupler 65 degrees. For higher orders of multiplication increased bias and more drive are needed.

A typical circuit using a 6CL6 pentode tube is shown in Fig. 6-14A. The input circuit is tuned to the driving frequency while the output tank is set for the desired harmonic. If such a multiplier were to be operated directly into an antenna, additional selectivity would be necessary to prevent the radiation of harmonic energy (other than the desired frequency).

Push-Push Multipliers

A two-tube circuit which works well at even harmonics, but not at the fundamental or odd harmonics, is known as the **push-push** circuit. The grids are connected in push-pull while the plates are connected in parallel. The efficiency of a doubler using this circuit approaches that of a straight amplifier.

This arrangement has an advantage in some applications. If the heater of one tube is turned off, its grid-plate capacitance, being the same as that of the remaining tube, serves to neutralize the circuit. Thus provision is made for either straight amplification at the fundamental with a single tube, or doubling frequency with two tubes.

Push-Pull Multiplier

A single- or parallel-tube multiplier will deliver output at either even or odd multiples of the exciting frequency. A push-pull stage does not work as a doubler or quadrupler but it will work as a tripler.

Transistor Multipliers

A transistor develops harmonic energy with good efficiency, often causing harmonic-output problems in straight-through amplifiers. Two harmonic-generating modes are present, parametric multiplication and multiplication caused by the nonlinear characteristic presented by the base-collector junction. Transistors may be used in single-ended, push-pull, or push-push circuits. A typical push-pull tripler is shown in Fig. 6-14B. A small amount of forward bias has been added to the bases of the 2N2102s to reduce the amount of drive required. If a high level of drive is available, the bias circuit may be omitted.

A number of integrated circuits can be employed as frequency multipliers. The circuit at C uses a Motorola MC1496G (or the Signetics S5596, or Fairchild μA796) as a doubler. The input signal is balanced out in the IC, so only the desired second harmonic of the input frequency appears at the output. With suitable bypass capacitors this doubler can be used from audio to vhf.

DRIVERS

Pentode tubes are usually chosen for the driver stages of tube transmitters because they provide high amplification, often without requiring neutralization. Many of the receiving-type pentodes and smaller TV sweep tubes may be employed. The 6CL6, 6GK6, 12BY7A, 6BA6, 6AU6, and 6DC6 are often chosen. In cw and fm service the driver stage is operated Class C, while for ssb operation the Class-A mode is preferred to keep distortion to a minimum (third-order products at least 50 dB down). In ssb exciters alc voltage is often applied to a driver stage, in which case a semiremote-cutoff tube is desirable. Sharp-cutoff types are not acceptable because of a rapid increase in distortion as alc voltage drives the grid increasingly negative.

A typical tube driver stage is shown in Fig. 6-15 at A. The output load is a parallel-resonant circuit. Often a bandpass network is used so that the stage does not have to be tuned by a panel control. Also, coupling with a bandpass transformer provides a higher order of attenuation of harmonic and spurious signals. At Fig. 6-15B, a 2N3632 medium-power transistor serves as a Class-C driver. Note that this circuit is not suitable for ssb service.

Broadband Driver

Transistor circuits often require complex interstage coupling networks, because of the low input and output impedance characteristics of bipolar devices. Designing a solid-state multiband hf transmitter often requires some very complex band-switch arrangements. To eliminate this problem, the current trend is to use a broadband multistage driver that covers 3.5 to 30 MHz, for example, without switching or tuning adjustments. A typical circuit, similar to that used in Signal/One's CX-7 transceiver, is shown in Fig.

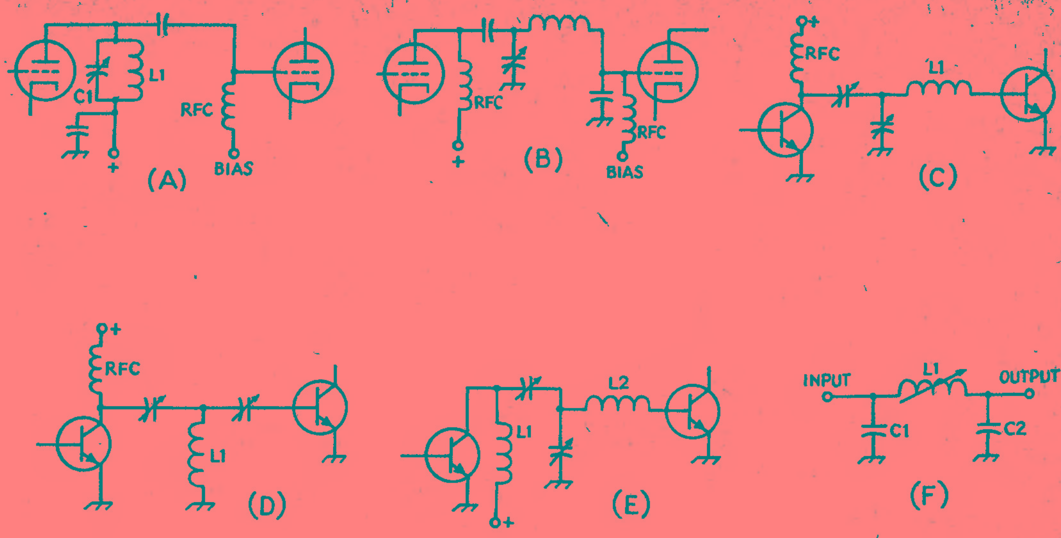


Fig. 6-16 — Interstage coupling networks for (A, B) tubes, (C-E) transistor stages, and (F) a grounded-grid amplifiar.

6-17. Only a few millivolts of ssb or cw drive will provide sufficient output to drive a 4CX250B operating Class AB_1. Interstage coupling is accomplished with broadband toroidal transformers. Feedback is added from the collector to the emitter of each bipolar-transistor stage to improve linearity. Output impedance of the broadband driver is approximately 390 ohms.

Interstage Coupling

To achieve the maximum transfer of power between the driver and the succeeding amplifier stage, the output impedance of the driver must be matched to the input impedance of the following amplifier. Some form of rf coupling or impedance-matching network is needed. The capacitive system of Fig. 6-16A is the simplest of all coupling systems. In this circuit, the plate tank circuit of the driver, C1L1, serves also as the grid tank of the amplifier. Although it is used more frequently than any other system, it is less flexible and has certain limitations that must be taken into consideration.

The two stages cannot be separated physically any appreciable distance without involving loss in transferred power, radiation from the coupling lead and the danger of feedback from this lead. Since both the output capacitance of the driver tube and the input capacitance of the amplifier are across the single circuit, it is sometimes difficult to obtain a tank circuit with a sufficiently low Q to provide an efficient circuit at the higher frequencies. The coupling can be varied by altering the capacitance of the coupling capacitor, C2. The driver load impedance is the sum of the amplifier grid resistance and the reactance of the coupling capacitor in series, the coupling capacitor serving simply as a series reactor. The driver load resistance increases with a decrease in the capacitance of the coupling capacitor.

When the amplifier grid impedance is lower than the optimum load resistance for the driver, a transforming action is possible by tapping the grid down on the tank coil, but this is not recommended because it invariably causes an increase in vhf harmonics and sometimes sets up a parasitic circuit.

So far as coupling is concerned, the Q of the circuit is of little significance. However, the other considerations discussed earlier in connection with tank-circuit Q should be observed.

Pi-Network Interstage Coupling

A pi-section tank circuit, as shown in Fig. 6-16B, may be used as a coupling device between screen-grid amplifier stages. The circuit can also be considered a coupling arrangement with the grid of the amplifier tapped down on the circuit by means of a capacitive divider. In contrast to the tapped-coil method mentioned previously, this system will be very effective in reducing vhf harmonics, because the output capacitor provides a direct capacitive shunt for harmonics across the amplifier grid circuit.

To be most effective in reducing vhf harmonics, the output capacitor should be a mica capacitor connected directly across the tube-socket terminals. Tapping down on the circuit in this manner also helps to stabilize the amplifier. Since the coupling to the grid is comparatively loose under any condition, it may be found that it is impossible to utilize the full power capability of the driver stage. If sufficient excitation cannot be obtained, it may be necessary to raise the plate voltage of the

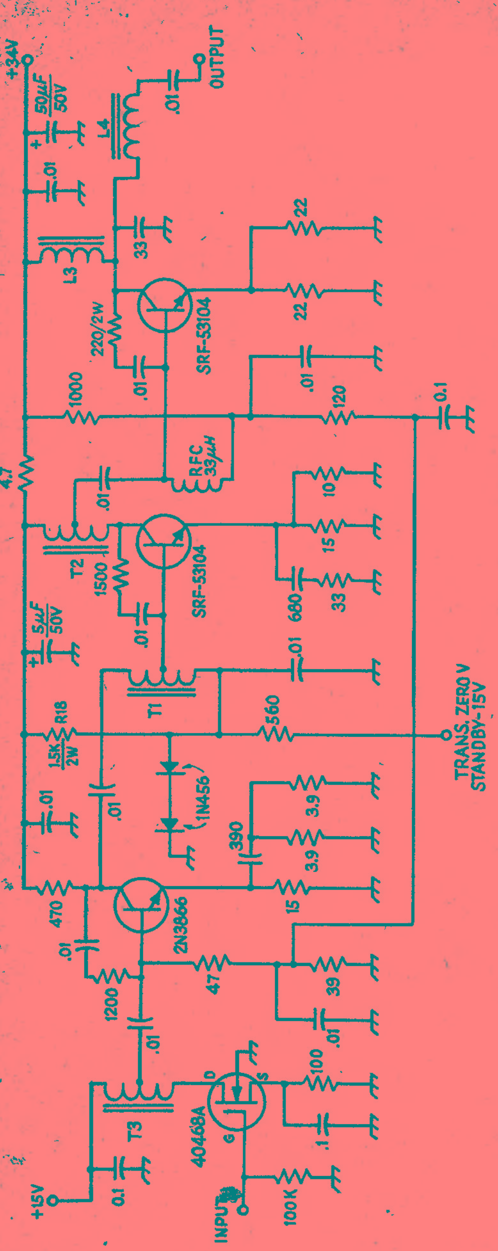


Fig. 6-17 — A solid-state broadband driver for 3 to 30 MHz. The design of transformers T1, T2 and T3 is covered later in the chapter.

driver, if this is permissible. Otherwise a larger driver tube may be required. As shown in Fig. 6-16B, parallel driver plate feed and amplifier grid feed are necessary.

