Contents

17.1 High Power, Who Needs It?
17.2 Types of Power Amplifiers
- 17.2.1 Why a "Linear" Amplifier?
- 17.2.2 Solid State vs Vacuum Tubes

17.3 Vacuum Tube Basics
- 17.3.1 Thermionic Emission
- 17.3.2 Components of a Vacuum Tube
- 17.3.3 Tube Nomenclature
- 17.3.4 Tube Mounting and Cooling Methods
- 17.3.5 Vacuum Tube Configurations
- 17.3.6 Classes of Operation in Tube Amplifiers
- 17.3.7 Understanding Tube Operation

17.4 Tank Circuits
- 17.4.1 Tank Circuit Q
- 17.4.2 Tank Circuit Efficiency
- 17.4.3 Tank Output Circuits

17.5 Transmitting Device Ratings
- 17.5.1 Plate, Screen and Grid Dissipation
- 17.5.2 Tank Circuit Components
- 17.5.3 Other Components

17.6 Sources of Operating Voltages
- 17.6.1 Tube Filament or Heater Voltage
- 17.6.2 Vacuum-Tube Plate Voltage
- 17.6.3 Grid Bias
- 17.6.4 Screen Voltage For Tubes

17.7 Tube Amplifier Cooling
- 17.7.1 Blower Specifications
- 17.7.2 Cooling Design Example
- 17.7.3 Other Considerations

17.8 Amplifier Stabilization
- 17.8.1 Amplifier Neutralization
- 17.8.2 VHF and UHF Parasitic Oscillations
- 17.8.3 Reduction of Distortion

17.9 Design Example: A High Power Vacuum Tube HF Amplifier
- 17.9.1 Tube Capabilities
- 17.9.2 Input and Output Circuits
- 17.9.3 Filament Supply
- 17.9.4 Plate Choke and DC Blocking
- 17.9.5 Tank Circuit Design
- 17.9.6 Checking Operation

17.10 Solid-State Amplifiers
- 17.10.1 Solid State vs Vacuum Tubes
- 17.10.2 Classes of Operation
- 17.10.3 Modeling Transistors
- 17.10.4 Impedance Transformation — "Matching Networks"
- 17.10.5 Combiners and Splitters
- 17.10.6 Amplifier Stabilization

17.11 A New 250-W Broadband Linear Amplifier
- 17.11.1 The 1.8 to 55 MHz PA — Detailed Description
- 17.11.2 Control and Protection
- 17.11.3 Low-Pass Filter
- 17.11.4 Power Supply
- 17.11.5 Amplifier Cooling
- 17.11.6 Metering and Calibration
- 17.11.7 Chassis
- 17.11.8 Performance

17.12 Tube Amplifier Projects
- *Project*: 3CX1500D7 RF Linear Amplifier
- *Project*: A 6-Meter Kilowatt Amplifier

17.13 References and Bibliography

Chapter 17

RF Power Amplifiers

Amateur Radio operators typically use a very wide range of transmitted power. With the milliwatts used by the QRP operator or the full legal power limit of 1.5 kW, amateur activities are successfully and enjoyably carried out by the amateur fraternity. This chapter covers RF power amplifiers beyond the 100-150 W level of the typical transceiver. The sections on tube-type amplifiers were prepared by John Stanley, K4ERO. Later sections on solid-state amplifiers were contributed by Richard Frey, K4XU. Roger Halstead, K8RI, contributed material on amplifier tuning and the use of surplus components in amplifier construction.

17.1 High Power, Who Needs It?

There are certain activities where higher power levels are recommended, almost required, for the greatest success. A good example is contest operation. While there are some outstanding operators who choose lower power as a sort of handicap and enjoy being competitive in spite of that disadvantage, sooner or later, the high-power stations have the biggest scores.

Another operator likely to run the legal limit is the avid DXer, although it is quite possible to work a lot of DX with a more modest station. There's plenty of DX activity on 40, 80 and 160 meters — where power becomes more important. On 160 meters, it's a real challenge to work 100 countries without the benefit of an amplifier. On 20 through 10 meters, power is not as important when skywave propagation permits working the world with a few watts. For skywave signals, power is much less important than band conditions and patient monitoring. On 10, 6 and 2 meter ground wave paths, power determines range.

Antennas are probably more important than raw power in both DX and contest operation in that the antenna helps both on receive and on transmit. Operating skill is more important than either.

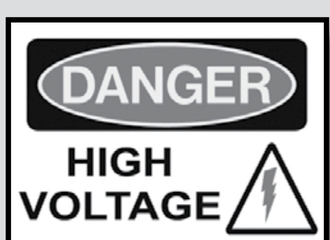
Another useful and important area where higher power may be needed is in net operations. The net control station needs to be heard by all potential net participants, some of whom may be hindered by a noisy location or limited antenna options or operating mobile. For general operations, rag chewing and casual contacts, the need for high power will depend on many factors and can best be learned by experience. As a quick rule of thumb, if you hear a lot of stations that don't seem to hear you, you probably need to run more power. If you operate on bands where noise levels are high (160 and 75/80 meters) or at times when signals are weak then you may find that running the legal limit makes operations more enjoyable. On the other hand, many stations will find that they can be heard fine with the standard 100 W transmitter or even with lower power.

Power requirements also depend on the mode being used. Some digital modes, such as PSK31, work very well with surprisingly low power. CW is more power efficient than SSB voice. Least effective is full carrier AM, which is still used by vintage equipment lovers. Once you have determined that higher power will enhance your operations, you should study the material in this chapter no matter whether you plan to build or buy your amplifier.

Note that many power amplifiers are capable of exceeding the legal limit, just as most automobiles are capable of exceeding posted speed limits. That does not mean that every operator with an amplifier capable of more than 1.5 kW output is a scofflaw. Longer tube life and a cleaner signal result from running an amplifier below its maximum rating. FCC rules call for an accurate way to determine output power, especially when you are running close to the limit.

Safety First!

YOU CAN BE KILLED by coming in contact with the high voltages inside a commercial or homebrew RF amplifier. Please don't take foolish chances. Remember that *you cannot go wrong* by treating each amplifier as potentially lethal! For a more thorough treatment of this all-important subject, please review the applicable sections of the *Safety* chapter in this *Handbook.*



17.2 Types of Power Amplifiers

Power amplifiers are categorized by their power level, intended frequencies of operation, device type, class of operation and circuit configuration. Within each of these categories there are almost always two or more options available. Choosing the most appropriate set of options from all those available is the fundamental concept of design.

17.2.1 Why a "Linear" Amplifier?

The amplifiers commonly used by amateurs for increasing their transmitted power are often referred to as "linears," rather than amplifiers or linear amplifiers. What does this mean and why is it important?

The active device in amplifiers, either tube or transistor, is like a switch. In addition to the "on" and "off" states of a true switch, the active device has intermediate conditions where it presents a finite value of resistance, neither zero nor infinity. As discussed in more detail in the **RF Techniques** chapter, active devices may be operated in various *classes of operation*. Class A operation never turns the device full on or off; it is always somewhere in between. Class B turns the device fully off for about half the time, but never fully on. Class C turns the device off for about 66% of the time, and almost achieves the fully on condition. Class D switches as quickly as possible between the on and off conditions. Other letters have been assigned to various rapid switching methods that try to do what Class D does, only better. Class E and beyond use special techniques to enable the device to make the switching transition as quickly as possible.

During the operating cycle, the highest efficiency is achieved when the active device spends most of its time in the on or off condition and the least in the resistive condition. For this reason, efficiency increases as we go from Class A to B to C to D.

A *linear amplifier* is one that produces an output signal that is identical to the input signal, except that it is stronger. Not all amplifiers do this. Linear amplifiers use Class A, AB or B operation. They are used for modes such as SSB where it is critical that the output be a close reproduction of the input.

The Class C amplifiers used for FM transmitters are *not* linear. A Class C amplifier, properly filtered to remove harmonics, reproduces the frequencies present in the input signal, but the *envelope* of the signal is distorted or even flattened completely. (See the **Modulation** chapter for more information on waveforms, envelopes and other signal characteristics.)

An FM signal has a constant amplitude, so it carries no information in the envelope. A CW signal does carry information in the amplitude variations. Only the on and off states must be preserved, so a Class C amplifier retains the information content of a CW signal. However, modern CW transmitters carefully shape the pulses so that key clicks are reduced to the minimum practical value. A Class C amplifier will distort the pulse shape and make the key clicks worse. Therefore, except for FM, a linear amplifier is recommended for all amateur transmission modes.

Some digital modes, such as RTTY using FSK, are a form of FM and can also use a nonlinear Class C, D or E amplifier. If these signals are not clean, however, a Class C amplifier may make them worse. Also, Class C or even D and E can be used for very slow CW, for very simple low-power CW transmitters or on uncrowded bands where slightly worse key clicks are not so serious. After all, Class C was used for many years with CW operation.

Class of operation as it relates to tube-type amplifier design is discussed in more detail in a later section of this chapter.

ACHIEVING LINEAR AMPLIFICATION

How is linear amplification achieved? Transistors and tubes are capable of being operated in a linear mode by restricting the input signal to values that fall on the linear portion of the curve that relates the input and output power of the device. Improper bias and excessive drive power are the two most common causes of distortion in linear amplifiers. All linear amplifiers can be improperly biased or overdriven, regardless of the power level or whether transistors or tubes are used.

Fig 17.1 shows a simple circuit capable of operating in a linear manner. For linear operation, the bias on the base of the transistors must be such that the circuit begins to produce an output signal even with very small input signal values. As shown in **Fig 17.2**, when properly adjusted for linear operation, the amplifier's output signal faithfully tracks the input signal. Without proper bias on the transistors, there will be no output signal until the input voltage goes above a threshold. As shown in Fig 17.2B, the improperly adjusted amplifier suddenly switches on and produces output when the input signal reaches 0.5 V.

Some tubes are designed for *zero bias* operation. This means that an optimum bias current is inherent in the design of the tube when it is operated with the correct plate voltage. Other types of tubes and all transistors must have bias applied with circuits made for that purpose.

All amplifiers have a limit to the amount of power they can produce, even if the input power is very large. When the output power runs up against this upper limit (that is, when additional drive power results in no more output power), *flat topping* occurs and the output is distorted, as shown in **Fig 17.3**. Many amplifiers use *automatic level control* (ALC) circuits to provide feedback between the amplifier and the companion transceiver or transmitter.

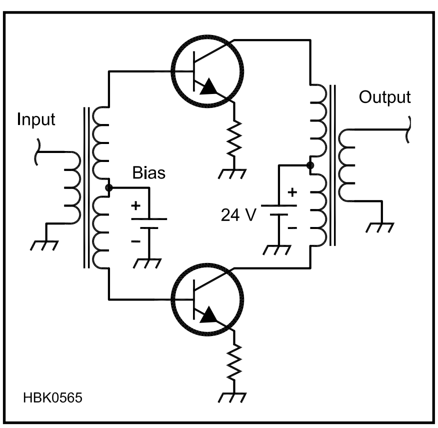


Fig 17.1 — This simple circuit can operate in a linear manner if properly biased.

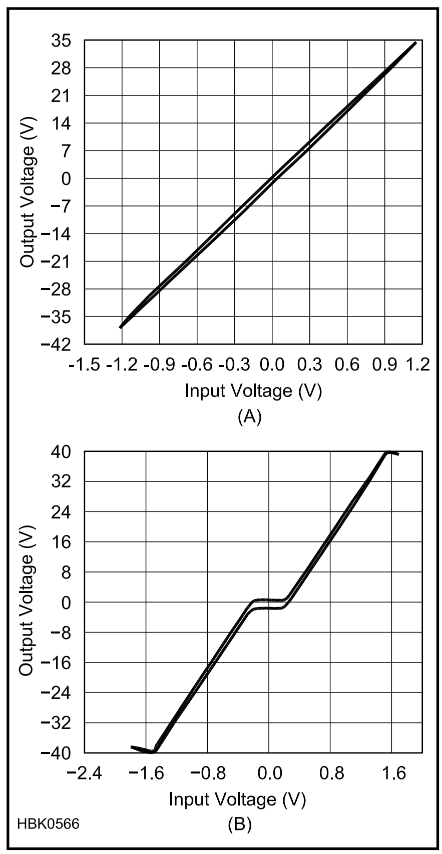


Fig 17.2 — Input versus output signals from an amplifier, as observed with the X-Y display on an oscilloscope. At A, an amplifier with proper bias and input voltage. At B, the same amplifier with improper bias and high input voltage.

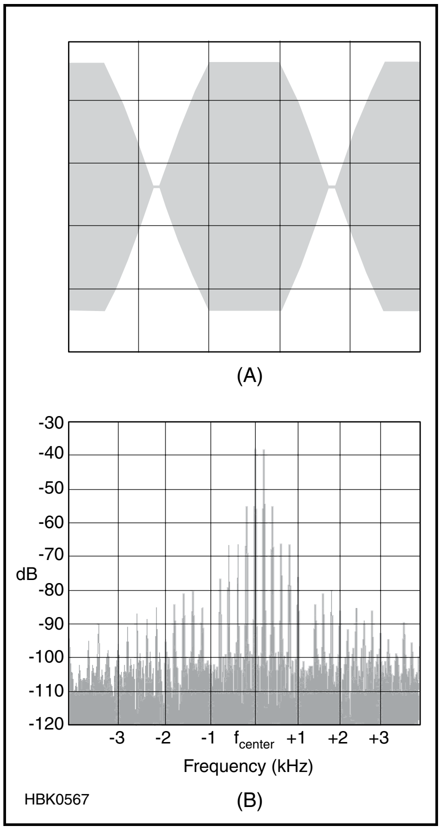


Fig 17.3 — Improper operation of a linear using a two-tone test will show peak clipping on an oscilloscope (A) and the presence of additional frequencies on a spectrum analyzer (B). When these patterns appear, your "linear" has become a "non-linear" amplifier. Users of adjacent channels will not be happy. More information on transmitter testing may be found in the Test Equipment and Measurements chapter.

When adjusted properly, ALC will control the transmitter output power, preventing the worst effects of overdriving the amplifier. Even with ALC, however, overdriving can occur.

17.2.2 Solid State vs Vacuum Tubes

With the exception of high-power amplifiers, nearly all items of amateur equipment manufactured commercially today use solid-state (semiconductor) devices exclusively. Semiconductor diodes, transistors and integrated circuits (ICs) offer several advantages in designing and fabricating equipment. Solid-state equipment is smaller, offers broadband (no-tune-up) operation, and is easily manufactured using PC boards and automated (lower cost) processes.

Based on all these facts, it might seem that there would be no place for vacuum tubes in a solid-state world. Transistors and ICs do have significant limitations, however, especially in a practical sense. Individual present-day transistors cannot generally handle the combination of current and voltage needed nor can they safely dispose of the amount of heat dissipated for RF amplification to high power levels. Pairs of transistors, or even pairs of pairs, are usually employed in practical power amplifier designs at the 100 W level and beyond. Sometimes various techniques of power combination from multiple amplifiers must be used.

Tube amplifiers can be more economical to build for a given output power. Vacuum tubes operate satisfactorily at surface temperatures as high as 150-200 ºC, so they may be cooled by simply blowing sufficient ambient air past or through their relative large cooling surfaces. The very small cooling surfaces of power transistors should be held to 75-100 ºC to avoid drastically shortening their life expectancy. Thus, assuming worst-case 50 ºC ambient air temperature, the large cooling surface of a vacuum tube can be allowed to rise 100-150 ºC above ambient, while the small surface of a transistor must not be allowed to rise more than about 50 ºC.

Furthermore, RF power transistors are much less tolerant of electrical abuse than are most vacuum tubes. An overvoltage spike lasting only microseconds can — and is likely to — destroy RF power transistors. A comparable spike is unlikely to have any effect on a tube. So the important message is this: designing with RF power transistors demands caution to ensure that adequate thermal and electrical protection is provided.

Even if one ignores the challenge of the RF portions of a high-power solid-state amplifier, there is the dc power supply to consider. A solid-state amplifier capable of delivering 1 kW of RF output might require regulated (and transient-free) 50 V at more than 40 A. Developing that much current can be challenging. A vacuum tube amplifier at the same power level might require 2000 to 3000 V, unregulated, at less than 1 A.

At the kilowatt level, the vacuum tube is still a viable option for amateur constructors because of its cost-effectiveness and ease of equipment design. Because tube amplifiers and solid-state amplifiers are quite different in many ways, we shall treat them in different sections of this chapter. Also, the author of the solid-state section presents a slightly different perspective on the tube-vs-solid-state discussion.

17.3 Vacuum Tube Basics

The term *vacuum tube* describes the physical construction of the devices, which are usually tubular and have a vacuum inside. The British call them electron valves which describes the operation of the devices, since they control the flow of electrons, like a water valve controls the flow of water.

17.3.1 Thermionic Emission

Metals are electrical conductors because the electrons in them readily move from one atom to the next under the influence of an electrical field. It is also possible to cause the electrons to be emitted into space if enough energy is added to them. Heat is one way of adding energy to metal atoms, and the resulting flow of electrons into space is called *thermionic emission*. As each electron leaves the metal surface, it is replaced by another provided there is an electrical connection from outside the tube to the heated metal.

In a vacuum tube, the emitted electrons hover around the surface of the metal unless acted upon by an electric field. If a positively charged conductor is placed nearby, the electrons are drawn through the vacuum and arrive at that conductor, thus providing a continuous flow of current through the vacuum tube.

17.3.2 Components of a Vacuum Tube

A basic vacuum tube contains at least two parts: a *cathode* and a *plate*. The electrons are emitted from the *cathode*. The cathode can be *directly heated* by passing a large dc current through it, or it can be located adjacent to a heating element (*indirectly heated*). Although ac currents can also be used to directly heat cathodes, if any of the ac voltage mixes with the signal, ac hum will be introduced into the output. If the ac heater supply voltage can be obtained from a center tapped transformer, and the center tap is connected to the signal ground, hum can be minimized.

The difficulty of producing thermionic emission varies with the metal used for the cathode, and is called the "work function" of that metal. An ideal cathode would be made of a metal with a low work function that can sustain high temperatures without melting. Pure tungsten was used in early tubes as it could be heated to a very high temperature. Later it was learned that a very thin layer of thorium greatly increased the emission. Oxides of metals with low work functions were also developed. In modern tubes, thoriated-tungsten is used for the

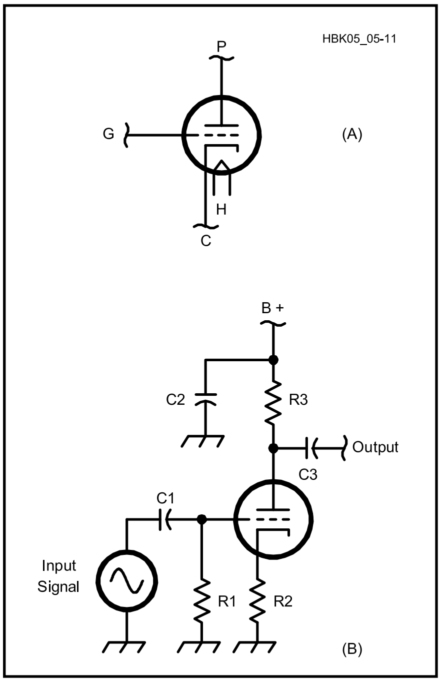


Fig 17.4 — Vacuum tube triode. (A) Schematic symbol detailing heater (H), cathode (C), grid (G) and plate (P). (B) Audio amplifier circuit using a triode. C1 and C3 are dc blocking capacitors for the input and output signals to isolate the grid and plate bias voltages. C2 is a bypass filter capacitor to decrease noise in the plate bias voltage, B+. R1 is the grid bias resistor, R2 is the cathode bias resistor and R3 is the plate bias resistor. Note that although the cathode and grid bias voltages are positive with respect to ground, they are still negative with respect to the plate.

higher power tubes and oxide-coated metals are commonly used at lower power levels.

Filament voltage is important to the proper operation of a tube. If too low, the emission will not be sufficient. If too high, the useful life of the tube will be greatly shortened. It is important to know which type of cathode is being used. Oxide-coated cathodes can be run at 5 to 10% under their rated voltage with an increase in their life, provided performance is satisfactory. They should never be run above the nominal value. Thoriated-tungsten tubes can also gain operational life by running with reduced filament voltage, and as they approach the end of their useful life, can be run at higher than the nominal value as necessary to maintain emission. Carelessly running either type at higher than nominal voltage when new is sure to lead to shorter than normal life with no performance advantage.

Every vacuum tube needs a receptor for the emitted electrons. After moving though the vacuum, the electrons are absorbed by the *plate,* also called the *anode.* This two-element tube — anode and cathode — is called a *diode*. The diode tube is similar to a semiconductor diode: it allows current to pass in only one direction. If the plate goes negative relative to the cathode, current cannot flow because electrons are not emitted from the plate. Years ago, in the days before semiconductors, tube diodes were used as rectifiers.

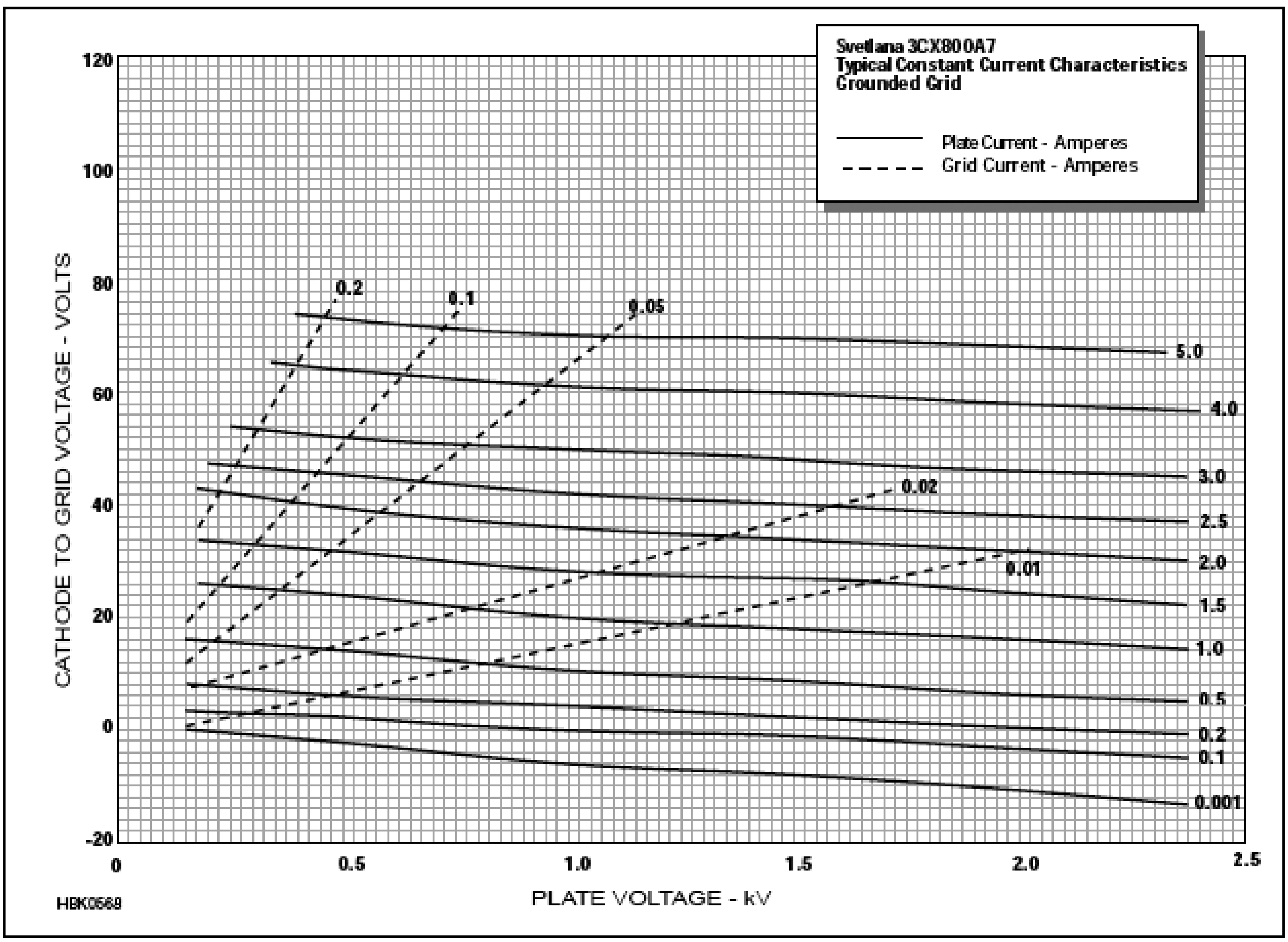


Fig 17.5 — Characteristic curves for a 3CX800A7 triode tube. Grid voltage is plotted on the left, plate voltage along the bottom. The solid lines are plate current, the dashed lines are grid current. This graph is typical of characteristic curves shown in this chapter and used with the *TubeCalculator* program described in the text and included on the *Handbook* CD.

TRIODES

To amplify signals, a vacuum tube must also contain a control *grid*. This name comes from its physical construction. The grid is a mesh of wires located between the cathode and the plate. Electrons from the cathode pass between the grid wires on their way to the plate. The electrical field that is set up by the voltage on these wires affects the electron flow from cathode to plate. A negative grid voltage repels electrons, blocking their flow to the plate. A positive grid voltage enhances the flow of electrons to the plate. Vacuum tubes containing a cathode, a grid and a plate are called *triode* tubes (*tri* for three components). See **Fig 17.4**.

The input impedance of a vacuum tube amplifier is directly related to the grid current. Grid current varies with grid voltage, increasing as the voltage becomes more positive. When the grid voltage is negative, no grid current flows and the input impedance of a tube is nearly infinite. When the grid is driven positive, it draws current and thus presents a lower input impedance, and requires significant drive power. The load placed across the plate of the tube strongly affects its output power and efficiency. An important part of tube design involves determining the optimum load resistance. These parameters are plotted as *characteristic curves* and are used to aid the design process. **Fig 17.5** shows an example.

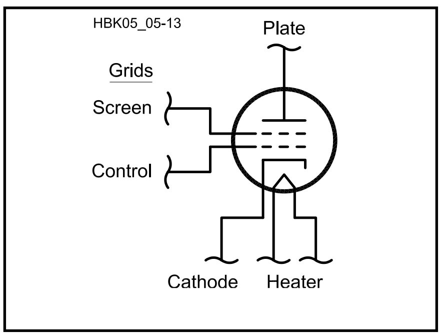


Fig 17.6 — Vacuum tube tetrode. Schematic symbol detailing heater (H), cathode (C), the two grids: control and screen, and the plate (P).

Since the elements within the vacuum tube are conductors that are separated by an insulating vacuum, the tube is very similar to a capacitor. The capacitance between the cathode and grid, between the grid and plate, and between the cathode and plate can be large enough to affect the operation of the amplifier at high frequencies. These capacitances, which are usually on the order of a few picofarads, can limit the frequency response of an amplifier and can also provide signal feedback paths that may lead to unwanted oscillation. Neutralizing circuits are sometimes used to prevent such oscillations. Techniques for neutralization are presented later in this chapter.

TETRODES

The grid-to-plate capacitance is the chief source of unwanted signal feedback. Therefore tubes were developed with a second grid, called a *screen grid*, inserted between the original grid (now called a *control grid*) and the plate. Such tubes are called tetrodes (having four elements). See **Fig 17.6**. This second grid is usually tied to RF ground and acts as a screen between the grid and the plate, thus preventing energy from feeding back, which could cause instability. Like the control grid, the screen grid is made of a wire mesh and electrons pass through the spaces between the wires to get to the plate.

The screen grid carries a high positive voltage with respect to the cathode, and its proximity to the control grid produces a strong electric field that enhances the attraction of electrons from the cathode. The gain of a tetrode increases sharply as the screen voltage is increased. The electrons accelerate toward the screen grid and most of them pass through the spaces and continue to accelerate until they reach the plate. In large tubes this is aided by careful alignment of the screen wires with the grid wires. The effect of the screen can also be seen in the flattening of the tube curves. Since the screen shields the grid from the plate, the plate current vs plate voltages becomes almost flat, for a given screen and grid voltage. **Fig 17.7** shows characteristic curves for a typical tetrode, and curves for many more tube types are included on the *Handbook* CD.

A special form of tetrode concentrates the electrons flowing between the cathode and the plate into a tight beam. The decreased electron-beam area increases the efficiency of the tube. *Beam tetrodes* permit higher plate currents with lower plate voltages and large power outputs with smaller grid driving power. The 6146 is an example of a beam power tube.

PENTODES

Another unwanted effect in vacuum tubes is the so-called *secondary emission*. The electrons flowing within the tube can have so much energy that they are capable of dislodging electrons from the metal atoms in the grids and plate. Secondary emission can cause a grid, especially the screen grid, to lose more electrons than it absorbs. Thus while a screen usually draws current from its supply, it occasionally pushes current into the supply. Screen supplies must be able to absorb as well as supply current.

A third grid, called the *suppressor grid*, can be added between the screen grid and the plate. This overcomes the effects of secondary emission in tetrodes. A vacuum tube with three grids is called a *pentode* (penta for five components). The suppressor grid is connected to a low voltage, often to the cathode.

17.3.3 Tube Nomenclature

Vacuum tubes are constructed with their elements (cathode, grid, plate) encased in an envelope to maintain the vacuum. Tubes with glass envelopes, such as the classic transmit-

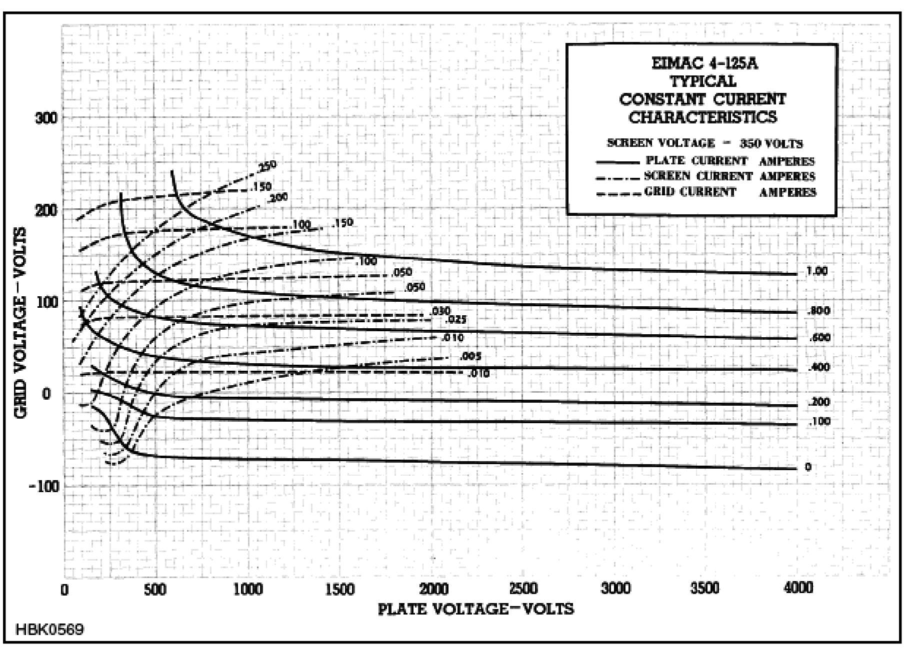


Fig 17.7 — Characteristic curves for a 4-125 tetrode tube.

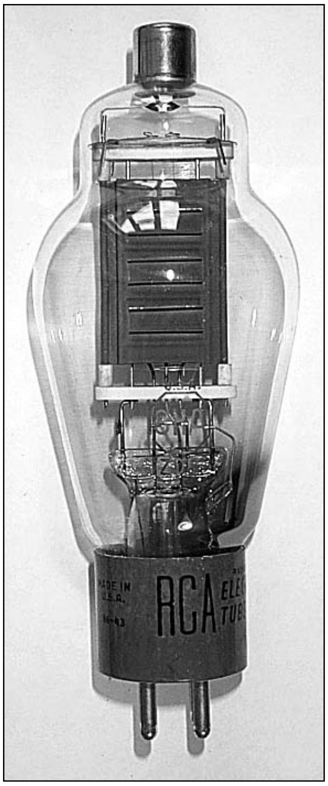


Fig 17.8 — The RCA 811A is an example of a transmitting triode with a glass envelope. (Photo courtesy the Virtual Valve Museum, www.tubecollector.org.)

Fig 17.9 — Modern power tubes, such as this 4CX1000A tetrode, tend to use metal and ceramic construction.

ting tube shown in **Fig 17.8**, are most familiar. Over time, manufacturers started exploring other, more rugged, methods for making high power transmitting tubes. Modern power tubes tend to be made of metal parts separated by ceramic insulating sections (**Fig 17.9**).

Because of their long history, vacuum tube types do not all follow a single logical system of identification. Many smaller tubes types begin with an indication of the filament voltage, such as the 6AU6 or 12AT7. Other tubes such as the 811 (Fig 17.8) and 6146 were assigned numbers in a more or less chronological order, much as transistors are today. Some glass envelope power tubes follow a numbering system that indicates number of tube elements and plate dissipation — 3-500Z and 4-1000A are two common examples in amateur circles.

Some power tubes follow the 3CX and 4CX numbering system. The first number indicates a triode (3) or tetrode (4) and the C indicates ceramic/metal construction. The X indicates cooling type: X for air, W for water and V for vapor cooling. The cooling type is followed by the plate dissipation. (The tubes that amateurs use typically have three or four numbers indicating plate dissipation; those used in commercial and broadcast service can have much higher numbers.) Thus a 4CX250 is a ceramic, air cooled 250 W tetrode. A 3CX1200 is a ceramic, air cooled, 1200 W triode. Often these tubes have additional characters following the plate dissipation to indicate special features or an upgraded design. For example, a 4CX250R is a special version of the 4CX250 designed for AB1 linear operation. A 4CX1500B is an updated version of the 4CX1500A.

During the heyday of tube technology, some tube types were developed with several tubes in the same glass envelope, such as the 12AX7 (a dual triode). Except for a very few devices used in the specialty audio market, tubes of this type are no longer made.

17.3.4 Tube Mounting and Cooling Methods

Most tubes mount in some kind of socket so that they can be easily replaced when they reach the end of their useful life. Connections to the tube elements are typically made through pins on the base. The pins are arranged or keyed so that the tube can be inserted into the socket only one way and are sized and spaced to handle the operating voltages and currents involved. Tubes generally use a standard base or socket, although a great many different bases developed over the years. Tube data sheets show pinouts for the various tube elements, just like data sheets for ICs and transistors. See the **Component Data and References** chapter for base diagrams of some popular transmitting tubes.

For transmitting tubes, a common arrangement is for filament and grid connections to be made through pins in the main base, while the plate connection is made through a large pin or post at the top of the tube for easier connection to the high voltage supply and tank circuit. This construction is evident in the examples shown in Figs 17.8 and 17.9. To reduce stray reactances, in some older glass tubes the grid used a separate connection.

Heat dissipation from the plate is one of the major limiting factors for vacuum tube power amplifiers. Most early vacuum tubes were encased in glass, and heat passed through it as infrared radiation. If more cooling was needed, air was simply blown over the outside of the glass. Modern ceramic tubes suitable for powers up to 5 kW are usually cooled by forcing air directly through an external anode. These tubes require a special socket that allows free flow of air. The large external anode and cooling fins may be seen in the example in Fig 17.9. Conduction through an insulating block and water cooling are other options, though they are not often seen in amateur equipment. Practical amplifier cooling methods are discussed in detail later in this chapter.

17.3.5 Vacuum Tube Configurations

Just as the case with solid-state devices described in the **Analog Basics** chapter, any of the elements of the vacuum tube can be common to both input and output. A common plate connection — called a cathode follower and similar to the emitter follower — was once used to reduce output impedance (current gain) with little loss of voltage. This application is virtually obsolete.

Most modern tube applications use either the common cathode or the common grid connection. Fig 17.4B shows the common cathode connection, which gives both current and voltage gain. The common grid (often

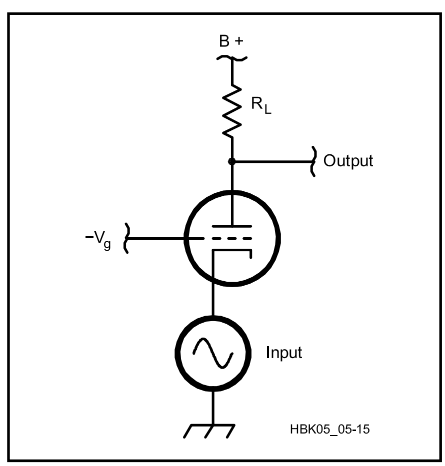


Fig 17.10 — Grounded grid amplifier schematic. The input signal is connected to the cathode, the grid is biased to the appropriate operating point by a dc bias voltage, $-V_G$, and the output voltage is obtained by the voltage drop through R_L that is developed by the plate current, I_P.

called *grounded grid*) connection shown in **Fig 17.10** gives only voltage gain. Thus the common cathode connection is capable of 20 to 30 dB of gain in a single stage, whereas the grounded grid connection typically gives 10 to 15 dB of gain.

The input impedance of a grounded grid stage is low, typically less than a few hundred ohms. The input impedance of a grounded cathode stage is much higher.

17.3.6 Classes of Operation in Tube Amplifiers

Class of operation was discussed briefly in the previous section describing the need for linear operation of RF power amplifiers for most Amateur Radio modes except FM. The class of operation of an amplifier stage is defined by its conduction angle, the angular portion of each RF drive cycle, in degrees, during which plate current flows. The conduction angle is determined by the bias on the device, and to a lesser extent on the drive level. These, in turn, determine the amplifier's efficiency, linearity and operating impedances. Refer to **Fig 17.11** for the following discussion.

Class A is defined as operation where plate current is always flowing. For a sine wave this means during 360° of the wave. Class A has the best linearity, but poor efficiency.

Class B is when the bias is set so that the tube is cut off for negative input signals, but current flows when the input signal is positive. Thus for a sine wave, conduction occurs during ½ of the cycle, or 180°. Class B is linear only when two devices operate in push-pull, so as to provide the missing half of the wave, or when a tuned circuit is present to restore the missing half by "flywheel" action (discussed later in this chapter).

Class AB is defined as operation that falls between Class A and Class B. For a sine wave, the conduction angle will be more than 180°, but less than 360°. In practice Class AB amplifiers usually fall within the gray area shown in the center area of the graph. Like Class B, a push pull connection or a tuned circuit are needed for linear operation. Class AB is less efficient that class B, but better than Class A. The linearity is better than class B but worse than class A.

Class AB vacuum tube amplifiers are further defined as class AB1 or AB2. In class AB1, the grid is not driven positive, so no grid current flows. Virtually no drive power is required. In Class AB2,the grid is driven positive at times with respect to the cathode and some grid current flows. Drive power and output both increase as compared to AB1. Most linear amplifiers used in the Amateur service operate Class AB2, although for greater linearity some operate Class AB1 or even Class A.

Class C is when conduction angle is less than 180º — typically 120º to 160º for vacuum tube amplifiers or within the gray area to the left in Fig 17.11. The tube is biased well beyond cutoff when no drive signal is applied. Output current flows only during positive crests in the drive cycle, so it consists of relatively narrow pulses at the drive frequency. Efficiency is high, but nonlinear operation results. Class C amplifiers always use tuned circuits at the input and output. Attempts to achieve extreme efficiency with very narrow pulses (small conduction angles) require very high drive power, so a point of diminishing returns is eventually reached.

Classes D through H use various switched mode techniques. These are used almost exclusively with solid state circuits.

17.3.7 Understanding Tube Operation

Vacuum tubes have complex current transfer characteristics, and each class of operation produces different RMS values of RF current through the load impedance. As described earlier, tube manufacturers provide characteristic curves that show how the tube behaves as operating parameters (such as plate and grid voltage and current) vary. See Figs 17.5 and 17.7 for examples of characteristic curves for two different transmitting tubes. The use of tube curves provides the best way to gain insight into the characteristics of a given tube. Before designing with a tube, get a set of these curves and study them thoroughly.

Because of the complexity and interaction among the various parameters, computer-aided design (CAD) software is useful in analyzing tube operation. One such program, *TubeCalculator*, is included on the *Handbook* CD, along with curves for many popular transmitting tubes. This software makes it easier to do analysis of a given operation with the tube you have chosen. **Fig 17.12** shows a *TubeCalculator* screen with an example of "constant current" curves for a typical tube used in high power amplifiers and tables showing values for the various operating parameters. The curves shown have grid voltage on the vertical axis and plate voltage on the horizontal axis. Older tubes and some newer ones use a slightly different format in which plate voltage is plotted on the horizontal axis and plate current on the vertical. *TubeCalculator* allows analysis using either type.

The tube is the heart of any amplifier. Using the software to arrive at the desired operating parameters is a major step toward understanding and designing an amplifier. The second most important part of the design is the tuning components. Before they can be designed, the required plate load resistance must be determined and *TubeCalculator* will do that. In addition it will give insight into what happens when a tube is under driven or over driven, when the bias is wrong, or when the load resistance incorrect.

If a tube is to be used other than with nominal voltages and currents, analysis using the tube curves is the only solution short of trial

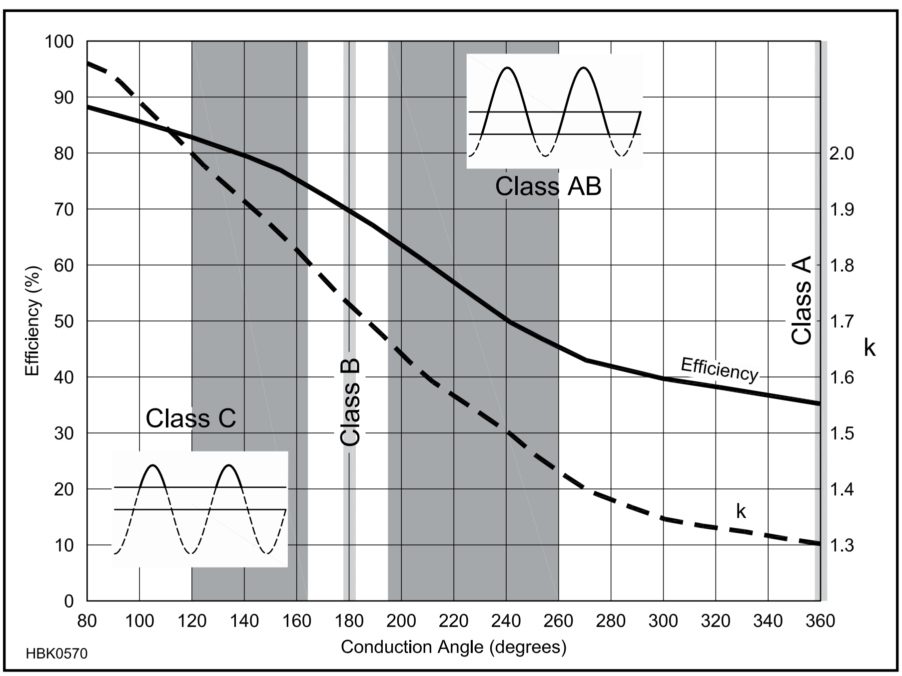


Fig 17.11 — Efficiency and K for various classes of operation. Read the solid line to determine efficiency. Read the dashed line for K, which is a constant used to calculate the plate load required, as explained in the text.

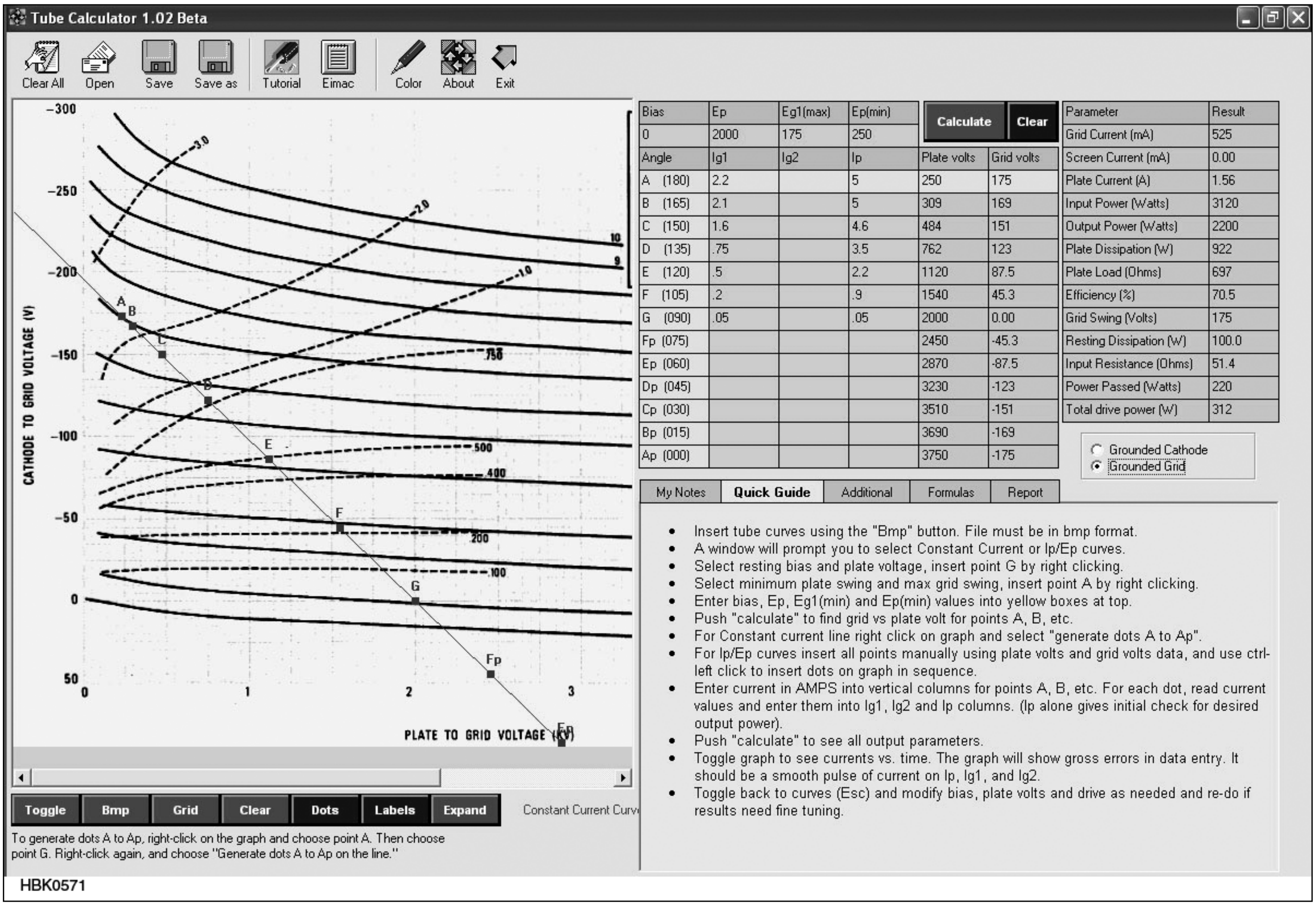


Fig 17.12 — Main screen of *TubeCalculator* program (included on the *Handbook* CD). A characteristic curve plot is loaded in the window on the left side of the screen.

and error. Trial and error is not a good idea because of the high voltages and currents found in high power amplifiers. It's best to conduct your analysis using curves and software, or else stick to operating the tube very close to the voltages, currents, drive levels and load values specified by the manufacturer.

ANALYZING OPERATING PARAMETERS

Characteristic curves allow a detailed look at tube operation as voltages and currents vary. For example, you can quickly see how much negative grid voltage is required to set the plate current to any desired value, depending also on the plate voltage and screen voltages. You can also see how much grid voltage is needed to drive the plate current to the maximum desired value.

With RF power amplifiers, both the grid voltage and the plate voltage will be sinusoidal and will be 180° out of phase. With constant current curves, an operating line can be drawn that will trace out every point of the operating cycle. This will be a straight line connecting two points. One point will be at the intersection of the peak plate voltage and the peak negative grid voltage. The other point will be at the intersection of the peak positive grid voltage and the minimum plate voltage.

If we plot the plate current along this line as a function of time, it will be seen that it changes in a nonlinear fashion; the exact shape of which depends on the class of operation. For class AB2 operation, which is the most commonly used in linear RF power amplifiers, it will look something like the plot shown in **Fig 17.13**. This rather complicated waveform is not easily evaluated using simple formulas, but it can be analyzed by taking the current values vs time and applying averaging techniques. This method was developed by Chaffee in 1936 and popularized by Eimac, the company that developed many of the power tubes used today.

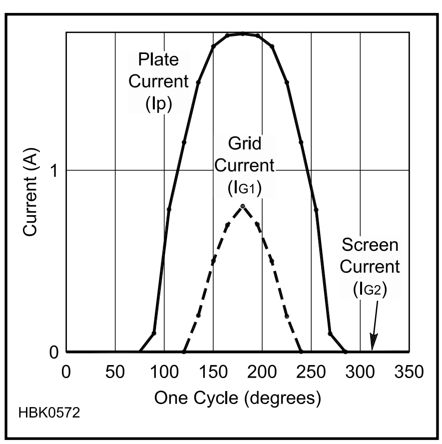


Fig 17.13 — Class AB2 plate and grid current over one cycle as plotted by the *TubeCalculator* program. This plot is for a triode, so there is no curve for screen (G2) current.

With *TubeCalculator*, data can be extracted from the curves and can be converted into many useful operating parameters such as input power, output power, grid power required, grid and plate dissipation and required load resistance.

MANUAL METHODS FOR TUBE PARAMETER SELECTION

For those not wishing to use a computer for design, most tube manufacturers will supply a table of typical operating values. A summary of some of this information is available in the **Component Data and References** chapter. These values have already been determined both from the tube characteristics and actual operational tests. As long as your proposed operation is close to the typical values in terms of voltages and currents, the typical values can provide you the desired load resistance and expected output power and efficiency. The typical operating parameters may also include a suggested optimum load resistance. If not, we can use well established "rules of thumb."

The optimum load resistance for vacuum-tube amplifiers can be approximated by the ratio of the dc plate voltage to the dc plate current at maximum signal, divided by a constant appropriate to each class of operation. The load resistance, in turn, determines the maximum power output and efficiency the amplifier can provide. The approximate value for tube load resistance is

$$R_L = \frac{V_p}{K \times I_p} \tag{1}$$

where

R_L = the appropriate load resistance, in ohms

V_P = the dc plate potential, in V

I_P = the dc plate current, in A

K = a constant that approximates the RMS current to dc current ratio appropriate for each class. For the different classes of operation: Class A, K ≈ 1.3; Class AB, K ≈ 1.5-1.7; Class B, K ≈ 1.57-1.8; Class C, K ≈ 2. The way in which K varies for different conduction angles is shown in Fig 17.11 (right scale).

Once we determine the optimum load resistance value for the tube(s) to be used we are ready to design the output networks for the amplifier. After tube selection, this is the most important part of the total design.

17.4 Tank Circuits

Usually we want to drive a transmission line, typically 50 Ω, with the output of our amplifier. An output network is used to transform that 50 Ω impedance to the optimum load resistance for the tube. This transformation is accomplished by resonant output networks which also serve to reduce harmonics to a suitable level. The **Electrical Fundamentals** chapter of this *Handbook* gives a detailed analysis of the operation of resonant circuits. We summarize here only the most important points.

Resonant circuits have the ability to store energy. Capacitors store electrical energy in the electric field between their plates; inductors store energy in the magnetic field induced by the coil winding. These circuits are referred to as *tank circuits* since they act as storage "tanks" for RF energy. This energy is continuously passed back and forth between the inductive storage and the capacitive storage. It can be shown mathematically that the "alternating" current and voltage produced by this process are sinusoidal in waveform with a frequency of

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

which, of course, is the resonant frequency of the tank circuit.

17.4.1 Tank Circuit Q

In order to quantify the ability of a tank circuit to store energy, a quality factor, Q, is defined. Q is the ratio of energy stored in a system during one complete RF cycle to energy lost.

$$Q = 2\pi \frac{W_S}{W_L} \tag{3}$$

where

W_S = is the energy stored

W_L = the energy lost to heat and the load

The Flywheel Effect

The operation of a resonant tank is sometimes referred to as the "flywheel" effect. A flywheel does illustrate certain functions of a resonant tank, but a flywheel alone is nonresonant; that is, it has no preferred frequency of operation. A better representation of a resonant tank is found in the *balance wheel* as used in old fashioned mechanical watches and small clocks.

A weighted wheel rotates back and forth, being returned to its center position by a spiral spring, sometimes called a hairspring. The balance wheel stores energy as inertia and is analogous to an inductor. The hairspring also stores energy and is similar in operation to a capacitor. The spring has a adjustment for the frequency of operation or resonance (1 Hz). An escapement mechanism gives the wheel a small kick with each tick of the watch to keep it going. For the purpose of our illustration, imagine energy from the balance wheel being used to drive the clock hands, which is not exactly how real clocks work. Most of the energy would go into the gears that drive the hands. Some energy would be wasted in heat due to friction.

A plot of the radial velocity of the balance wheel will give a sine function. It thus converts the pulses from the escapement into smooth sinusoidal motion. In a similar way, pulses of current in the plate of a tube are smoothed into a sine wave in the "tank" circuit, which is so called because it stores electrical energy, just as a tank stores water.

The plate voltage is a sine wave, even though the plate current is made up of pulses. The frequency at which pulses are added must match the natural resonant frequency of the tank which is adjusted with the "plate tune" capacitor.

Efficient operation of the tank itself occurs when the losses are small compared to the energy transferred to the output. In a similar manner, lowering the losses in a balance wheel by using jeweled bearings makes the watch more efficient. The amount of energy stored in the balance wheel should also be kept as low as practical since excess "oscillation" of the wheel wastes energy to the air and bearings. Likewise, the "circulating currents" in the tank must be limited to reduce heating from the inevitable losses in the components.

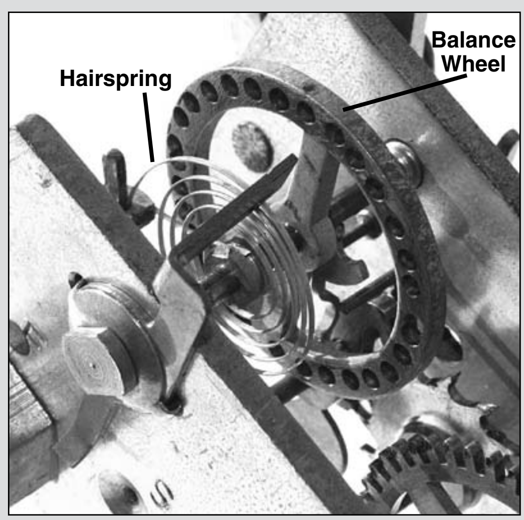


Fig 17.A1 — A balance wheel and hairspring in an old clock illustrate the flywheel effect in tank circuits.

A load connected to a tank circuit has exactly the same effect on tank operation as circuit losses. Both consume energy. It just happens that energy consumed by circuit losses becomes heat rather than useful output. When energy is coupled out of the tank circuit into a load, the loaded Q (Q_L) is:

$$Q_L = \frac{X}{R_{Loss} + R_{Load}} \quad (4)$$

where R_{Load} is the load resistance. Energy dissipated in R_{Loss} is wasted as heat. And X represents the reactance of the inductor or the capacitor, assumed to be equal at resonance. Ideally, all the tank circuit energy should be delivered to R_{Load}. This implies that R_{Loss} should be as small as possible.

17.4.2 Tank Circuit Efficiency

The efficiency of a tank circuit is the ratio of power delivered to the load resistance (R_{Load}) to the total power dissipated by losses (R_{Load} and R_{Loss}) in the tank circuit. Within the tank circuit, R_{Load} and R_{Loss} are effectively in series and the circulating current flows through both. The power dissipated by each is proportional to its resistance. The loaded tank efficiency can, therefore, be defined as

$$\text{Tank Efficiency} = \frac{R_{Load}}{R_{Load} + R_{Loss}} \times 100 \quad (5)$$

where efficiency is stated as a percentage. The loaded tank efficiency can also be expressed as

$$\text{Tank Efficiency} = \left(1 - \frac{Q_L}{Q_U}\right) \times 100 \quad (6)$$

where

Q_L = the tank circuit loaded Q, and

Q_U = the unloaded Q of the tank circuit.

For practical circuits, Q_U is very nearly the Q of the coil with switches, capacitors and parasitic suppressors making a smaller contribution. It follows, then, that tank efficiency can be maximized by keeping Q_L low which keeps the circulating current low and the I^2R losses down. Q_U should be maximized for best efficiency; this means keeping the circuit losses low. With a typical Q_L of 10, about 10% of the stored energy is transferred to the load in each cycle. This energy is replaced by energy supplied by the tube. It is interesting to contemplate that in a typical amplifier which uses a Q_L of 10 and passes 1.5 kW from the tube to the output, the plate tank is storing about 15 kW of RF energy. This is why component selection is very important, not only for low loss, but to resist the high voltages and currents.

Resonant circuits are always used in the plate circuit. When the grid is used as the input (common cathode), both matching and a tuned circuit may be used or else a low impedance load is connected from grid to ground with a broad matching transformer or network. The "loaded grid" reduces gain, but improves stability. In grounded grid operation, a tuned circuit may not be needed in the cathode circuit, as the input Z may be close to 50 Ω, but a tuned network may improve the match and usually improves the linearity. These resonant circuits help to ensure that the voltages on grid and plate are sine waves. This wave-shaping effect is the same thing as harmonic rejection. The reinforcing of the fundamental frequency and rejection of the harmonics is a form of filtering or selectivity.

The amount of harmonic suppression is dependent upon circuit loaded Q_L, so a dilemma exists for the amplifier designer. A low Q_L is desirable for best tank efficiency, but yields poorer harmonic suppression. High Q_L keeps amplifier harmonic levels lower at the expense of some tank efficiency. At HF, a compromise value of Q_L can usually be chosen such that tank efficiency remains high and harmonic suppression is also reasonable. At higher frequencies, tank Q_L is not always readily controllable, due to unavoidable stray reactances in the circuit. Unloaded Q_U can always be maximized, however, regardless of frequency, by keeping circuit losses low.

17.4.3 Tank Output Circuits

THE PI NETWORK

The pi network with the capacitors to ground and the inductor in series is commonly used for tube type amplifier matching. This acts like a low pass filter, which is helpful for getting rid of harmonics. Harmonic suppression of a pi network is a function of the impedance transformation ratio and the Q_L of the circuit. Second-harmonic attenuation is approximately 35 dB for a load impedance of 2000 Ω in a pi network with a Q_L of 10. In addition to the low pass effect of the pi network, at the tube plate the third harmonic is already typically 10 dB lower and the fourth approximately 7 dB below that. A typical pi network as used in the output circuit of a tube amplifier is shown in **Fig 17.14**. The **RF and AF Filters** chapter describes harmonic filters that can also greatly reduce harmonics. These are typically not switched but left in the circuit at all times. With such a filter, the requirements for reducing the harmonics on the higher bands with the amplifier pi network is greatly reduced.

The formulas for calculating the component values for a pi network, for those who wish to use them, are included on the CD that accompanies this *Handbook*, along with tabular data for finished designs. The input variables are desired plate load resistance, output impedance to be matched (usually 50 Ω), the desired loaded Q (typically 10 or 12) and the frequency. With these inputs one can calculate the values of the components C1, L1 and C2. These components are usually referred to as the plate tune capacitor, the tank inductor and the loading capacitor. In a multi-frequency amplifier, the coil inductance is changed with a band switch and the capacitors are adjusted to the correct value for the band in question.

Tank circuit component values are most easily found using computer software. The program *PI-EL Design* by Jim Tonne, W4ENE, is included on the *Handbook CD* and is illustrated here. With this software all of the components for a pi or pi-L network (described in the next section) can be quickly calculated. Since there are so many possible variables, especially with a pi-L network, it is impractical to publish graphical or tabular data to cover all cases. Therefore, the use of this software is highly recommended for those designing output networks. The software allows many "what-if" possibilities to be quickly checked and an optimum design found.

THE PI-L NETWORK

There are some advantages in using an additional inductor in the output network, effectively changing it from a pi network to a pi-L network as shown in the bottom right corner of **Fig 17.15**. The harmonic rejection is increased, as shown in **Fig 17.16**, and the

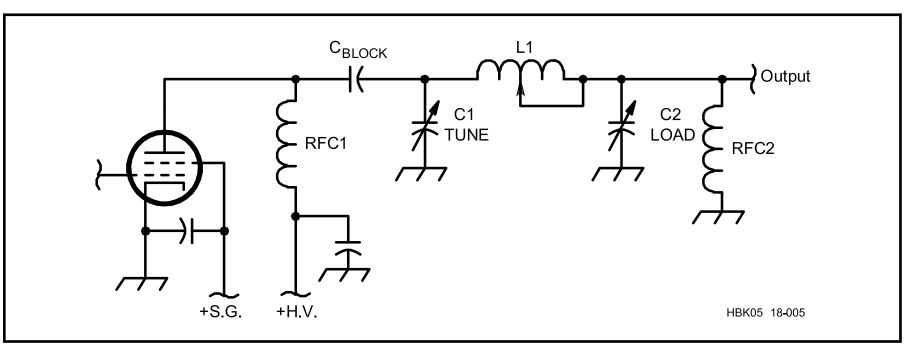


Fig 17.14 — A pi matching network used at the output of a tetrode power amplifier. RFC2 is used for protective purposes in the event C_{BLOCK} fails.

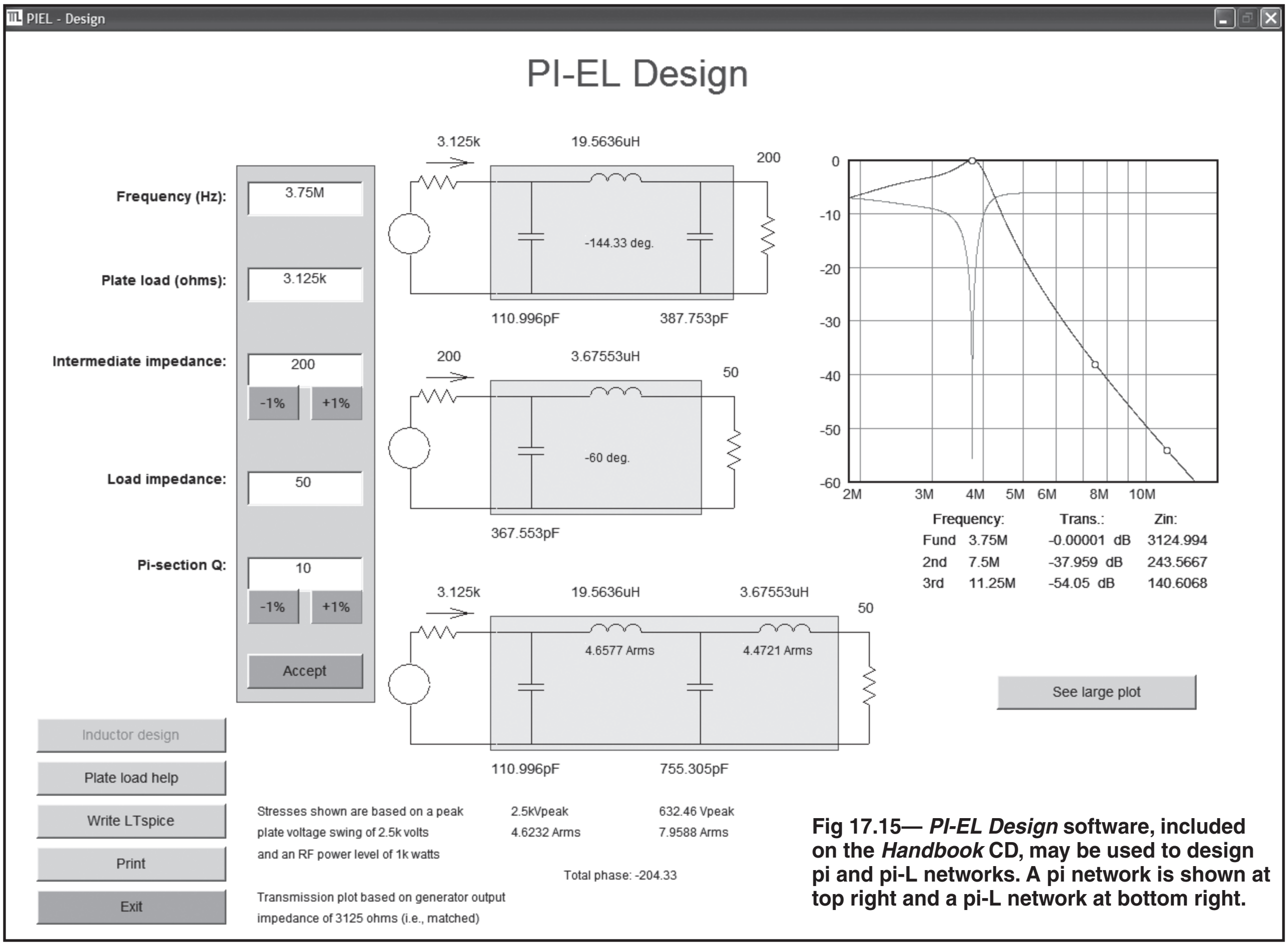


Fig 17.15— *PI-EL Design* software, included on the *Handbook* CD, may be used to design pi and pi-L networks. A pi network is shown at top right and a pi-L network at bottom right.

component values may become more convenient. Alternatively, the Q can be reduced to lower losses while retaining the same harmonic rejection as the simple pi. This can reduce maximum required tuning capacitance to more easily achievable values.

With a pi-L design, there are many more options than with the simple pi network. This is because the intermediate impedance can take on any value we wish to assign between the output impedance (usually 50 Ω) and the desired plate resistance. This intermediate impedance need not be the same for each frequency, providing the possibility of further optimizing the design. For that reason, it is especially desirable that the software be used instead of using the chart values when a pi-L is contemplated. Using this software, one can quickly determine component values for the required load resistance and Q values for any frequency as well as plotting the harmonic rejection values. Even the voltage and current ratings of the components are calculated.

Further analysis can be done using various versions of *SPICE*. A popular spice version is *LTspice* available from Linear Technologies and downloadable for free on their Web site, **www.linear.com**. The *PI-EL Design* software mentioned above generates files for *LTspice* automatically. *PI-EL Design* assumes that the

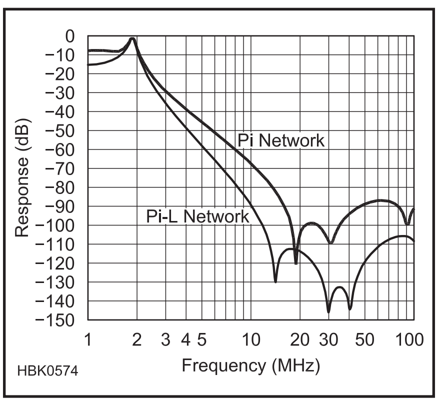


Fig 17.16 — Relative harmonic rejection of pi and pi-L circuits

blocking capacitor has negligible reactance and the RF choke has infinite reactance. There are times when these assumptions may not be valid. The effect of these components and changes to compensate for them can easily be evaluated using *LTspice*. It can also evaluate the effects of parasitic and stray effects, which all components have. The deep nulls in the response curves in Fig 17.16 are caused by the stray capacitance in the inductors. These can be used to advantage but, if not understood, can also lead to unexpected results. See the **RF Techniques** chapter for more information.

MANUAL METHODS FOR TANK DESIGN

For those who wish to try designing a pi network without a computer, pi designs in chart form are provided. These charts (**Fig 17.17** to **Fig 17.19**) give typical values for the pi network components for various bands and desired plate load resistance. For each value of load resistance the component values can be read for each band. For bands not shown, an approximate value can be reached by interpolation between the bands shown. It's not necessary to be able to read these values to high precision. In practice, unaccounted-for stray capacitance and inductance will likely make the calculated values only an approximate starting point. For those desiring greater precision, this data is available in tabular form on the *Handbook* CD along with the formulas for calculating them.

Several things become obvious from these charts. The required capacitance is reduced and the required inductance increased when a higher load resistance is used. This means that an amplifier with higher plate voltage and lower plate current will require smaller

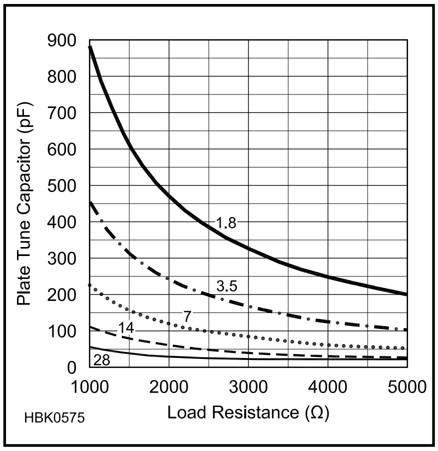


Fig 17.17 — Plate tuning capacitor values for various bands and values of load resistance. Figs 17.17 through 17.19 may be used for manual design of a tank circuit, as explained in the text.

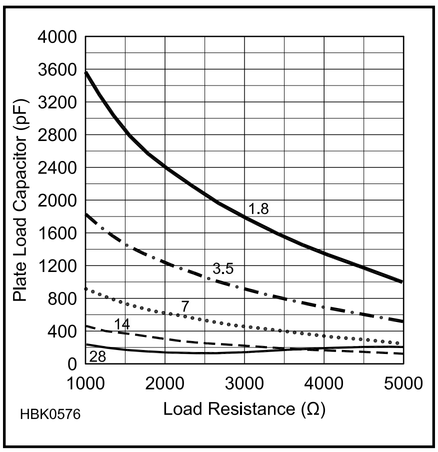


Fig 17.18 — Plate loading capacitor values for various bands and values of load resistance.

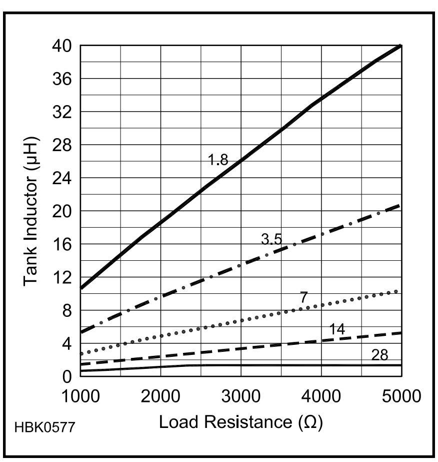


Fig 17.19 — Tank inductor values for various bands and values of load resistance.

capacitor values and larger inductors. The capacitors will, of course, also have to withstand higher voltages, so their physical size may or may not be any smaller. The inductors will have less current in them so a smaller size wire may be used. It will be obvious that for an amplifier covering many bands, the most challenging parts of the design are at the frequency extremes. The 1.8 MHz band requires the most inductance and capacitance and the 28 MHz band requires the least. Often, the output capacitance of the tube plus the minimum value of the plate capacitor along with assorted stray capacitance will put a lower limit on the effective plate capacitance that can be achieved. These charts assume that value to be 25 pF and the other components are adjusted to account for that minimum value. An inevitable trade-off here is that the Q of the tank will be higher than optimum on the highest bands.

When tuning to the lower frequency bands, the maximum value of the capacitor, especially the loading capacitor, will be a limiting factor. Sometimes, fixed mica or ceramic transmitting rated capacitors will be switched in parallel with the variable capacitor to reach the total value required.

COMPONENT SELECTION FOR THE PI-L NETWORK

For those wishing to use manual methods to design a pi-L network, there are look-up tables on the CD along with the mathematical formulas from which they are derived. **Table 17.1** is an abbreviated version that shows the general trends. Values shown are for 1.8 MHz. Other bands can be approximated by dividing all component values by the frequency ratio. For example, for 18 MHz, divide all component values by 10. For 3.6 MHz, divide values by two, and so on.

For a given Q_L, the pi-L circuit has better harmonic rejection than the pi circuit. This allows the designer to use a lower Q_L, resulting in lower losses and lower capacitor values, which is an advantage on the lower frequencies. However, the inductor values will be higher, and the lower capacitor values may be unachievable at higher frequencies. With the pi-L circuit there are many more variables to work with and, thus, one can try many different possibilities to make the circuit work within the limits of the components that are available.

PROBLEMS AT VHF AND HIGHER FREQUENCIES

As the size of a circuit approaches about 5% of a wavelength, components begin to seriously depart from the pure inductance or capacitance we assume them to be. Inductors begin to act like transmission lines. Capacitors often exhibit values far different from their marked values because of stray internal reactances and lead inductance. Therefore, tuned circuits are frequently fabricated in the form of striplines or other transmission lines in order to circumvent the problem of building "pure" inductances and capacitances. The choice of components is often more significant than the type of network used.

The high impedances encountered in VHF tube-amplifier plate circuits are not easily matched with typical networks. Tube output capacitance is usually so large that most matching networks are unsuitable. The usual practice is to resonate the tube output capacitance with a low-loss inductance connected in series or parallel. The result can be a very high-Q tank circuit. Component losses must be kept to an absolute minimum in order to achieve reasonable tank efficiency. Output impedance transformation is usually performed by a link inductively coupled to the tank circuit or by a parallel transformation of the output resistance using a series capacitor.

Since high values of plate load impedance call for low values of plate tuning capacitance, one might be tempted to add additional tubes in parallel to reduce the required load impedance. This only adds to the stray plate capacitance, and the potential for parasitic oscillations is increased. For these reasons, tubes in parallel are seldom used at VHF. Push-pull circuits offer some advantages, but with modern compact ceramic tube types, most VHF amplifiers use a single tube along with distributed type tuned networks. Other approaches are discussed later in this chapter.

"COLD TUNING" AN AMPLIFIER

Because of the high voltage and current involved, as well as the danger of damaging an expensive tube or other component, it is prudent to "cold tune" an amplifier before applying power to it. This can actually be done early in the construction as soon as tank components and the tube are in place. Cold tuning requires some test equipment, but is not difficult or time consuming. Only if you have a problem in getting the tuning right will it take much time, but that is exactly the case in which you would not want to turn on the power without having discovered that there is a problem. With cold tuning, you can also add and remove additional components, such as the RF choke, and see how much it affects the tuning.