Coupling Transistor Stages

In stages using bipolar power transistors, the input circuit must provide a match between the driver collector and the PA base. The latter exhibits a very low impedance. The input impedance of an rf power transistor is between several tenths of an ohm and several ohms. Generally, the higher the power rating of the device, the lower the input impedance. The base connection also has a reactive component which is capacitive at low frequencies and inductive at higher frequencies. At some frequency, usually between 50 and 150 MHz, the base lead will be self-resonant. The input impedance will vary with drive level, which makes a cut-and-try adjustment of the interstage network necessary.

An interstage network must provide the proper impedance transformation while tuning out reactance in the transistors. The reactive components of the base and collectors of power transistors are of such magnitude that they must be included in any network calculations. Fig. 6-16 shows several networks capable of interstage matching in a multistage transistor amplifier. At C, a T network is pictured. The value of the inductor is chosen so that its reactance is much greater than the capacitive reactance of the second transistor's base circuit. The capacitive divider provides the impedance match between the collector and the base.

The circuit of 6-16D is also basically a T network in which both the inductor and second capacitor are chosen to have reactance of a greater magnitude than the base-emitter capacitance of the second transistor. The circuits of C and D require that the collector of the driver transistor be shunt fed through a high-impedance rf choke. Fig. 6-16E shows a coupling network that eliminates the need for a choke. Here the collector of the driver transistor is parallel-tuned and the base-emitter junction of the following stage is series-tuned.

The remaining circuit, Fig. 6-16F, shows the pi-section network that is often used to match the 50-ohm output of an exciter to a grounded-grid power amplifier. A Q of 1 or 2 is chosen so that the circuit will be broad enough to operate across an amateur band without retuning. The network is designed for a 50-ohm input impedance and to match an output load of 30 to 150 ohms (the impedance range of the cathode of typical grounded-grid stages). Typical *LC* values are given in the construction projects presented later in this chapter.

RF POWER AMPLIFIER CIRCUITRY

Tube Operating Conditions

In addition to proper tank and output-coupling circuits, an rf amplifier must be provided with suitable operating voltages and an rf driving or excitation voltage. All rf amplifier tubes require a voltage to operate the filament or heater (ac is usually permissible), and a positive dc voltage between the plate and filament or cathode (plate voltage). Most tubes also require a negative dc voltage (biasing voltage) between control grid (grid No. 1) and filament or cathode. Screen-grid tubes require in addition a positive voltage (screen voltage or grid No. 2 voltage) between screen and filament or cathode.

Biasing and plate voltages may be fed to the tube either in series with or in parallel with the associated rf tank circuit as discussed in the chapter on electrical laws and circuits.

It is important to remember that true plate, screen or biasing voltage is the voltage between the particular electrode and filament or cathode. Only when the cathode is directly grounded to the chassis may the electrode-to-chassis voltage be taken as the true voltage. The required rf driving voltage is applied between grid and cathode.

Power Input and Plate Dissipation

Plate power input is the dc power input to the plate circuit (dc plate voltage X dc plate current). Screen power input likewise is the dc screen voltage X the dc screen current.

Plate dissipation is the difference between the rf power delivered by the tube to its loaded plate tank circuit and the dc plate power input. The screen, on the other hand, does not deliver any output power, and therefore its dissipation is the same as the screen power input.

TRANSMITTING-TUBE RATINGS

Tube manufacturers specify the maximum values that should be applied to the tubes they produce. They also publish sets of typical operating values that should result in good efficiency and normal tube life.

Maximum values for all of the most popular transmitting tubes will be found in the tables of transmitting tubes in the last chapter. Also included are as many sets of typical operating values as space permits. However, it is recommended that the amateur secure a transmitting-tube manual from the manufacturer of the tube or tubes he plans to use.

CCS and ICAS Ratings

The same transmitting tube may have different ratings depending upon the manner in which the tube is to be operated, and the service in which it is to be used. These different ratings are based primarily upon the heat that the tube can safely dissipate. Some types of operation, such as with grid or screen modulation, are less efficient than others, meaning that the tube must dissipate more heat. Other types of operation, such as cw or single-sideband phone are intermittent in nature, resulting in less average heating than in other modes where there is a continuous power input to the tube during transmissions. There are also different ratings for tubes used in transmitters that are in almost constant use (CCS – Continuous Commercial Service), and for tubes that are to be used in transmitters that average only a few hours of daily operation (ICAS – Intermittent Commercial and Amateur Service). The latter are the ratings used by amateurs who wish to obtain maximum output with reasonable tube life.

Maximum Ratings

Maximum ratings, where they differ from the values given under typical operating values, are not normally of significance to the amateur except in special applications. No single maximum value should be used unless all other ratings can simultaneously be held within the maximum values. As an example, a tube may have a maximum plate-voltage rating of 2000, a maximum plate-current rating of 300 mA, and a maximum plate-power-input rating of 400 watts. Therefore, if the maximum plate voltage of 2000 is used, the plate current should be limited to 200 mA (instead of 300 mA) to stay within the maximum power-input rating of 400 watts.

SOURCES OF TUBE ELECTRODE VOLTAGES

Filament or Heater Voltage

The heater voltage for the indirectly heated cathode-type tubes found in low-power classifications may vary 10 percent above or below rating without seriously reducing the life of the tube. But the voltage of the higher-power filament-type tubes should be held closely between the rated voltage as a minimum and 5 percent above rating as a maximum. Make sure that the plate power drawn from the power line does not cause a drop in filament voltage below the proper value when plate power is applied.

Thoriated-type filaments lose emission when the tube is overloaded appreciably. If the overload

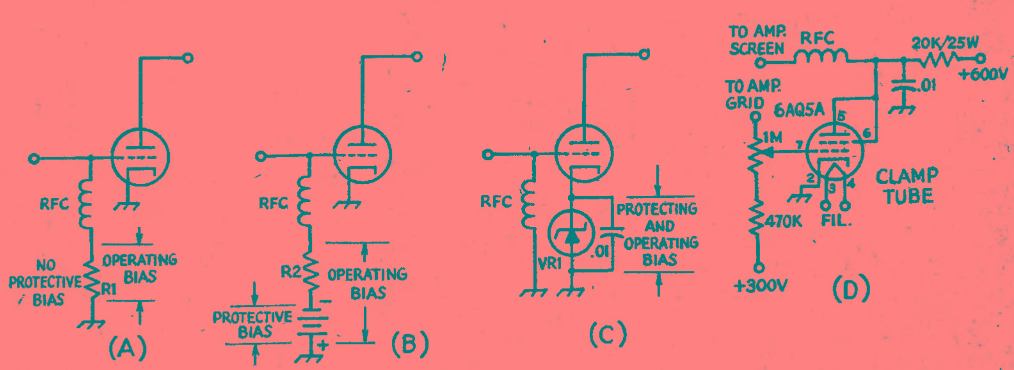


Fig. 6-18 – (A-C) Various systems for obtaining protective end operating bias. (D) Screen clamper circuit for protecting power tetrodes.

has not been too prolonged, emission sometimes may be restored by operating the filament at rated voltage with all other voltages removed for a period of 10 minutes, or at 20 percent above rated voltage for a few minutes.

Plate Voltage

Dc plate voltage for the operation of rf amplifiers is most often obtained from a transformer-rectifier-filter system (see power-supply chapter) designed to deliver the required plate voltage at the required current. However, batteries or other dc-generating devices are sometimes used in certain types of operation (see portable-mobile chapter).

Bias and Tube Protection

Several methods of obtaining bias are shown in Fig. 6-18. At A, bias is obtained by the voltage drop across a resistor in the grid dc return circuit when rectified grid current flows. The proper value of resistance may be determined by dividing the required biasing voltage by the dc grid current at which the tube will be operated. Then, so long as the rf driving voltage is adjusted so that the dc grid current is the recommended value, the biasing voltage will be the proper value. The tube is biased only when excitation is applied, since the voltage drop across the resistor depends upon grid-current flow. When excitation is removed, the bias falls to zero. At zero bias most tubes draw power far in excess of the plate-dissipation rating. So it is advisable to make provision for protecting the tube when excitation fails by accident, or by intent as it does when a preceding stage in a cw transmitter is keyed.