There are always stray capacitances and inductances in larger sized equipment, so there is a good chance that your carefully designed circuits may not be quite right. Even with commercial equipment, you may want to become aware of the limitations of the tuning ranges and the approximate settings for the dials for each band. The equipment manual may provide this information, but what if the

Table 17.1
Pi-L Values for 1.8 MHz

Plate Load Resistance (Ω)	*Plate Tune Capacitor (pF)*	*Plate Inductor (µH)*	*Plate Load Capacitor (pF)*	*Output Inductor (µH)*	*Intermediate Resistance (Ω)*	*Loaded Q (Q_L)*	*Harmonic Attenuation (3rd/5th, dB)*
1000	883	10.5	3550	None (pi)	None (pi)	12	–30/–42
1000	740	15	2370	7.7	200	12	–38/–54
1000	499	21.5	1810	7.7	200	8	–35/–51
2200	376	26.4	1938	7.7	200	12	–39/–55
2200	255	37.7	1498	7.7	200	8	–36/–52
5000	180	50.4	1558	7.7	200	12	–39/–55
5000	124	71	1210	7.7	200	8	–36/–53
5000	114	83	928	11.7	400	8	–39/–55

amplifier you have is a bit out of calibration? In all of these cases, cold tests provide cheap insurance against damage caused by bad tuning and, at the same time, give you practice in setting up.

The first step is to ensure that the equipment is truly cold by removing the power plug from the wall. Since you will be working around the high voltage circuits, you may want to remove fuses or otherwise ensure that power cannot come on. In a well-designed amplifier there will be interlocks that prevent turn on and, perhaps, also short out the plate voltage. If these are in a place that affects the RF circuits, they may have to be temporarily removed. Just be sure to put them back when done.

The adjustment of the plate circuit components is the most important. The easiest way to check those is to attach a resistor across the tube from plate to ground. The resistor should be the same value as the design load resistance and must be noninductive. Several series resistors in the 500 Ω range, either carbon film or the older carbon composition type will work, but *not* wire wound. The tube must remain in its socket and all the normal connections to it should be in place. Covers should be installed, at least for the final tests, since they may affect tuning.

Connect a test instrument to the 50 Ω output. This can be an antenna bridge or other 50 Ω measuring device. Examples of suitable test equipment would be a vector network analyzer, preferably with an impedance step up transformer, a vector impedance meter or an RX meter, such as the Boonton 250A. If the wattage of the resistor across the plate can stand it, it can even be a low power transmitter.

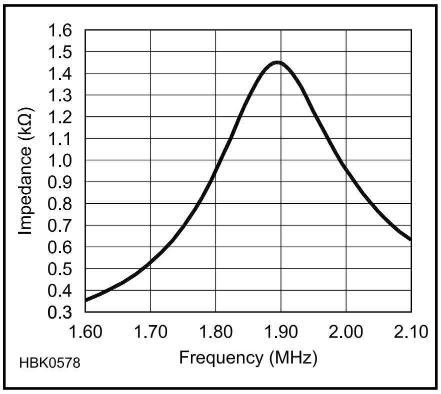


Fig 17.20 — Impedance as would be measured at the plate of an amplifier tube with the output network tuned to 1.9 MHz and the output terminated in its proper load.

Select the correct band and tune the plate and load capacitors so that the instrument at the output connector shows 50 Ω or a low SWR at the 50 Ω point. Since this type of circuit is bilateral, you now have settings that will be the same ones you need to transform the 50 Ω load to look like the resistive load you used at the tube plate for the test. If you have an instrument capable of measuring relatively high impedances at RF frequencies, you can terminate the output with 50 Ω and measure the impedance on the plate. It will look something like **Fig 17.20**.

A similar test will work for the input, although it is sometimes more difficult to know what the input impedance at the tube will be. In fact, it will change somewhat with drive level so a low power cold test may not give a full picture. However, just like the case of the plate circuit, you can put your best estimate of the input impedance at the grid using a noninductive resistor. Then, tune the input circuits for a match.

An old timer's method for tuning these circuits, especially the plate circuit, is to use a dip meter. These are getting pretty hard to find these days and, in any case, only show resonance. You won't know if the transformation ratio between plate and output is correct, but it is better than nothing. At least, if you can't get a dip at the proper frequency, you will know that something is definitely wrong.

17.5 Transmitting Device Ratings

17.5.1 Plate, Screen and Grid Dissipation

The ultimate factor limiting the power-handling capability of a tube is often (but not always) its maximum plate dissipation rating. This is the measure of how many watts of heat the tube can safely dissipate, if it is cooled properly, without exceeding critical temperatures. Excessive temperature can damage or destroy internal tube components or vacuum seals — resulting in tube failure. The same tube may have different voltage, current and power ratings depending on the conditions under which it is operated, but its safe temperature ratings must not be exceeded in any case! Important cooling considerations are discussed in more detail later in this chapter.

The efficiency of a power amplifier may range from approximately 25% to 85%, depending on its operating class, adjustment and circuit losses. The efficiency indicates how much of the dc power supplied to the stage is converted to useful RF output power; the rest is dissipated as heat, mostly by the plate. The *TubeCalculator* program will calculate the dissipation of the plate as one of its outputs. Otherwise, it can be determined by multiplying the plate voltage (V) times the plate current (A) and subtracting the output power (W).

For a class AB amplifier, the resting dissipation should also be noted, since with no RF input, *all* of the dc power is dissipated in the plate. Multiply plate voltage times the resting plate current to find this resting dissipation value. Screen dissipation is simply screen voltage times screen current. Grid dissipation is a bit more complicated since some of the power into the grid goes into the bias supply, some is passed through to the output (when grounded grid is used) and some is dissipated in the grid. Some tubes have very fragile grids and cannot be run with any grid current at all.

Almost all vacuum-tube power amplifiers in amateur service today operate as linear amplifiers (Class AB or B) with efficiencies of approximately 50% to 65%. That means that a useful power output of approximately 1 to 2 times the plate dissipation generally can be achieved. This requires, of course, that the tube is cooled enough to realize its maximum plate dissipation rating and that no other tube rating, such as maximum plate current or grid dissipation, is exceeded.

Type of modulation and duty cycle also influence how much output power can be achieved for a given tube dissipation. Some types of operation are less efficient than others, meaning that the tube must dissipate more heat. Some forms of modulation, such as CW or SSB, are intermittent in nature, causing less average heating than modulation formats in which there is continuous transmission (RTTY or FM, for example).

Power-tube manufacturers use two different rating systems to allow for the variations in service. CCS (Continuous Commercial Service) is the more conservative rating and is used for specifying tubes that are in constant use at full power. The second rating system is based on intermittent, low-duty-cycle operation, and is known as ICAS (Intermittent Commercial and Amateur Service). ICAS ratings are normally used by commercial manufacturers and individual amateurs who wish to obtain maximum power output consistent with reasonable tube life in CW and SSB service. CCS ratings should be used for FM, RTTY and SSTV applications. (Plate power transformers for amateur service are also rated in CCS and ICAS terms.).

MAXIMUM RATINGS

Tube manufacturers publish sets of maximum values for the tubes they produce. No maximum rated value should ever be exceeded. As an example, a tube might have a maximum plate-voltage rating of 2500 V, a maximum plate-current rating of 500 mA, and a maximum plate dissipation rating of 350 W. Although the plate voltage and current ratings might seem to imply a safe power input of 2500 V × 500 mA = 1250 W, this is true only if the dissipation rating will not be exceeded. If the tube is used in class AB2 with an expected efficiency of 60%, the maximum safe dc power input is

$$P_{IN} = \frac{100 P_D}{100 - N_D} = \frac{100 \times 350}{100 - 60} = 875 \text{ W}$$

Table 17.2
Typical Tank-Capacitor Plate Spacings

Spacing Inches	*Peak Voltage*	*Spacing Inches*	*Peak Voltage*	*Spacing Inches*	*Peak Voltage*
0.015	1000	0.07	3000	0.175	7000
0.02	1200	0.08	3500	0.25	9000
0.03	1500	0.125	4500	0.35	11000
0.05	2000	0.15	6000	0.5	13000

17.5.2 Tank Circuit Components

CAPACITOR RATINGS

The tank capacitor in a high-power amplifier should be chosen with sufficient spacing between plates to preclude high-voltage breakdown. The peak RF voltage present across a properly loaded tank circuit, without modulation, may be taken conservatively as being equal to the dc plate voltage. If the dc supply voltage also appears across the tank capacitor, this must be added to the peak RF voltage, making the total peak voltage twice the dc supply voltage. At the higher voltages, it is usually desirable to design the tank circuit so that the dc supply voltages do not appear across the tank capacitor, thereby allowing the use of a smaller capacitor with less plate spacing. Capacitor manufacturers usually rate their products in terms of the peak voltage between plates. Typical plate spacings are given in **Table 17.2**.

Output tank capacitors should be mounted as close to the tube as possible to allow short low inductance leads to the plate. Especially at the higher frequencies, where minimum circuit capacitance becomes important, the capacitor should be mounted with its stator plates well spaced from the chassis or other shielding. In circuits in which the rotor must be insulated from ground, the capacitor should be mounted on ceramic insulators of a size commensurate with the plate voltage involved and — most important of all, from the viewpoint of safety to the operator — a well-insulated coupling should be used between the capacitor shaft and the knob. The section of the shaft attached to the control knob should be well grounded. This can be done conveniently by means of a metal shaft bushing at the panel.

COIL RATINGS

Tank coils should be mounted at least half their diameter away from shielding or other large metal surfaces, such as blower housings, to prevent a marked loss in Q. Except perhaps at 24 and 28 MHz, it is not essential that the coil be mounted extremely close to the tank capacitor. Leads up to 6 or 8 inches are permissible. It is more important to keep the tank capacitor, as well as other components, out of the immediate field of the coil.

The principal practical considerations in designing a tank coil usually are to select a conductor size and coil shape that will fit into available space and handle the required power without excessive heating. Excessive power loss as such is not necessarily the worst hazard in using too-small a conductor. It is not uncommon for the heat generated to actually unsolder joints in the tank circuit and lead to physical damage or failure. For this reason it's extremely important, especially at power levels above a few hundred watts, to ensure that all electrical joints in the tank circuit are secured mechanically as well as soldered.

Table 17.3 shows recommended conductor sizes for amplifier tank coils, assuming loaded tank circuit Q of 15 or less on the 24 and 30 MHz bands and 8 to 12 on the lower frequency bands. In the case of input circuits for screen-grid tubes where driving power is quite small, loss is relatively unimportant and almost any physically convenient wire

Table 17.3
Copper Conductor Sizes for Transmitting Coils for Tube Transmitters

Power Output (W)	*Band (MHz)*	*Minimum Conductor Size*
1500	1.8-3.5	10
	7-14	8 or ⅛"
	18-28	6 or 3/16"
500	1.8-3.5	12
	7-14	10
	18-28	8 or ⅛"
150	1.8-3.5	16
	7-14	12
	18-28	10

*Whole numbers are AWG;
fractions of inches are tubing ODs.

size and coil shape is adequate.

The conductor sizes in Table 17.3 are based on experience in continuous-duty amateur CW, SSB and RTTY service and assume that the coils are located in a reasonably well ventilated enclosure. If the tank area is not well ventilated and/or if significant tube heat is transferred to the coils, it is good practice to increase AWG wire sizes by two (for example, change from #12 to #10) and tubing sizes by 1⁄16 inch.

Larger conductors than required for current handling are often used to maximize unloaded Q, particularly at higher frequencies. Where skin depth effects increase losses, the greater surface area of large diameter conductors can be beneficial. Small-diameter copper tubing, up to 3⁄8 inch outer diameter, can be used successfully for tank coils up through the lower VHF range. Copper tubing in sizes suitable for constructing high-power coils is generally available in 50-ft rolls from plumbing and refrigeration equipment suppliers. Silver-plating the tubing may further reduce losses. This is especially true as the tubing ages and oxidizes. Silver oxide is a much better conductor than copper oxide, so silver-plated tank coils maintain their low-loss characteristics even after years of use. (There is some debate in amateur circles about the benefits of silver plating.).

At VHF and above, tank circuit inductances do not necessarily resemble the familiar coil. The inductances required to resonate tank circuits of reasonable Q at these higher frequencies are small enough that only strip lines or sections of transmission line are practical. Since these are constructed from sheet metal or large diameter tubing, current-handling capabilities normally are not a relevant factor.

17.5.3 Other Components

RF CHOKES

The characteristics of any RF choke vary with frequency. At low frequencies the choke presents a nearly pure inductance. At some higher frequency it takes on high impedance characteristics resembling those of a parallel-resonant circuit. At a still higher frequency it goes through a series-resonant condition, where the impedance is lowest — generally much too low to perform satisfactorily as a shunt-feed plate choke. As frequency increases further, the pattern of alternating parallel and series resonances repeats. Between resonances, the choke will show widely varying amounts of inductive or capacitive reactance.

In most high-power amplifiers, the choke is directly in parallel with the tank circuit, and is subject to the full tank RF voltage. See **Fig 17.21A**. If the choke does not present a sufficiently high impedance, enough power will be absorbed by the choke to burn it out. To avoid this, the choke must have a sufficiently high reactance to be effective at the lowest frequency (at least equal to the plate load resistance) and yet have no series resonances near any of the higher frequency bands. A resonant-choke failure in a high-power amplifier can be very dramatic and damaging!

Thus, any choke intended for shunt-feed use should be carefully investigated. The best way would be to measure its reactance to ground with an impedance measuring instrument. If the dip meter is used, the choke must be shorted end-to-end with a direct, heavy braid or strap. Because nearby metallic objects affect the resonances, it should be mounted in its intended position, but disconnected from the rest of the circuit. A dip meter coupled an inch or two away from one end of the choke nearly always will show a deep, sharp dip at the lowest series-resonant frequency and shallower dips at higher series resonances.

Any choke to be used in an amplifier for the 1.8 to 28 MHz bands requires careful (or at least lucky!) design to perform well on all amateur bands within that range. Most simply put, the challenge is to achieve sufficient inductance that the choke doesn't "cancel" a large part of tuning capacitance at 1.8 MHz. At the same time, try to position all its series resonances where they can do no harm. In general, close wind enough #20 to #24 magnet wire to provide about 135 µH inductance on a 3⁄4 to 1-inch diameter cylindrical form of ceramic, Teflon or fiberglass. This gives a reactance of 1500 Ω at 1.8 MHz and yet yields a first series resonance in the vicinity of 25 MHz. Before the advent of the 24 MHz band this worked fine. But trying to "squeeze" the resonance into the narrow gaps between the 21, 24 and/or 28-MHz bands is quite risky unless sophisticated instrumentation is available. If the number of turns on the choke is selected to place its first series resonance at 23.2 MHz, midway between 21.45 and 24.89 MHz, the choke impedance will typically be high enough for satisfactory operation on the 21, 24 and 28 MHz bands. The choke's first series resonance should be measured very carefully as described above using a dip meter and calibrated receiver or RF impedance bridge, with the choke mounted in place on the chassis.

Investigations with a vector impedance meter have shown that "trick" designs, such as using several shorter windings spaced along the form, show little if any improvement in choke resonance characteristics. Some commercial amplifiers circumvent the problem by band switching the RF choke. Using a larger diameter (1 to 1.5 inches) form does move the first series resonance somewhat higher for a given value of basic inductance. Beyond that, it is probably easiest for an all-band amplifier to add or subtract enough turns to move the first resonance to about 35 MHz and settle for a little less than optimum reactance on 1.8 MHz.

However, there are other alternatives. If one is willing to switch the choke when changing bands, it is possible to have enough inductance for 1.8 to 10 MHz, with series resonances well above 15 MHz. Then for 14 MHz and above, a smaller choke is used which has its resonances well above 30 MHz. Providing an extra pole on the band switch is, of course, the trade-off. This switch must withstand the full plate voltage. Switches suitable for changing bands for the pi network would handle this fine.

Another approach is to feed the high-voltage dc through the main tank inductor, putting the RF choke at the loading capacitor,

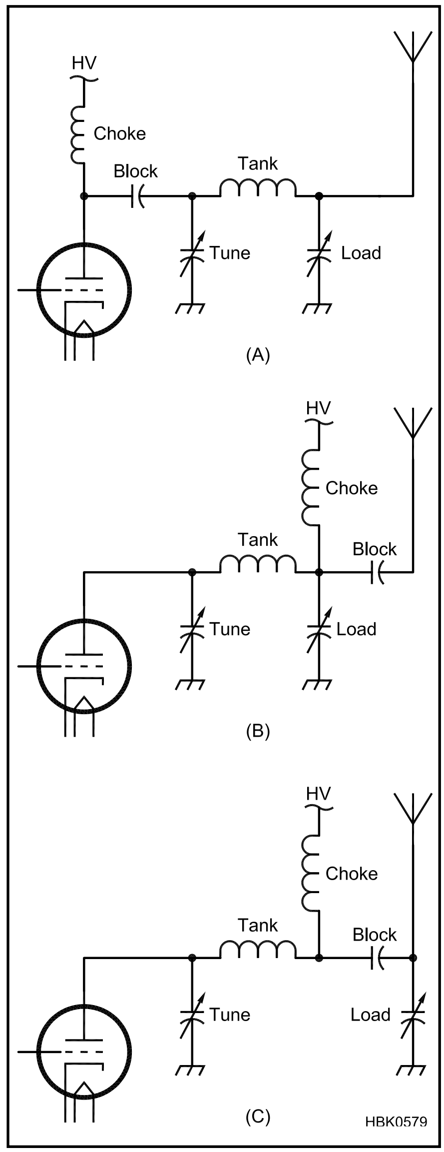


Fig 17.21 — Three ways of feeding dc to a tube via an RF choke. See text for a discussion of the tradeoffs.

instead of at the tube. (See Fig 17.21B) This puts a much lower RF voltage on the choke and, thus, not as much reactance is required for satisfactory rejection of the RF voltage. However, this puts both dc and RF voltages on the plate and loading capacitors which may be beyond their ratings. The blocking capacitor can be put before the loading capacitor, as in Fig 17.21C. This removes the dc from the loading capacitor, which typically has a lower voltage rating than the plate capacitor, but puts high current in the blocker.

Yet another method involves using hollow tubing for the plate tank and passing the dc lead through it. This lowers the RF voltage on the choke without putting dc voltage on the tuning components. This method works best for higher power transmitters where the tuning inductor can be made of ⅛ inch or larger copper tubing.

BLOCKING CAPACITORS

A series capacitor is usually used at the input of the amplifier output circuit. Its purpose is to block dc from appearing on matching circuit components of the antenna. As mentioned in the section on tank capacitors, output-circuit voltage requirements are considerably reduced when only RF voltage is present.

To provide a margin of safety, the voltage rating for a blocking capacitor should be at least 25% to 50% greater than the dc voltage applied. A large safety margin is desirable, since blocking capacitor failure can bring catastrophic results. The worse case is when dc is applied to the output of the transmitter and even to the antenna, with potentially fatal results. Often an RF choke is placed from the RF output jack to ground as a safety backup. A shorted blocker will blow the power supply fuse.

To avoid affecting the amplifier's tuning and matching characteristics, the blocking capacitor should have a low impedance at all operating frequencies. If it presents more than 5% of the plate load resistance, the pi components should be adjusted to compensate. Use of a *SPICE* analysis provides a useful way to see what adjustments might be required to maintain the desired match.

The capacitor also must be capable of handling, without overheating or significantly changing value, the substantial RF current that flows through it. This current usually is greatest at the highest frequency of operation where tube output capacitance constitutes a significant part of the total tank capacitance. A significant portion of circulating tank current, therefore, flows through the blocking capacitor. When using the connection of the RF choke shown in Fig 17.21C, the entire circulating current must be accommodated.

Transmitting capacitors are rated by their manufacturers in terms of their RF current-carrying capacity at various frequencies. Below a couple hundred watts at the high frequencies, ordinary disc ceramic capacitors of suitable voltage rating work well in high-impedance tube amplifier output circuits. Some larger disk capacitors rated at 5 to 8 kV also work well for higher power levels at HF. For example, two inexpensive Centralab type DD-602 discs (0.002 µF, 6 kV) in parallel have proved to be a reliable blocking capacitor for 1.5-kW amplifiers operating at plate voltages to about 2.5 kV. At very high power and voltage levels and at VHF, ceramic "doorknob" transmitting capacitors are needed for their low losses and high current handling capabilities. When in doubt, adding additional capacitors in parallel is cheap insurance against blocking capacitor failure and also reduces the impedance. So-called "TV doorknobs" may break down at high RF current levels and should be avoided.

The very high values of Q_L found in many VHF and UHF tube-type amplifier tank circuits often require custom fabrication of the blocking capacitor. This can usually be accommodated through the use of a Teflon "sandwich" capacitor. Here, the blocking capacitor is formed from two parallel plates separated by a thin layer of Teflon. This capacitor often is part of the tank circuit itself, forming a very low-loss blocking capacitor. Teflon is rated for a minimum breakdown voltage of 2000 V per mil of thickness, so voltage breakdown should not be a factor in any practically realized circuit. The capacitance formed from such a Teflon sandwich can be calculated from the information presented elsewhere in this *Handbook* (use a dielectric constant of 2.1 for Teflon). In order to prevent any potential irregularities caused by dielectric thickness variations (including air gaps), Dow-Corning DC-4 silicone grease should be evenly applied to both sides of the Teflon dielectric. This grease has properties similar to Teflon, and will fill in any surface irregularities that might cause problems.

17.6 Sources of Operating Voltages

17.6.1 Tube Filament or Heater Voltage

A power vacuum tube can use either a directly heated filament or an indirectly heated cathode. The filament voltage for either type should be held within 5% of rated voltage. Because of internal tube heating at UHF and higher, the manufacturers' filament voltage rating often is reduced at these higher frequencies. The derated filament voltages should be followed carefully to maximize tube life.

Series dropping resistors may be required in the filament circuit to attain the correct voltage. Adding resistance in series will also reduce the inrush current when the tube is turned on. Cold tungsten has much lower resistance than when hot. Circuits are available that both limit the inrush current at turn on and also regulate the voltage against changes in line voltage.

The voltage should be measured with a true RMS meter at the filament pins of the tube socket while the amplifier is running. The filament choke and interconnecting wiring all have voltage drops associated with them. The high current drawn by a power-tube heater circuit causes substantial voltage drops to occur across even small resistances. Also, make sure that the plate power drawn from the power line does not cause the filament voltage to drop below the proper value when plate power is applied.

Thoriated filaments lose emission when the tube is overloaded appreciably. If the overload has not been too prolonged, emission, sometimes, may be restored by operating the filament at rated voltage, with all other voltages removed, for a period of 30 to 60 minutes. Alternatively, you might try operating the tube at 20% above rated filament voltage for five to ten minutes.

17.6.2 Vacuum-Tube Plate Voltage

DC plate voltage for the operation of RF amplifiers is most often obtained from a transformer-rectifier-filter system (see the **Power Supplies** chapter) designed to deliver the required plate voltage at the required current. It is not unusual for a power tube to arc over internally (generally from the plate to the screen or control grid) once or twice, especially soon after it is first placed into service. The flashover by itself is not normally dangerous to the tube, provided that instantaneous maximum plate current to the tube is held to a safe value and the high-voltage plate supply is shut off very quickly.

A good protective measure against this is the inclusion of a high-wattage power resistor in series with the plate high-voltage circuit. The value of the resistor, in ohms, should be approximately 10 to 15 times the no-load plate voltage in KV. This will limit peak fault current to 67 to 100 A. The series resistor should be rated for 25 or 50-W power dissipation; vitreous enamel coated wire-wound resistors have been found to be capable of handling repeated momentary fault-current surges without damage. Aluminum-cased resistors (Dale) are not recommended for this application. Each resistor also must be large enough to safely handle the maximum value of normal plate current; the wattage rating required may be calculated from $P = I^2R$. If the total filter capacitance exceeds 25 µF, it is a good idea to use 50-W resistors in any case. Even at high plate-current levels, the addition of the resistors does little to affect the dynamic regulation of the plate supply.

Since tube (or other high-voltage circuit) arcs are not necessarily self-extinguishing, a fast-acting plate overcurrent relay or primary circuit breaker is also recommended to quickly shut off ac power to the HV supply when an arc begins. Using this protective system, a mild HV flashover may go undetected, while a more severe one will remove ac power from the HV supply. (The cooling blower should remain energized, however, since the tube may be hot when the HV is removed due to an arc.) If effective protection is not provided, however, a "normal" flashover, even in a new tube, is likely to damage or destroy the tube, and also frequently destroys the rectifiers in the power supply as well as the plate RF choke. A power tube that flashes over more than about 3 to 5 times in a period of several months likely is defective and will have to be replaced before long.

17.6.3 Grid Bias

The grid bias for a linear amplifier should be highly filtered and well regulated. Any ripple or other voltage change in the bias circuit modulates the amplifier. This causes hum and/or distortion to appear on the signal. Since most linear amplifiers draw only small amounts of grid current, these bias-supply requirements are not difficult to achieve.

Fixed bias for class AB1 tetrode and pentode amplifiers is usually obtained from a variable-voltage regulated supply. Voltage adjustment allows setting bias level to give the desired resting plate current. **Fig 17.22A** shows a simple Zener-diode-regulated bias supply. The dropping resistor is chosen to allow approximately 10 mA of Zener current. Bias is then reasonably well regulated for all drive conditions up to 2 or 3 mA of grid current. The potentiometer allows bias to be adjusted between Zener and approximately 10 V higher. This range is usually adequate to allow for variations in the characteristics of different tubes. Under standby conditions, when it is desirable to cut off the tube entirely, the Zener ground return is interrupted so the full bias supply voltage is applied to the grid.

In Fig 17.22B and C, bias is obtained from the voltage drop across a Zener diode in the cathode (or filament center-tap) lead. Operating bias is obtained by the voltage drop across D1 as a result of plate (and screen) current flow. The diode voltage drop effectively raises the cathode potential relative to the grid. The grid is, therefore, negative with respect to the cathode by the Zener voltage of the diode. The Zener-diode wattage rating should be twice the product of the maximum cathode current times the rated Zener voltage. Therefore, a tube requiring 15 V of bias with a maximum cathode current of 100 mA would dissipate 1.5 W in the Zener diode. To allow a suitable safety factor, the diode rating should be 3 W or more. The circuit of Fig 17.22C illustrates how D1 would be used with a cathode driven (grounded grid) amplifier as opposed to the grid driven example at B.

In all cases, the Zener diode should be bypassed by a 0.01-µF capacitor of suitable voltage. Current flow through any type of diode generates shot noise. If not bypassed, this noise would modulate the amplified signal, causing distortion in the amplifier output.

17.6.4 Screen Voltage For Tubes

Power tetrode screen current varies widely with both excitation and loading. The current may be either positive or negative, depending on tube characteristics and amplifier operating conditions. In a linear amplifier, the screen voltage should be well regulated for all values of screen current. The power output from a tetrode is very sensitive to screen voltage, and any dynamic change in the screen potential can cause distorted output. Zener diodes are commonly used for screen regulation.

Fig 17.23 shows a typical example of a regulated screen supply for a power tetrode amplifier. The voltage from a fixed dc supply is dropped to the Zener stack voltage by the current-limiting resistor. A screen bleeder resistor is connected in parallel with the Zener stack to allow for the negative screen current developed under certain tube operating conditions. Bleeder current is chosen to be roughly 10 to 20 mA greater than the expected maximum negative screen current, so that screen voltage is regulated for all values of current between maximum negative screen current and maximum positive screen current. For external-anode tubes in the 4CX250 family, a typical screen bleed-

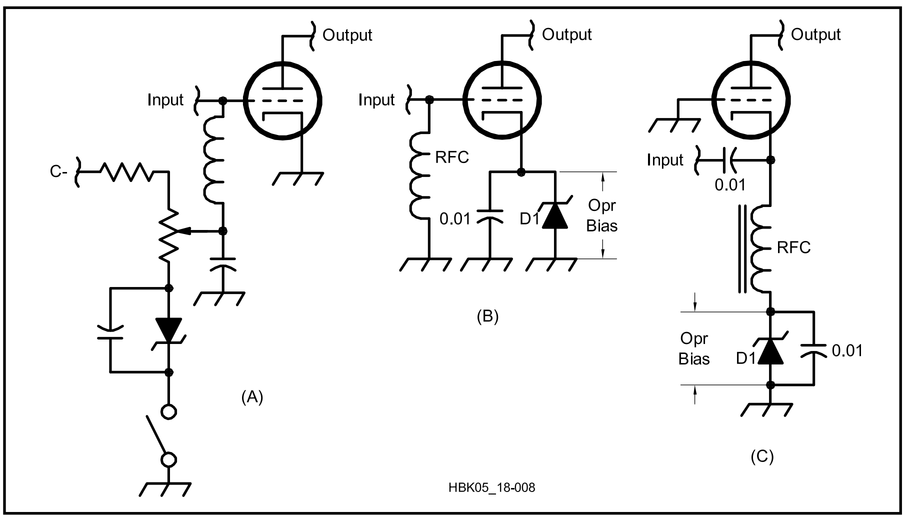


Fig 17.22 — Various techniques for providing operating bias with tube amplifiers.

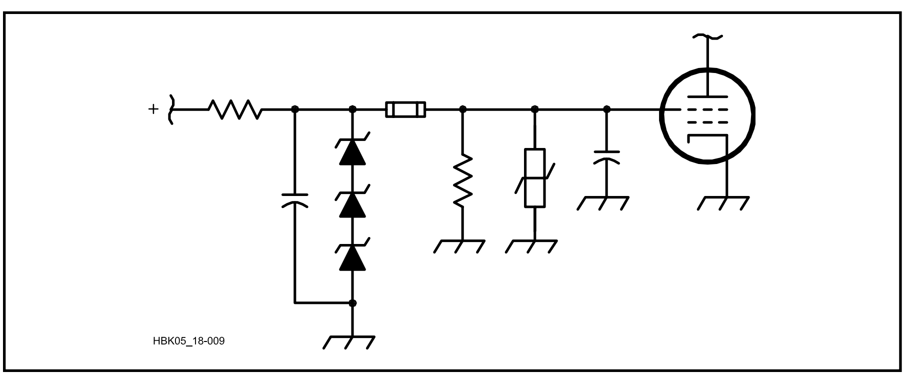


Fig 17.23 — A Zener-regulated screen supply for use with a tetrode. Protection is provided by a fuse and a varistor.

er current value would be 20 mA. For the 4CX1000 family, a screen-bleeder current of 70 mA is required.

Screen voltage should never be applied to a tetrode unless plate voltage and load also are applied; otherwise, the screen will act like an anode and will draw excessive current. Perhaps the best way to insure this is to include logic circuits that will not allow the screen supply to turn on until it senses plate voltage. Supplying the screen through a series-dropping resistor from the plate supply affords a measure of protection, since the screen voltage only appears when there is plate voltage. Alternatively, a fuse can be placed between the regulator and the bleeder resistor. The fuse should not be installed between the bleeder resistor and the tube because the tube should never be operated without a load on the screen. Without a load, the screen potential tends to rise to the anode voltage. Any screen bypass capacitors or other associated circuits are likely be damaged by this high voltage.

In Fig 17.23, a varistor is connected from screen to ground. If, because of some circuit failure, the screen voltage should rise substantially above its nominal level, the varistor will conduct and clamp the screen voltage to a low level. If necessary to protect the varistor or screen dropping resistors, a fuse or overcurrent relay may be used to shut off the screen supply so that power is interrupted before any damage occurs. The varistor voltage should be approximately 30% to 50% higher than normal screen voltage.

17.7 Tube Amplifier Cooling

Vacuum tubes must be operated within the temperature range specified by the manufacturer if long tube life is to be achieved. Tubes having glass envelopes and rated at up to 25-W plate dissipation may be used without forced-air cooling if the design allows a reasonable amount of convection cooling. If a perforated metal enclosure is used, and a ring of ¼ to ⅜-inch-diameter holes is placed around the tube socket, normal convective airflow can be relied on to remove excess heat at room temperatures.

For tubes with greater plate dissipation ratings, and even for very small tubes operated close to maximum rated dissipation, forced-air cooling with a fan or blower is needed. Most manufacturers rate tube-cooling requirements for continuous-duty operation. Their literature will indicate the required volume of airflow, in cubic feet per minute (CFM), at some particular back pressure. Often, this data is given for several different values of plate dissipation, ambient air temperature and even altitude above sea level.

One extremely important consideration is often overlooked by power-amplifier designers and users alike: a tube's plate dissipation rating is only its maximum potential capability. The power that it can actually dissipate safely depends directly on the cooling provided. The actual power capability of virtually all tubes used in high-power amplifiers for amateur service depends on the volume of air forced through the tube's cooling structure.

Table 17.4
Specifications of Some Popular Tubes, Sockets and Chimneys

Tube	*CFM*	*Back Pressure (inches)*	*Socket*	*Chimney*
3-500Z	13	0.13	SK-400, SK-410	SK-416
3CX800A7	19	0.50	SK-1900	SK-1906
3CX1200A7	31	0.45	SK-410	SK-436
3CX1200Z7	42	0.30	SK-410	—
3CX1500/8877	35	0.41	SK-2200, SK-2210	SK-2216
4-400A/8438	14	0.25	SK-400, SK-410	SK-406
4-1000A/8166	20	0.60	SK-500, SK-510	SK-506
4CX250R/7850	6.4	0.59	SK602A, SK-610, SK-610A SK-611, SK-612, SK-620, SK-620A, SK-621, SK-630	
4CX400/8874	8.6	0.37	SK1900	SK606
4CX400A	8	0.20	SK2A	—
4CX800A	20	0.50	SK1A	—
4CX1000A/8168	25	0.20	SK-800B, SK-810B, SK-890B	SK-806
4CX1500B/8660	34	0.60	SK-800B, SK-1900	SK-806
4CX1600B	36	0.40	SK3A	CH-1600B

These values are for sea-level elevation. For locations well above sea level (5000 ft/1500 m, for example), add an additional 20% to the figure listed.

17.7.1 Blower Specifications

This requirement usually is given in terms of cubic feet of air per minute (CFM), delivered into a back pressure, representing the resistance of the tube cooler to air flow, stated in inches of water. Both the CFM of airflow required and the pressure needed to force it through the cooling system are determined by ambient air temperature and altitude (air density), as well as by the amount of heat to be dissipated. The cooling fan or blower must be capable of delivering the specified airflow into the corresponding back pressure. As a result of basic air flow and heat transfer principles, the volume of airflow required through the tube cooler increases considerably faster than the plate dissipation, and back pressure increases even faster than airflow.

Table 17.5
Blower Performance Specifications

Wheel Dia	*Wheel Width*	*RPM*	*Free Air CFM*	*CFM for Back Pressure (inches) 0.1*	*0.2*	*0.3*	*0.4*	*0.5*	*Stock No.*
2"	1"	3340	12	9	6	—	—	—	1TDN2
2¹⁵⁄₁₆"	1½"	3388	53	52	50	47	41	23	1TDN5
3"	1⅞"	3036	50	48	44	39	32	18	1TDN7
3"	1⅞"	3010	89	85	78	74	66	58	1TDP1
3¹⁵⁄₁₆"	1¹⁵⁄₁₆"	3016	75	71	68	66	61	56	1TDP3
3¾"	1⅞"	2860	131	127	119	118	112	105	1TDP5

Representative sample of Dayton squirrel cage blowers. More information and other models available from Grainger Industrial Supply (**www.grainger.com**).

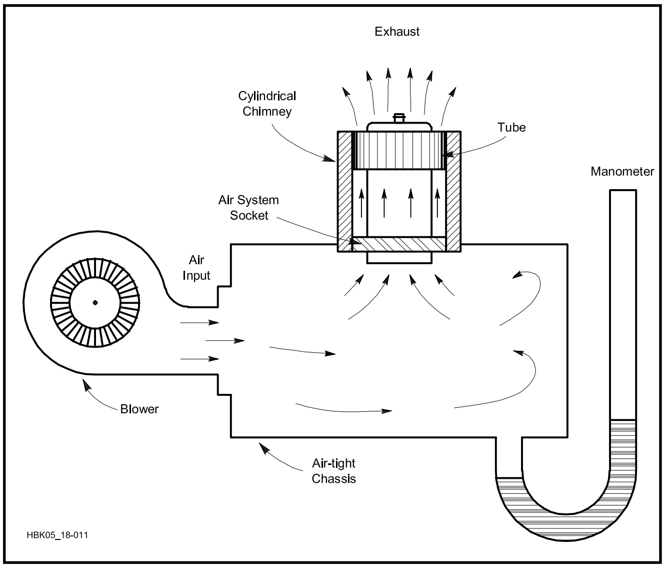


Fig 17.24 — Air is forced into the chassis by the blower and exits through the tube socket. The manometer is used to measure system back pressure, which is an important factor in determining the proper size blower.