If the maximum cw ratings shown in the tube tables are to be used, the input should be cut to zero when the key is open. Aside from this, it is not necessary that plate current be cut off completely but only to the point where the rated dissipation is not exceeded. In this case plate-modulated phone ratings should be used for cw operation, however.

With triodes this protection can be supplied by obtaining all bias from a source of fixed voltage, as shown in Fig. 6-18B. It is preferable, however, to use only sufficient fixed bias to protect the tube and obtain the balance needed for operating bias from a grid leak. The grid-leak resistance is calculated as above, except that the fixed voltage is subtracted first.

Fixed bias may be obtained from dry batteries or from a power pack (see power-supply chapter). If dry batteries are used, they should be checked periodically, since even though they may show normal voltage, they eventually develop a high internal resistance.

In Fig. 16-8D, bias is obtained from the voltage drop across a Zener diode in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across VR1 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cutoff protective bias cannot be obtained.

The voltage of the cathode biasing Zener diode VR1 should be chosen for the value which will give the correct operating bias voltage with rated grid, plate and screen currents flowing with the amplifier loaded to rated input. When excitation is removed, the input to most types of tubes will fall to a value that will prevent damage to the tube, at least for the period of time required to remove plate voltage. A disadvantage of this biasing system is that the cathode rf connection to ground depends upon a bypass capacitor.

Screen Voltage

For cw and fm operation, and under certain conditions of phone operation (see amplitude-modulation chapter), the screen may be operated from a power supply of the same type used for plate supply, except that voltage and current ratings should be appropriate for screen requirements. The screen may also be operated through a series resistor or voltage-divider from a source of higher voltage, such as the plate-voltage supply, thus making a separate supply for the screen unnecessary. Certain precautions are necessary, depending upon the method used.

It should be kept in mind that screen current varies widely with both excitation and loading. If the screen is operated from a fixed-voltage source, the tube should never be operated without plate voltage and load, otherwise the screen may be damaged within a short time. Supplying the screen through a series dropping resistor from a higher-voltage source, such as the plate supply, affords a measure of protection, since the resistor causes the screen voltage to drop as the current increases, thereby limiting the power drawn by the screen. However, with a resistor, the screen voltage may vary considerably with excitation, making it necessary to check the voltage at the screen terminal under actual operating conditions to make sure that the screen voltage is normal. Reducing excitation will cause the screen current to drop, increasing the voltage; increasing excitation will have the opposite effect. These changes are in addition to those caused by changes in bias and plate loading, so if a screen-grid tube is operated from a series resistor or a voltage divider, its voltage should be checked as one of the final adjustments after excitation and loading have been set.

An approximate value for the screen-voltage dropping resistor may be obtained by dividing the voltage *drop* required from the supply voltage (difference between the supply voltage and rated screen voltage) by the rated screen current in decimal parts of an ampere. Some further adjustment may be necessary, as mentioned above, so an adjustable resistor with a total resistance above that calculated should be provided.

Protecting Screen-Grid Tubes

Considerably less grid bias is required to cut off an amplifier that has a fixed-voltage screen supply than one that derives the screen voltage through a high value of dropping resistor. When a "stiff" screen voltage supply is used, the necessary grid cutoff voltage may be determined from an inspection of the tube curves or by experiment.

When the screen is supplied from a series dropping resistor, the tube can be protected by the use of a clamper tube, as shown in Fig. 6-18D. The grid-leak bias of the amplifier tube with excitation is supplied also to the grid of the clamper tube. This is usually sufficient to cut off the clamper tube. However, when excitation is removed, the clamper-tube bias falls to zero and it draws enough current through the screen dropping resistor usually to limit the input to the amplifier to a safe value. If complete screen-voltage cutoff is desired, a Zener diode may be inserted in the screen lead. The regulator diode voltage rating should be high enough so that it will cease conducting when excitation is removed.

Feeding Excitation to the Grid

The required rf driving voltage is supplied by an oscillator generating a voltage at the desired frequency, either directly or through intermediate amplifiers, mixers, or frequency multipliers.

As explained in the chapter on vacuum-tube fundamentals, the grid of an amplifier operating under Class C conditions must have an exciting voltage whose peak value exceeds the negative biasing voltage over a portion of the excitation cycle. During this portion of the cycle, current will flow in the grid-cathode circuit as it does in a diode circuit when the plate of the diode is positive in respect to the cathode. This requires that the rf driver supply power. The power required to develop the required peak driving voltage across the grid-cathode impedance of the amplifier is the rf driving power.

The tube tables give approximate figures for the grid driving power required for each tube under various operating conditions. These figures, however, do not include circuit losses. In general, the driver stage for any Class C amplifier should be capable of supplying at least three times the driving power shown for typical operating conditions at frequencies up to 30 MHz and from three to ten times at higher frequencies.

Since the dc grid current relative to the biasing voltage is related to the peak driving voltage, the dc grid current is commonly used as a convenient indicator of driving conditions. A driver adjustment that results in rated dc grid current when the dc bias is at its rated value, indicates proper excitation to the amplifier when it is fully loaded.

In coupling the grid input circuit of an amplifier to the output circuit of a driving stage the objective is to load the driver plate circuit so that the desired amplifier grid excitation is obtained without exceeding the plate-input ratings of the driver tube.

Driving Impedance

The grid-current flow that results when the grid is driven positive in respect to the cathode over a portion of the excitation cycle represents an average resistance across which the exciting voltage must be developed by the driver. In other words, this is the load resistance into which the driver plate circuit must be coupled. The approximate grid input resistance is given by:

$$\textit{Input impedance}\ (\text{ohms}) = \frac{\textit{driving power (watts)}}{\textit{dc grid current}\ (\text{mA})^2} \times 620{,}000$$

For normal operation, the driving power and grid current may be taken from the tube tables. Since the grid input resistance is a matter of a few thousand ohms, an impedance step-up is necessary if the grid is to be fed from a low-impedance transmission line

TRANSISTOR RATINGS

Transistor ratings are similar in some respects to the maximum limits given for tubes. However, solid-state devices are generally not so forgiving of overload; they can quickly be ruined if a voltage or current parameter of the device is exceeded. All semiconductors undergo irreversible changes if their temperature is allowed to go above a critical limit.

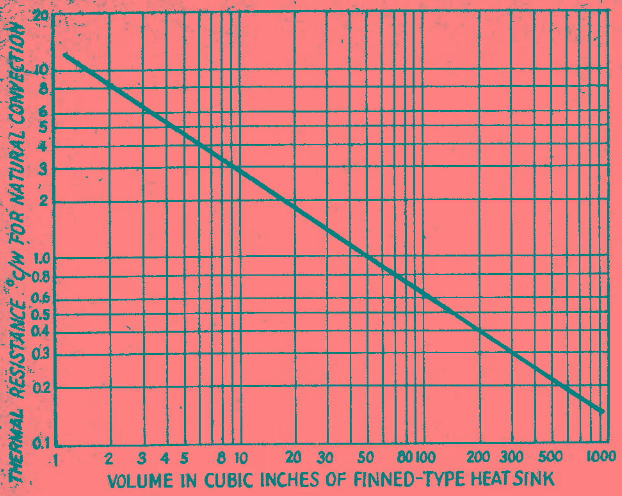
Voltage Rating

In general, the higher the collector-emitter voltage rating of a transistor the less the chance of damage when used as an rf power amplifier. A mismatched load, or the loss of the load entirely, causes high voltages to appear between the collector and emitter of the transistor. If the maximum rating is exceeded, the transistor may break down and pass reverse current. Transistor manufacturers are now including a resistance in series with the emitter lead of each of the many junctions that make up the power transistor as break-down protection. This technique is called ballasting or balanced emitters. Another way to protect a power transistor is to include a Zener diode from collector to emitter. The break-down voltage rating of the diode should be above the peak rf voltage to be developed in the circuit, but below the maximum rating of the power device.

Current and Heat

The current that a power device can stand is related to its ability to dissipate heat. A transistor is physically small, so high-power models must use effective heat radiators, called heat sinks, to insure that the operating temperature is kept to a moderate value even when large currents are flowing through the device.

Cooling considerations for practical solid-state amplifiers are outlined below. Manufacturer's specification sheets describe a safe operating area for an individual power transistor. Also, transistors are rated in terms of power output, rather than input, so it should be remembered that a device specified to deliver 80 watts of output power will probably be running 160 watts or more input. Transistor amplifiers pass an appreciable amount of driver power to the output, as do grounded-grid tube stages, and this fact must also be taken into account by the circuit designer.



(A)

Device Case	5W	10W	25W	50W	100W
TO-5	17.2	7.2	1.2	.71	.35
TO-44	1.2	9.2	.44	n/a	n/a

(B)

Fig. 6-19 — (A) Graph to determine the thermal resistance of a heat sink of a given size. The heat sink volume may be computed by multiplying cross-sectional area by height. (B) Approximate thermal resistance needed for proper cooling of two types of transistor cases when operated at the proper levels given.

COOLING

Tubes

Vacuum tubes must be operated within the temperature range specified by the manufacturer if long tube life is to be achieved. Tubes with glass envelopes rated at up to 25 watts of plate dissipation may be run without forced-air cooling, if a moderate amount of cooling by convection can be arranged. If a cane-metal enclosure is used, and a ring of 1/4-inch diameter holes are placed around the tube socket, normal air flow can be relied upon to remove excess heat at room temperatures.