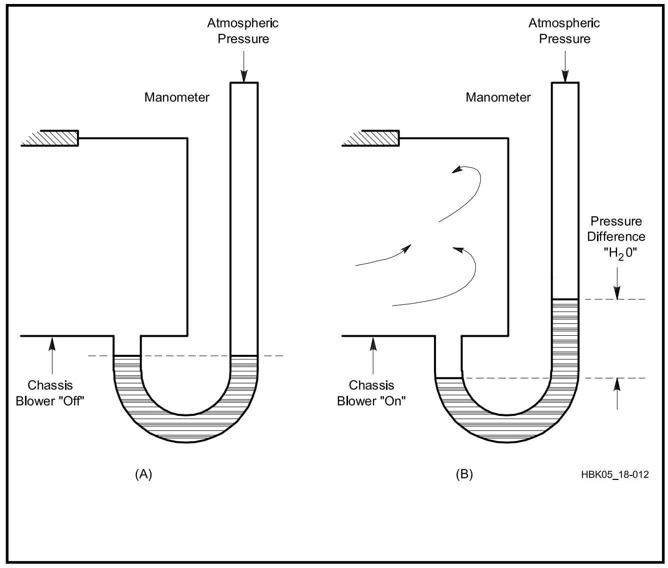


Fig 17.25 — At A the blower is "off" and the water will seek its own level in the manometer. At B the blower is "on" and the amount of back pressure in terms of inches of water can be measured as indicated.

In addition, blower air output decreases with increasing back pressure until, at the blower's so-called "cutoff pressure," actual air delivery is zero. Larger and/or faster-rotating blowers are required to deliver larger volumes of air at higher back pressure.

Values of CFM and back pressure required to realize maximum rated plate dissipation for some of the more popular tubes, sockets and chimneys (with 25 °C ambient air and at sea level) are given in **Table 17.4**. Back pressure is specified in inches of water and can be measured easily in an operational air system as indicated in **Figs 17.24** and **17.25**. The pressure differential between the air passage and atmospheric pressure is measured with a device called a *manometer*. A manometer is nothing more than a piece of clear tubing, open at both ends and fashioned in the shape of a "U." The manometer is temporarily connected to the chassis and is removed after the measurements are completed. As shown in the diagrams, a small amount of water is placed in the tube. At Fig 17.25A, the blower is "off" and the water seeks its own level, because the air pressure (ordinary atmospheric pressure) is the same at both ends of the manometer tube. At B, the blower is "on" (socket, tube and chimney in place) and the pressure difference, in terms of inches of water, is measured. For most applications, a standard ruler used for measurement will yield sufficiently accurate results.

Table 17.5 gives the performance specifications for a few of the many Dayton blowers, which are available through Grainger Industrial Supply (**www.grainger.com**). Other blowers having wheel diameters, widths and rotational speeds similar to any in Table 17.5 likely will have similar flow and back pressure characteristics. If in doubt about specifications, consult the manufacturer. Tube temperature under actual operating conditions is the ultimate criterion for cooling adequacy and may be determined using special crayons or lacquers that melt and change appearance at specific temperatures. The setup of Fig 17.25, however, nearly always gives sufficiently accurate information.

17.7.2 Cooling Design Example

As an example, consider the cooling design of a linear amplifier to use one 3CX800A7 tube to operate near sea level with the air temperature not above 25 °C. The tube, running 1150-W dc input, easily delivers 750-W continuous output, resulting in 400-W plate dissipation ($P_{DIS} = P_{IN} - P_{OUT}$). According to the manufacturer's data, adequate tube cooling at 400 W P_D requires at least 6 CFM of air at 0.09 inches of water back pressure. In Table 17.5, a Dayton no. 1TDN2 will do the job with a good margin of safety.

If the same single tube were to be operated at 2.3 kW dc input to deliver 1.5 kW output (substantially exceeding its maximum electrical ratings!), P_{IN} would be about 2300 W and $P_D \approx 800$ W. The minimum cooling air required would be about 19 CFM at 0.5 inches of water pressure — doubling P_{DIS}, more than tripling the CFM of air flow required and increasing back pressure requirements on the blower by a factor of 5.5!

However, two 3CX800A7 tubes are needed to deliver 1.5 kW of continuous maximum legal output power in any case. Each tube will operate under the same conditions as in the single-tube example above, dissipating 400 W. The total cooling air requirement for the two tubes is, therefore, 12 CFM at about 0.09 inches of water, only two-thirds as much air volume and one-fifth the back pressure required by a single tube. While this may seem surprising, the reason lies in the previously mentioned fact that both the airflow required by a tube and the resultant back pressure increase much more rapidly than P_D of the tube. Blower air delivery capability, conversely, decreases as back pressure is increased. Thus, a Dayton 1TDN2 blower can cool two 3CX800A7 tubes dissipating 800 W total, but a much larger (and probably noisier) no. 1TDN7 would be required to handle the same power with a single tube.

In summary, three very important considerations to remember are these:

- A tube's actual safe plate dissipation capability is totally dependent on the amount of cooling air forced through its cooling system. Any air-cooled power tube's maximum plate dissipation rating is meaningless unless the specified amount of cooling air is supplied.
- Two tubes will always safely dissipate a given power with a significantly smaller (and quieter) blower than is required to dissipate the same power with a single tube of the same type. A corollary is that a given blower can virtually always dissipate more power when cooling two tubes than when cooling a single tube of the same type.
- Blowers vary greatly in their ability to deliver air against back pressure so blower selection should not be taken lightly.

17.7.3 Other Considerations

A common method for directing the flow of air around a tube involves the use of a pressurized chassis. This system is shown in Fig 17.24. A blower attached to the chassis forces air around the tube base, often through holes in its socket. A chimney is used to guide air leaving the base area around the tube envelope or anode cooler, preventing it from dispersing and concentrating the flow for maximum cooling.

A less conventional approach that offers a significant advantage in certain situations is shown in **Fig 17.26**. Here the anode compartment is pressurized by the blower. A special chimney is installed between the anode heat exchanger and an exhaust hole in the compartment cover. When the blower pressurizes the anode compartment, there are two parallel paths for airflow: through the anode and its chimney, and through the air system socket. Dissipation, and hence cooling air required, generally is much greater for the anode than for the tube base. Because high-volume anode airflow need not be forced through restrictive air channels in the base area, back pressure may be very significantly reduced with certain tubes and sockets. Only airflow actually needed is bled through the base area. Blower back pressure requirements may sometimes be reduced by nearly half through this approach.

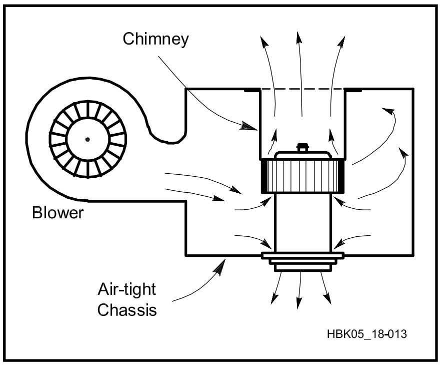


Fig 17.26 — Anode compartment pressurization may be more efficient than grid compartment pressurization. Hot air exits upwards through the tube anode and through the chimney. Cool air also goes down through the tube socket to cool tube's pins and the socket itself.

Table 17.4 also contains the part numbers for air-system sockets and chimneys available for use with the tubes that are listed. The builder should investigate which of the sockets listed for the 4CX250R, 4CX300A, 4CX1000A and 4CX1600A best fit the circuit needs. Some of the sockets have certain tube elements grounded internally through the socket. Others have elements bypassed to ground through capacitors that are integral parts of the sockets.

Depending on your design philosophy and tube sources, some compromises in the cooling system may be appropriate. For example, if glass tubes are available inexpensively as broadcast pulls, a shorter life span may be acceptable. In such a case, an increase of convenience and a reduction in cost, noise, and complexity can be had by using a pair of "muffin" fans. One fan may be used for the filament seals and one for the anode seal, dispensing with a blower and air-system socket and chimney. The airflow with this scheme is not as uniform as with the use of a chimney. The tube envelope mounted in a cross flow has flow stagnation points and low heat transfer in certain regions of the envelope. These points become hotter than the rest of the envelope. The use of multiple fans to disturb the cross airflow can significantly reduce this problem. Many amateurs have used this cooling method successfully in low-duty-cycle CW and SSB operation but it is not recommended for AM, SSTV or RTTY service.

The true test of the effectiveness of a forced air cooling system is the amount of heat carried away from the tube by the air stream. The power dissipated can be calculated from the airflow temperatures. The dissipated power is

$$P_D = 0.543\ Q_A\ (T_2 - T_1) \qquad (7)$$

where

P_D = the dissipated power, in W
Q_A = the air flow, in CFM (cubic feet per minute).
T_1 = the inlet air temperature, °C (normally quite close to room temperature).
T_2 = the amplifier exhaust temperature, °C.

The exhaust temperature can be measured with a cooking thermometer at the air outlet. The thermometer should not be placed inside the anode compartment because of the high voltage present.

17.8 Amplifier Stabilization

Purity of emissions and the useful life (or even survival) of a tube depend heavily on stability during operation. Oscillations can occur at the operating frequency, or far from it, because of undesired positive feedback in the amplifier. Unchecked, these oscillations pollute the RF spectrum and can lead to over-dissipation and subsequent failure. Each type of oscillation has its own cause and its own cure.

17.8.1 Amplifier Neutralization

An RF amplifier, especially a linear amplifier, can easily become an oscillator at various frequencies. When the amplifier is operating, the power at the output side is large. If a fraction of that power finds it way back to the input and is in the proper phase, it can be re-amplified, repeatedly, leading to oscillation. An understanding of this process can be had by studying the sections on feedback and oscillation in the **Analog Basics** chapter. Feedback that is self-reinforcing is called "positive" feedback, even though its effects are undesirable. Even when the positive feedback is insufficient to cause actual oscillation, its presence can lead to excessive distortion and strange effects on the tuning of the amplifier and it, therefore, should be eliminated or at least reduced. The deliberate use of "negative" feedback in amplifiers to increase linearity is discussed briefly elsewhere in this chapter.

The power at the output of an amplifier will couple back to the input of the amplifier through any path it can find. It is a good practice to isolate the input and output circuits of an amplifier in separate shielded compartments. Wires passing between the two compartments should be bypassed to ground if possible. This prevents feedback via paths external to the tube.

However, energy can also pass back through the tube itself. To prevent this, a process called neutralization can be used. Neutralization seeks to prevent or to cancel out any transfer of energy from the plate of the tube back to its input, which will be either the grid or the cathode. An effective way to neutralize a tube is to provide a grounded shield between the input and the output. In the grounded grid connection, the grid itself serves this purpose. For best neutralization, the grid should be connected through a low inductance conductor to a point that is at RF ground. Ceramic external tube types may have multiple low inductance leads to ground to enhance the shielding effect. Older glass type tubes may have significant inductance inside the tube and in the socket, and this will limit the effectiveness of the shielding effect of the grid. Thus, using a grounded grid circuit with those tube types does not rule out the need for further efforts at neutralization, especially at the higher HF frequencies.

When tetrodes are used in a grounded cathode configuration, the screen grid acts as an RF shield between the grid and plate. Special tube sockets are provided that provide a very low inductance connection to RF ground. These reduce the feed through from plate to grid to a very small amount, making the effective grid-to-plate capacitance a tiny fraction of one picofarad. If in doubt about amplifiers that will work over a large frequency range, use a network analyzer or impedance measuring instrument to verify how well grounded a "grounded" grid or screen really is. If at some frequencies the impedance is more than an ohm or two, a different grounding configuration may be needed.

For amplifiers to be used at only a single frequency, a series resonant circuit can be used at either the screen or grid to provide nearly perfect bypassing to ground. Typical values for a 50 MHz amplifier are shown in **Fig 17.27**.

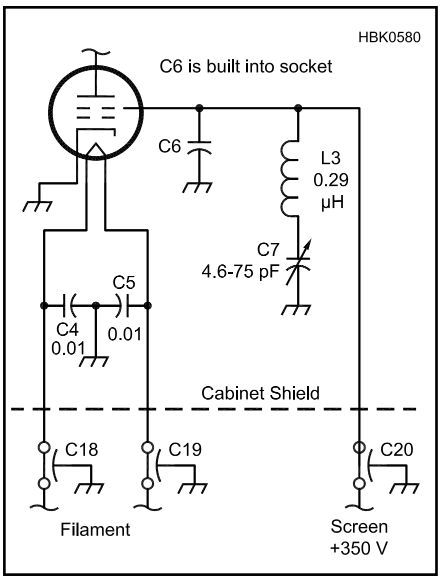


Fig 17.27 — A series-resonant circuit can be used to provide nearly perfect screen or grid bypassing to ground. This example is from a single-band 50 MHz amplifier.

For some tubes, at a certain frequency, the lead inductance to ground can just cancel the grid-to-plate capacitance. Due to this effect, some tube and socket combinations have a naturally self neutralizing frequency based on the values of screen inductance and grid-to-plate capacitance. For example, the "self neutralizing frequency" of a 4-1000 is about 30 MHz. This effect has been utilized in some VHF amplifiers.

BRIDGE NEUTRALIZATION

When the shielding effect of a grid or screen bypassed to ground proves insufficient, other circuits must be devised to cancel out the remaining effect of the grid-to-plate or grid-to-cathode capacitance. These, in effect, add an additional path for negative feedback that will combine with the undesired positive feedback and cancel it. The most commonly used circuit is the "bridge neutralization" circuit shown in **Fig 17.28**. This method gets its name from the fact that the four important capacitance values can be redrawn as a bridge circuit, as shown in **Fig 17.29**. Clearly when the bridge is properly balanced, there is no transfer of energy from the plate to the grid tanks. Note that four different capacitors are part of the bridge. C_{gp} is characteristic of the chosen tube, somewhat affected by the screen or grid bypass mentioned earlier. The other components must be chosen properly so that bridge balance is achieved. C1 is the neutralizing capacitor. Its value should be adjustable to the point where

$$\frac{C1}{C3} = \frac{C_{gp}}{C_{IN}} \tag{8}$$

where

C_{gp} = tube grid-plate capacitance
C_{IN} = tube input capacitance

The tube input capacitance must include all strays directly across the tube. C3 is not simply a bypass capacitor on the ground side of the grid tank, but rather a critical part of the bridge. Hence, it must provide a stable value of capacitance. Sometimes, simple bypass capacitors are of a type which change their value drastically with temperature. These are not suitable in this application.

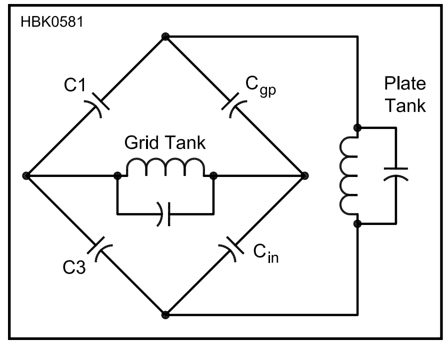


Fig 17.29 — The "bridge neutralization" circuit of Fig 17.28 redrawn to show the capacitance values.

Neutralization adjustment is accomplished by applying energy to the output of the amplifier, and measuring the power fed through the input. Conversely, the power may be fed to the input and the output power measured. *with the power off*, the neutralization capacitor C1 is adjusted for minimum feed through, while keeping the output tuning circuit and the input tuning (if used) at the point of maximum response. Since the bridge neutralization circuit is essentially broad band, it will work over a range of frequencies. Usually, it is adjusted at the highest anticipated frequency of operation, where the adjustment is most critical.

BROADBAND TRANSFORMER

Another neutralizing method is shown in **Fig 17.30**, where a broadband transformer provides the needed out of phase signal. C4 is adjusted so that the proper amount of negative feedback is applied to the input to just cancel the feedback via the cathode to plate capacitance. Though many 811A amplifiers have been built without this neutralization, its use makes tuning smoother on the higher bands. This circuit was featured in June 1961 *QST* and then appeared in the *RCA Transmitting Tube Handbook*. Amplifiers featuring this basic circuit are still being manufactured in 2009 and are a popular seller. Many thousands of hams have built such amplifiers as well.

An alternate method of achieving stable operation is to load the grid of a grounded cathode circuit with a low value of resistance. A convenient value is 50 Ω as it provides a match for the driver. This approach reduces the power being fed back to the grid from the output to a low enough level that good stability is achieved. However, the amplifier

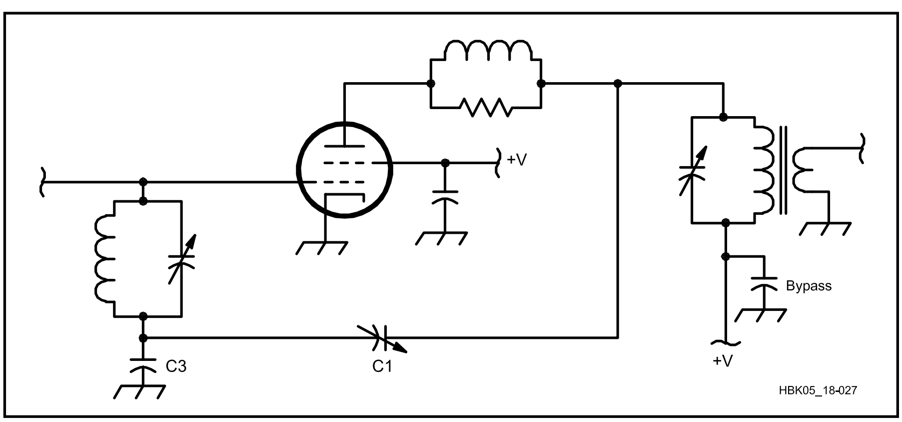


Fig 17.28 — A neutralization circuit uses C1 to cancel the effect of the tube internal capacitance.

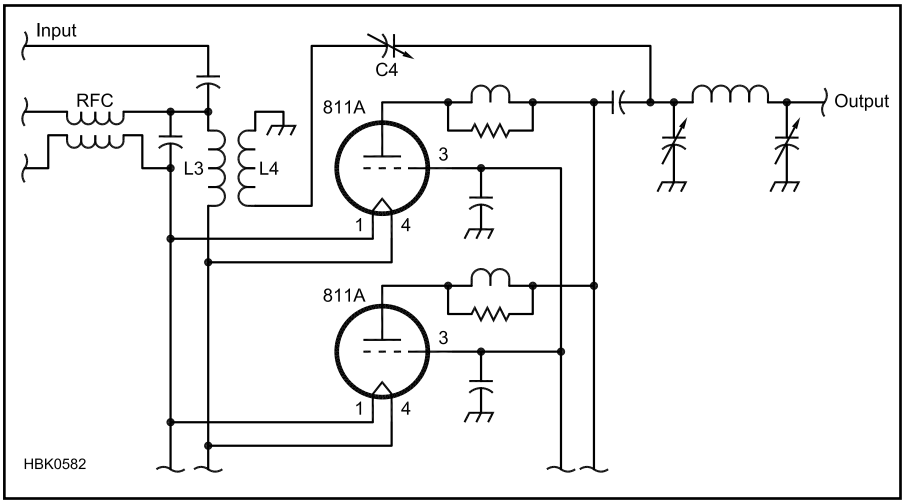


Fig 17.30 — In this neutralizing method, a broadband transformer (L3, L4) provides the needed out-of-phase signal. L3 is 6 turns of #14 wire closewound, ½ in. dia. L4 is 5 turns of insulated wire over L3. C4 is 6 pF with 0.06-in. spacing. This circuit was originally featured in June 1961 *QST* and is still found in modern amplifiers using 811A tubes.

gain will be much less than without the grid load. Also, unlike the grounded grid circuit, where much of the power applied to the input feeds through to the output, with this "loaded grid" approach, the input power is lost in the load, which must be able to dissipate such power. Distortion may be low in that the driver stage sees a very constant load. In addition, no tuning of the input is required.

17.8.2 VHF and UHF Parasitic Oscillations

RF power amplifier circuits contain parasitic reactances that have the potential to cause so-called parasitic oscillations at frequencies far above the normal operating frequency. Nearly all vacuum-tube amplifiers designed for operation in the 1.8 to 29.7-MHz frequency range exhibit tendencies to oscillate somewhere in the VHF-UHF range — generally between about 75 and 250 MHz depending on the type and size of tube. A typical parasitic resonant circuit is shown in **Fig 17.31**. Stray inductance between the tube plate and the output tuning capacitor forms a high-Q resonant circuit with the tube's C_{OUT}. C_{OUT} normally is much smaller (higher X_C) than any of the other circuit capacitances shown. The tube's C_{IN} and the tuning capacitor C_{TUNE} essentially act as bypass capacitors, while the various chokes and tank inductances shown have high reactances at VHF. Thus, the values of these components have little influence on the parasitic resonant frequency.

Oscillation is possible because the VHF resonant circuit is an inherently high-Q parallel-resonant tank that is not coupled to the external load. The load resistance at the plate is very high and thus, the voltage gain at the parasitic frequency can be quite high, leading to oscillation. The parasitic frequency, f_r, is approximately:

$$f_r = \frac{1000}{2\,\pi\sqrt{L_P C_{OUT}}} \qquad (9)$$

where

f_r = parasitic resonant frequency in MHz

L_P = total stray inductance between tube plate and ground via the plate tuning capacitor (including tube internal plate lead) in μH.

C_{OUT} = tube output capacitance in pF.

In a well-designed HF amplifier, L_P might be in the area of 0.2 μH and C_{OUT} for an 8877 is about 10 pF. Using these figures, the equation above yields a potential parasitic resonant frequency of

$$f_r = \frac{1000}{2\,\pi\sqrt{0.2 \times 10}} = 112.5 \text{ MHz}$$

For a smaller tube, such as the 3CX800A7 with C_{OUT} of 6 pF, f_r = 145 MHz. Circuit details affect f_r somewhat, but these results do, in fact, correspond closely to actual parasitic oscillations experienced with these tube types. VHF-UHF parasitic oscillations can be prevented (*not* just minimized!) by reducing the loaded Q of the parasitic resonant circuit so that gain at its resonant frequency is insufficient to support oscillation. This is possible with any common tube, and it is especially

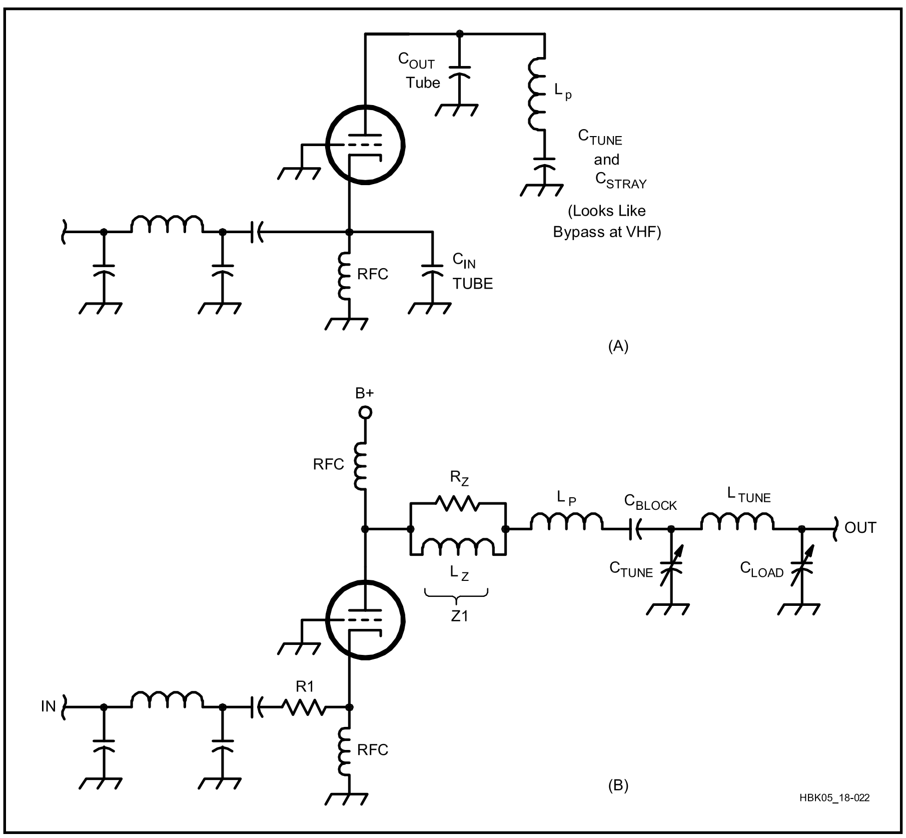


Fig 17.31 — At A, typical VHF/UHF parasitic resonance in plate circuit. The HF tuning inductor in the pi network looks like an RF choke at VHF/UHF. The tube's output capacitance and series stray inductance combine with the pi-network tuning capacitance and stray circuit capacitance to create a VHF/UHF pi network, presenting a very high impedance to the plate, increasing its gain at VHF/UHF. At B, Z1 lowers the Q and therefore gain at parasitic frequency.

easy with modern external-anode tubes like the 8877, 3CX800A7 and 4CX800A.

PARASITIC SUPPRESSORS

Z1 of Fig 17.31B is a parasitic suppressor. Its purpose is to add loss to the parasitic circuit and reduce its Q enough to prevent oscillation. This must be accomplished without significantly affecting normal operation. L_z should be just large enough to constitute a significant part of the total parasitic tank inductance (originally represented by L_P), and located right at the tube plate terminal(s). If L_z is made quite lossy, it will reduce the Q of the parasitic circuit as desired.

The inductance and construction of L_z depend substantially on the type of tube used. Popular glass tubes like the 3-500Z and 4-1000A have internal plate leads made of wire. This significantly increases L_P when compared to external-anode tubes. Consequently, L_z for these large glass tubes usually must be larger in order to constitute an adequate portion of the total value of L_P. Typically a coil of 3 to 5 turns of #10 wire, 0.25 to 0.5 inches in diameter and about 0.5 to 1 inches long is sufficient. For the 8877 and similar tubes it usually is convenient to form a "horseshoe" in the strap used to make the plate connection. A "U" about 1-inch wide and 0.75 to 1 inch deep usually is sufficient. In either case, L_z carries the full operating-frequency plate current; at the higher frequencies this often includes a substantial amount of circulating tank current, and L_z must be husky enough to handle it without overheating even at 29 MHz. Fig **17.32** shows a typical parasitic suppressor.

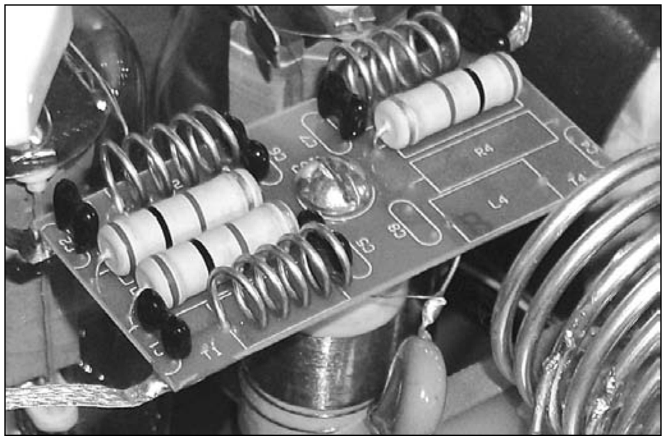


Fig 17.32 — Typical parasitic suppressor.

Regardless of the form of L_z, loss may be introduced as required by shunting L_z with one or more suitable non-inductive resistors. In high-power amplifiers, two composition or metal film resistors, each 100 Ω, 2 W, connected in parallel across L_z usually are adequate. For amplifiers up to perhaps 500 W a single 47-Ω, 2-W resistor may suffice. The resistance and power capability required to prevent VHF/UHF parasitic oscillations, while not overheating as a result of normal plate circuit current flow, depend on circuit parameters. Operating-frequency voltage drop across L_z is greatest at higher frequencies, so it is important to use the minimum necessary value of L_z in order to minimize power dissipation in R_z.

The parasitic suppressors described above very often will work without modification, but in some cases it will be necessary to experiment with both L_z and R_z to find a suitable combination. Some designers use nichrome or other resistance wire for L_z.

In exceptionally difficult cases, particularly when using glass tetrodes or pentodes, additional parasitic suppression may be attained by connecting a low value resistor (about 10 to 15 Ω) in series with the tube input, near the tube socket. This is illustrated by R1 of Fig 17.31B. If the tube has a relatively low input impedance, as is typical of grounded-grid amplifiers and some grounded-cathode tubes with large C_{IN}, R1 may dissipate a significant portion of the total drive power.

TESTING TUBE AMPLIFIERS FOR VHF-UHF PARASITIC OSCILLATIONS

Every high-power amplifier should be tested, before being placed in service, to insure that it is free of parasitic oscillations. For this test, nothing is connected to either the RF input or output terminals, and the band switch is first set to the lowest-frequency range. If the input is tuned and can be band switched separately, it should be set to the highest-frequency band. The amplifier control system should provide monitoring for both grid current and plate current, as well as a relay, circuit breaker or fast-acting fuse to quickly shut off high voltage in the event of excessive plate current. To further protect the tube grid, it is a good idea to temporarily insert in series with the grid current return line a resistor of approximately 1000 Ω to prevent grid current from soaring in the event a vigorous parasitic oscillation breaks out during initial testing.

Apply filament and bias voltages to the amplifier, leaving plate voltage off and/or cutoff bias applied until any specified tube warm-up time has elapsed. Then apply the lowest available plate voltage and switch the amplifier to transmit. Some idling plate current should flow. If it does not, it may be necessary to increase plate voltage to normal or to reduce bias so that at least 100 mA or so does flow. Grid current should be zero. Vary the plate tuning capacitor slowly from maximum capacitance to minimum, watching closely for any grid current or change in plate current, either of which would indicate a parasitic oscillation. If a tunable input network is used, its capacitor (the one closest to the tube if a pi circuit) should be varied from one extreme to the other in small increments, tuning the output plate capacitor at each step to search for signs of oscillation. If at any time either the grid or plate current increases to a large value, shut off plate voltage immediately to avoid damage! If moderate grid current or changes in plate current are observed, the frequency of oscillation can be determined by loosely coupling an RF absorption meter or a spectrum analyzer to the plate area. It will then be necessary to experiment with parasitic suppression measures until no signs of oscillation can be detected under any conditions. This process should be repeated using each band switch position.

When no sign of oscillation can be found, increase the plate voltage to its normal operating value and calculate plate dissipation (idling plate current times plate voltage). If dissipation is at least half of, but not more than its maximum safe value, repeat the previous tests. If plate dissipation is much less than half of maximum safe value, it is desirable (but not absolutely essential) to reduce bias until it is. If no sign of oscillation is detected, the temporary grid resistor should be removed and the amplifier is ready for normal operation.

LOW-FREQUENCY PARASITIC OSCILLATIONS

The possibility of self-oscillations at frequencies lower than VHF is significantly lower than in solid-state amplifiers. Tube amplifiers will usually operate stably as long as the input-to-output isolation is greater than the stage gain. Proper shielding and dc-power-lead bypassing essentially eliminate feedback paths, except for those through the tube itself.

On rare occasions, tube-type amplifiers will oscillate at frequencies in the range of about 50 to 500 kHz. This is most likely with high-gain tetrodes using shunt feed of dc voltages to both grid and plate through RF chokes. If the resonant frequency of the grid RF choke and its associated coupling capacitor occurs close to that of the plate choke and its blocking capacitor, conditions may support a tuned-plate tuned-grid oscillation. For example, using typical values of 1 mH and 1000 pF, the expected parasitic frequency would be around 160 kHz.

Make sure that there is no low-impedance, low-frequency return path to ground through inductors in the input matching networks in series with the low impedances reflected by a transceiver output transformer. Usually, oscillation can be prevented by changing choke or capacitor values to insure that the input resonant frequency is much lower than that of the output.

17.8.3 Reduction of Distortion

As mentioned previously, a common cause of distortion in amplifiers is over drive (flat topping). The use of automatic level control (ALC) is a practical way of reducing the ill effects of flat topping while still being assured of having a strong signal. This circuit detects the voltage applied to the input of the amplifier. Other circuits are based on detecting the onset of grid current flow. In either case, when the threshold is reached, the ALC circuit applies a negative voltage to the ALC input of the transceiver and forces it to cut back on the driving power, thus keeping the output power within set limits. Most transceivers also apply an ALC signal from their own output stage, so the ALC signal from the amplifier will add to or work in parallel with that. See **Fig 17.33** for a representative circuit.

Some tube types have inherently lower distortion than others. Selection of a tube specifically designed for linear amplifier service, and operating it within the recommended voltage and current ranges is a good start. In addition, the use of tuned circuits in the input circuits when running class AB2 will help by maintaining a proper load on the driver stages over the entire 360° cycle, rather then letting the load change as the tube begins to draw grid current. Another way to accomplish this is with the "loaded grid," the use of a rather low value of resistance from the grid to cathode. Thus, when grid current flows, the change in impedance is less drastic, having been swamped by the resistive load.

For applications where the highest linearity is desired, operating class A will greatly reduce distortion but at a high cost in efficiency. Some solid-state amateur transceivers have provision for such operation. The use of negative feedback is another way of greatly reducing distortion. High efficiency is maintained, but there is a loss of overall gain. Often, two stages of gain are used and the feedback applied around both stages. In this way, gain can be as high as desired, and both stages are compensated for any inherent nonlinearities. Amplifiers using RF negative feedback can achieve values of intermodulation distortion (IMD) as much as 20 dB lower than amplifiers without feedback.

It must be remembered that distortion tends to be a cumulative problem, with each nonlinear part of the transmission chain adding its part. It is not worthwhile to have a super clean transceiver if it is followed by an amplifier with poor linearity. In the same way, a very good linear will look bad if the transceiver driving it is poor. It is even possible to have a clean signal out of your amplifier, but have it spoiled by a ferrite core tuner inductor or balun that is saturated.

Distortion in a linear amplifier is usually measured with a spectrum analyzer while transmitting a two tone test. If the spectrum analyzer input is overloaded, this can also produce apparent distortion in the amplifier. Reducing the level so that the analyzer is not clipping the input signal is necessary to see the true distortion in the amplifier chain. Use of test equipment for various types of measurements is covered in the **Test Equipment and Measurements** chapter.

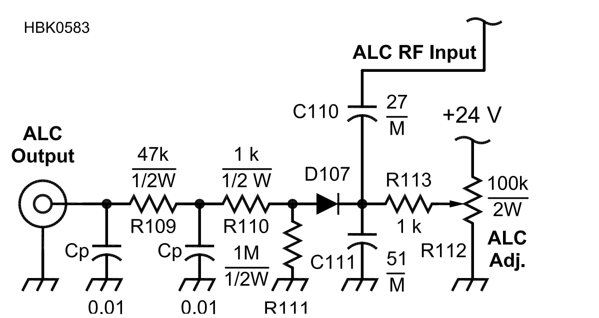


Fig 17.33 — An amplifier ALC circuit can be used to automatically limit the drive power from the transceiver to a safe level.

17.9 Design Example: A High Power Vacuum Tube HF Amplifier

Most popular HF transceivers produce approximately 100 W output. The EIMAC 8877 can deliver 1500 W output for approximately 60 W of drive when used in a grounded grid circuit. Grounded-grid operation is usually the easiest tube amplifier circuit to implement. Its input impedance is relatively low, often close to 50 Ω. Input/output shielding provided by the grid and negative feedback inherent in the grounded-grid circuit configuration reduce the likelihood of amplifier instability and provide excellent linearity without critical adjustments. Fewer supply voltages are needed in this configuration compared to others, often just high-voltage dc for the plate and low-voltage ac for the filament.

17.9.1 Tube Capabilities

The first step in the amplifier design process is to verify that the tube is actually capable of producing the desired results while remaining within manufacturer's ratings. The plate dissipation expected during normal operation of the amplifier is computed first. Since the amplifier will be used for SSB, a class of operation producing linear amplification must be used. Class AB2 provides a very good compromise between linearity and good efficiency, with effective efficiency typically exceeding 60%. Given that efficiency, an input power of 2500 W is needed to produce the desired 1500 W output. Operated under these conditions, the tube will dissipate about 1000 W — well within the manufacturer's specifications, provided adequate cooling airflow is supplied.

The grid in modern high-mu triodes is a relatively delicate structure, closely spaced to the cathode and carefully aligned to achieve high gain and excellent linearity. To avoid shortening tube life or even destruction of the tube, the specified maximum grid dissipation must not be exceeded for more than a few milliseconds under any conditions. For a given power output, the use of higher plate voltages tends to result in lower grid dissipation. It is important to use a plate voltage that is high enough to result in safe grid current levels at maximum output. In addition to maximum ratings, manufacturers' data sheets often provide one or more sets of "typical operation" parameters. This makes it even easier for the builder to achieve optimum results.

According to typical operation data, the 8877, operating at 3500 V, can produce 2075 W of RF output with excellent linearity and 64 W of drive. Operating at 2700 V it can deliver 1085 W with 40 W of drive. To some extent, the ease and cost of constructing a high-power amplifier, as well as its ultimate reliability, are enhanced by using the lowest plate voltage that will yield completely satisfactory performance. Working with various load lines on the characteristic curves shows that the 8877 can comfortably deliver 1.5 kW output with a 3100 V plate supply and 50 to 55 W of drive. Achieving 1640 W output power at this plate voltage requires 800 mA of plate current — well within the 8877's maximum rating of 1.0 A.

17.9.2 Input and Output Circuits

The next step in the design process is to calculate the optimum plate load resistance at this plate voltage and current for Class AB2 operation and design an appropriate output-matching network. Using the operating line shown in **Fig 17.34**, the *TubeCalculator* program gives these values (**Fig 17.35**). The simple formula for load resistance shown earlier in this chapter (equation 1) gives a similar value, about 2200 Ω.