For tubes with greater plate dissipation, or those operated with plate currents in excess of the manufacturer's ratings (often the case with TV sweep tubes) forced air cooling with a fan or blower is needed. Fans, especially those designed for cooling hi-fi cabinets, are preferred because they operate quietly. However, all fans lose their ability to move air when excessive back pressure exists. For applications where a stream of air must be directed through a tube socket, a blower is usually required. Blowers vary in their ability to work against back pressure, so this specification should be checked when selecting a particular model. Some air will always leak around the socket and through other holes in a chassis, so the blower chosen should have a capacity which is 30 to 50 percent beyond that called for by the tube manufacturer.

An efficient blower is required when using the external-anode tubes, such as the 4X150A. Such tubes represent a trade-off which allows high-power operation with a physically small device at the expense of increased complexity in the cooling system. Other types of external-anode tubes are now being produced for conductive cooling. An electrical insulator which is also an excellent thermal conductor, such as AlSiMag, couples the tube to a heat sink. Requirements for the heat dissipator are calculated in the same way as for power transistors, as outlined below. Similar tubes are made with special anode structures for water or vapor cooling, allowing high-power operation without producing an objectionable noise level from the cooling system.

Transistor Cooling

Bipolar power transistors usually have the collector connected directly to the case of the device, as the collector must dissipate most of the heat generated when the transistor is in operation. However, even the larger case designs cannot conduct heat away fast enough to keep the operating temperature of the device functioning within the safe area, the maximum temperature that a device can stand without damage. Safe area is usually specified in a device data sheet, often in graphical form. Germanium power transistors may be operated at up to 100 degrees C while the silicon types may be run at up to 200 degrees C. Leakage currents in germanium devices can be very high at elevated temperatures; thus, for power applications silicon transistors are preferred.

A thermal sink, properly chosen, will remove heat at a rate which keeps the transistor junction temperature in the safe area. For low-power applications a simple clip-on heat sink will suffice, while for 100-watts of input power a massive cast-aluminum finned radiator will be necessary. In general, the case temperature of a power transistor must be kept below the point at which it will produce a burn when touched.

Heat-Sink Design

Simple heat sinks, made as described in the Construction Practices chapter, can be made more effective (by 25 percent or more) by applying a coat of flat-black paint. Finned radiators are most effective when placed where maximum air flow can be achieved — outside a case with the fins placed vertically. The size of a finned heat sink required to give a desired thermal resistance, a measure of the ability to dissipate heat, is shown in Fig. 6-19A. Fig. 6-19B is a simplified chart of the thermal resistance needed in a heat sink for transistors in TO-5 and TO-44 cases. These figures

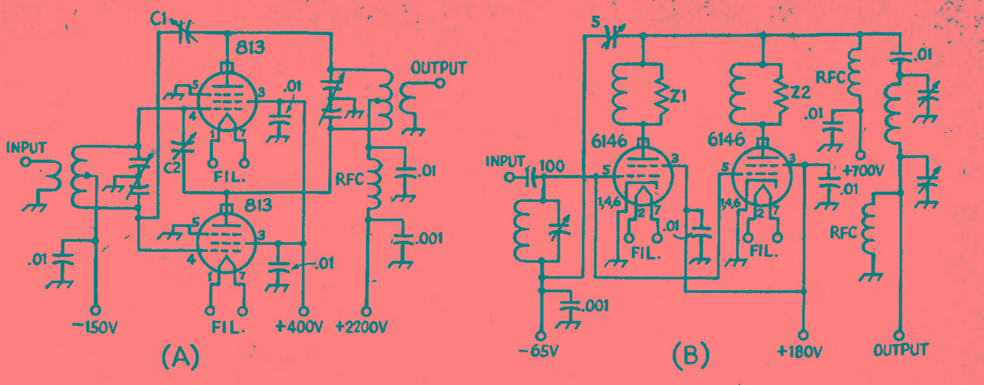


Fig. 6-20 — Typical (A) push-pull and (B) parallel amplifier circuits.

are based on several assumptions, so they can be considered a *worst-case* situation. Smaller heat sinks may be usable.

The thermal design of solid-state circuits has been covered in *QST* for April, 1972. The surface contact between the transistor case and the heat sink is extremely important. To keep the sink from being "hot" with dc, a mica insulator is usually employed between the transistor case and the heat dissipator. Newer types of transistors have a case mounting bolt insulated from the collector so that it may be connected directly to the heat sink. Whatever the arrangement, the use of a conductive compound such as silicone grease (Corning PC-4) is recommended between the transistor and the sink. For high-power designs, it may be desirable to add a small cooling fan, providing a stream of air across the heat sink, to keep the size of the heat dissipator within reasonable limits. Even a light air flow greatly increases the radiator's ability to dispose of excess heat.

OUTPUT POWER FROM TRANSMITTERS

CW or FM: In a cw or fm transmitter, any class of amplifier can be used as an output or intermediate amplifier. (For reasonable efficiency, a frequency multiplier *must* be operated Class C.) Class-C operation of the amplifier gives the highest efficiency (65 to 75 percent), but it is likely to be accompanied by appreciable harmonics and consequent TVI possibilities. If the excitation is keyed in a cw transmitter, Class-C operation of subsequent amplifiers will, under certain conditions, introduce key clicks not present on the keyed excitation (see chapter on Code Transmission). The peak envelope power (PEP) input or output of any cw (or fm) transmitter is the "key-down" input or output.

A-M: In an amplitude-modulated phone transmitter, plate modulation of a Class-C output amplifier results in the highest output for a given input to the output stage. The efficiency is the same as for cw or fm with the same amplifier, from 65 to 75 percent. (In most cases the manufacturer rates the *maximum allowable input* on plate-modulated phone at about 2/3 that of cw or fm.) A plate-modulated stage running 100 watts input will deliver a carrier output of from 65 to 75 watts, depending upon the tube, frequency and circuit factor. The PEP output of any a-m signal is four times the carrier output power, or 260 to 300 watts for the 100-watt input example.

Grid- (control or screen) modulated output amplifiers in a-m operation run at a carrier efficiency of 30 to 35 percent, and a grid-modulated stage with 100 watts input has a carrier output of 30 to 35 watts. (The PEP output, four times the carrier output, is 120 to 140 watts.)

Running the legal input limit in the United States, a plate-modulated output stage can deliver a carrier output of 650 to 750 watts, while a screen- or control-grid-modulated output amplifier can deliver only a carrier of 300 to 350 watts.

SSB: Only *linear* amplifiers can be used to amplify ssb signals without distortion, and this limits the choice of output amplifier operation to Classes A, AB_1, AB_2, and B. The efficiency of operation of these amplifiers runs from about 20 to 65 percent. In all but Class-A operation the indicated (by plate-current meter) input will vary with the signal, and it is not possible to talk about relative inputs and outputs as readily as it is with other modes. Therefore linear amplifiers are rated by PEP (input or output) at a given distortion level, which indicates not only how much ssb signal they will deliver but also how effective they will be in amplifying an a-m signal.

LINEAR AMPLIFIERS FOR A-M: In considering the practicality of adding a linear output amplifier to an existing a-m transmitter, it is necessary to know the carrier output of the a-m transmitter and the PEP output rating of the linear amplifier. Since the PEP output of an a-m signal is four times the carrier output, it is obvious that a linear with a PEP output rating of only four times the carrier output of the a-m transmitter is no amplifier at all. If the linear amplifier has a PEP output rating of 8 times the a-m transmitter carrier output, the output power will be doubled and a 3-dB improvement will be obtained. In most cases a 3-dB change is *just discernible* by the receiving operator.

By comparison, a linear amplifier with a PEP output rating of four times an existing ssb, cw or fm transmitter will *quadruple* the output, a 6-dB improvement, It should be noted that the linear amplifier must be rated for the mode (ssb, cw or fm) with which it is to be used.

GROUNDED-GRID AMPLIFIERS: The preceding discussion applies to vacuum-tube amplifiers connected in a grounded-cathode or grounded-grid circuit. However, there are a few points that apply only to grounded-grid amplifiers.

A tube operated in a given class (AB_1, B, C) will require more driving power as a grounded-grid amplifier than as a grounded-cathode amplifier. This is not because the grid losses run higher in the grounded-grid configuration but because some of the driving power is coupled directly through the tube and appears in the plate load circuit. Provided enough driving power is available, this increased requirement is of no concern in cw or linear operation. In a-m operation, however, the fed-through power prevents the grounded-grid amplifier from being fully modulated (100 percent).

AMPLIFIER CIRCUITS

Parallel and Push-Pull Amplifiers

The circuits for parallel-tube amplifiers are the same as for a single tube, similar terminals of the tubes being connected together. The grid impedance of two tubes in parallel is half that of a single tube. This means that twice the grid tank capacitance shown in Fig. 6-20B should be used for the same *Q*.

The plate load resistance is halved so that the plate-tank capacitance for a single tube (Fig. 6-24) also should be doubled. The total grid current will be doubled, so to maintain the same grid bias, the grid-leak resistance should be half that used for a single tube. The required driving power is doubled. The capacitance of a neutralizing capacitor should be doubled and the value of the screen dropping resistor should be cut in half.

In treating parasitic oscillation, it is often necessary to use a choke in each plate lead, rather than one in the common lead to avoid building in a push-pull type of vhf circuit, a factor in obtaining efficient operation at higher frequencies.

Two or more transistors are often operated in parallel to achieve high output power, because several medium-power devices often cost less than a single high-power type. When parallel operation is used, precautions must be taken to insure that equal drive is applied to each transistor. Otherwise, one transistor may "hog" most of the drive and exceed its safe ratings.

A basic push-pull circuit is shown in Fig. 6-20A. Amplifiers using this circuit are cumbersome to bandswitch and consequently are not very popular below 30 MHz. However, since the push-pull configuration places tube input and output capacitances in series, the circuit is often used at 50 MHz and higher.

In the circuit shown at A two 813s are used. Cross neutralization is employed, with C1 connected from the plate of one tube to the grid of the second, while C2 is attached in the reverse order.

GROUNDED-GRID AMPLIFIERS

Fig. 6-21A shows the input circuit of a grounded-grid triode amplifier. In configuration it is similar to the conventional grounded-cathode circuit except that the grid, instead of the cathode, is at ground potential. An amplifier of this type is characterized by a comparatively low input impedance and a relatively high driver power requirement. The additional driver power is not consumed in the amplifier but is "fed through" to the plate circuit where it combines with the normal plate output power. The total rf power output is the sum of the driver and amplifier output powers less the power normally required to drive the tube in a grounded-cathode circuit.

Positive feedback is from plate to cathode through the plate-cathode capacitance of the tube. Since the grounded-grid is interposed between the plate and cathode, this capacitance is small, and neutralization usually is not necessary.

In the grounded-grid circuit the cathode must be isolated for rf from ground. This presents a practical difficulty especially in the case of a filament-type tube whose filament current is large. In plate-modulated phone operation the driver power fed through to the output is not modulated.

The chief application for grounded-grid amplifiers in amateur work below 30 MHz is in the case where the available driving power far exceeds the power that can be used in driving a conventional grounded-cathode amplifier.

Screen-grid tubes are also used sometimes in grounded-grid amplifiers. In some cases, the screen

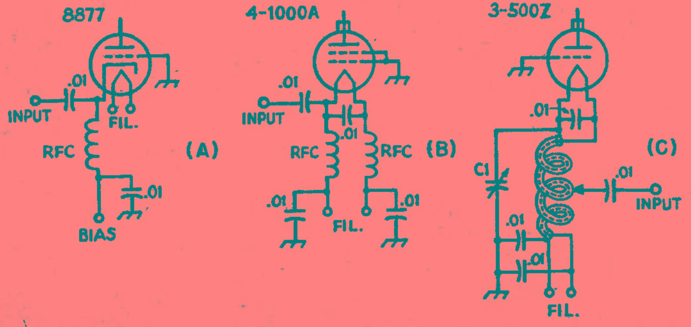


Fig. 6-21 — Input circuits for triode or triode-connected power tubes operated grounded grid.

Fig. 6-22 — A 30-A filement choke for a grounded-grid power amplifier consisting of 28 turns of No. 10 enam. wire on a 1/2-inch diameter ferrite rod 7 inches long.

is simply connected in parallel with the grid, as in Fig. 6-21B and the tube operates as a high-μ triode. In other cases, the screen is bypassed to ground and operated at the usual dc potential, as shown at C. Since the screen is still in parallel with the grid for rf, operation is very much like that of a triode except that the positive voltage on the screen reduces driver-power requirements.

In indirectly-heated cathode tubes, the low heater-to-cathode capacitance will often provide enough isolation to keep rf out of the heater transformer and the ac lines. If not, the heater voltage must be applied through rf chokes.

In a directly-heated cathode tube, the filament must be maintained above rf ground. This can be done by using a pair of filament chokes or by using the input tank circuit, as shown in Fig. 6-21C. In the former method, a double solenoid (often wound on a ferrite core) is generally used, although separate chokes can be used. When the tank circuit is used, the tank inductor is wound from two (insulated) conductors in parallel or from an insulated conductor inside a tubing outer conductor. A typical filament choke is shown in Fig. 6-22.

The input impedance of a grounded-grid power stage is usually between 30 and 150 ohms. For circuits similar to those shown in Figs. 6-21A and B some form of input tuning network is needed. A high-*C*, low-*Q* parallel-resonant or pi-section network will suffice. The input network provides benefit other than impedance matching — a reduction in the IM distortion produced by the stage when amplifying an ssb signal. A typical input circuit is shown in Fig. 6-16F. When an amplifier is built for single-band operation, a tank circuit similar to that shown in Fig. 6-21C may be employed. Proper input matching is achieved by tapping the input down on the coil.

TRANSISTOR CIRCUITS

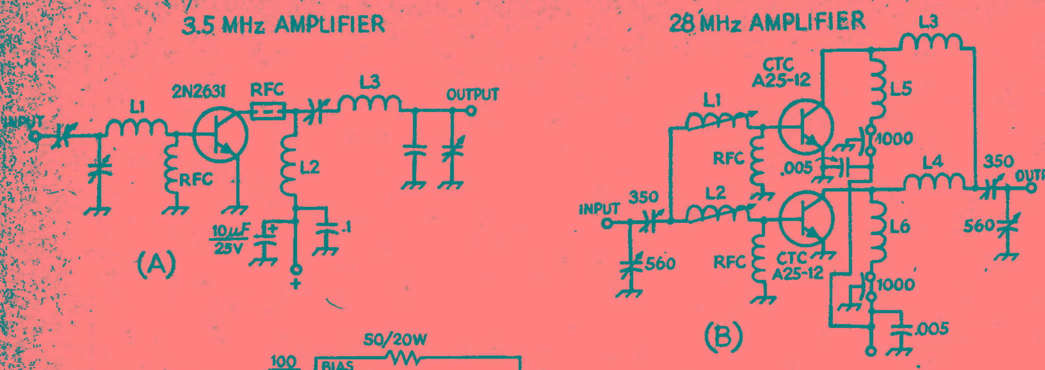
A transistor amplifier requires some means for impedance matching at the input and output of the stage. For conventional narrow-band amplifier designs, impedance matching is achieved with tuned networks (pi, L or T sections or combinations thereof). To simplify band-switching requirements, broadband amplifiers with four octaves or more of bandwidth are desirable. Wide bandwidths are achieved by using a special form of transmission-line transformer for interstage and output coupling that is described later in this chapter.

Most solid-state Class-C amplifiers are operated with both the base and emitter leads connected to dc ground. Thus, the transistor is practically off when no driving signal is present. The distortion of the drive signal by such an amplifier is appreciable. However, with cw, fm, or collector-modulated a-m, the harmonics produced are removed from the desired frequency by at least a factor of 2. Thus, harmonic energy can be reduced or eliminated by using appropriate filters.

Fig. 6-23A shows a basic Class-C transistor amplifier. The base input is held at dc ground through a radio-frequency choke. A second choke, consisting of two ferrite beads (collector lead), eliminates a tendency to vhf parasitic oscillation. At B, parallel-connected transistors are operated Class C. Adjustment of L1 and L2 provide equal levels of drive. The devices chosen for this circuit are designed for 30- to 50-MHz operation. Below 14 MHz some form of degenerative feedback will be needed to prevent self oscillation, as the gain of the transistors is quite high at lower frequencies.

For ssb operation transistors must be forward biased at the base. The lowest distortion results with Class-A operation, but, efficiency is poor. The best trade off between low distortion and high efficiency is Class-B operation, even though operation in this region introduces some severe requirements for the bias circuit. Whenever a transistor is forward biased, thermal runaway can be a problem. Also, ssb drive varies in amplitude causing large variations in the transistor base current. For best linearity, the dc base-bias voltage should remain constant as the rf drive level is varied. This situation is in conflict with the conditions needed to prevent thermal runaway. Exotic schemes have been designed to provide the proper base bias for Class-B ssb amplification. However, a simple diode circuit such as shown in Figs. 6-23C and D can provide the required dc stability with protection against thermal damage. The ballasted type of transistors are preferred for these circuits. Typical choices for Class-B ssb service are the 2N5941, 2N2942, 2N3375, 2N5070, 2N5071, and the 2N5093. The design of suitable broadband transformers for the circuits of Fig. 6-29 is covered later in this chapter.