Several different output networks might be used to transform the nominal 50-Ω resistance of the actual load to the 2200-Ω load resistance required by the 8877, but experience shows that pi and pi-L networks are most practical. Each can provide reasonable harmonic attenuation, is relatively easy to build mechanically and uses readily available components. The pi-L gives significantly greater harmonic attenuation than the pi and usually is the better choice — at least in areas where there is any potential for TVI or cross-band interference. In a multiband amplifier, the extra cost of using a pi-L network is the "L" inductor and its associated band switch section.

Parameter	Result
Grid Current (mA)	34.2
Screen Current (mA)	0.00
Plate Current (A)	0.800
Input Power (Watts)	2470
Output Power (Watts)	1640
Plate Dissipation (W)	831
Plate Load (Ohms)	2210
Efficiency (%)	66.3
Grid Swing (Volts)	80.0
Resting Dissipation (W)	620
Input Resistance (Ohms)	62.5
Power Passed (Watts)	48.8
Total drive power (W)	51.6

Fig 17.35 — Typical operating parameters for the 8877 triode used in the design example detailed in the text.

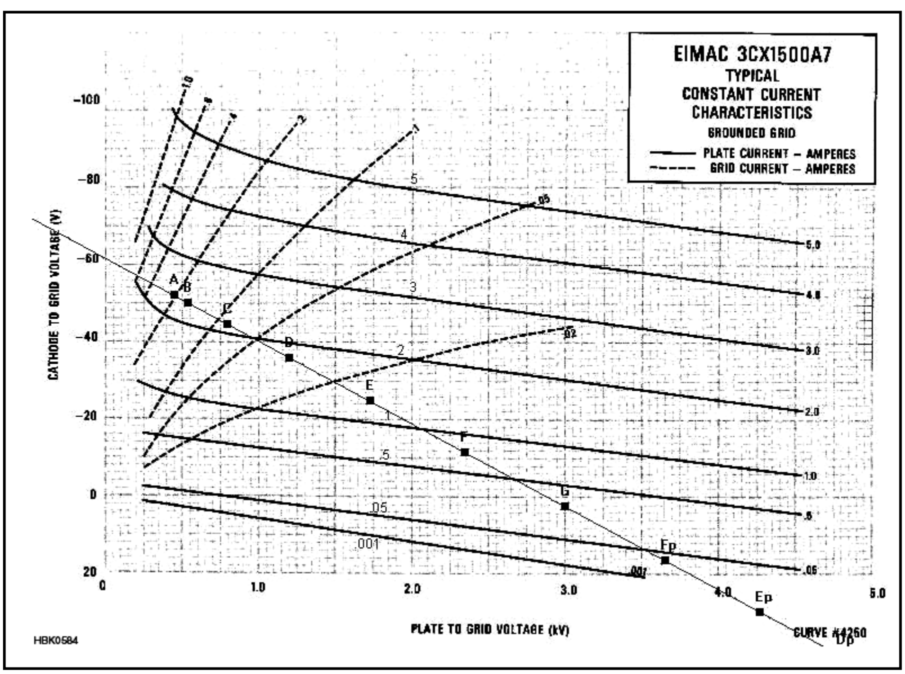


Fig 17.34 — Characteristic curves for the 8877 triode used in the design example detailed in the text.

The input impedance of a grounded-grid 8877 is typically on the order of 50 to 55 Ω, shunted by input capacitance of about 38 pF. While this average impedance is close enough to 50 Ω to provide negligible input SWR, the instantaneous value varies greatly over the drive cycle — that is, it is nonlinear. This nonlinear impedance is reflected back as a nonlinear load impedance at the exciter output, resulting in increased intermodulation distortion, reduced output power, and often meaningless exciter SWR meter indications. In addition, the tube's parallel input capacitance, as well as parasitic circuit reactances, often are significant enough at 28 MHz to create significant SWR.

A tank circuit at the amplifier input can solve both of these problems by tuning out the stray reactances and stabilizing (linearizing) the tube input impedance through its flywheel effect. The input tank should have a loaded Q of at least two for good results. A Q of five results in a further small improvement in linearity and distortion, but at the cost of a narrower operating bandwidth. Using the *PI-EL Design* software, one can quickly determine values for C1, L1 and C2 for the various bands as well as the bandwidth that various values of Q will provide. Since the 3.5 to 4 MHz band is the widest in terms of percentage bandwidth, using a lower Q for that band seems wise. If we wish to cover two bands with the same switch position, for example the 24 and 28 MHz bands, that will also call for a lower Q. For the 40, 30 and 20 meter bands alone, a rather high Q would work fine. Remember to subtract the input capacitance of the tube from the calculated value of C2.

Fig 17.36 illustrates these input and output networks applied in the amplifier circuit. The schematic shows the major components in the amplifier RF section, but with band-switching and cathode dc return circuits omitted for clarity. C1 and C2 and L1 form the input pi network. C3 is a blocking capacitor to isolate the exciter from the cathode dc potential. Note that when the tube's average input resistance is close to 50 Ω, as in the case of the 8877, a simple parallel-resonant tank often can successfully perform the tuning and flywheel functions, since no impedance transformation is necessary. In this case, it is important to minimize stray lead inductance

Tuning Your Vacuum-Tube Amplifier

Today most hams are used to a wide variety of amplifiers using triodes that are tuned for maximum power output. Unfortunately, not all amplifiers or tubes are rugged enough for this approach. For example, amplifiers that use tetrodes employ a different tuning procedure in which tuning for maximum power may result in a destroyed control grid or screen grid! Procedures vary depending not only on the ratings of the tetrode, but the voltages as well. When tuning any amplifier, monitor grid currents closely and do not exceed the specified maximum current as those are the easiest elements of the tube to damage.

For commercial amplifiers, "Read the Manual" as the manufacturer's directions may not follow the following procedures exactly and can be quite different in some cases. In all cases, the last tuning adjustments should be made at full power output, not at reduced power, because the characteristics of the tube change with different power levels. Operating the amplifier at high power after tuning at low power can result in spurious emissions or over-stressing the output network components.

Begin by making sure you have all band-switching controls set properly. If the amplifier TUNE and LOAD controls (sometimes referred to as "Plate Tune" and "Output", respectively) have recommended settings on a particular band, start at those settings. If your amplifier can be set to a TUNE mode, do so. Set the initial amount of drive (input power) from the exciter — read the amplifier manual or check the tube's specifications if a manual is not available. The exciter output should be one-half or more of full power so that the exciter's ALC systems function properly.

Tuning a triode-based, grounded-grid amplifier is the simplest: Tune for maximum output power without exceeding the tube ratings, particularly grid current, or the legal output power limit. The typical procedure is, while monitoring grid current, to increase drive until the plate current equals about one-quarter to one-half of the target current (depending on the tube and grid bias) while monitoring the output on a wattmeter or the internal power meter. Adjust the TUNE control then advance the LOAD control for maximum output. Repeat the sequence of peaking TUNE then increasing LOAD until no more output power can be obtained without exceeding the ratings for the tube or the legal power limit. If necessary, increase drive and re-peak both the TUNE and LOAD controls.

Operating somewhat differently, a grid-driven tetrode (or pentode) amplifier operating near its designed output power uses the TUNE control for peaking output power and the LOAD control for increasing, but not exceeding, the maximum allowable screen current. Generally, the first part of tetrode amplifier tuning is the same as for a triode amplifier with both the TUNE and LOAD controls adjusted for maximum power output while monitoring screen and control grid current. After the initial tuning step, the LOAD control is used to peak the screen current. Maximum power should coincide with maximum screen current. The screen current is just a better indicator. As with the triode amplifier, if drive needs to be increased, readjust the TUNE and LOAD controls.

For both triode and tetrode amplifiers, once the procedures above have been completed, try moving the TUNE control a small amount higher or lower and re-peaking the LOAD control. Also known as "rocking" the TUNE control, this small variation can find settings with a few percent more output power or better efficiency.

Once tuning has been completed, it is a good idea to mark the settings of the TUNE and LOAD controls for each band. This reduces the amount of time for on-the-air or dummy-load tuning, reducing stress to the tube and interference to other stations. Usually, a quick "fine-tune" adjustment is all that is required for maximum output. The set of markings also serves as a diagnostic tool, should the settings for maximum power suddenly shift. This indicates a change in the antenna system, such as a failing connector or antenna. — *Roger Halstead, K8RI*

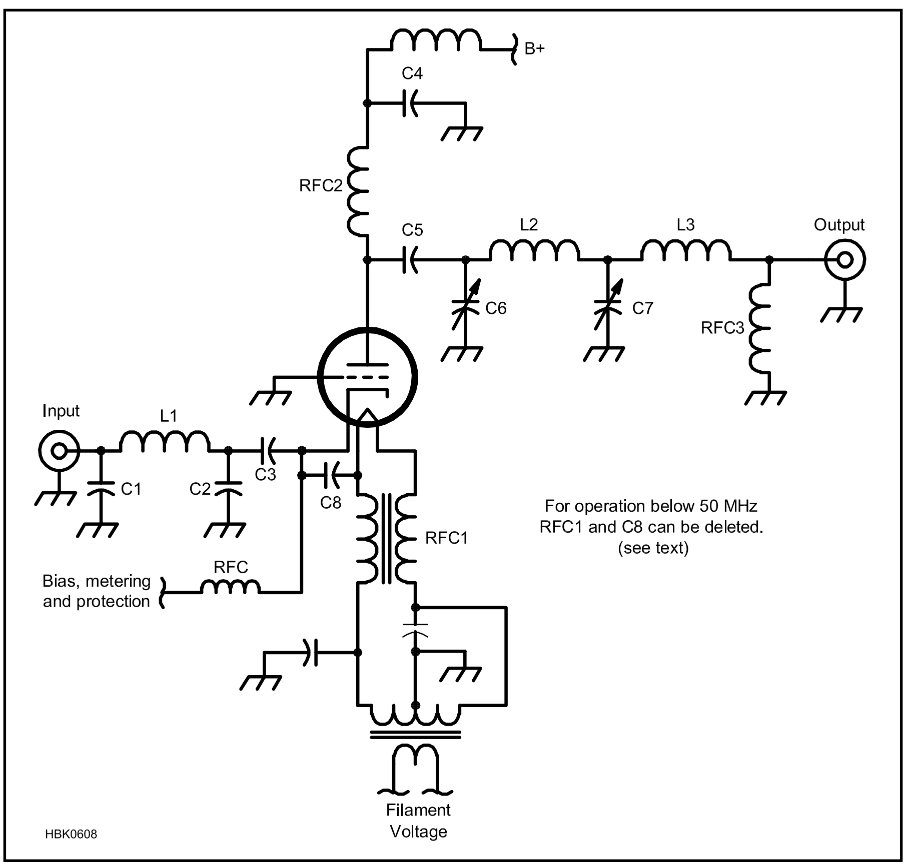


Fig 17.36 — A simplified schematic of a grounded-grid amplifier using a pi network input and pi-L network output.

between the tank and tube to avoid undesired impedance transformation.

17.9.3 Filament Supply

The filament or "heater" in indirectly heated tubes such as the 8877 must be very close to the cathode to heat the cathode efficiently. A capacitance of several picofarads exists between the two. Particularly at very high frequencies, where these few picofarads represent a relatively low reactance, RF drive intended for the cathode can be capacitively coupled to the lossy filament and dissipated as heat. To avoid this, above about 50 MHz, the filament must be kept at a high RF impedance above ground. The high impedance (represented by choke RFC1 in Fig 17.36) minimizes RF current flow in the filament circuit so that RF dissipated in the filament becomes negligible. The choke's low-frequency resistance should be kept to a minimum to lessen voltage drops in the high-current filament circuit.

The choke most commonly used in this application is a pair of heavy-gauge insulated wires, bifilar-wound over a ferrite rod. The ferrite core raises the inductive reactance throughout the HF region so that a minimum of wire is needed, keeping filament-circuit voltage drops low. The bifilar winding technique assures that both filament terminals are at the same RF potential.

Below 30 MHz, the use of such a choke seldom is necessary or beneficial, but actually can introduce another potential problem. Common values of cathode-to-heater capacitance and heater-choke inductance often are series resonant in the 1.8 to 29.7 MHz HF range. A capacitance of 5 pF and an inductance of 50 µH, for example, resonate at 10.0 MHz; the actual components are just as likely to resonate near 7 or 14 MHz. At resonance, the circuit constitutes a relatively low impedance shunt from cathode to ground, which affects input impedance and sucks out drive signal. An unintended resonance like this near any operating frequency usually increases input SWR and decreases gain on that one particular band. While aggravating, the problem rarely completely disables or damages the amplifier, and so is seldom pursued or identified.

Fortunately, the entire problem is easily avoided — below 30 MHz the heater choke can be deleted. At VHF-UHF, or wherever a heater isolation choke is used for any reason, the resonance can be moved below the lowest operating frequency by connecting a sufficiently large capacitance (about 1000 pF) between the tube cathode and one side of the heater. It is good practice also to connect a similar capacitor between the heater terminals. It also would be good practice in designing other VHF/UHF amplifiers, such as those using 3CX800A7 tubes, unless the builder can ensure that the actual series resonance is well outside of the operating frequency range.

17.9.4 Plate Choke and DC Blocking

Plate voltage is supplied to the tube through RFC2. C5 is the plate blocking capacitor. The output pi-L network consists of tuning capacitor C6, loading capacitor C7, pi coil L2, and output L coil L3. RFC3 is a high-inductance RF choke placed at the output for safety purposes. Its value, usually 100 µH to 2 mH, is high enough so that it appears as an open circuit across the output connector for RF. However, should the plate blocking capacitor fail and allow high voltage onto the output matching network, RFC3 would short the dc to ground and blow the power-supply fuse or breaker. This prevents dangerous high voltage from appearing on the feed line or antenna. It also prevents electrostatic charge — from the antenna or from blocking capacitor leakage — from building up on the tank capacitors and causing periodic dc discharge arcs to ground. If such a dc discharge occurs while the amplifier is transmitting, it can trigger a potentially damaging RF arc.

17.9.5 Tank Circuit Design

The output pi-L network must transform the nominal 50-Ω amplifier load to a pure resistance of 2200 Ω. We previously calculated that the 8877 tube's plate must see 2200 Ω for optimum performance. In practice, real antenna loads are seldom purely resistive or exactly 50 Ω; they often exhibit SWRs of 2:1 or greater on some frequencies. It's desirable that the amplifier output network be able to transform any complex load impedance corresponding to an SWR up to about 2:1 into a resistance of 2200 Ω. The network also must compensate for tube C_{OUT} and other stray plate-circuit reactances, such as those of interconnecting leads and the plate RF choke. These reactances, shown in **Fig 17.37**, must be taken into account when designing the matching networks. Because the values of most stray reactances are not accurately known, the most satisfactory approach is to estimate them, and then allow sufficient flexibility in the matching network to accommodate modest errors.

Fig 17.37 shows the principal reactances in the amplifier circuit. C_{OUT} is the actual tube output capacitance of 10 pF plus the stray capacitance between its anode and the enclosure metalwork. This stray C varies with layout; we will approximate it as 5 pF, so C_{OUT} is roughly 15 pF. L_{OUT} is the stray inductance of leads from the tube plate to the tuning capacitor (internal to the tube as well as external circuit wiring.) External-anode tubes like the 8877 have essentially no internal plate leads, so L_{OUT} is almost entirely external. It seldom exceeds about 0.3 µH and is not very significant below 30 MHz. L_{CHOKE} is the reactance presented by the plate choke, which usually is significant only below 7 MHz. C_{STRAY} represents the combined stray capacitances to ground of the tuning capacitor stator and of interconnecting RF plate circuit leads. In a well-constructed, carefully thought out power amplifier, C_{STRAY} can be estimated to be approximately 10 pF. Remaining components C_{TUNE}, C_{LOAD}, and the two tuning inductors, form the pi-L network proper.

The values for the output network components can be calculated using the *PI-EL Design* software, taken from the graphical charts in Figs 17.17 to 17.19, or from tables included on the *Handbook* CD. For pi networks, a Q of 12 is a good compromise between harmonic suppression and circuit losses. In practice, it often is most realistic and practical with both pi and pi-L output networks to accept somewhat higher Q values on the highest HF frequencies — perhaps as large as 18 or even 20 at 28 MHz. When using a pi-L on the 1.8 and 3.5 MHz bands, it often is desirable to choose a moderately lower Q , perhaps 8 to 10, to permit using a more reasonably-sized plate tuning capacitor.

CIRCUIT REACTANCES

The calculated output network values must be adjusted to allow for circuit reactances outside the pi-L proper. First, low-frequency component values should be examined. At 3.5 MHz, assuming that total tuning capacitance C1 is 140 pF, we know that three other stray reactances are directly in parallel with C_{TUNE} (assuming that L_{OUT} is negligible at the operating frequency as it should be). The tube's internal and external plate capacitance to ground, C_{OUT}, is about 15 pF. Strays in the RF circuit, C_{STRAY}, are roughly 10 pF.

The impedance of the plate choke, X_{CHOKE}, is also in parallel with C_{TUNE}. Plate chokes with self-resonance characteristics suitable for use in amateur HF amplifiers typically have inductances of about 90 µH. At 3.5 MHz this is an inductive reactance of +1979 Ω. This appears in parallel with the tuning capacitance, effectively canceling an equal value of capacitive reactance. At 3.5 MHz, an X_C of 1979 Ω corresponds to 23 pF of capacitance — the amount by which tuning capacitor C_{TUNE} must be increased at 3.5 MHz to compensate for the effect of the plate choke.

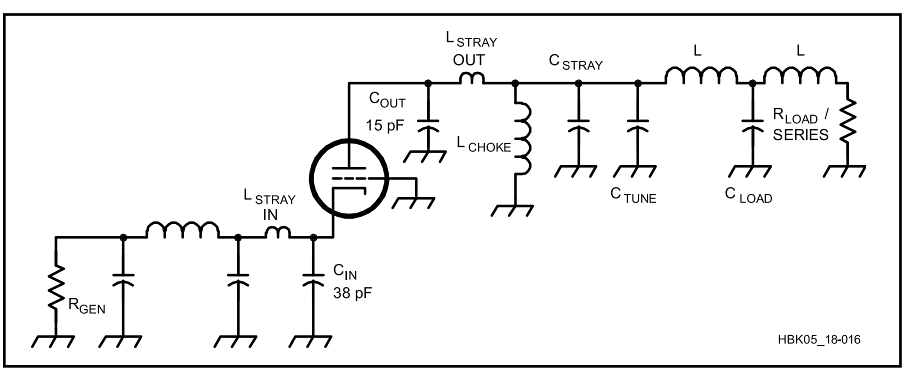


Fig 17.37 — The effective reactances for the amplifier in Fig 17.36.

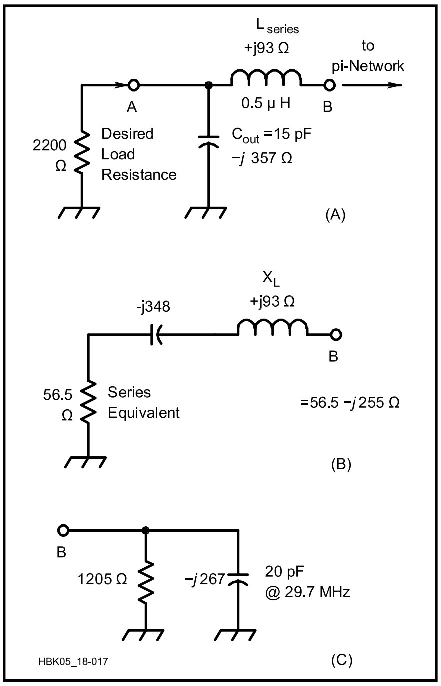


Fig 17.38 — The effect of adding a small inductor in series with the tube plate to aid matching at high frequencies. See text for details.

If the pi-L network requires an effective capacitance of 140 pF at its input at 3.5 MHz, subtracting the 25 pF provided by C_{OUT} and C_{STRAY} and adding the 23 pF canceled by X_{CHOKE}, the actual value of C_{TUNE} must be 140 – 25 + 23 = 138 pF. It is good practice to provide at least 10% extra capacitance range to allow matching loads having SWRs up to 2:1. So, if 3.5 MHz is the lower frequency limit of the amplifier, a variable tuning capacitor with a maximum value of at least 150 to 160 pF should be used.

PERFORMANCE AT HIGH FREQUENCIES

Component values for the high end of the amplifier frequency range also must be examined, for this is where the most losses will occur. At 29.7 MHz we can assume a minimum pi-L input capacitance of 35 pF. Since C_{OUT} and C_{STRAY} contribute 25 pF, C_{TUNE} must have a minimum value no greater than 10 pF. A problem exists, because this value is not readily achievable with a 150 to 160-pF air variable capacitor suitable for operation with a 3100 V plate supply. Such a capacitor typically has a minimum capacitance of 25 to 30 pF. Usually, little or nothing can be done to reduce the tube's C_{OUT} or the circuit C_{STRAY}, and in fact the estimates of these may even be a little low. If 1.8 MHz capability is desired, the maximum tuning capacitance will be at least 200 to 250 pF, making the minimum- capacitance problem at 29.7 MHz even more severe.

There are three potential solutions to this dilemma. We could accept the actual minimum value of pi-L input capacitance, around 50 to 55 pF, realizing that this will raise the pi-L network's loaded Q to about 32. This results in very large values of circulating tank current. To avoid damage to tank components — particularly the band switch and pi inductor — from heat due to I^2R losses, it will be necessary to either use oversize components or reduce power on the highest-frequency bands. Neither option is appealing.

A second potential solution is to reduce the minimum capacitance provided by C_{TUNE}. We could use a vacuum variable capacitor with a 300-pF maximum and a 5-pF minimum capacitance. These are rated at 5 to 15 kV, and are readily available. This reduces the minimum effective circuit capacitance to 30 pF, allowing use of the pi-L values for a Q of 12 on all bands from 1.8 through 29.7 MHz. While brand new vacuum variables are quite expensive, suitable models are widely available in the surplus and used markets for prices not much higher than the cost of a new air variable. A most important caveat in purchasing a vacuum capacitor is to ensure that its vacuum seal is intact and that it is not damaged in any way. The best way to accomplish this is to "hi-pot" test the capacitor throughout its range, using a dc or peak ac test voltage of 1.5 to 2 times the amplifier plate supply voltage. For all-band amplifiers using plate voltages in excess of about 2500 V, the initial expense and effort of securing and using a vacuum-variable input tuning capacitor often is well repaid in efficient and reliable operation of the amplifier.

A third possibility is the use of an additional inductance connected in series between the tube and the tuning capacitor. In conjunction with C_{OUT} of the tube, the added inductor acts as an L network to transform the impedance at the input of the pi-L network up to the 2200-Ω load resistance needed by the tube. This is shown in **Fig 17.38A**. Since the impedance at the input of the main pi-L matching network is reduced, the loaded Q for the total capacitance actually in the circuit is lower. With lower Q, the circulating RF currents are lower, and thus tank losses are lower.

C_{OUT} in Fig 17.38 is the output capacitance of the tube, including stray C from the anode to metal enclosure. X_L is the additional series inductance to be added. As determined previously, the impedance seen by the tube anode must be a 2200 Ω resistance for best linearity and efficiency, and we have estimated C+ of the tube as 15 pF. If the network consisting of C_{OUT} and X_L is terminated at A by 2200 Ω, we can calculate the equivalent impedance at point B, the input to the pi-L network, for various values of series X_L. The pi-L network must then transform the nominal 50-Ω load at the transmitter output to this equivalent impedance.

IMPEDANCE TRANSFORMATIONS

We work backwards from the plate of the tube towards the C_{TUNE} capacitor. First, calculate the series-equivalent impedance of the parallel combination of the desired 2200-Ω plate load and the tube X_{OUT} (15 pF at 29.7 MHz = – *j* 357 Ω). The series-equivalent impedance of this parallel combination is 56.5 – *j* 348 Ω, as shown in Fig 17.38B. Now suppose we use a 0.5 μH inductor, having an impedance of + *j* 93 Ω at 29.7 MHz, as the series inductance X_L. The resulting series-equivalent impedance is 56.5 – *j*348 + *j*93, or 56.5 – *j*255 Ω. Converting back to the parallel equivalent gives the network of Fig 17.38C: 1205 Ω resistance in parallel with –*j*267 Ω, or 20 pF at 29.7 MHz. The pi-L tuning network must now transform the 50-Ω load to a resistive load of 1205 Ω at B, and absorb the shunt capacitance of 20 pF.

Using the *PI-EL Design* software or pi network formulas on the *Handbook* CD, R1 = 1205 Ω and Q = 15 at 29.7 MHz, yields a required total capacitance of about 67 pF at 29.7 MHz. Note that for the same loaded Q for a 2200-Ω load line without the series inductor, the capacitance was about 36 pF. When the 20 pF of transformed input capacitance is subtracted from the 67 pF total needed, the amount of capacitance is 47 pF. If the minimum capacitance in C_{TUNE} is 25 pF and the stray capacitance is 10 pF, then there is a margin of 47– 35 = 12 pF beyond the minimum capacitance for handling SWRs greater than 1:1 at the load.

The series inductor should be a high-Q coil wound from copper tubing to keep losses low. This inductor has a decreasing, yet significant effect, on progressively lower frequencies. A similar calculation to the above should be made on each band to determine the transformed equivalent plate impedance, before calculating the network values. The impedance-transformation effect of the additional inductor decreases rapidly with decreasing frequency. Below 21 MHz, it usually may be ignored and pi-L network values calculated for R1 = 2200 Ω.

The nominal 90-μH plate choke remains in parallel with C_{TUNE}. It is rarely possible to calculate the impedance of a real HF plate choke at frequencies higher than about 5 MHz because of self-resonances. However, as mentioned previously, the choke's reactance should be sufficiently high that the calculations are not seriously affected if the choke's first series-resonance is at 23.2 MHz.

This amplifier is made operational on multiple bands by changing the values of inductance at L2 and L3 for different bands. The usual practice is to use inductors for the lowest operating frequency, and short out part of each inductor with a switch, as necessary, to provide the inductance needed for each individual band. Wiring to the switch and the switch itself add stray inductance and capacitance to the circuit. To minimize these effects at the higher frequencies, the unswitched 10-m L2 should be placed closest to the high-impedance end of the network at C6. Stray capacitance associated with the switch then is effectively in parallel with C7, where the impedance level is around 300 Ω. The effects of stray capacitance are relatively insignificant at this low impedance level. This configuration also minimizes the peak RF voltage that the switch insulation must withstand.

17.9.6 Checking Operation

After the input and output networks are designed, cold tuning as described earlier will confirm that all of the tuned circuits are working properly. These tests are well worth doing before any power is applied to the amplifier. The band switch itself will have significant inductance especially on the higher frequencies. To determine the proper taps for the various bands on the tank inductor, start with the highest frequency and verify that the calculated number of turns gives the frequency range desired, moving the tap as needed to allow for the inductance of the band switch. As each tap is located, it should be securely wired with strap or braid and the process repeated for successively lower bands. Once the cold tuning looks good, proceed to the tests for parasitics.

You should now be ready to apply full power to the amplifier and see how it performs.

17.10 Solid-State Amplifiers

17.10.1 Solid State vs Vacuum Tubes

Solid-state amplifiers have become the norm in transceivers over the past 20 years, but their use in external high power amplifiers has not. The primary reason is economic. It is more expensive to generate a kilowatt or more with transistors because they are smaller and have less dissipation capability. This is changing. The increasing demand for auto-tuning amplifiers, an expensive undertaking with tube amplifiers, is raising the demand for solid-state amplifiers and lowering their price in the process. A number of solid-state Amateur Radio amplifiers in the 500-1000 W power class are available. **Fig 17.39** shows one of many available examples.

Vacuum tubes will not be around forever. Commercial broadcasting and industrial RF plasma, the principal commercial users of vacuum tubes, have gone solid state for all new designs. The higher acquisition cost of solid-state amplifiers is offset by their lower operating cost, which leads to a lower lifetime cost. The remaining commercial demand is the replacement market for failed tubes in existing equipment. Lower demand for vacuum tubes reduces factory volumes, driving up costs. Many tube manufacturers have gone out of business or have been absorbed by their competitors. Most of the tube types available today are not manufactured by the same company that designed them. Tube design is a withering art because there is so little demand for it.

On the bright side, new and more powerful transistors are being designed all the time. There is actually a commercially available 1500-W solid-state amplifier that uses only two transistors — Tokyo Hy-Power's HL-2.5KFX.

17.10.2 Classes of Operation

This topic applies to transistors as well as tubes, and it was covered earlier in this chapter and also in the **RF Techniques** chapter. In communications amplifiers, Class A is used mainly for driver stages where linearity is desired and efficiency is not a concern.

Class B is usually passed over in favor of the more linear Class AB. Class AB offers increased linearity, mainly less crossover distortion, for a very small (perhaps 1 or 2%) reduction in efficiency. It is the most commonly used class of operation for linear power amplifiers that must cover a wide range of frequencies. Broadband solid-state Class AB amplifiers typically achieve 50 to 60% efficiency.

Class C is used where efficiency is important and linearity and bandwidth or harmonics are not. FM transmitters are the most com-

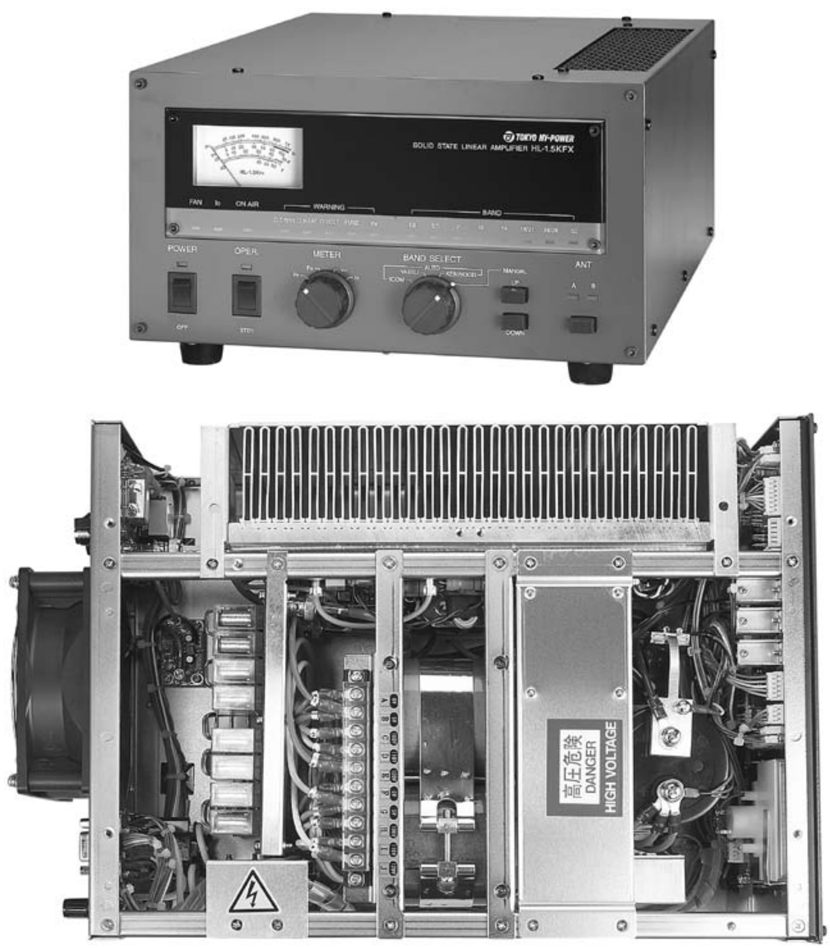


Fig 17.39 — Solid-state amateur power amplifiers such as this 1000-W class Tokyo Hy-Power HL-1.5KFX are much more common today than they were just a decade ago.

mon application in communications. Single-band tuned amplifiers can be as much as 80% efficient. However, in a single-ended amplifier they require a tank circuit. Class C amplifiers are *not* suitable for on-off keyed modes like CW without extensive pre-distortion of the driving signal to prevent key clicks.

Class D and E are most efficient, up to 95% in practical circuit applications. But these both require a tuned tank circuit to achieve this efficiency. They are not linear; the output is essentially either on or off. They can be quite effectively used for linear amplification by the process of EER — envelope elimination and restoration — but it is always in a single-band circuit. On-off keying can be employed if the power supply is keyed with a properly shaped envelope. EER is difficult to do well and requires very complex circuitry. Without careful system design, EER results in poor SSB performance.

17.10.3 Modeling Transistors

The design method using performance curves that was detailed earlier in this chapter is more applicable to vacuum tube amplifier design than solid state. The most common method used with solid state is electronic design analysis (EDA, also called computer-aided design, or CAD) using electronic models. *SPICE* or S-parameter models are available for some high power transistors, and simple amplifiers can be readily designed with the aid of an appropriate analysis program. (See the **Computer-Aided Circuit Design** chapter for more information on *SPICE* and related modeling techniques.).

Full-featured circuit design and analysis programs are expensive, and the resulting designs are only as good as the accuracy of the models used. A complicating factor is that any design relies heavily on models for all the passive components in the circuit. While passive part models can be obtained for some commercial components, many others — such as ferrite-loaded transformers — must be designed also. It is not unusual for the electronic design to take much longer and cost more than the benefits are worth.

As detailed in the **Computer-Aided Circuit Design** chapter, many of the EDA vendors offer free or inexpensive "student versions" of their products. These are fully capable up to a certain level of circuit complexity. Although they usually are not big enough to analyze a whole amplifier, student versions are still useful for looking at parts of it.

Electronic design is very useful for getting the circuit design "in the ballpark." The design will be close enough that it will work when built, and any necessary fine tuning can be done easily once it is constructed. Computer modeling is very useful for evaluating the stresses on the circuit's passive components so they can be properly sized. Another very helpful use of CAD is in the development of the output filters. W4ENE's *SVC Filter Designer* program included with the *Handbook* CD is exceptional in this regard.

17.10.4 Impedance Transformation — "Matching Networks"

Aside from the supply voltage, there is little difference between the operation of a tube amplifier and a transistor amplifier. Each amplifies the input signal, and each will only work into a specific load impedance. In a tube amplifier, the proper plate load impedance is provided by an adjustable pi or pi-L plate tuning network, which also transforms the impedance down to 50 Ω.

A single-transistor amplifier can be made in the same way, and in fact most single-band VHF amplifier "bricks" are. A tuned matching network provides the proper load impedance for the transistor and transforms it to 50 Ω. The major difference is that the proper load impedance for a transistor, at any reasonable amount of power, is much *lower* than 50 Ω. For vacuum tubes it is much *higher* than 50 Ω.

BROADBAND TRANSFORMERS

Broadband transformers are often used in matching to the input impedance or optimum load impedance in a power amplifier. Multi-octave power amplifier performance can be achieved by appropriate application of these transformers. The input and output transformers are two of the most critical components in a broadband amplifier. Amplifier efficiency, gain flatness, input SWR, and even linearity all are affected by transformer design and application.

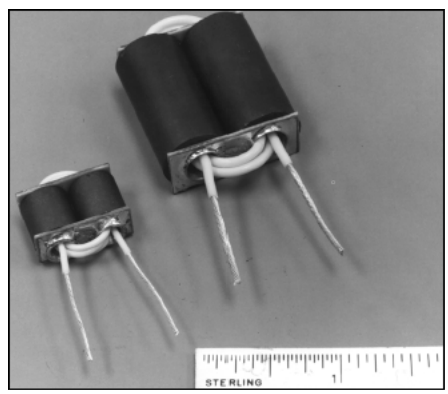
There are two basic RF transformer types, as described in the **RF Techniques** and **Transmission Lines** chapters: the conventional transformer and the transmission-line transformer. More information on RF transformers is included on the *Handbook* CD as well.

The conventional transformer is wound much the same way as a power transformer. Primary and secondary windings are wound around a high-permeability core, usually made from a ferrite or powdered-iron material. Coupling between the secondary and primary is made as tight as possible to minimize leakage inductance. At low frequencies, the coupling between windings is predominantly magnetic. As the frequency rises, core permeability decreases and leakage inductance increases; transformer losses increase as well.

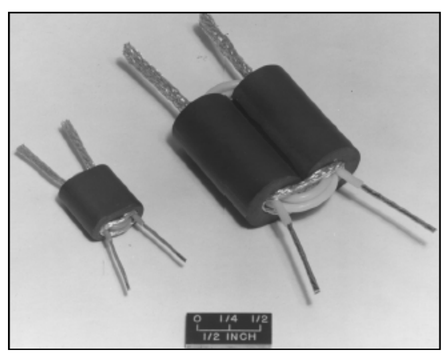
Typical examples of conventional transformers are shown in **Fig 17.40**. In Fig 17.40A, the primary winding consists of brass or copper tubes inserted into ferrite sleeves. The tubes are shorted together at one end by a piece of copper-clad circuit board material, forming a single turn loop. The secondary winding is threaded through the tubes. Since the low-impedance winding is only a single turn, the transformation ratio is limited to the squares of integers — 1, 4, 9, 16 and so on.