The circuits at 6-23C and D are similar except for the choice of the active device. Both designs were developed by K7QWR. The base-bias circuit maintains a steady voltage while supplying current that varies by a factor of 100 to 1 with drive. The gain versus frequency of both circuits follows the



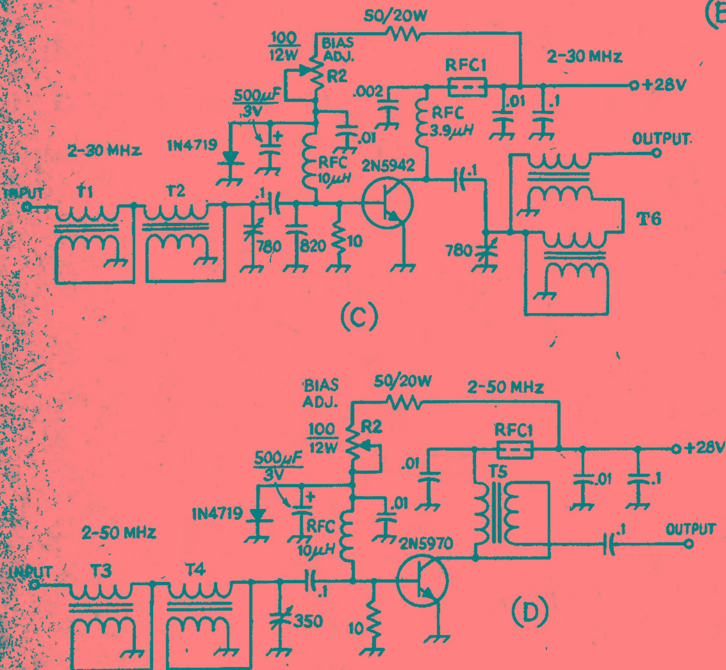


Fig. 6-23 – Some typical transistor power-amplifier circuits. At C, R1 is adjusted for a collector current of 40 mA with no drive, while R2 at D is set for 20 mA collector current with no input. Broadband transformers used consist of the following:

T1, T3, T5 – 6 turns of 2 twisted pairs of No. 26 enam. wire on a Stackpole 57-9322 No. 11 toroid core, connected for 4:1. (See table 6-A.)

T2, T4 – 4 turns of 4 twisted pairs of No. 26 enam. wire on a Stackpole 57-9322 No. 11 toroid core, connected for 4:1.

T6 – 10 turns of 3 twisted pairs of No. 28 enem. wire on two Stackpole 57-9074 No. 11 toroid cores, connected for 9:1

power-output curves of the transistors used, changing from 25 dB at 2 MHz to 13 dB at 30 MHz. IMD is typically 30 dB or more down with either circuit.

RF POWER-AMPLIFIER TANKS AND COUPLING

TANK Q

Rf power amplifiers used in amateur transmitters are operated under Class-C or -AB conditions (see chapter on tube fundamentals). The main objective, of course, is to deliver as much fundamental power as possible into a load, R without exceeding the tube ratings. The load resistance R may be in the form of a transmission line to an antenna, or the input circuit of another amplifier. A further objective is to minimize the harmonic energy (always generated by an amplifier) fed into the load circuit. In attaining these objectives, the Q of the tank circuit is of importance. When a load is coupled inductively, the Q of the tank circuit will have an effect on the coefficient of coupling necessary for proper loading of the amplifier. In respect to all of these factors, a tank Q of 10 to 20 is usually considered optimum. A much lower Q will result in less efficient operation of the amplifier tube, greater harmonic output, and greater difficulty in coupling inductively to a load. A much higher Q will result in higher tank current with increased loss in the tank coil. Efficiency of a tank circuit is determined by the ratio of loaded Q to unloaded Q by the relationship:

$$Eff. = 100\left(1 - \frac{Q_L}{Q_U}\right)$$

where Q_L is the loaded Q and Q_U is the unloaded Q.

The Q is determined (see chapter on electrical laws and circuits) by the L/C ratio and the load resistance at which the tube is operated. The tube load resistance is related, in approximation, to the ratio of the dc plate voltage to dc plate current at which the tube is operated and can be computed from:

Class-A Tube:

$$R_L = \frac{\text{Plate Volts}}{1.3 \times \text{Plate Current}}$$

Class-B Tube:

$$R_L = \frac{\text{Plate Volts}}{1.57 \times \text{Plate Current}}$$

Class-C Tube:

$$R_L = \frac{\text{Plate Volts}}{2 \times \text{Plate Current}}$$

Transistor:

$$R_L = \frac{(\text{Collector Volts})^2}{2 \times \text{Power Output (Watts)}}$$

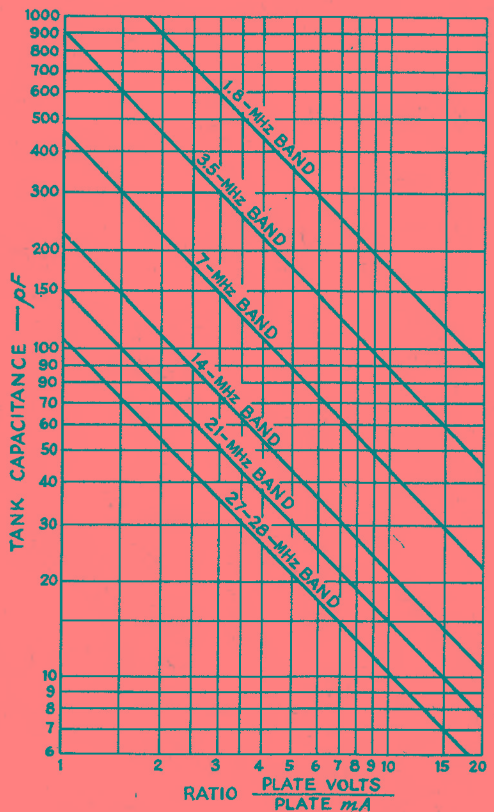


Fig. 6-24 — Chart showing plate tank capacitance required for a Q of 10. Divide the tube plate voltage by the plate current in milliamperes. Select the vertical line corresponding to the answer obtained. Follow this vertical line to the diagonal line for the band in question, and thence horizontally to the left to read the capacitance. For a given ratio of plate voltage/plate current, doubling the capacitance shown doubles the Q. When a split-stator capacitor is used in a balanced circuit, the capacitance of each section may be one half the value given by the chart.

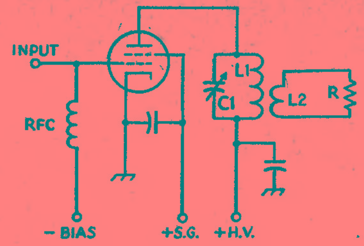


Fig. 6-25 — Inductive-link output coupling circuits.

C1 — Plate tank capacitor — see text and Fig. 6-24 for capacitance.

L1 — To resonate at operating frequency with C1. See *LC* chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator.*

L2 — Reactance equal to line impedance. See reactance chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator.*

R — Representing load.

Parallel-Resonant Tank

The amount of C that will give a Q of 10 for various ratios is shown in Fig. 6-24. For a given plate-voltage/plate-current ratio, the Q will vary directly as the tank capacitance, twice the capacitance doubles the Q, etc. For the same Q, the capacitance of *each section* of a split-stator capacitor in a balanced circuit should be half the value shown.

These values of capacitance include the output capacitance of the amplifier tube, the input capacitance of a following amplifier tube if it is coupled capacitively, and all other stray capacitances. At the higher plate-voltage/plate-current ratios, the chart may show values of capacitance, for the higher frequencies, smaller than those attainable in practice. In such a case, a tank Q higher than 10 is unavoidable.

INDUCTIVE-LINK COUPLING

Coupling to Flat Coaxial Lines

When the load R in Fig. 6-25 is located for convenience at some distance from the amplifier, or when maximum harmonic reduction is desired, it is advisable to feed the power to the load through a low-impedance coaxial cable. The shielded construction of the cable prevents radiation and makes it possible to install the line in any convenient manner without danger of unwanted coupling to other circuits.

If the line is more than a small fraction of a wavelength long, the load resistance at its output end should be adjusted, by a matching circuit if necessary, to match the impedance of the cable. This reduces losses in the cable and makes the coupling adjustments at the transmitter independent of the cable length. Matching circuits for use between the cable and another transmission line are discussed in the chapter on transmission lines, while the matching adjustments when the load is the grid circuit of a following amplifier are described elsewhere in this chapter.

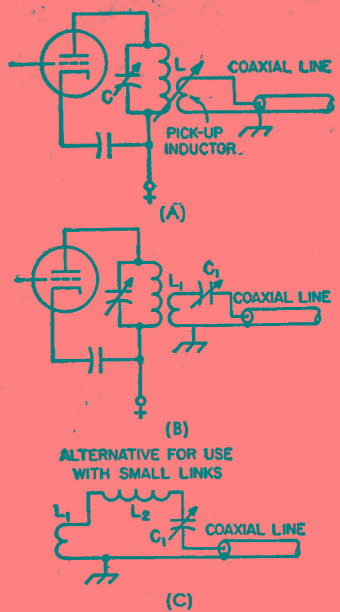


Fig. 6-26 – With flat transmission lines, power transfer is obtained with looser coupling if tha line input is tuned to resonance. C1 and L1 should resonate at the operating frequency. See table for maximum usable value of C1. If circuit does not resonate with maximum C1 or less, inductance of L1 must be increased or added in series et L2.

Tabla 6-A

Capacitance in pF Required for Coupling to Flat Coaxial Lines with Tuned Coupling Circuit[1]

Frequency Band MHz	*Characteristic Impedance of Line* 52 *ohms*	75 *ohms*
3.5	450	300
7	230	150
14	115	75
21	80	50
28	60	40

[1] Capacitance values are maximum usable.
Note: Inductance in circuit must be adjusted to resonate at operating frequency.

Assuming that the cable is properly terminated, proper loading of the amplifier will be assured, using the circuit of Fig. 6-26A, if

1) The plate tank circuit has reasonably higher value of Q. A value of 10 is usually sufficient.

2) The inductance of the pickup or link coil is close to the optimum value for the frequency and type of line used. The optimum coil is one whose self-inductance is such that its reactance at the operating frequency is equal to the characteristic impedance, Z_0, of the line.