The lowest effective transformer frequency is determined by the inductance of the one-turn winding. It should have a reactance, at the lowest frequency of intended operation, at least four times greater than the impedance it is connected to. The coupling coefficient between the two windings is a function of the primary tube diameter and its length, and the diameter and insulation thickness of the wire used in the high-impedance winding. High impedance ratios, greater than 36:1, should use large-diameter secondary windings. Miniature coaxial cable (using only the braid as the conductor) works well. Another use for coaxial cable braid is illustrated in Fig 17.40B. Instead of using tubing for the primary winding, the secondary winding is threaded through copper braid. Performance of the two units is almost identical.

The cores used must be large enough so the



(A)



(B)

Fig 17.40 — The two methods of constructing the transformers outlined in the text. At A, the one-turn loop is made from brass tubing; at B, a piece of coaxial cable braid is used for the loop.

core material will not saturate at the power level applied to the transformer. Core saturation can cause permanent changes to the core permeability, as well as overheating. Transformer nonlinearity also develops at core saturation. Harmonics and other distortion products are produced, clearly an undesirable situation. Multiple cores can be used to increase the power capabilities of the transformer.

Transmission-line transformers are similar to conventional transformers, but can be used over wider frequency ranges. In a conventional transformer, high-frequency performance deterioration is caused primarily by leakage inductance, which rises with frequency. In a transmission-line transformer, the windings are arranged so there is tight capacitive coupling between the two. A high coupling coefficient is maintained up to considerably higher frequencies than with conventional transformers.

The upper frequency limit of the transmission line transformer is limited by the length of the lines. As the lines approach ¼ wavelength, they start to exhibit resonant line effects and the transformer action becomes erratic.

MATCHING NETWORKS AND TRANSFORMERS

The typical tube amplifier tank circuit is an impedance transforming network in a pi or pi-L configuration. With reasonable loaded Q, it also functions as a low-pass filter to reduce the output signal harmonic levels below FCC minimums.

If a transistor amplifier uses a broadband transformer, it must be followed by a separate low-pass filter to achieve FCC harmonic suppression compliance. This is one reason broadband transistor amplifiers are operated in push-pull pairs. The balance between the circuit halves naturally discriminates against even harmonics, making the filtering job easier, especially for the second harmonic. The push-pull configuration provides double the power output when using two transistors, with very little increase in circuit complexity or component count. Push-pull pairs are also easier to match. The input and output impedance of a push-pull stage is twice that of a single-ended stage. The impedance is low, and raising it usually makes the matching task easier

The transistor's low-impedance operation provides the opportunity to use a simple broadband transformer to provide the transformation needed from the transistor's load impedance up to 50 Ω. This low impedance also swamps out the effects of the device's output capacitance and, with some ferrite loading on the transformer, the amplifier can be made to operate over a very wide bandwidth without adjustment. This is not possible with tubes.

On the other hand, a tube amplifier with its variable output network can be adjusted for the actual output load impedance. The transistor amplifier with its fixed output network cannot be adjusted and is therefore much less forgiving of load variations away from 50 Ω. Circuits are needed to protect the transistor from damage by mismatched loads. These protection circuits generally operate "behind the scenes" without any operator intervention. They are essential for the survival of any transistor amplifier operating in the real world.

CALCULATING PROPER LOAD IMPEDANCE

The proper load impedance for a single transistor (or a tube for that matter) is defined by $R = E^2/P$. Converting this from RMS to peak voltage, the formula changes to $R = E^2/2P$. If two devices are used in push-pull (with twice the power and twice the impedance) the formula becomes $R = 2E^2/P$.

There is a constraint. Transformer impedance ratios are constrained to the square of their turns ratios to integer values, 1, 4, 9, 16 and so on. Real-world transformers quickly lose their bandwidth at ratios larger than 25:1 due to stray capacitance and leakage inductance. A design solution must be found which uses one of these ratios. We will use the ubiquitous 100-W, 12-V transceiver power amplifier as an example. Using the push-pull formula, the required load impedance is $2 \times 12.5^2/100 = 3.125\ \Omega$. The required transformer ratio is 50/3.125 = 16, which gives a turns ratio of 4:1.

Transistor manufacturers, recognizing the broadband transformer constraint, have developed devices that operate effectively using automotive and military battery voltages and practical transformer ratios: 65 W devices for 12 V operation and 150 W devices for 48 V. Bigger 50-V devices have been designed (for example, the MRF154) that will put out 600 W, but practical transformer constraints limit 50-V push-pull output power to 900 W. These device and transformer limitations on output power can be overcome by combining the outputs of several PA modules.

17.10.5 Combiners and Splitters

With some exceptions, practical solid state amplifiers have an upper power limit of about 500 W. This is a constraint imposed by the available devices and, to some extent, the ability to cool them. As devices are made more powerful by increasing the area of silicon die, the power density can become so high that only water cooling can provide adequate heat removal. Large devices also have large parasitic capacitances that make securing a broadband match over several octaves very difficult. By building a basic amplifier cell or "brick" and then combining several cells together, transmitter output powers are only limited by the complexity. Combiners and splitters have losses and add cost, so there are practical limits.

Broadband combiners usually take the form of an N-way 0° hybrid followed by an N:1 impedance transformer. The square ratio rule applies here too because the output impedance of a broadband combiner with N input ports is 50/N Ω — so combiners are usually 2, 4, 9 and 16-way. The higher the ratio, the lower the bandwidth will be.

Many types of combiners have been developed. The most common is the 4-way. It is easy to construct and has very good bandwidth. Most of the commercial "1-kW" broadband amplifiers use a 4-way combiner to sum the output of four 300-W push-pull modules operating on 48 V. Every combiner has loss. It may only be a few percent, but this represents a considerable amount of heat and loss of efficiency for a kilowatt output. This is a case where 4 × 300 does not make 1200.

The combiner approach to make a 1-kW solid-state amplifier uses a large number of individual parts. A comparable 1-kW tube amplifier requires relatively few. This makes the high-powered solid-state unit more expensive and potentially less reliable.

There is an alternative. If we had higher voltage transistors, we could use the same output transformer network configuration to get more output because power rises with the square of the operating voltage. There are high voltage transistors that can operate on 200 V or more. The problem is that these transistors must be capable of handling the corresponding higher power dissipation. The downside of making bigger, more powerful devices is an increase in parasitic capacitance. These bigger transistors become harder to drive and to match over broad bandwidths. However, the circuit simplicity and elimination of the combiner and its losses makes the high voltage approach quite attractive.

17.10.6 Amplifier Stabilization

Purity of emissions and the useful life (or even survival) of the active devices in a tube or transistor circuit depend heavily on stability during operation. Oscillations can occur at the operating frequency, or far from it, because of undesired positive feedback in the amplifier. Unchecked, these oscillations pollute the RF spectrum and can lead to tube or transistor over-dissipation and subsequent failure. Each type of oscillation has its own cause and its own cure.

In a linear amplifier, the input and output circuits operate on the same frequency. Un-

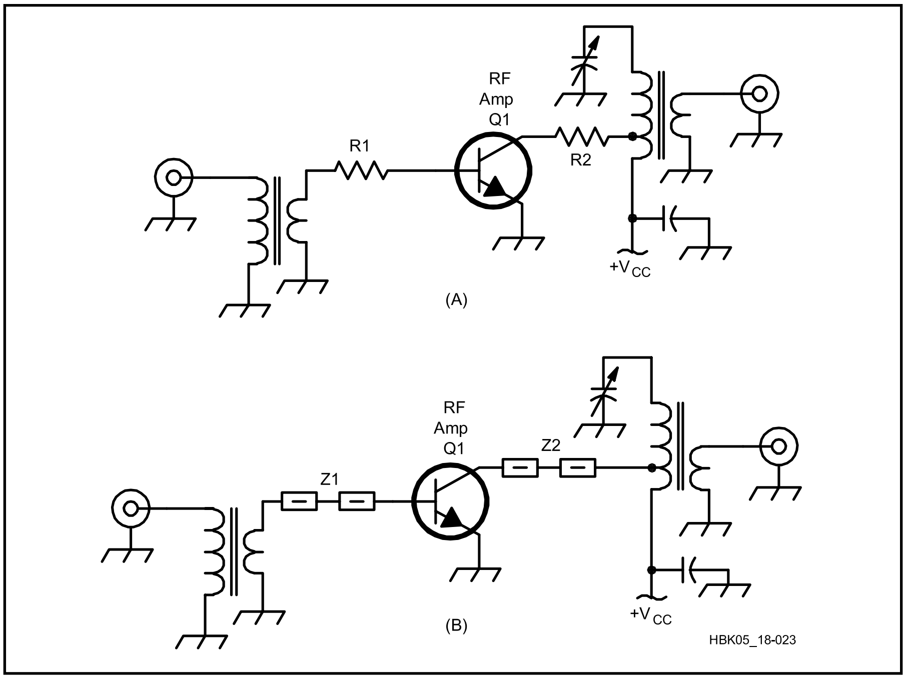


Fig 17.41 — Suppression methods for VHF and UHF parasitics in solid-state amplifiers.

less the coupling between these two circuits is kept to a small enough value, sufficient energy from the output may be coupled in phase back to the input to cause the amplifier to oscillate. 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. A high degree of 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 input and output circuits are on opposite sides of the chassis. Individual shielded compartments for the input and output circuitry add to the isolation. Transistor circuits are somewhat more forgiving, since all the impedances are relatively low. However, the high currents found on most amplifier circuit boards can easily couple into unintended circuits. Proper layout, the use of double-sided circuit boards (with one side used as a ground plane and low-inductance ground return), and heavy doses of bypassing on the dc supply lines often are sufficient to prevent many solid-state amplifiers from oscillating.

PARASITIC OSCILLATIONS

In low-power solid-state amplifiers, parasitic oscillations can be prevented by using a small amount of resistance in series with the base or collector lead, as shown in **Fig 17.41A**. The value of R1 or R2 typically should be between 10 and 22 Ω. The use of both resistors is seldom necessary, but an empirical determination must be made. R1 or R2 should be located as close to the transistor as practical.

At power levels in excess of approximately 0.5 W, the technique of parasitic suppression shown in Fig 17.41B is effective. The voltage drop across a resistor would be prohibitive at the higher power levels, so one or more ferrite beads placed over connecting leads can be substituted (Z1 and Z2). A bead permeability of 125 presents a high impedance at VHF and above without affecting HF performance. The beads need not be used at both circuit locations. Generally, the terminal carrying the least current is the best place for these suppression devices. This suggests that the resistor or ferrite beads should be connected in the base lead of the transistor.

C3 of **Fig 17.42** can be added to some power amplifiers to dampen VHF/UHF parasitic oscillations. The capacitor should be low in reactance at VHF and UHF, but must present a high reactance at the operating frequency. The exact value selected will depend upon the collector impedance. A reasonable estimate is to use an X_C of 10 times the collector impedance at the operating frequency. Silver-mica or ceramic chip capacitors are suggested for this application. An additional advantage is the resultant bypassing action for VHF and UHF harmonic energy in the collector circuit. C3 should be placed as close to the collector terminal as possible, using short leads.

The effects of C3 in a broadband amplifier are relatively insignificant at the operating frequency. However, when a narrow-band collector network is used, the added capacitance of C3 must be absorbed into the network design in the same manner as the C_{OUT} of the transistor.

LOW-FREQUENCY PARASITIC OSCILLATIONS

Bipolar transistors exhibit a rising gain characteristic as the operating frequency is lowered. To preclude low-frequency instabilities because of the high gain, shunt and degenerative feedback are often used. In the regions where low-frequency self-oscillations are most likely to occur, the feedback increases by nature of the feedback network, reducing the amplifier gain. In the circuit of Fig 17.42, C1 and R3 provide negative feedback, which increases progressively as the frequency is lowered. The network has a small effect at the desired operating frequency but has a pronounced effect at the lower frequencies. The values for C1 and R3 are usually chosen experimentally. C1 will usually be between 220 pF and 0.0015 µF for HF-band amplifiers while R3 may be a value from 51 to 5600 Ω.

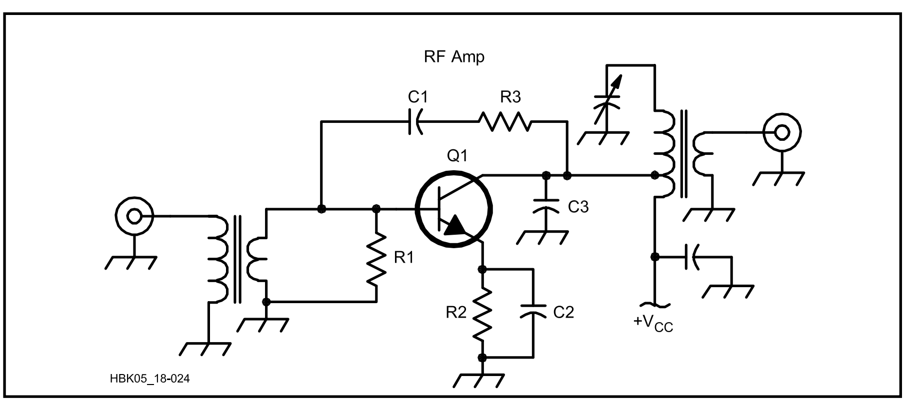


Fig 17.42 — Illustration of shunt feedback in a transistor amplifier. C1 and R3 make up the feedback network.

R2 of Fig 17.42 develops emitter degeneration at low frequencies. The bypass capacitor, C2, is chosen for adequate RF bypassing at the intended operating frequency. The impedance of C2 rises progressively as the frequency is lowered, thereby increasing the degenerative feedback caused by R2. This lowers the amplifier gain. R2 in a power stage is seldom greater than 10 Ω, and may be as low as 1 Ω. It is important to consider that under some operating and layout conditions R2 can cause instability. This form of feedback should be used only in those circuits in which unconditional stability can be achieved.

R1 of Fig 17.42 is useful in swamping the input of an amplifier. This reduces the chance for low-frequency self oscillations, but has an effect on amplifier performance in the desired operating range. Values from 3 to 27 Ω are typical. When connected in shunt with the normally low base impedance of a power amplifier, the resistors lower the effective device input impedance slightly. R1 should be located as close to the transistor base terminal as possible, and the connecting leads must be kept short to minimize stray reactances. The use of two resistors in parallel reduces the amount of inductive reactance introduced compared to a single resistor.

17.11 A New 250-W Broadband Linear Amplifier

The amplifier described here and shown in **Fig 17.43** is neither revolutionary nor daring. It uses commonly available parts — no special parts and no flea market specials. It is based on well-proven commercial designs and "best design practices" acquired over the past 30 years as solid-state technology has matured. This section will serve as a detailed design example for solid-state power amplifiers as well as a practical project that you can build.

Fig 17.43 — This 250-W amplifier for 160 through 6 meters provides a detailed design example as well as a practical project. Additional photos and information about the interior layout may be found on the *Handbook* CD-ROM.

A block diagram for the project is given in **Fig 17.44**. The amplifier is built on three PC boards — a PA module, a low-pass filter assembly, and a board for control, protection and metering circuitry. The *Handbook* CD includes *ExpressPCB* files for these boards, and the artwork can be used to have boards made in small quantities (see **www.expresspcb.com** for details).

The basic PA configuration has been in the *Handbook* for several years. It is intended to be a "ham-proof" external amplifier for QRP transceivers that put out 15 W or less. It is designed for a gain of 30×, or 15 dB. Drive power of less than 10 W will provide 250 W output from 1.8 through 50 MHz. The amplifier provides exceptionally linear performance, necessary for high quality SSB and PSK modes, and is rugged enough to withstand the most rigorous contest environment.

Amplifier design tends to focus on the RF section, but a successful stand-alone solid-state amplifier is equally dependent on its control system. The control requirements for a tube amplifier are well known, while those for solid-state amplifiers are not. This is mostly because the functions of a solid-state amplifier's control system are generally transparent to the user. Parameters are monitored and protection is applied without any operator intervention. This must be. While tubes are fairly forgiving of abuse, semiconductors can heat so quickly that intervention *must* be automatic or they can be destroyed.

Transistors are sensitive to heat, so cooling and temperature compensation are critical to a successful design. Transistors require a heat sink. Power amplifier tubes have large surface areas and are cooled by air blown on or through them. Transistors are small. Mounting them on a heat sink increases their thermal mass and provides a much larger surface area so the heat dissipated in the devices can be removed either by convection or forced air. The thermal design of an amplifier is just as important as the electrical design. More information on thermal design may be found in the **Analog Basics** chapter.

Silicon's thermal coefficient causes the bias current to increase as the device heats up if the bias source is fixed. The increased current causes even more heating and can lead to thermal runaway. For stable Class AB linear operation, the gate bias for a MOSFET or bipolar transistor must track the temperature of the device. The control circuit typically uses another silicon device such as a diode thermally coupled to the amplifier heat sink near the transistor to sense the temperature and adjust the bias to maintain a constant bias current.

Transistor power amplifiers are designed to operate into 50 Ω. Operation into a VSWR other than 1:1 will cause an increase in device dissipation and other stress. The success of the solid-state transceiver is due to its integrated PA protection system. The temperature of the heatsink, the load VSWR, the output power and the supply current are all monitored by the control system. If any of these exceed their threshold limits, the RF drive is reduced by the transceiver's ALC system.

An external solid-state PA protection system must perform the same functions, but the driver's ALC circuit is not always available so other means must be used to protect the PA. This is usually accomplished simply by taking the amplifier out of the circuit. An indicator then tells the operator which condition caused the fault so appropriate action can be taken. Access to the driver's ALC system would make this protection task more automatic, smoother and less troublesome, but no two transceiver models have the same ALC characteristic. This makes the design of a universal ALC interface more difficult.

17.11.1 The 1.8 to 55 MHz PA — Detailed Description

Fig 17.45 shows the power amplifier (PA) schematic. Two Microsemi VRF151 MOSFETs are used in this amplifier. The circuit topology is a 4:1 transmission line transformer type, rather than a "tube and sleeve" type common in many PA designs and discussed earlier. This style offers more bandwidth, necessary to provide performance on 6 meters. Typical gain is 15 dB; 10-W drive will easily provide 250-W output with a 48-V supply. There is a lot of latitude in this design. It can be operated on an unregulated supply. As long as the maximum unloaded voltage does not exceed 65 V, the transistors will not be overstressed. Other devices such as the MRF151, SD2931 or BLF177 would probably also work but have not been tested. They will require a regulated power supply, however.

FEEDBACK — TWO KINDS

The amplifier's gain is controlled by two

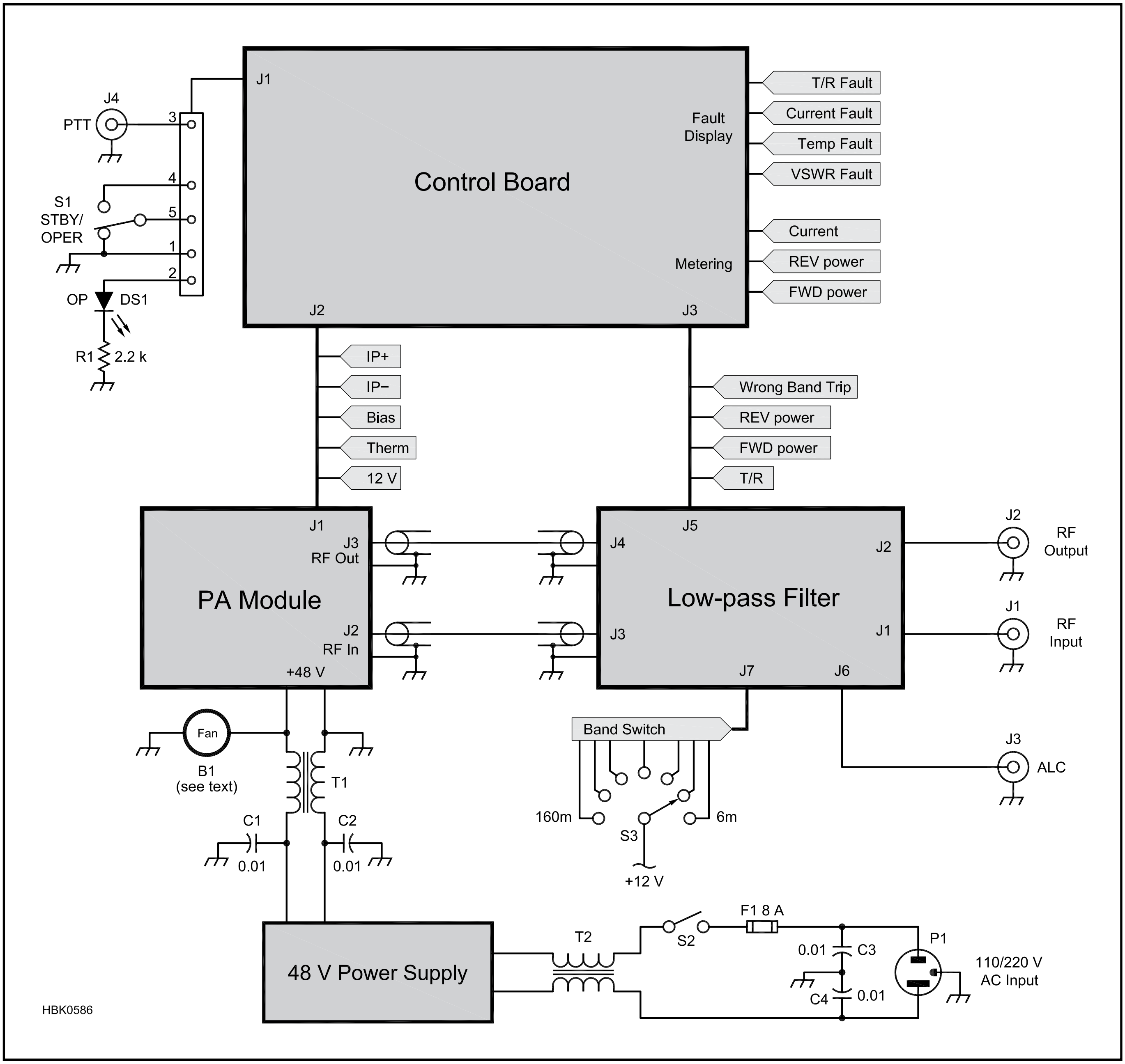


Fig 17.44 — Block diagram of the 250-W solid-state amplifier. It is built on three PC boards, which are described in the text and accompanying diagrams. T1 and T2 are 2 turns through a Fair-Rite 2643540002 ferrite bead if needed to suppress RFI from the switching power supply.

kinds of feedback. Shunt feedback (from drain to gate) is provided by the link on T2 through resistors R5 and R6. It tends to lower the input impedance but it also helps to keep the gain constant over frequency and improve the linearity. Series feedback is provided by the 0.05 Ω of resistance in each source. This increases the input impedance, cuts the gain by 3 dB, and most importantly, it has a huge effect on the linearity.

Without any feedback at all, the amplifier would have more than 30 dB gain (×1000) at some frequencies, tending to make it unstable — prone to parasitic oscillation. And the linearity would be terrible, –20 dB or so IMD products. It would also be very sensitive to load changes. The input SWR is 1.2:1 on 160 meters and rises to 1.5:1 on 6 meters. The amplifier's gain is 15 dB ±0.5 dB over the same frequency range.

PA INPUT

On the input side, T1 is a sleeve-and-tube RF transformer. Its small size and tight coupling make it suitable for very wide bandwidth operation. When properly compensated, this type is able to provide a match to a 12.5-Ω load over a very wide frequency range from 1 to 100 MHz. The problem is that the input impedance of the two transistors is not a flat resistive load.

The gate of a MOSFET is essentially a high-Q capacitor. A voltage greater than its threshold (V_{th}) applied between the gate and source will control the conductivity of the drain-to-source path. No power can be dis-

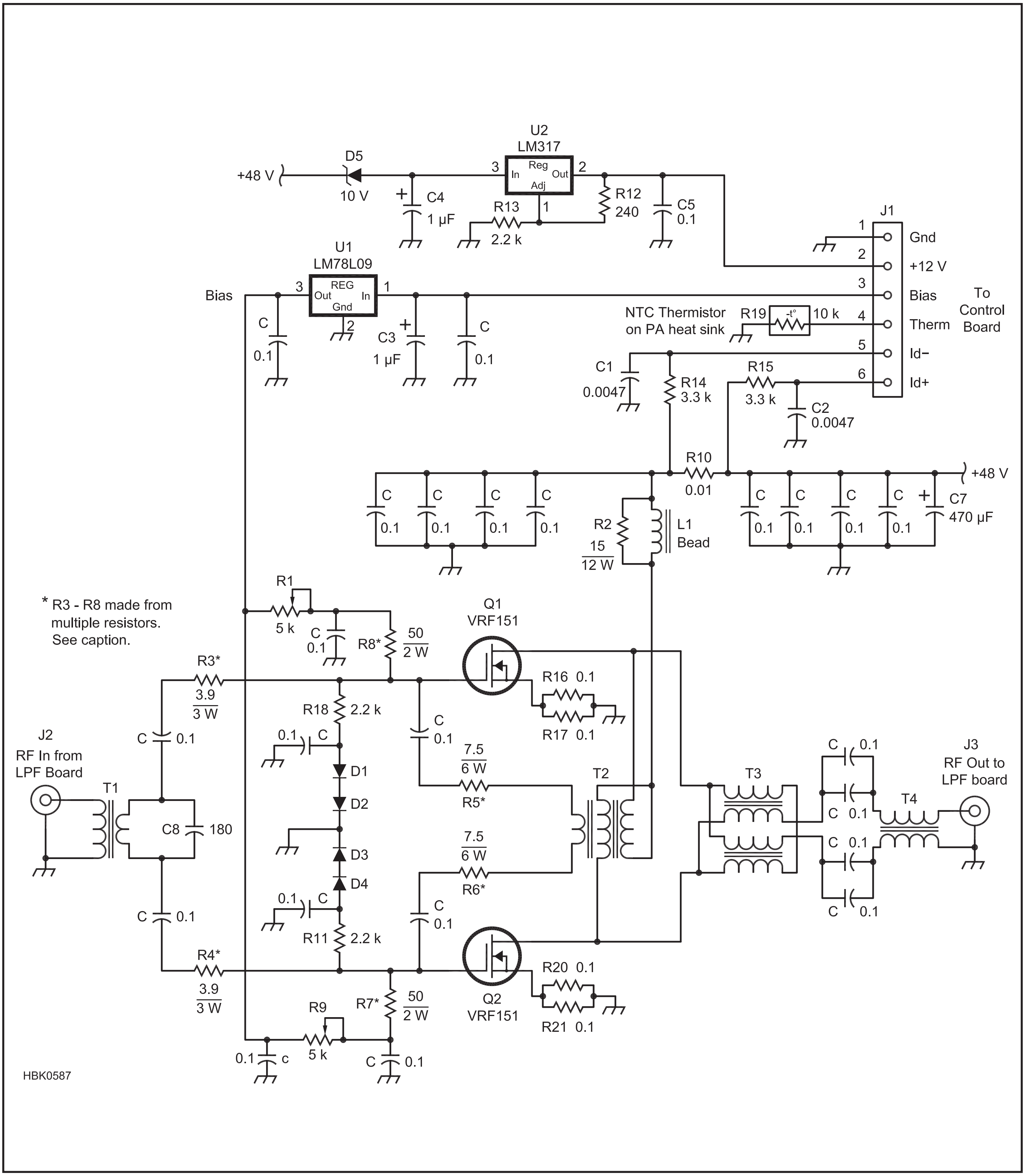


Fig 17.45 — Schematic diagram of the 250-W amplifier PA module. A complete parts list may be found on the *Handbook* CD.

C — 0.1 µF, 100 V X7R 1206 SMT.
D1-D4 — SMT silicon PN-junction diodes such as 1N4148 or equivalent.
L1 — 2t on Fair-Rite 2643480102 ferrite bead.
R2 — 15 Ω, 12 W.
R3, R4 — Three 12 Ω, 2 W SMT resistors in parallel.
R5, R6 — Two 15 Ω, 3 W resistors in parallel.
R7, R8 — Two 100 Ω, 1 W SMT resistors in parallel.
R19 — 10 kΩ, 5% NTC thermistor, SMT (DigiKey 541-1150-1-ND).
T1 — 2t #22 wire on CCI RF400-0 core (or two Fair-Rite 2643006302 cores)
T2 — Primary, 1t #22 wire; secondary 8t #22 wire bifilar wound, on Fair-Rite 5961004901 core.
T3 — 2 × 3t 25 Ω coax on Fair-Rite 2861010002 core (see text).
T4 — 3t RG-188 coax on Fair-Rite 2643665802 core (on cable to LPF board, not on PC board).
Q1, Q2 — Microsemi VRF151 MOSFET.

sipated in a purely reactive element, so we cannot match to a capacitor. Fortunately, all real reactive elements have losses and it is this loss that we would match to in a MOSFET gate for single-band operation. But that is no good here because we want a broadband match.

There are several ways to match a MOSFET over a broad frequency range. The most common is to swamp the gate capacitors with resistors. If the resistor value is lower than the impedance of the transistor gate capacitance, it dominates what the transformer sees as a load at the lower end of the bandwidth. The gate capacitor impedance decreases at higher frequency so some series resistance is added, R3 and R4, so that there is always a minimum real part to the load on the secondary of T1.

For a 1:1 input SWR, T1 wants to be terminated by a 12.5 Ω load. More than half of this is provided by R3 and R4. The rest is through shunt loads R7 and R8, plus the impedance of the output passed through the feedback network. T1 has no center tap so a balanced load is forced by the action of resistors R7 and R8. Having "soft" center taps on both the input and output absorbs any differences between transistors and greatly improves the network's RF balance. This in turn improves the cancellation of even harmonics at the output. The gate impedance is raised by the effect of the source resistors, further improving the match.

DC FEED TRANSFORMER

In addition to providing the link for the feedback, T2 also acts as the dc feed choke. It is wound with two parallel bifilar #22 wires and a single turn for the feedback. At dc, the current flows in opposite directions through each #22 wire so the net current is zero and the core does not saturate with dc. At RF, the choke with its ferrite core provides at least 50 Ω of inductive reactance making it essentially invisible to the RF signals across it. As the feedback transformer it provides a 1⁄16 sample of the drain-drain voltage to the gate feedback loop.

It is not often mentioned in the literature, but the dc feed transformer T2 acts as a 180° hybrid combiner (**Fig 17.46**). A hybrid combiner has four ports: two inputs, the sum and the difference port. The sum of the two input signals appears on the sum port and the difference between them (differences due to voltage or phase) appears at the difference port. In this case T2 is terminated with a 1:4 balanced transformer that brings the output impedance up to 50 Ω. The sum port is across the whole secondary. The two input ports are between each end of the secondary and ground. The difference port is between the center tap and ground.

No two transistors or their circuit layouts are exactly equal. In a push-pull circuit

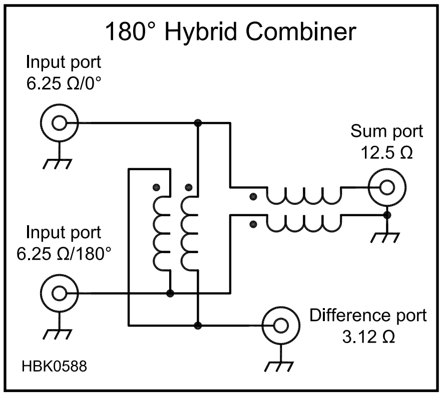


Fig 17.46 — Dc feed transformer T2 acts as a 180° hybrid combiner.

these slight differences gives rise to imbalance between sides which gives rise to even harmonics at the output. By not placing the usual heavy RF bypassing at the center (the difference port) of the bifilar winding in T2 the difference between the two drain voltages shows up across L1 and is dissipated in the parallel resistor R2. This enables the amplifier to achieve better than 40 dB suppression of even harmonics, helps the efficiency, and greatly improves its stability when driving mismatched loads.

THE PA OUTPUT TRANSFORMERS

The output circuit of the PA uses a 4:1 impedance transformer. We have two VRF151 transistors that will each put out 150 W if properly cooled. That means 300 W output power for the pair operating on 48 V. The classic formula relating drain-to-drain load impedance to the supply and output power is

$$R_L = \frac{2\,V_{dd}^{\;2}}{P_O} \qquad (10)$$

For 48 V this gives 15.36 Ω. But the drain cannot swing all the way to zero because of the finite on-resistance of the MOSFET. The actual available swing is typically 90% of the supply voltage so a more practical formula is

$$R_L = \frac{2\,(0.9\,V_{dd})^2}{P_O} \qquad (11)$$

A 4:1 transformer gives 12.5 Ω. This is a nice load for these devices.

The 4:1 impedance ratio is performed by a simple transmission line transformer. The impedance of the coax used should be the geometric mean between the two impedances to be matched:

$$Z_0 = \sqrt{50 \times 12.5} = 25\,\Omega$$

Coax with this characteristic impedance is not a common stock item but it is available as p/n D260-4118-0000 from Communications Concepts, Inc. (**www.communication-concepts.com**). Two feet are required. An acceptable alternative is #22 shielded 600-V Teflon-insulated wire such as Belden 83305-E whose physical dimensions result in approximately the same characteristic impedance.

These two coax lines are connected in parallel on the drain end and in series on the 50 Ω output end. Any voltage on the input is put in series at the output, giving a 2× voltage ratio or a 4× impedance ratio, exactly what we want. If coiled up in separate coils, the transformer will work without any ferrite. The two coils cannot be allowed to couple so they cannot be on the same form.

In order to get a wider frequency response, the inductance of these coax coils is increased by winding them on a ferrite core. The core used here has two holes, so independent coils can be wound on each side without any coupling between them and it makes a nice neat package. Separate cores would work just as well. The ferrite core is type 61 material with a permeability of μ_i = 125. It is a binocular bead but could be replaced by two 0.5 dia × 1-inch long sleeves of the same material. The ferrite "load" on the coax makes its outside shield a high impedance from end to end, while inside the shield the coax maintains its 25 Ω characteristic impedance between center and shield. As long as the coax lines of T3 are wound as two non-coupled coils, the amplifier will operate from 21 to 80 MHz without any ferrite at all.

The advantage of the transmission line type of RF transformer is that it does not have the leakage reactance that plagues the tube-and-sleeve type of transformer used on many solid-state amplifiers. Simply stated, a parasitic leakage inductance is introduced in series with the primary due to incomplete coupling of the flux between the primary and secondary winding. This increases the apparent impedance of the low-Z side of the transformer as the frequency increases. The output impedance of the transistors decreases with frequency — a double hit of mismatch that causes the gain to drop off quickly.

There is one disadvantage to the transmission line transformer. It has a balanced input and output. Sometimes designers will ground one side of the output and rely on the ferrite loading to decouple the ground side of the output. This has a negative effect on the balance of the amplifier, and on those even harmonics we want to minimize. It also doubles the flux stress in the ferrite causing it to heat more. The solution is T4, a simple current balun — four passes of the 50 Ω output line through a toroid of type 61 ferrite. With T4 in place, a balanced load on the output of T3 is maintained, the even harmonics are suppressed, and the efficiency is 5 to 10% better on most bands.

PA LAYOUT

The PA board (**Fig 17.47**) was designed with all parts mounted on the top surface

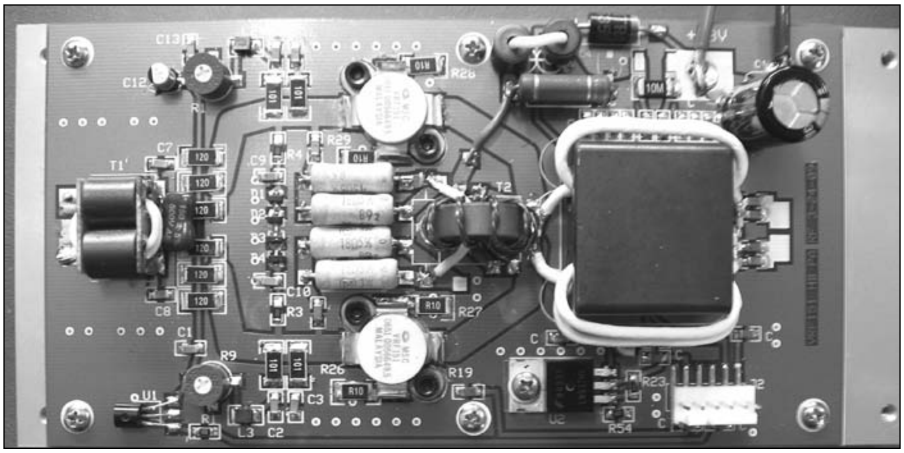


Fig 17.47 — The power amplifier module PC board assembly mounted on the heat sink.

— no through-hole parts at all. The back side of the PC board is a continuous ground plane and is mounted directly to the heatsink without spacers.

17.11.2 Control and Protection

The control board appears far more complicated than the PA but in reality, it is just a few analog and logic ICs. The various control and protection circuits are shown in **Figs 17.48** and the LED displays and drivers (also on the control board) are shown in **Fig 17.49**. This circuitry monitors several parameters, displays them, and if necessary, puts the amplifier into standby if one of them goes out of range. This control system could be used on any amplifier. All solid-state amplifiers need similar protection. The amplifier is protected for:

1. Over temperature, by a thermistor on the heatsink and setting a limit.
2. Over current, by measuring the PA current and setting a maximum limit.
3. Over SWR, by monitoring the reflected power and setting a maximum limit.
4. Selection of a low-pass filter lower than the frequency in use.