3) It is possible to make the coupling between the tank and pickup coils very tight.

The second in this list is often hard to meet. Few manufactured link coils have adequate inductance even for coupling to a 50-ohm line at low frequencies.

If the line is operating with a low SWR, the system shown in Fig. 6-26A will require tight coupling between the two coils. Since the secondary (pickup coil) circuit is not resonant, the leakage reactance of the pickup coil will cause some detuning of the amplifier tank circuit. This detuning effect increases with increasing coupling, but is usually not serious. However, the amplifier tuning must be adjusted to resonance, as indicated by the plate-current dip, each time the coupling is changed.

Tuned Coupling

The design difficulties of using "untuned" pickup coils, mentioned above, can be avoided by using a coupling circuit tuned to the operating frequency. This contributes additional selectivity as well, and hence aids in the suppression of spurious radiations.

If the line is flat the input impedance will be essentially resistive and equal to the Z_0 of the line. With coaxial cable, a circuit of reasonable Q can be obtained with practicable values of inductance and capacitance connected in series with the line's input terminals. Suitable circuits are given in Fig. 6-26 at B and C. The Q of the coupling circuit often may be as low as 2, without running into difficulty in getting adequate coupling to a tank circuit of proper design. Larger values of Q can be used and will result in increased ease of coupling, but as the Q is increased the frequency range over which the circuit will operate without readjustment becomes smaller. It is usually good practice, therefore, to use a coupling-circuit Q just low enough to permit operation, over as much of a band as is normally used for a particular type of communication, without requiring retuning.

Capacitance values for a Q of 2 and line impedances of 52 and 75 ohms are given in the accompanying table. These are the *maximum* values that should be used. The inductance in the circuit should be adjusted to give resonance at the

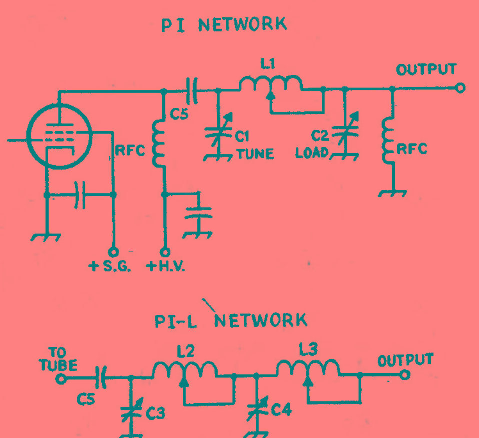


Fig. 6-27 – Pi and pi-L output-coupling networks.

operating frequency. If the link coil used for a particular band does not have enough inductance to resonate, the additional inductance may be connected in series as shown in Fig. 6-26C.

Characteristics

In practice, the amount of inductance in the circuit should be chosen so that, with somewhat loose coupling between L1 and the amplifier tank coil, the amplifier plate current will increase when the variable capacitor, C1, is tuned through the value of capacitance given by the table. The coupling between the two coils should then be increased until the amplifier loads normally, without changing the setting of C1. If the transmission line is flat over the entire frequency band under consideration, it should not be necessary to readjust C1 when changing frequency, if the values given in the table are used. However, it is unlikely that the line actually will be flat over such a range, so some readjustment of C1 may be needed to compensate for changes in the input impedance of the line. If the input impedance variations are not large, C1 may be used as a loading control, no changes in the coupling between L1 and the tank coil being necessary.

The degree of coupling between L1 and the amplifier tank coil will depend on the coupling-circuit Q. With a Q of 2, the coupling should be tight – comparable with the coupling that is typical of "fixed-link" manufactured coils. With a swinging link it may be necessary to increase the Q of the coupling circuit in order to get sufficient power transfer. This can be done by increasing the L/C ratio.

PI AND PI-L OUTPUT TANKS

A pi-section and pi-L tank circuit may also be used in coupling to an antenna or transmission line, as shown in Fig. 6-27. The optimum values of capacitance and inductance are dependent upon values of amplifier power input and output load resistance.

Values for L and C may be taken directly from the charts of Fig. 6-28 if the output load resistance is the usual 52 ohms. It should be borne in mind that these values apply only where the output load is resistive, i.e., where the antenna and line have been matched. Fig. 6-28 and 6-28A were provided by W6FFC.

Output-Capacitor Ratings

The voltage rating of the output capacitor will depend upon the SWR. If the load is resistive, receiving-type air capacitors should be adequate for amplifier input powers up to 2 kW PEP when

	TUBE LOAD IMPEDANCE (OPERATING Q)									
	MHz	1500(12)	2000(12)	2500(12)	3000(12)	3500(12)	4000(12)	5000(13)	6000(14)	8000(16)
C1	3.5	420	315	252	210	180	157	126	114	99
	7	190	143	114	95	82	71	57	52	45
	14	93	70	56	47	40	35	28	25	22
	21	62	47	37	31	27	23	19	17	15
	28	43	32	26	21	18	16	13	12	10
C2	3.5	2117	1776	1536	1352	1203	1079	875	862	862
	7	942	783	670	583	512	451	348	341	341
	14	460	382	326	283	247	217	165	162	162
	21	305	253	216	187	164	144	109	107	107
	28	210	174	148	128	111	97	72	70	70
L1	3.5	5.73	7.46	9.17	10.86	12.53	14.19	17.48	19.18	21.98
	7	3.14	4.09	5.03	5.95	6.86	7.77	9.55	10.48	12.02
	14	1.60	2.08	2.56	3.03	3.49	3.95	4.85	5.33	6.11
	21	1.07	1.39	1.71	2.02	2.34	2.64	3.25	3.56	4.09
	28	0.77	1.01	1.24	1.46	1.69	1.91	2.34	2.57	2.95

	TUBE LOAD IMPEDANCE (OPERATING Q)									
	MHz	1500(12)	2000(12)	2500(12)	3000(12)	3500(12)	4000(12)	5000(12)	6000(12	8000(12)
C3	3.5	406	305	244	203	174	152	122	102	76
	7	188	141	113	94	81	71	56	47	35
	14	92	69	55	46	40	35	28	23	17
	21	62	46	37	31	26	23	18	15	12
	28	43	32	26	21	18	16	13	11	8
C4	3.5	998	859	764	693	638	593	523	472	397
	7	430	370	329	298	274	255	225	203	171
	14	208	179	159	144	133	123	109	98	83
	21	139	119	106	96	89	82	73	65	55
	28	95	81	72	66	60	56	50	45	38
L2	3.5	7.06	9.05	10.99	12.90	14.79	16.67	20.37	24.03	31.25
	7	3.89	4.97	6.03	7.07	8.10	9.12	11.13	13.11	17.02
	14	1.99	2.54	3.08	3.61	4.13	4.65	5.68	6.69	8.68
	21	1.33	1.69	2.05	2.41	2.76	3.10	3.78	4.46	5.78
	28	0.96	1.22	1.48	1.74	1.99	2.24	2.73	3.22	4.17
L3	3.5	4.45	4.45	4.45	4.45	4.45	4.45	4.45	4.45	4.45
	7	2.44	2.44	2.44	2.44	2.44	2.44	2.44	2.44	2.44
	14	1.24	1.24	1.24	1.24	1.24	1.24	1.24	1.24	1.24
	21	0.83	0.83	0.83	0.83	0.83	0.83	0.83	0.83	0.83
	28	0.60	0.60	0.60	0.60	0.60	0.60	0.60	0.60	0.60

Fig. 6-28 – Chart to determine the values of L and C needed for e pi (A) and pi-L (B) network to match a range of input impedances to a 50-ohm load.

R1 Ohms	F MHz	C1 pF	L1 µH	C2 pF	R2 Ohms	Q Qual.
50	3.5	2600	0.94	4153	10	2.9
50	7.0	1179	0.49	1678	10	2.6
50	14.0	579	0.25	801	10	2.5
50	21.0	384	0.16	528	10	2.5
50	29.7	266	0.12	351	10	2.5
50	3.5	2098	1.27	2811	20	2.3
50	7.0	952	0.67	1220	20	2.1
50	14.0	467	0.34	593	20	2.1
50	21.0	310	0.23	393	20	2.0
50	29.7	214	0.16	268	20	2.0
50	3.5	2098	1.43	2533	30	2.3
50	7.0	952	0.76	1131	30	2.1
50	14.0	467	0.38	553	30	2.1
50	21.0	310	0.26	367	30	2.0
50	29.7	214	0.18	253	30	2.0
50	3.5	2098	1.55	2290	40	2.3
50	7.0	952	0.83	1033	40	2.1
50	14.0	467	0.42	506	40	2.1
50	21.0	310	0.28	336	40	2.0
50	29.7	214	0.20	232	40	2.0
50	3.5	2098	1.66	2098	50	2.3
50	7.0	952	0.88	952	50	2.1
50	14.0	467	0.45	467	50	2.1
50	21.0	310	0.30	310	50	2.0
50	29.7	214	0.21	214	50	2.0
50	3.5	2098	1.66	2098	50	2.3
50	7.0	952	0.88	952	50	2.1
50	14.0	467	0.45	467	50	2.1
50	21.0	310	0.30	310	50	2.0
50	29.7	214	0.21	214	50	2.0
75	3.5	1399	2.21	1630	50	2.3
75	7.0	634	1.17	731	50	2.1
75	14.0	311	0.59	358	50	2.1
75	21.0	207	0.40	238	50	2.0
75	29.7	143	0.28	164	50	2.0
100	3.5	1049	2.72	1337	50	2.3
100	7.0	476	1.43	591	50	2.1
100	14.0	234	0.72	288	50	2.1
100	21.0	155	0.48	191	50	2.0
100	29.7	107	0.35	131	50	2.0

R1 Ohms	F MHz	C1 pF	L1 µH	C2 pF	R2 Ohms	Q Qual.
125	3.5	839	3.19	1124	50	2.3
125	7.0	381	1.67	488	50	2.1
125	14.0	187	0.84	237	50	2.1
125	21.0	124	0.56	157	50	2.0
125	29.7	86	0.40	107	50	2.0
150	3.5	699	3.62	957	50	2.3
150	7.0	317	1.89	405	50	2.1
150	14.0	156	0.95	196	50	2.1
150	21.0	103	0.64	129	50	2.0
150	29.7	71	0.45	88	50	2.0
175	3.5	599	4.03	816	50	2.3
175	7.0	272	2.09	333	50	2.1
175	14.0	133	1.05	159	50	2.1
175	21.0	89	0.70	105	50	2.0
175	29.7	61	0.50	70	50	2.0
200	3.5	569	4.26	822	50	2.5
200	7.0	258	2.22	334	50	2.3
200	14.0	127	1.12	160	50	2.2
200	21.0	84	0.74	105	50	2.2
200	29.7	58	0.53	70	50	2.2
225	3.5	543	4.48	827	50	2.7
225	7.0	246	2.34	335	50	2.4
225	14.0	121	1.18	160	50	2.4
225	21.0	80	0.79	106	50	2.4
225	29.7	55	0.56	70	50	2.3
250	3.5	520	4.68	831	50	2.9
250	7.0	236	2.45	336	50	2.6
250	14.0	116	1.23	160	50	2.5
250	21.0	77	0.82	106	50	2.5
250	29.7	53	0.59	70	50	2.5
275	3.5	499	4.86	834	50	3.0
275	7.0	227	2.56	336	50	2.7
275	14.0	111	1.29	160	50	2.7
275	21.0	74	0.86	106	50	2.7
275	29.7	51	0.61	70	50	2.6
300	3.5	481	5.04	836	50	3.2
300	7.0	218	2.66	337	50	2.9
300	14.0	107	1.34	160	50	2.8
300	21.0	71	0.89	106	50	2.8
300	29.7	49	0.64	70	50	2.8

Fig. 6-28A — The following data is for a pi network with a *Q* of 2 at the top of each band. Tha *Q* shown is thet for the same inductor at the bottom of the band. The capacitors are shown for the bottom of the band to indicate the maximum capacitance needed. If the transformation ratio exceeds 70 percent of maximum, the *Q* has baen automatically recalculated in order to retain the characteristics of a pi network and that new value shown. Do not forget which end of the network represents 50 ohms!

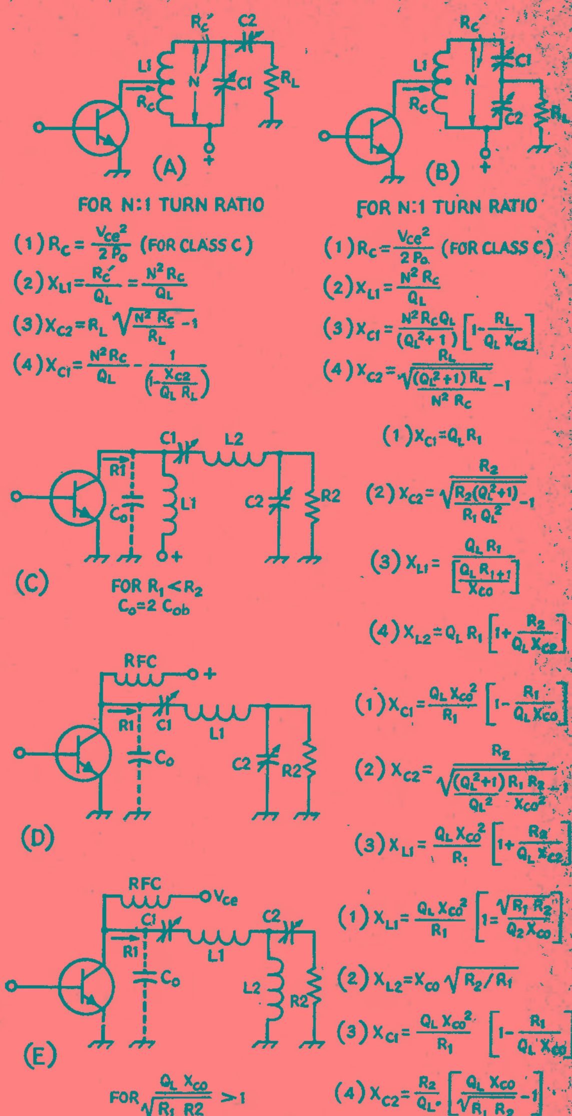
feeding 52- 75-ohm loads. In obtaining the larger capacitances required for the lower frequencies, it is common practice to switch one or more fixed capacitors in parallel with the variable air capacitor. While the voltage rating of a mica or ceramic capacitor may not be exceeded in a particluar case, capacitors of these types are limited in current-carrying capacity. Postage-stamp silver-mica capacitors should be adequate for amplifier inputs over the range from about 70 watts at 28 MHz to 400 watts at 14 MHz and lower. The larger mica capacitors (CM-45 case) having voltage ratings of 1200 and 2500 volts are usually satisfactory for inputs varying from about 350 watts at 28 MHz to 1 kW at 14 MHz and lower. Because of these current limitations, particularly at the higher frequencies, it is advisable to use as large an air capacitor as practicable, using the micas only at the lower frequencies. Broadcast-receiver replacement-type capacitors can be obtained reasonably. Their voltage insulation should be adequate for inputs of 1000 watts or more.

TRANSISTOR OUTPUT CIRCUITS

Since rf power transistors have a low output impedance (on the order of 5 ohms or less), the problem of coupling the transistor to the usual 50-ohm load is the reverse of the problem with a vacuum-tube amplifier. The 50-ohm load must be transformed to a low resistance.

Figs 6-29A and B show two types of parallel-tuned circuits used to couple the load to the collector circuit. The collector is tapped down on the inductor in both cases. C1 provides tuning

Fig. 6-29 — Typical transistor output-matching networks.



for the collector and C2 adjusts the coupling to the load to achieve the proper impedance transformation. The use of the tapped connection to the inductor helps to maintain the loaded Q of the circuit while minimizing variations in tuning with changes in the junction capacitance of the transistor.

Circuits of Figs. 6-29C through E are not dependent upon coupling coefficient of a tapped coil for load-impedance transformation, making them more suitable for use at hf than either A or B. The collector-emitter capacitance (C_o) of the transistor is a major factor in the calculations used to design these circuits. Unfortunately C_o is not constant, so cut-and-try adjustments are usually necessary to optimize a particular circuit.

Early tests of transistor rf power amplifiers should be made with low voltage, a dummy load and no drive. Some form of output indicator should be included. When it has been established that no instability exists, the drive can be applied in increments and adjustment made for maximum output. The amplifier should never be operated at high voltage and no load.

BROADBAND COUPLING

The techniques of broadband-transformer construction use transmission-line elements. A transformer consists of a short transmission line (one-eighth wavelength or less) made from a twisted-wire pair, coaxial or strip line, wound on a high-permeability toroid core to improve the low-frequency characteristics. At vhf the core may be omitted. Only discrete impedance transformations are possible; typical ratios are 9/4:1, 4:1, 9:1, 16:1, and 25:1. The higher ratios are difficult to achieve in practice, so several 4:1 transformers are employed for a large transformation ratio as shown in Fig. 6-23. Hybrid transformers, providing the 180-degree phase shift for input and output matching to push-pull stages, may also be made using broadband techniques.

Large toroid cores are not required for moderate power levels. A one-half inch diameter core is sufficient for operation at 100 watts at the low impedance levels found in transistor circuits. Because the current is high it is important to keep the resistance of the conductors low. Multiconductor leads (3 or 4 strands of No. 26 enam., twisted) or the flat enam. strip used for transformer windings) are suitable. Some typical designs are shown in Table 6-II.

STABILIZING AMPLIFIERS

A straight amplifier operates with its input and output circuits tuned to the same frequency. Therefore, unless the coupling between these two circuits is brought to the necessary minimum, the amplifier will oscillate as a tuned-plate tuned-grid circuit. Care should be used in arranging components and wiring of the two circuits so that there will be negligible opportunity for coupling external to the tube or transistor itself. Complete shielding between input and output circuits usually is required. All rf leads should be kept as short as possible and particular attention should be paid to the rf return paths from input and output tank circuits to emitter or cathode. In general, the best arrangement using a tube is one in which the cathode connection to ground, and the plate tank circuit are on the same side of the chassis or other shielding. The "hot" lead from the input tank (or driver plate tank) should be brought to the socket through a hole in the shielding. Then when the grid tank capacitor or bypass is grounded, a return path through the hole to cathode will be encouraged, since transmission-line characteristics are simulated.