Each of these fault trips results in lighting an error LED and forcing the amplifier into the standby position, out of the RF path. There is also an ALC level detector that generates a negative-going feedback voltage for the driver when the RF drive goes above the level corresponding to maximum power. If this PA were part of a transceiver, the several faults described above would generate inputs into the ALC system and turn back the drive rather than taking it off the air. We do not always have that luxury so the best course is to take it off line until the cause can be fixed.

THERMAL COMPENSATION

MOSFETs are sensitive to temperature. If a fixed bias is used on the gate to set the quiescent bias at 100 mA when the device is cold, the current will increase as the device heats up. In some devices, it will cause thermal runaway. The hotter it gets, the more current it draws, causing even more heating and so on. The solution is to sense the temperature of the device and reduce the gate bias as it heats up. The VRF151 is relatively insensitive compared to similar high power RF MOSFETS.

The compensation system is quite simple, effective and foolproof. It relies on the thermal characteristic of silicon diodes that as a diode heats up, the forward voltage across it goes down approximately 2.4 mV/°C. Two diodes in series are used at the bottom of each gate voltage divider (D1-D4, any silicon PN-junction diode in a suitable SMT package will work). Mounted on the PA board, they heat up along with the transistors and reduce the gate voltage by a proportional amount. The 100 mA of bias at 25 °C is less than 150 mA at 200 °C. This compensation system may not work as well if other MOSFET types are substituted because they have different thermal coefficients of V_{th} and may require more aggressive thermal compensation. Gate bias voltage is provided via a PTT-activated 9-V regulator.

A simple way to check or adjust compensation is to place the amplifier module, board and heatsink, on an electric frying pan. Set it to 212 °F (100 °C) and monitor the drain current. It should stay within 150% of the cold setting. Be patient. This takes a while because the response time of this arrangement is quite long. Under some operating conditions, it is possible for the transistor to get very hot before the heat travels to the diodes. For this reason, we have additional means to protect the transistors.

OVER-CURRENT PROTECTION

One simple protection method is to limit the power into the PA module by limiting the maximum supply current. The drain current is sensed across shunt resistor R10. The sense voltage is amplified by U2A and sent to the current meter. It also goes to comparator U2B where it is monitored and compared to the limit voltage set by R13. If it exceeds this limit, the comparator trips the fault latch U1, lights the OC LED, and opens the PTT.

OVER-TEMPERATURE PROTECTION

The heat sink temperature is monitored directly by a negative temperature coefficient thermistor, R19, mounted on the PA assembly. It is the lower half of a voltage divider sensed by U3A and fed to comparator U3B. When the thermal sense voltage exceeds the limit set by R14, the U1 fault latch is tripped and the PTT line opened. Any time the fault latch is tripped it lights an LED to indicate the cause of the fault so the operator can address the problem. Cycling the OPERATE - STANDBY switch resets the fault latch and restores normal operation.

VSWR PROTECTION

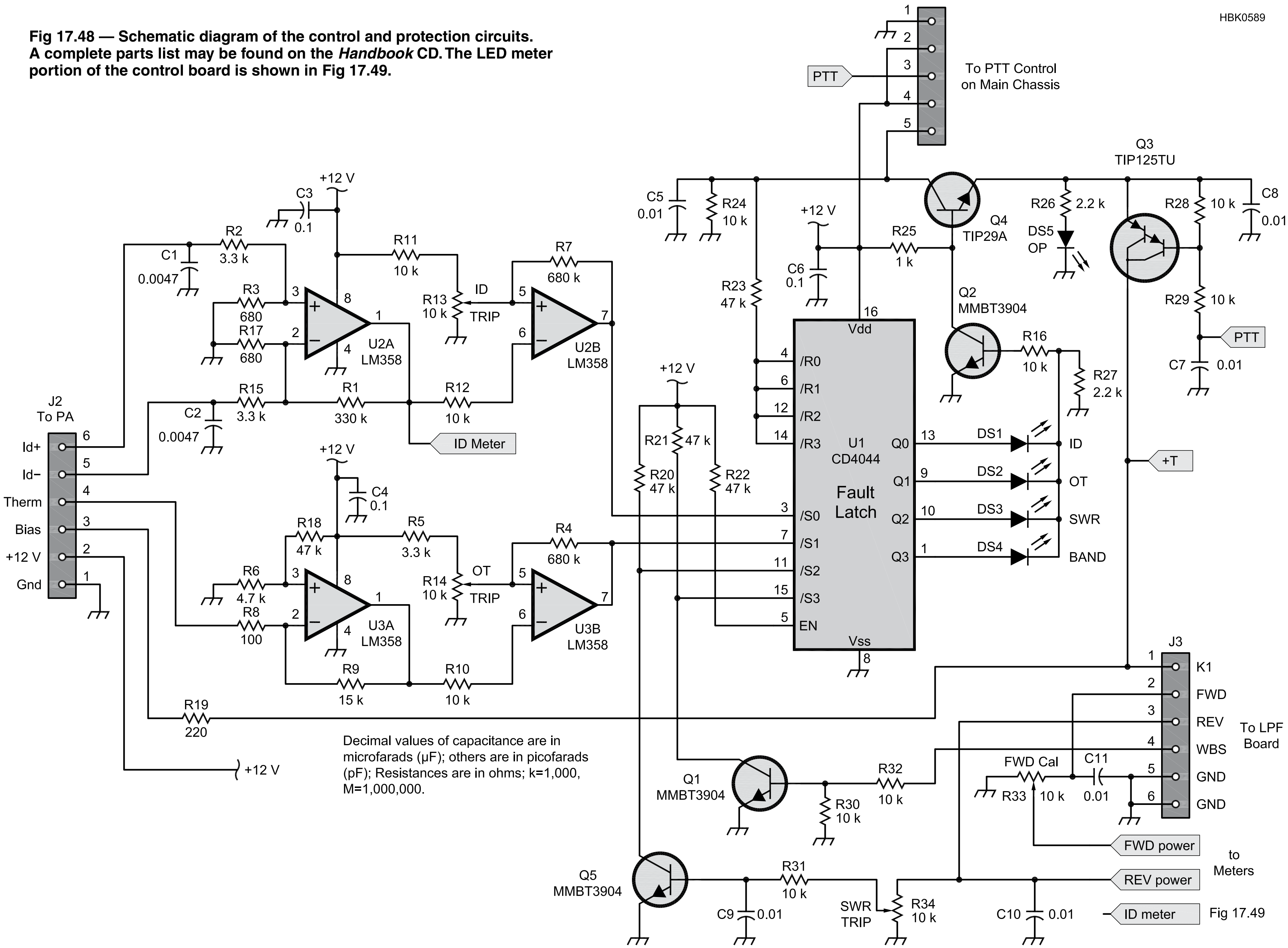
The amplifier is protected for high VSWR. The amplifier is designed to operate into 50 Ω. It acts somewhat like a constant voltage source. It will dutifully try to put the same voltage across whatever load it is given, even a short. The SWR bridge produces a voltage proportional to both the output power and the relative mismatch between the load and 50 Ω.

T1 on the LPF board is the heart of a dual directional coupler that provides both the VSWR detection and the forward power monitoring. It is really two transformers on one two-hole core, Fair-Rite # 2843010402. Each has 17 turns of #28 AWG as the secondary and a single "turn" of #22 AWG Teflon hookup wire as the primary.

T2 is a single directional coupler that looks at the power reflected from the low-pass filter. Its secondary has 20 turns of #28 AWG wire, and the primary is a single pass of #22 AWG wire. In order to minimize the impedance bump it places in the RF signal path between the PA and the LPF, the core is mounted in a small window cut in the board. This allows the toroid to be mounted so the primary wire can go straight through the center of the toroid. Detailed photos of all the transformers are contained on the accompanying CD.

Several load conditions can produce the same value of indicated VSWR. At one mismatch load condition the amp might be trying to put out way too much power. At a different reflection coefficient, it might see a very high reflected voltage. This could raise the peak voltage on the drains past the voltage breakdown limit of the MOSFET. The voltage from the detector on the reflected power port of the SWR bridge is brought to the meter and to a comparator that will trip the fault latch when the voltage is past a limit. Again the PTT is opened. The operator can either lower the SWR by improving load match or

Fig 17.48 — Schematic diagram of the control and protection circuits. A complete parts list may be found on the *Handbook* CD. The LED meter portion of the control board is shown in Fig 17.49.



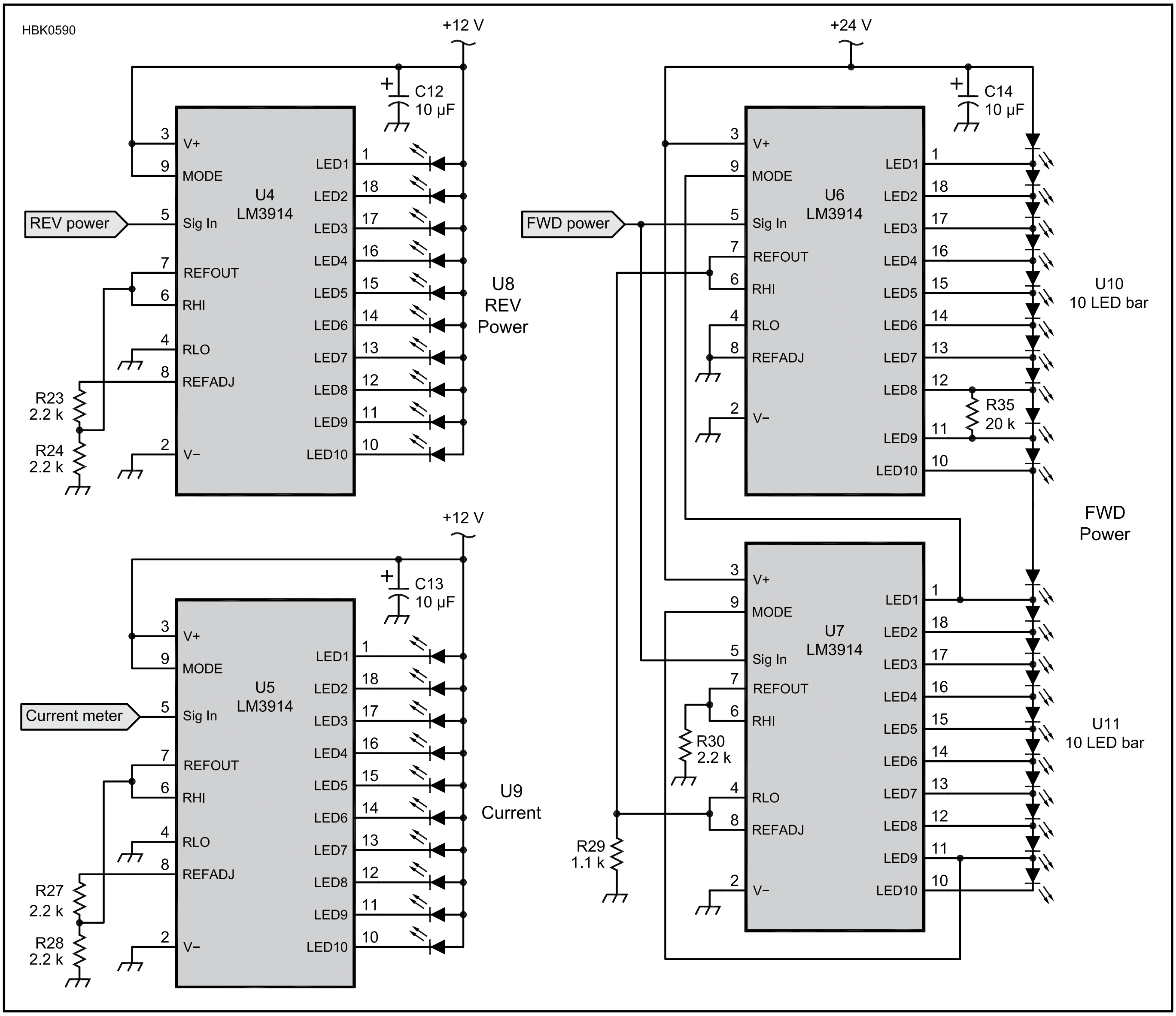


Fig 17.49 — Schematic for the LED meters, which are also located on the control board with the circuitry shown in Fig 17.48. A complete parts list may be found on the *Handbook* CD.

reduce the output power to limit the reflected power from the bad match. The amplifier is happy either way.

Transformer T1 on the LPF board is the heart of the directional coupler used for VSWR protection and power metering. It consists of identical transformers wound on each side of a two-hole ferrite core. The secondary is formed from 17 turns of #28 AWG enameled wire. The primary is a single pass of #22 AWG insulated hookup wire though the hole. A picture of the transformer is contained in the companion CD.

BAND FAULT PROTECTION

One problem unique to external broadband amplifiers is that they generally do not care what frequency they amplify, but their low-

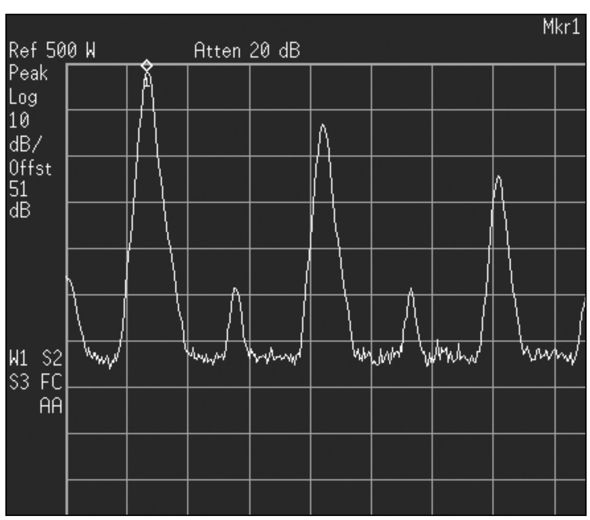
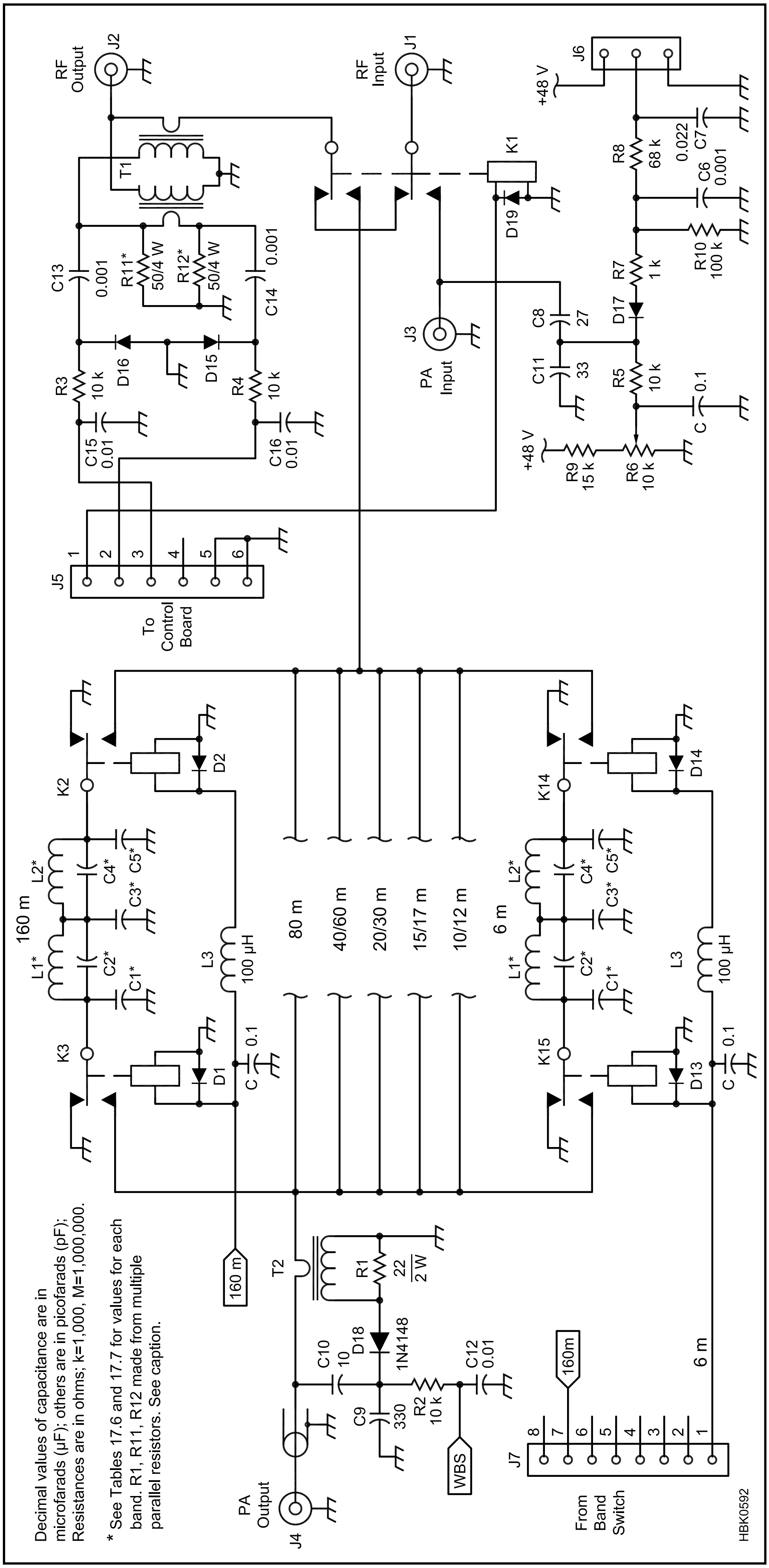


Fig 17.50 — The output of the 250 W amplifier without a low-pass filter. The even harmonics are suppressed by more than 40 dB. The odd harmonics are not suppressed by the push-pull balance and must be attenuated in the LPF. Harmonics after the LPF are all greater than 60 dB down.



pass filters on the output certainly do. The problem arises when the operator forgets to change the band switch on the amplifier when moving to a higher band. All the power from the amp is reflected back from the LPF and the amplifier is distressed — all that power with no load. The solution is to place a reflected power sensor between the PA and LPF. A monitor will see this condition and trip the fault latch for Wrong Band Selection.

T2 is single directional coupler used for reflected power only. The transformer is mounted in a small window cut in the board that allows the primary wire to go straight through the center of the toroid. This coupler normally sees all the reflected harmonic power from the LPF. The maximum harmonic power from the PA output is down 13 dB from the fundamental. See Fig. 17.50. When the wrong filter is selected, all of the power is reflected by the filter so the threshold is set to detect this 20× (13 dB) difference.

17.11.3 Low-Pass Filter

For amateur use, FCC §97.307(d) requires that harmonics be suppressed at least 43 dB for operation below 30 MHz, and at least 60 dB on 6 meter and higher frequency bands. The output signal from a broadband amplifier itself contains harmonics and needs to have a separate filter to meet the FCC's requirements for harmonic suppression. **Fig 17.50** shows the harmonic output of this amplifier without any low-pass filtering.

In a well designed solid-state push-pull amplifier, the second harmonic is 30 to 40 dB below the fundamental if the balance is good, but the third harmonic is only down 13 dB. This means the low-pass filter needs to supply about 10-13 dB of attenuation at the 2nd harmonic and 30 dB or more to the rest. On 6 meters the filter must provide at least 35 dB attenuation at the 2nd harmonic and 50 dB to the rest. None of this is difficult for a properly designed low-pass filter. In a commercial application, all harmonics must be at least 60 dB down. This usually requires

Fig 17.51 — Schematic for the low-pass filter board. The filters are described in detail in the text, and values are shown in Tables 17.6 and 17.7. This board also contains the TR relays, directional couplers and ALC circuitry. A complete parts list may be found on the *Handbook* CD.

K1 — DPDT, 12 V coil, 8 A contacts (Potter & Brumfield RTE24012F).

K2-K15 — SPDT, 12 V coil, 10 A contacts (Omron G5LA-14-12DC).

R1 — Two 47 W, 2 W resistors in parallel.

R11, R12 — Two 100 W, 2 W resistors in parallel.

T1 — Dual transformer (see text).

T2 — Secondary: 20t #28 AWG on Fair-Rite 5961000201 toroid. Primary: 1t #22 AWG (see text).

a more complicated filter, especially if continuous coverage is desired.

The largest board in the amplifier, shown schematically in **Fig 17.51**, contains the low-pass filters, TR antenna relay, SWR bridge and the ALC detector. Amateur bands are harmonically related, so a filter for 40 meters will not do anything to reduce the second harmonic on 80 meters. If an amplifier is going to cover 160 to 10 meters, it will need at least five filters. With 30, 17 and 12 meters, the number of filters is usually increased by one to provide better suppression of harmonics. When 6 meters is added another filter is required, and it must be able to bring all 6-meter harmonics to –60 dB. This requires a more complicated filter.

FILTER DESIGN

The HF band filters can easily meet their requirements with simple five-element, 0.044-dB ripple Cauer filters. These filters use an elliptic topology. The nulls can be arranged to provide specific treatment of the third harmonic. Also, the insertion loss is the lowest of the several common filter types. The *SVC Filter Designer* software provided on the *Handbook* CD was used to design the LPFs used here. (See the **RF and AF Filters** chapter for more information on Cauer and other filter types.).

Low-pass filters are precision-tuned circuits. If the values are not right, the filter will have high loss and/or high VSWR in the passband. The calculated capacitor values are rounded to the closest 5% standard values. This is done by the *SVC Filter Designer* program. In a 5th order Cauer filter, there are two parallel resonant circuits that set the nulls in the response. If the exact calculated C value is not used, its paired L must be adjusted so the desired null still hits at the proper frequency. **Table 17.6** gives all the LPF capacitor and inductor values, the corner frequency (F_c), and the frequencies of the nulls (F1 and F2). L1-C2 resonate at F1 and L2-C4 resonate at F2.

LPF Inductors

Inductor winding details are given in **Table 17.7**. The low-frequency coils are wound on Micrometals T80-2 powdered iron toroid cores. For 20/30 and 15/17 meters the mix is changed to T80-6. Use of toroid cores keeps the Q of the coils high, makes physically smaller coils and provides magnetic shielding. This construction also helps to prevent the various sections of the filter from "talking to each other" and causing "suck-outs" in the passband or "lumps" in the stop band.

The 10/12 and 6 meter coils are self-supporting air-wound types with no cores which gives the highest possible Q. Coil adjustment is done by compressing or spreading turns on the cores. On the high bands, the coils are 5% high when wound tight. Spreading the coils slightly brings them to the proper value.

As mentioned before, it is important that the nulls in the Cauer filter response occur at the right frequency. Since the capacitor values have been rounded to the closest 5% values, the value of each parallel inductor has to be tweaked to set the null on the proper frequency. This is easy to do with a network analyzer but rather difficult for the home builder because the nulls are at various frequencies up to 154 MHz.

LPF Capacitors

Selecting capacitors for the LPF is a bit trickier. At 300 W of RF, the requirement is 123 V RMS or 174 V peak. A 500-V capacitor will easily handle this. We are also looking for RF current handling capability. At 300 W of RF, the requirement is 2.5 A RMS into 50 Ω. If a capacitor carrying this current has an equivalent series resistance (ESR) of 0.4 Ω it will dissipate 2.5 W. Harmonic currents can increase the heating dramatically.

Ceramic capacitors come in several grades of dielectric quality — X7R, Z5U and NP0 (or C0G). The NP0 is called a "Class 1" dielectric (see the **Component Data and References** chapter for more on capacitor characteristics). NP0 (C0G) capacitors are more expensive and harder to find, but they are the only type suitable for use in RF power filters. The best way to make capacitors for a PA low-pass filter is to use several in parallel. This spreads the RF current across several units and permits obtaining odd values by combining standard value parts.

RF capacitors are very difficult to find, and all the required values are never in the dealer's stock. Several capacitor manufacturers were tried and most quoted 6 to 14 week lead times and had minimum order quantities of $100

Table 17.6
Low Pass Filters
5th order Cauer
Ap = 0.044, As = 40 dB

Band (m)	*Fc (MHz)*	*C1 (pF)*	*L1 (µH)*	*C2 (pF)*	*F1 (MHz)*	*C3 (pF)*	*L2 (µH)*	*C4 (pF)*	*F2 (MHz)*	*C5 (pF)*
6	57.4	36	0.1571	6.8	154	75	0.1245	20	100.85	27
10/12	30.9	68	0.307	12	82.9	150	0.2387	36	54.3	51
15/17	22.265	120	0.424	22	52.09	220	0.338	62	34.771	91
20/30	15.053	180	0.619	33	35.217	330	0.458	100	23.508	130
40/60	8.837	300	1.058	56	20.674	560	0.831	160	13.801	240
80	4.778	560	2.027	100	11.177	1000	1.517	300	7.461	430
160	2.243	1200	4.181	220	5.248	2200	3.329	620	3.503	910

Table 17.7
Low Pass Filter Inductor Winding Details

Band (m)	*L1 (nH)*	*Core Type*	*Wire Size (AWG)*	*No. of Turns*	*Inside Dia. (in.)*	*L2 (nH)*	*Core Type*	*Wire Size (AWG)*	*No. of Turns*	*Inside Dia. (in.)*
6	157	—	16	5	0.312	124	—	16	4	0.33
10/12	307	—	16	7	0.35	238	—	16	3	0.33
15/17	424	T80-6	18	10	—	338	T80-6	18	9	—
20/30	619	T80-6	18	12	—	458	T80-6	18	10	—
40/60	1058	T80-2	18	14	—	831	T80-2	18	12	—
80	2027	T80-2	20	19	—	1516	T80-2	20	17	—
160	4180	T80-2	20	28	—	3329	T80-2	20	24	—

All of the cores are wound with enameled copper wire. The size is as large as will fit to maximize coil Q.

to $500 of each value! The best ceramic RF capacitors are made by ATC. Fortunately they also had most of the values in stock and in reasonable minimum quantities. Several values were paralleled when a particular value was not available. They are quite expensive, about $3 each for 500 V units below 200 pF and $8 each for the larger units. The filter set uses 29 different values. A commercial amplifier manufacturer would have to make a large capital investment to stock an LPF production line, another barrier for a solid-state legal-limit amplifier.

An alternative capacitor solution would be to use 500-V silver-mica capacitors. These are suitable for all but the 10-meter and 6-meter filters where a single unit's RF current rating might be exceeded and the lead parasitics changes the net value considerably. While SMT silver-micas are available, they are also hard to find in all the values needed. The leaded versions are far more common and are made by several companies. The downside is that the LPF board was laid out for surface-mount capacitors, all mounted on the bottom side of the board. Using leaded through-hole parts will require laying out a new LPF board and it will have to be larger.

With the capacitors used, the filter is suitable for a full kW output in a system with an SWR protection circuit as implemented in this amplifier. The coils might need to be implemented on the next larger size core.

LPF Construction

The LPF board, shown in **Fig 17.52**, has a continuous ground plane on the top side. All the coils are mounted through the board. As noted, all the capacitors are leadless SMT types mounted on the bottom side. Plated-thru vias complete the circuit to the top side ground. This arrangement with the coils on the top and capacitors on the bottom provides the best space efficiency and gives repeatable filter performance. Repeatability is not so much a concern for the home builder, but critical for commercial equipment.

To prevent coupling between bands, the filters are not in order across the board. On the two high band filters, the axis of each coil is perpendicular to the next to prevent coupling between them. Finally, each filter is shorted out by the selection relays when not in use, which helps prevent "sneak paths" around the selected filter.

The filter selection relays are used in pairs. Each is rated for 10 A, overkill for this application. They are quite inexpensive, around $1 each in quantities of 25, and have very low loss and parasitic inductance. The 10/12 meter and 6 meter filter values were tweaked for lowest loss and best efficiency after the whole filter was built. Computer modeling of the stray capacitances and inductances proved tricky and in the end it was better to temporarily replace the end capacitors with trimmers and tune for best performance to find the final values and then replace them with fixed units.

ALC DETECTOR

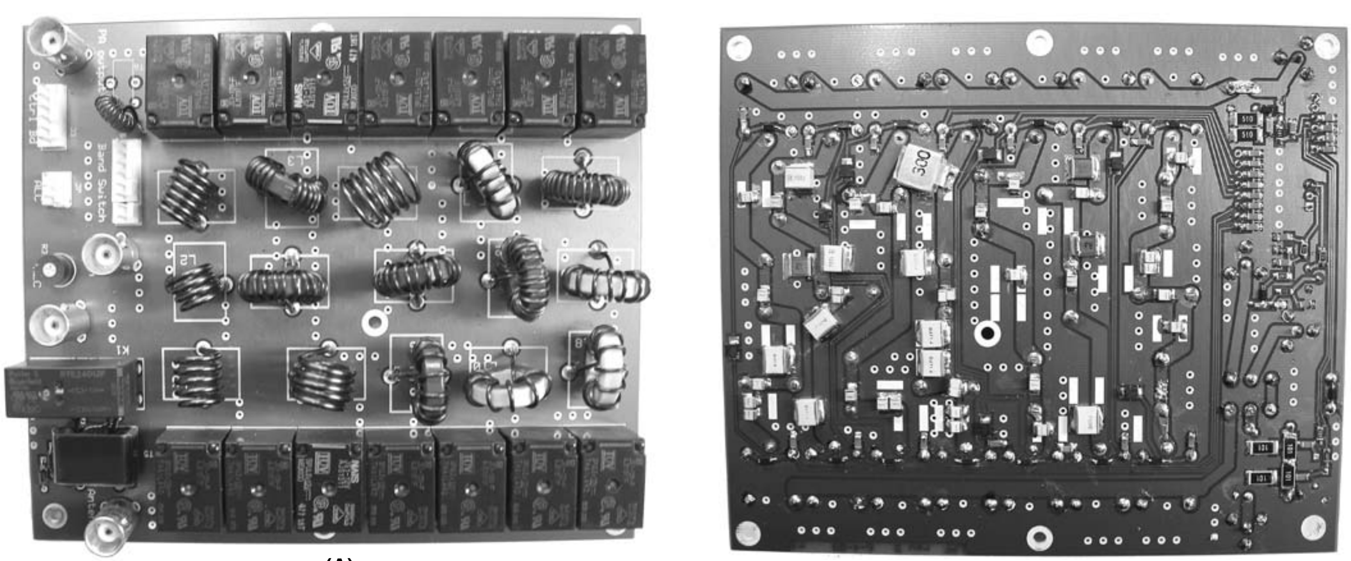
There is also an automatic level control (ALC) detector on the LPF board. ALC provides a feedback signal to the exciter to prevent overdriving the amplifier, and it was once a very common connection between the exciter and amplifier. Now that most solid-state transceivers have an output power control, it is used less often but still useful.

ALC interfacing is difficult because there is no standard ALC system specification among different transceiver brands and models. The detector here is a generic circuit with median values that generates a negative-going ALC signal. D17 rectifies the drive signal after a detection threshold, set by R69, has been exceeded. The attack time and filtering time constants are set by the RC network following the diode. Some adjustment of the R/C values may be required to work with a particular model of transceiver because of loop dynamic stability issues.

17.11.4 Power Supply

The PA requires 48 V at 10 A peak. It does not have to be regulated provided the following conditions are met. The maximum no-load voltage of 55 V is preferred, and it cannot exceed 60 V under any condition to provide a suitable safety margin. It must have at least 10,000 µF of filtering to prevent noticeable hum modulation. A simple power transformer feeding a 20-A bridge rectifier and a filter capacitor is sufficient. It may require a step-start circuit to ease the high current surge on the line power switch. See the **Power Supplies** chapter for more information.

The alternative is to use a switching supply. Supplies rated for 48 V and 500-600 W supplies regularly go for less than $100 at online auction. They are small and light. You may have to deal with RFI issues on some of the cheaper models, though. This usually just requires a ferrite bead and additional bypass capacitors on both the ac input lines and the dc output lines as well.



(A) (B)

Fig 17.52 — The assembled low-pass filter board with the coils, relays and other components mounted on the top side (A) and the LPF capacitors mounted on the bottom (B). The prototype used available capacitors — two sizes of ATC ceramic RF chip capacitors, a metal clad mica and three unencapsulated SMT mica units. All are serviceable, but the ATC ceramics work best.

Power output varies with the square of supply voltage. The amplifier will put out 75 W if run from 25 V. The gain changes very little, just the output power. This squared relationship makes it more sensitive to ripple modulation, so make sure that the power supply, no matter how it is made, has reasonable filtering. It does not need to be current-limited since the PA control has this built-in. It does need a suitable fuse, however. A 15-A low-voltage fuse will do nicely.

The power supply used in the test amplifier was purchased on eBay. It was called "500 W 48 V 10.4 A Switching Power Supply for Radio" and was less than $100 including shipping. It is 8.5-inch long, 2-inch high and 5.54-inch wide. It was the best value of all power supply options investigated. However, as in all compromise decisions, it was not without a downside as will be described later.

17.11.5 Amplifier Cooling

Cooling is the single most important requirement for a transistor's reliability. Tubes can get hot and angry, but they are made from high-melting-point materials so they can stand the heat. Transistors cannot. They will melt. The active area of a VRF151 die is 0.034 square inch, about the area inside the letter P on a computer keyboard. It must dissipate up to 150 W when key down. The heat dissipated in the transistor die is conducted through to the transistor's package base. Here it is conducted to the heatsink which, because of its large surface area, can effectively couple the dissipated heat into the surrounding air.

Fig 17.53 shows the VRF151s on the PC board and heatsink. There is no substitute for ensuring a good mechanical fit before mounting the transistors. The heatsink surface must be as flat as possible and the transistors must also be flat. A sheet of 600 grit sandpaper on a glass plate makes a good way to check the flatness of high power metal-backed transistors. A couple of light strokes across the paper will reveal any high spots on the bottom of the device. Keep rubbing until the back is all the same color. The plating on the back is not necessary once it is mounted — bare copper is fine.

Similarly, the heatsink must also be flat. Extruded aluminum heat sinks are notoriously "lumpy." A careful swipe across the width of the sink with a large flat file, perpendicular to the direction of the fins, will reveal any ridges and valleys. File it until they disappear. Be careful to keep the file clean or you can do more harm than good. In commercial practice, extruded heat sinks are usually milled flat before use.

The interface between the transistor and the heatsink is always greased with thermal heatsink compound. Thermal grease is a suspension of zinc oxide powder in silicone or mineral oil. It is very similar to the white sun screen paint used to protect your nose at the beach. Here it is the oil that does the work. The zinc oxide is simply a filler to keep the oil from running away. Oil is not a good conductor of heat but is much better than air. The thermal grease is used to fill any microscopic gaps and scratches between the sink and the transistor. Use only as much grease as needed to fill the gaps. Spread an even thin coat of grease on the bottom of the transistor with a knife blade then put the part on the sink. Before putting in the screws, wiggle the part around on the sink to force out any trapped air and to make the layer of grease is as thin as possible. You should be able to feel the sink grab the part as you move it. It does not want to "float" on the grease!

THERMAL DESIGN

The VRF151's maximum junction temperature rating is 200 °C, but device lifetime is seriously degraded at this temperature. Industry design standards typically aim for 150 °C absolute maximum, and typical operating temperature about 130 °C. The heat sink chosen for this amplifier is a 7-inch length of aluminum extrusion 3.25 inches wide and 1 inch high with nine longitudinal fins. This provides a lot of surface area, but in order to remove the 250 W of heat dissipated when the amplifier is running at maximum CW power, the heat sink must have air forced through it. It is impossible to dissipate this much heat by convection alone. The forced air cooling keeps the junction temperature below 150 °C at all times. The combination of a fairly large heat sink area and a relatively small fan allows the amplifier to "keep its cool" without making a lot of noise.

A single 3-inch 24-V dc fan (47 CFM) provides pressurized air across the sink. The closed chassis forms a pressurized plenum. The air comes in through the fan but can only leave the chassis after traveling down the length of the heat sink and out the 3 × 1 inch opening in the rear panel. Ducting the air this way makes very effective use of the fin area on the heat sink.

The original prototype of the amplifier had two smaller cooling fans mounted on the rear panel. When the amplifier was put in its place on the operating table, it was overheating because both the air intake and exhaust were on the same end of the chassis. It was sucking in its own hot air. A new chassis was bent up and this time the fans were mounted on one side of the cover. This eliminated the self-heating problem and allowed a much cleaner layout of the rear panel. It provided enough space to mount the fuse holder outside rather than inside. The final version uses a single high capacity 24-V fan. A 24-V 5-W Zener diode is placed in series to run it from the 48-V supply. A 48-V fan would be a better choice, but a suitable unit was not available from distributor stock, only by special order.

Note of caution: The amplifier cannot be run at power without the top cover in place because the chassis must be pressurized in order to force the cooling air thorough the fins of the heat sink.

The fan runs at full speed all the time. It would be more "operator friendly" to run the fan at a very low speed to start off with and use the temperature sensing circuit to increase the fan speed when the heat sink reaches a moderate temperature, say 100 °F. A second threshold point would run the fan at full speed when the sink temperature reaches 120 °F. This is a project for the next version. As it is, the fan does not require any controls. It will continue to run after the supply is switched off until it bleeds down the filter capacitors.

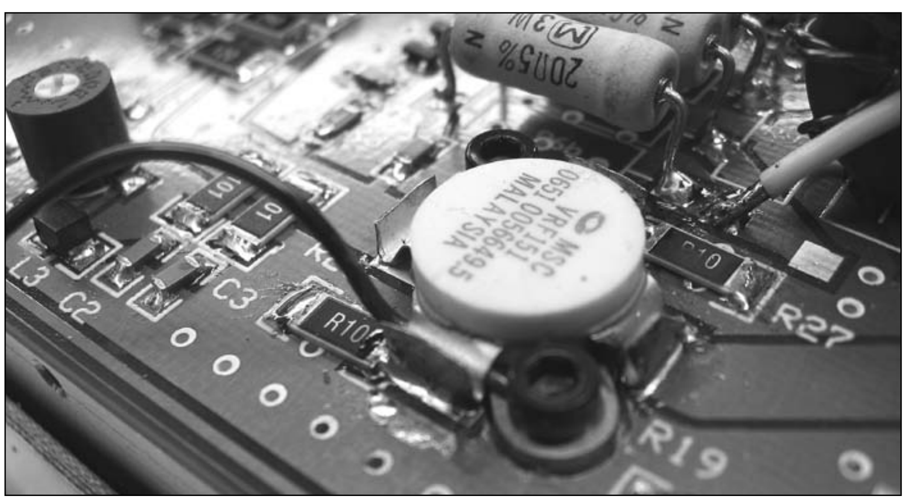


Fig 17.53 — It's important to make sure that the VRF151 power transistors mount flush to the heat sink, without gaps or air pockets that would impede heat transfer (see text for details). Once the transistors are mounted, a piece of wire is used as a lead forming device to bend the transistor leads down to the PC board pads for soldering.

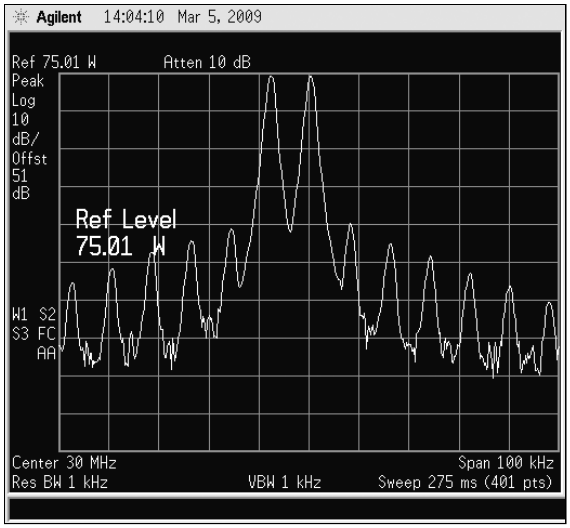


Fig 17.55 — Output of the 250 W amplifier during two-tone IMD testing. All IMD products are better than 38 dB down from either tone.

17.11.6 Metering and Calibration

The cost of quality moving-needle meters has gone beyond reasonable, in part because digital panel meters have become more popular. However, you need to be able to assess the health of the amplifier at a glance and digital meters are not good for this. LED bar graphs are very inexpensive and are used for metering in this amplifier. They are easy to read and because of their low cost we can have one for each parameter being measured. Output power is displayed on a 20 LED string (made from two 10 LED bars). There is a "hang" built in so the SSB voice peaks can be seen easily. The drain current and reflected power are each shown on 10 LED bars. The output of the thermistor on the heatsink is also available for display by changing a jumper or incorporating a selector switch.

Absolute meter calibration is not so important in an amplifier like this. As long as none of the displays overflow in normal operation, they will serve as a way to easily monitor the operation of the amplifier.

Calibration of the meters will require using external standards. An ammeter in series with the power supply will be enough to calibrate the current meter. It should be set to display 15 A at full scale. R1 is adjusted to light bar #10 at 15 A. It would work out exactly right if all the resistors were 1% values. Forward power similarly requires a calibrated power meter on the output. R33 should be adjusted for 250 W when 18 bars of the 20 are lit. The top two LEDs are red and the others green so the operator can readily tell when the drive is too high.

Calibrating the reflected power is a little more difficult. The simplest method requires a calibrated external wattmeter. Run 30 W of RF power from another source backward through the amplifier, from antenna to input, *with the operate switch in the standby position*. Set R34 so the SWR LED trips at this point. This corresponds to an SWR of 2:1 at 250 W output. The amplifier should be stable and totally reliable up to this mismatch. If your load has a higher VSWR than 2:1, you can back off on the drive and continue to operate. As long as the reflected power is less than 30 W, the amplifier is happy.

Protecting the amplifier from damage requires a combination of careful attention to the operating conditions and reliance on the automatic limits in the control system.

17.11.7 Chassis

As seen in **Fig 17.54**, the chassis is a double clamshell configuration made from aluminum sheet stock. The top cover is 0.031-inch thick and the chassis bottom is 0.062-inch thick. This provides both adequate strength and workability with simple hand tools. Aluminum 12 × 24 inch sheets are available from several suppliers. They were sheared to size and bent into U-shaped parts at a local sheet metal shop.

The power supply and the fans are mounted on the cover. It overhangs the chassis base by 1⁄16 inch on all sides. The top is attached by 4-40 threaded L-brackets, DigiKey part 612K-ND, that are riveted to the main chassis. A drawing of the chassis layout and front and rear panels is included on the *Handbook* CD.

17.11.8 Performance

When the design was nearly complete, it became apparent that things were not coming out right. There was plenty of cooling, but the available power supply is a 500-W 48-V unit that goes into sudden current limiting when the load exceeds 10.4 A. The low-pass filters have about 0.2 dB loss, or about 5%. If the output was increased to make up for the filter loss, the power supply would limit on some bands causing severe distortion on SSB and CW.

The maximum for the PA design itself is 300 W. Increasing its output past 300 W to make up for filter loss quickly degrades the IMD performance. It needs some headroom. So, in very un-amateur fashion, this amplifier is conservatively rated at 250 W output. This provides a clean signal and plenty of margin for wrong antenna selection, disconnected feed lines, and all the other things that can kill amplifiers that are run too close to their limit.

The PA will provide 250 W PEP for sideband or PSK and 250 W CW. The design goal for this PA was to make it reliable and at least as good as any competitive transceiver. The harmonics are –60 dB on HF and –70 dB on 6 meters. Transmit IMD is >38 dB down from either tone as shown in **Fig 17.55**.

Parasitics are not usually a problem in broadband amplifiers because of the feedback used. The prototype was tested into a 3:1 SWR load at all phase angles without breaking into parasitic oscillation anywhere.

This design project was undertaken to provide a design example for a practical solid-state amplifier for the 2010 *Handbook*. It is a continuing project.

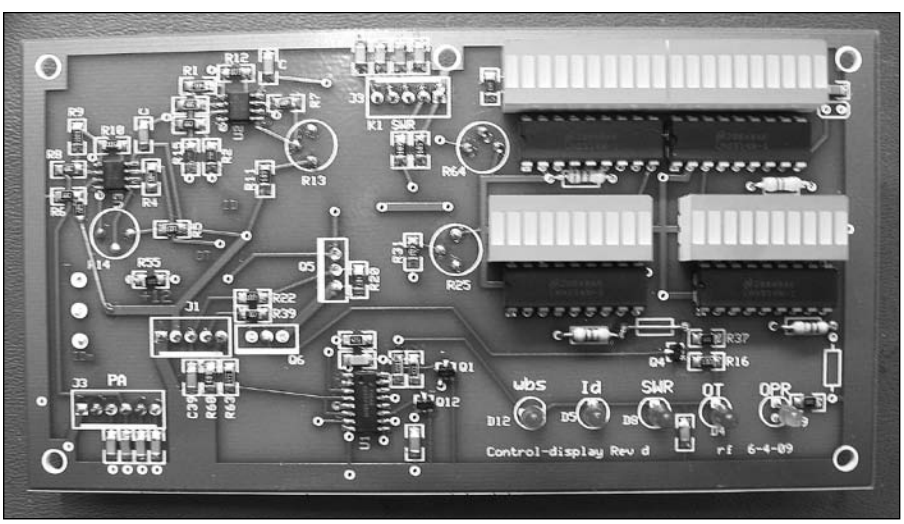


Fig 17.54 — The control and display circuitry mounts on a PC board that attaches to the front panel. LED bar graphs show forward and reflected power and current.

17.12 Tube Amplifier Projects

Project: 3CX1500D7 RF Linear Amplifier

The following describes a 10-to-160-meter RF linear amplifier that uses the new compact Eimac 3CX1500D7 metal ceramic triode. It was designed and constructed by Jerry Pittenger, K8RA.

The amplifier features instant-on operation and provides a solid 1500 W RF output with less than 100 W drive. Specifications for this rugged tube include 1500-W anode dissipation, 50-W grid dissipation and plate voltages up to 6000 V. A matching 4000-V power supply is included. The amplifier can be easily duplicated and provides full output in key-down service with no time constraints in any mode. **Fig 17.56** shows the RF deck and power supply cabinets.

DESIGN OVERVIEW

The Eimac 3CX1500D7 was designed as a compact, but heavy-duty, alternative to the popular lineup of a pair of 3-500Z tubes. It has a 5-V/30-A filament and a maximum plate dissipation of 1500 W, compared to the 1000-W dissipation for a pair of 3-500Zs. The 3CX1500D7 uses the popular Eimac SK410 socket and requires forced air through the anode for cooling. The amplifier uses a conventional grounded-grid design with an adjustable grid-trip protection circuit. See the RF Deck schematic in **Fig 17.57**.

Output impedance matching is accomplished using a pi-L tank circuit for good harmonic suppression. The 10 to 40-meter coils are hand wound from copper tubing, and they are silver plated for efficiency. Toroids are used for the 80- and 160-meter coils for compactness. The amplifier incorporates a heavy-duty shorting-type band-switch. Vacuum variable capacitors are used for pi-L tuning and loading.

A unique feature of this amplifier is the use of a commercial computer-controlled impedance-matching module at the input. This greatly simplifies the amplifier design by eliminating the need for complex ganged switches and sometimes frustrating setup adjustments. The AT-100AMP module kit available from W4RT Electronics (**www.w4rt.com**) is an acceptable tuning unit.

An adjustable ALC circuit is also included to control excess drive power. The amplifier metering circuits allow simultaneous monitoring of plate current, grid current, and a choice of RF output, plate voltage or filament voltage.

The blower was sized to allow full 1500-W plate dissipation (65 cfm at 0.45 inches H_2O hydrostatic backpressure). The design provides for blower mounting on the rear of the RF deck or optionally in a remote location

Fig 17.56—At the top, front panel view of RF Deck and Power Supply for 3CX1500D7 amplifier. At bottom, rear view of RF Deck and Power Supply.

to reduce ambient blower noise in the shack. The flange on the socket for connecting an air hose was ground off for better air flow. (This is not necessary.).

The power supply is built in a separate cabinet with casters and is connected to the RF deck using a 6-conductor control cable, with a separate high voltage (HV) cable. The power transformer has multiple primary taps (220/230/240 V ac) and multiple secondary taps (2300/2700/3100 V ac). No-load HV ranges can be selected from 3200 to 4600 V dc using different primary-secondary combinations. The amplifier is designed to run at 4000 V dc under load to maintain a reasonable plate resistance and component size. A step-start circuit is included to protect against current surge at turn on that can damage the diode bank. The power supply schematic is shown in **Fig 17.58** and a photo of the inside of the power supply is shown in **Fig 17.59**.

Both +12-V and +24-V regulated power supplies are included in the power supply. The +12 V is required for the computer-controlled input network and +24 V is needed for the output vacuum relay. The input and output relays are time sequenced to avoid amplifier drive without a 50-Ω load. Relay actuation from the exciter uses a low-voltage/low-current circuit to accommodate the amplifier switching constraints imposed by many new solid-state radios.

Much thought was put into the physical appearance of the amplifier. The goal was to obtain a unit that looks commercial and that would look good sitting on the operating table. To accomplish the desired look, commercial cabinets were used. Not only does this help obtain a professional look but it eliminates a large amount of the metal work required in construction. Careful attention was taken making custom meter scales and cabinet labeling. The results are evident in the pictures provided.

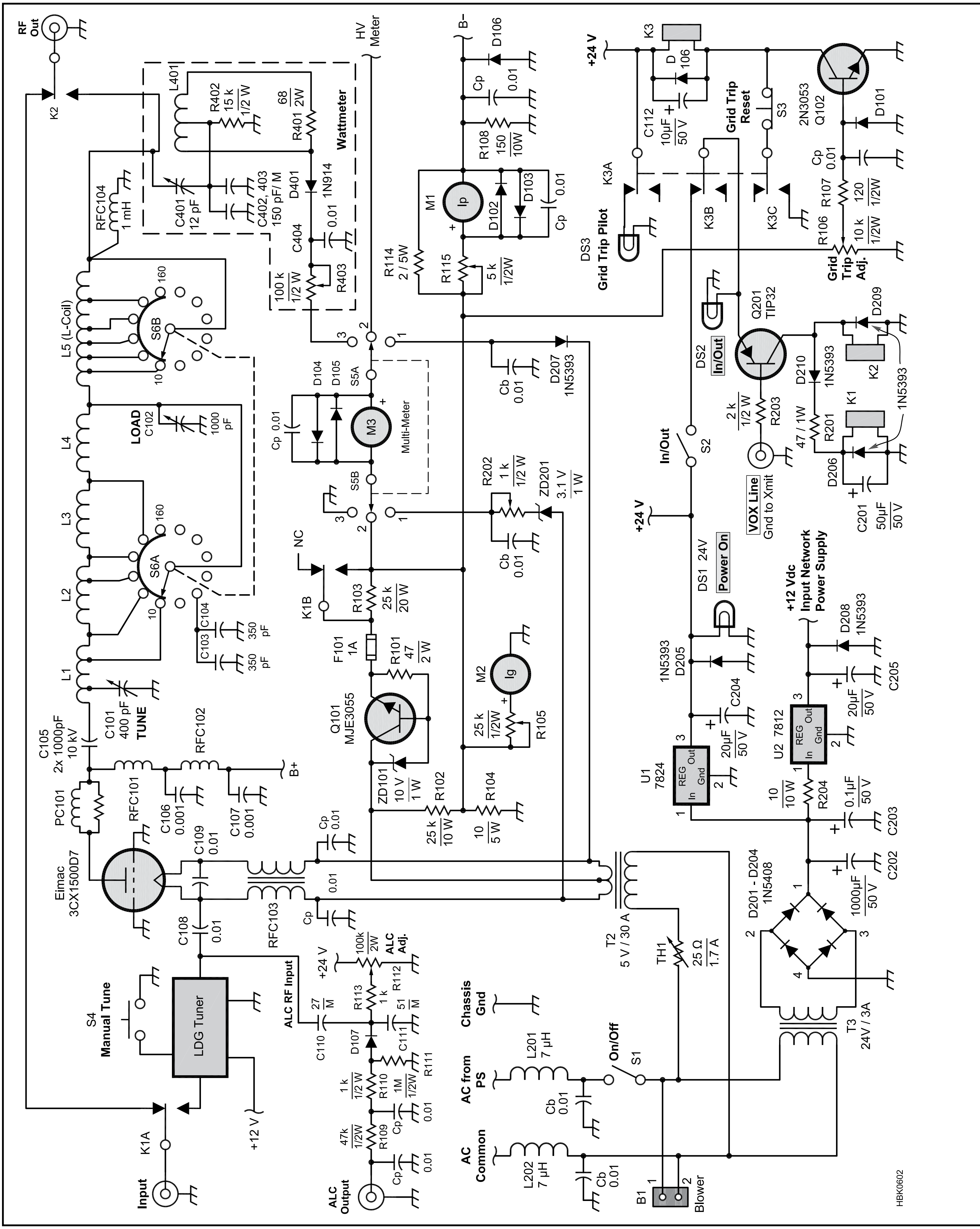
GENERAL CONSTRUCTION NOTES

The amplifier was constructed using basic shop tools and does not require access to a sophisticated metal shop or electronics test bench. Basic tools included a band saw, a jig saw capable of cutting thin aluminum sheet, a drill press and common hand tools. Some skill in using tools is needed to obtain good results and insure safety, but most people can accomplish this project with careful planning and diligence.

Metal work can be a laborious activity. Building cabinets is an art within itself. This part of the project can be greatly simplified by using commercial cabinets. However, commercial cabinets are expensive (~$250 each) and could be a place where some dollars could be saved.

The amplifier is built in modules. This breaks the project into logical steps and facilitates testing the circuits along the way. For example, modules include the HV power supply, LV power supply, input network, control circuits, tank circuit and wattmeter. Each module can be tested prior to being integrated into the amplifier.

The project also made extensive use of computer tools in the design stage. The basic layout of all major components was done using the *Visio* diagramming software package. The printed-circuit boards



RF Out
K2
HV Meter
B–
D106
+24 V
K3
D 106
2N3053 Q102
L401
R402 15 k 1/2 W
68 2W R401
Wattmeter
Cp 0.01
R108 150 10W
C112 10µF 50 V
Grid Trip Reset
S3
D101
Cp 0.01
RFC104 1 mH
C401 12 pF
C402, 403 150 pF/ M
D401 1N914
C404 0.01
M1
Ip
D102
D103
Cp 0.01
K3A
K3B
K3C
R107 120 1/2W
R106 10 k 1/2W
Grid Trip Adj.
DS3 Grid Trip Pilot
100 k 1/2 W R403
R114 2 / 5W
R115 5 k 1/2W
L5 (L-Coil)
S6B
160
10
Q201 TIP32
D209
DS2 In/Out
D210 1N5393
K2
1N5393
S5A
D104 D105
M3 Multi-Meter
Cp 0.01
D207 1N5393
Cb 0.01
2 k 1/2 W R203
47 / 1W R201
K1
LOAD C102 1000 pF
L4
In/Out S2
VOX Line Gnd to Xmit
D206
C201 50µF 50 V
L3
S5B
R202 1 k 1/2 W
ZD201 3.1 V 1 W
+24 V
DS1 24V Power On
+12 Vdc Input Network Power Supply
L2
S6A
NC
K1B
R103 25 k 20 W
C103 C104 350 pF 350 pF
L1
F101 1A
R101 47 2 W
M2 Ig
1N5393 D205
D208 1N5393
C105 2x 1000pF 10 kV
C101 400 pF TUNE
RFC102
Q101 MJE3055
25 k 1/2W R105
C204 20µF 50 V
C205 20µF 50 V
U1 7824
REG In Out Gnd
U2 7812
B+
PC101
RFC101
ZD101 10 V 1 W
R102 25 k 10 W
R104 10 5 W
R204 10 10 W
0.1µF 50 V C203
Eimac 3CX1500D7
C106 0.001
C109 0.01
C107 0.001
Cp 0.01
T2 5 V / 30 A
D201 - D204 1N5408
C202 1000µF 50 V
C108 0.01
RFC103
Cp 0.01
TH1 25 Ω 1.7 A
S4 Manual Tune
LDG Tuner
ALC RF Input
+24 V
100k 2W ALC Adj.
R113 1 k
R112
C110 27 M
C111 51 M
D107
Chassis Gnd
T3 24V / 3A
K1A
+12 V
R110 1 k 1/2 W
R111 1M 1/2W
AC from PS
L201 7 µH
On/Off S1
Cb 0.01
Input
R109 47k 1/2W
Cp 0.01
ALC Output
AC Common
L202 7 µH
Cb 0.01
B1
Blower
HBK0602

Fig 17.57 — Schematic of the RF deck and control circuitry.

B1 — Dayton 4C763 squirrel-cage blower.
Cb, Cp — 0.01 μF, 1 kV disc ceramic.
C101 — 400 pF, 10 kV Jennings vacuum variable, UCSL-400.
C102—1000 pF, 5 kV Jennings vacuum variable, UCSL-1000.
C103, C104—350 pF, 5 kV ceramic doorknob.
C105—two parallel 1000 pF, 10 kV ceramic doorknob capacitors (Ukrainian mfg).
C106, C107—0.001 μF, 7.5 kV disc ceramic.
C108, C109—0.01 μF, 3 kV transmitting mica (1 kV disc ceramics can be used).
C401—12 pF piston trimmer.
C403, C403—150 pF silver mica.
D101, D107, D205-D210—1N5393 (200 V, 1.5 A).
D102-D106, D201-D204—1N5408 (1 kV, 3 A).
K1—4PDT, 24 V dc KHP style (gold contacts).
K2—SPST vacuum relay, Kilovac H8/S4.
K3—4PDT, 24 V dc KHP style (gold contacts).
L1-L5—See Table 17.8.
L201, L202—Line chokes, 7 μH.
L401—24 t #22 enamel wire, center tapped on T50-6 core.
M1-M3—Simpson Designer Series, Model 523, 1 mA movement.
PC101—2 t ¾-inch diameter × 2-inch long, ½-inch brass strap with two 150 Ω, 2 W non-inductive carbon resistors in parallel.
Q101—MJE3055 TO-220 case on heat sink.
Q102—2N3053 TO-18 case.
R103—25 kΩ, 25 W wire-wound.
R104—10 Ω, 5 W.
R108—150 Ω, 10 W wire-wound.
R112—100 kΩ, 2 W potentiometer
R403—100 kΩ, 0.5 W trim pot.
RFC101—90 μH, 3 A Plate Choke, Peter W. Dahl p/n CKRF000100, (see text).
RFC102—14 t #18 enamel wire wound on 100 Ω, 2 W resistor.
RFC103—Bifilar 30 A filament choke, Peter W. Dahl p/n CKRF000080, (see text).
RFC104—1 mH, 300 mA RF choke.
S1-S2—Alco 164TL5 DPDT switch (only SPST contacts are used), www.alliedelec.com/.
S3-S4—Alco 164TL2 momentary DPDT (only SPST contacts are used; S3 wired as normally closed, S4 as normally open), www.alliedelec.com/.
S5—2 pole, 3 position rotary switch.
S6—RadioSwitch model 86, double-pole 12-position (30° indexing) with 6-finger wiper on each deck, p/n R862R1130001, www.multi-tech-industries.com.
T2—5 V, 30 A center-tapped transformer, Peter W. Dahl EI-150 × 1.5 core, primary 115/230 V ac, (see text).
TH1—Thermistor, Thermometrics CL-200 (Mouser 527-CL200).
ZD101—10 V, 1 W Zener 1N4740A.
ZD201—3.1 V, 1 W Zener.
Other parts:
Cabinet—Buckeye Shapeform DSC-1054-16 (10×17×16-inch H×W×D), www.buckeyeshapeform.com.
Chimney (Teflon)—A. Howell, KB8JCY, PO Box 5842, Youngstown, OH 44504.
LDG Tuner—AT-100AMP autotuner, see text.
Tube socket—Eimac SK-410.

were designed using a free layout program called *ExpressPCB* (**www.expresspcb.com**). Masks were developed and the iron-on transfer technique was used to transfer the traces to copper-clad board. The boards were then etched with excellent results. The layout underneath the RF Deck is shown in **Fig 17.60A** and the top side of the RF Deck is shown in Fig 17.60B.

Meter scales were made using an excel-lent piece of software called *Meter* by Jim Tonne, W4ENE (**www.tonnesoftware.com**). Also, K8RA wrote an *Excel* spread-sheet to calculate the pi-L tank parameters. A copy of the spreadsheet, *Meter Basic* software and *Express PCB* files for the PC boards are all included on the CD-ROM that accompanies this book.

Although using computer tools simplifies the design step, all design work can be done without the use of a computer. Be creative and use the tools and resources at hand! There are many different ways to construct this design. The key secret is diligence and not compromising until it is done right. Note that the tank coils in this amplifier were wound at least three times, the inside side panels were cut twice and many printed circuit boards ended in the trash before acceptable boards were fabricated.

Since this project was built, Peter W. Dahl has discontinued business, Dahl transformers are now available from Harbach Electronics (**www.harbachelectronics.com**). Contact Harbach to cross-reference the Peter Dahl part numbers in the parts list for T1, T2 and RFC103 with current Harbach stock or equivalent designs.

CABINET METAL WORK

By purchasing commercial cabinets, metal work required was minimized but not eliminated. The power supply components are very heavy. The transformer weighs about 70 pounds by itself. Therefore the base plate of the power supply cabinet needed to be reinforced. The original base plate for the cabinet was not used. One-eighth-inch plate was purchased from a local aluminum scrap company. Two pieces were sandwiched to provide a ¼-inch plate. Of course ¼-inch material could have been used but it was not available at the time of purchase.

The plate can be cut on a metal band saw using a guide or on a radial arm saw. Metal blades are readily available from Sears for both saws. If using a radial arm saw, multiple passes are required, lowering the blade slightly with each pass. Be sure to wear eye protection because the metal chips fly. The edges were then cleaned and straightened using a 4-inch belt grinder. If a belt sander is not available, a large file will work.

The two metal plates were held together with the mounting bolts on the four casters. The power supply base plate exactly matches the original base plate and fastened to the cabinet using the original tapped screw holes. All the heavy components are mounted on the base plate. The power supply must always be handled by lifting the base plate, since the cabinet does not have the structural integrity to bear the weight by itself.

The RF deck needed both a chassis plate and a front sub-panel. See Fig 17.60B. The sub-panel is used to mount the load and tune capacitors, the bandswitch and also provides RF shielding for the meters. Side plates were needed because of the cabinet configuration. The side plates, chassis plate and sub-panel all use 1⁄16-inch aluminum plate. After the side plates are cut and mounted to the cabinet sides, the chassis plate and front sub-panel are mounted using ½-inch aluminum angle to join the edges.

Cutting holes can often be a challenge. If a drill bit is the correct size, drilling a hole is easy, of course. But large-size round holes and square holes can be a challenge. This was especially true in this project since the front and rear panels are ⅛-inch aluminum plate.

The large meter holes can be cut using a hole saw on a drill press. For odd sizes, a "fly cutter" can be used. Fly cutters are available from Sears but a special warning is in order. These devices work well but are extremely dangerous. Make sure the cutting bit andthe placement into the drill chuck are secure.

Large square holes are required for the turn counters. Mark the square hole to be cut. Drill a hole in each corner. The hole must be at least the size of the saw blade if a jigsaw is used to finish the hole. Note that the jigsaw must have a removable straight blade. If a metal-cutting jigsaw is not available, a series of small holes can be drilled in a straight line on all four sides and the edges smoothed with a file. Almost any hole can be custom cut by making a hole the approximate size and finishing it to the exact dimension with a file. It is slow and laborious but it works.When using a file on panels, be very careful that the file does not slip out of the hole and put an undesired scratch in the panel!

Once panel holes are cut, carefully label the panels before mounting the components. Dry transfers are used on both the power supply and the RF deck. Dry transfers of all sizes and fonts are available at graphics art stores and hobby shops. The author has found that hobby shops carry an excellent selection of dry transfers in the model railroad section.

RF DECK CONTROL CIRCUITS

The control circuits in the amplifier are not complex due to the simplicity of the grounded-grid design and the instant-on capability of the 3CX1500D7 tube. 120 V ac is routed from the HV power supply to the RF Deck in the 6-conductor control cable. When the on/off switch (S1) is pressed,

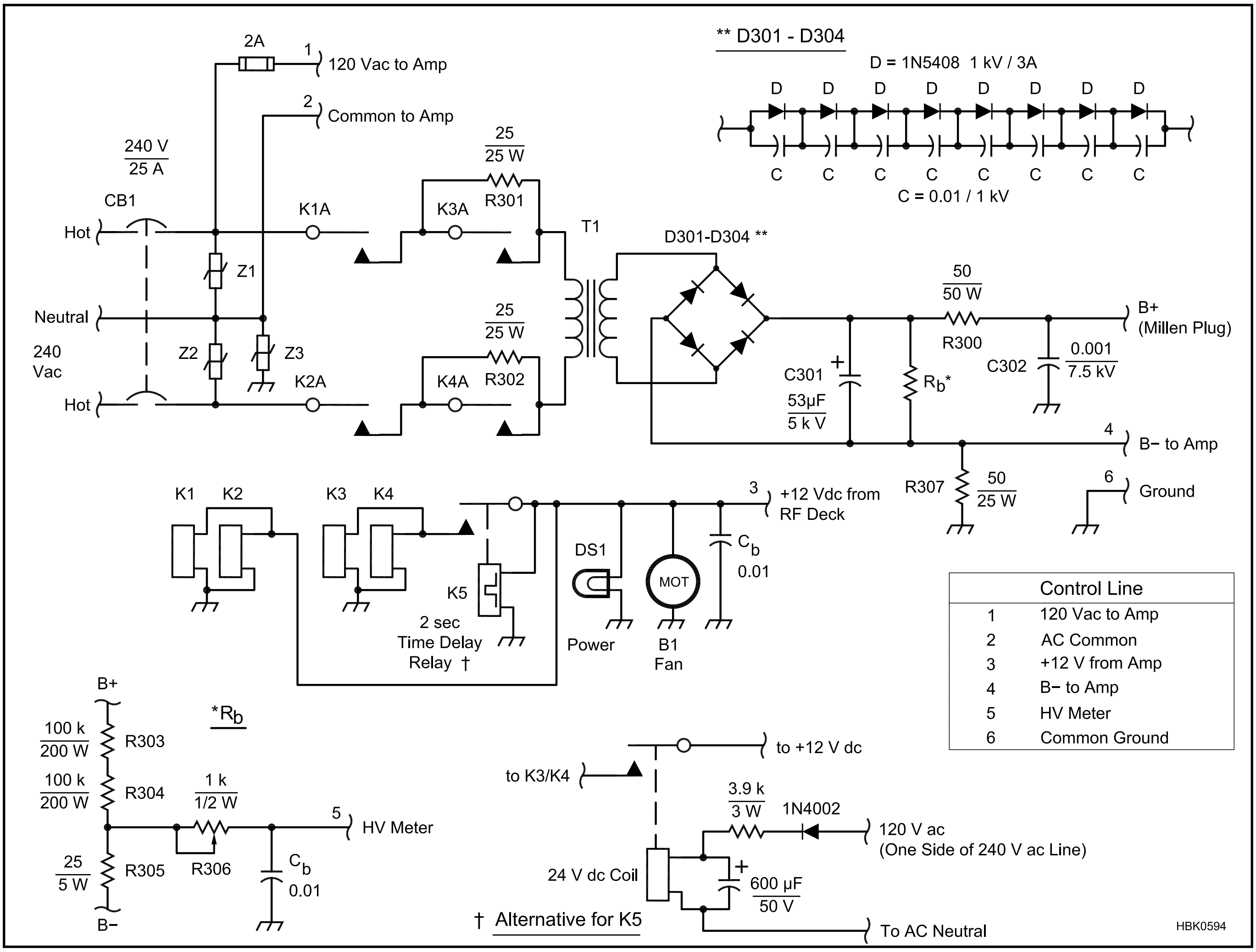


Fig 17.58—Schematic for power supply for 3CX1500D7 amplifier.

B1—12 V dc brushless fan, 2¼ inch (Mouser 432-31432).
DS1—12 V dc pilot lamp (Alco 164-TZ).
Cb—0.01 µF, 1 kV disc ceramic bypass capacitor.
C301—53 µF, 5 kV oil-filled.
CB1—2 pole 25 A, 240 V circuit breaker.
K1-K4—SPST solid-state relay 240 V ac, 25 A line voltage with 12 V dc input (the author used surplus Crydom relays but a readily available substitute is the Tyco/P&B SSR-240D25R).
K5—2 second trimpot time delay relay (surplus Bourns 3900H-1-125 or equiv; or FTD-12N03 3 second glass timer relay from Surplus Sales of Nebraska). Note: This part may be difficult to find. You can build the equivalent from a 24 V dc SPST relay, diode, resistor and capacitor as shown in the drawing inset. This technique is borrowed from the K6GT HV supply shown later in this chapter (Fig 17.63).
R303, R304—100 kΩ, 200 W wirewound resistor).
T1—Plate transformer, primary 220/230/240 V; secondary 2300/2700/3100 V at 1.5 A CCS (Peter W. Dahl Co P/N Pittenger PT-3100, see text).
Z1-Z3—130 V MOV.
Cabinet—Buckeye Shapeform DSC-1204-16 (12×18×16 inches HWD), www.buckeyeshapeform.com.

120 V ac is sent to the primary of the low voltage transformer (T3) and the filament transformer (T2). The surge current to the filament of the tube is suppressed by the thermistor (TH1) in one leg of the filament transformer primary. These are excellent current limiting devices that have a resistance of approximately 25 Ω cold but decrease to less then 1 Ω as they heat. Keep the thermistor in open air away from other components since they are designed to run hot.

The low voltage supply provides regulated +12 V dc and +28 V dc. The voltages are regulated using simple three-terminal regulators. Pilot lights are included in each push button switch, S1-S4, and a power indicator on the HV power supply. When the low-voltage power supply first comes on, +12 V dc is directed through the control cable back to the HV supply. High voltage is applied immediately to the instant-on tube. Therefore the amplifier is turned on and ready to go instantly—You don't have to listen to your friends working that rare one for three or four tense minutes while you wait for your amplifier to time in!

The amplifier is switched in and out of the circuit using a 4PDT KHP style relay (K1) for the input and a SPDT vacuum relay (K2) for the output. It is important to select the timing constants for the input relay (C201 and R201) so the input relay closes a few milliseconds after the vacuum relay. This avoids hot switching the output, which could fuse the vacuum-relay contacts. This is a balancing act since the brief time the input relay is open will present an open circuit to the exciter. Many modern radios now have exciter-timing circuits that close the amplifier relay circuit a few milliseconds before RF is transmitted.

It is recommended that timing components for the input relay be located in a place where they can be easily changed. Another approach

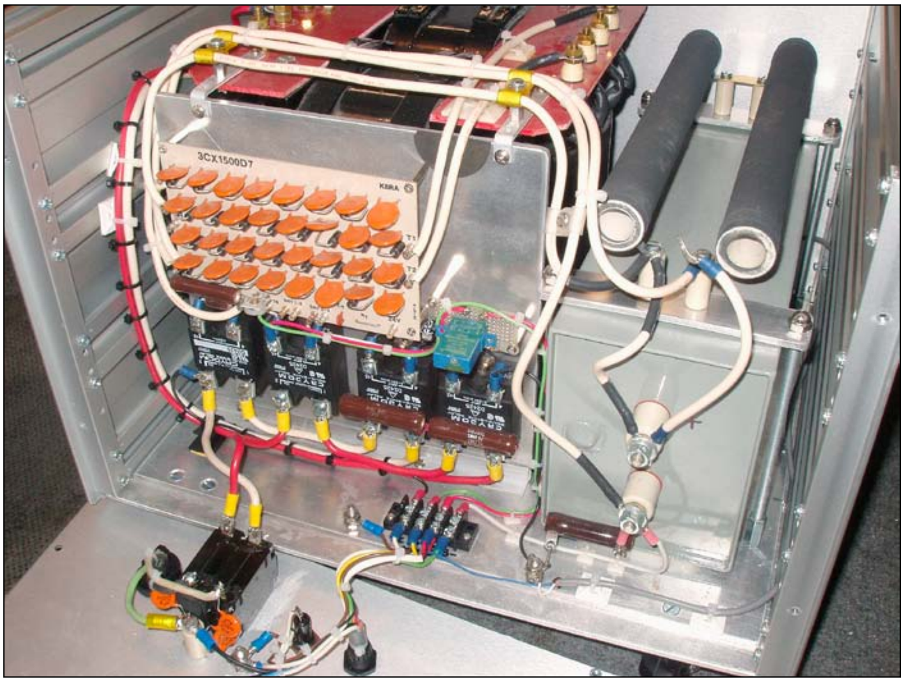


Fig 17.59—Inside view of the power supply, showing rectifier stack, control relays and HV filter capacitor with bleeder resistors. The heavy-duty high voltage transformer is at the upper left in this photo.

is to build a breadboard circuit that feeds the relay coils in parallel but places the contacts in series. Feed a low voltage through the contacts of the two relays and monitor the timing with a dual-trace oscilloscope. This technique allows precise timing of contact closure as the two relays work together. Note that different relays will need different timing-circuit component values. A set of contacts on input relay, K1, is used to short across bias resistor, R103. The resistor biases the tube to cutoff in standby.

Approximately +10 V bias is provided to the center tap of the filament transformer to limit the idle current of the tube to approximately 125 mA. The bias is developed using the three components D101, R101 and Q101. These components could be replaced with a single 10-V/50-W Zener diode. However, 50-W zeners are expensive and they are difficult to obtain. Using the circuit shown, the bias is provided by a common NPN transistor (Q101) and a one-watt zener (D101) you can obtain from RadioShack.

TUBE PROTECTION CIRCUIT

The main protection for the tube is a plate-current surge resistor and a grid-trip circuit. The current surge resistor (R308, 50 Ω/50 W) is in series with the B+ line and acts as a fuse should excessive current be drawn from the HV power supply. Ohm's law says that up to 1-A plate current can be drawn through the resistor and still stay within the 50-W rating of R308. However, let's assume a problem occurs and 5 A flows through the resistor. Resistor R3 must now dissipate 1250 W. The resistor will quickly fail and will shut down the HV to the 3CX1500D7 tube.

Q102 is a grid-trip circuit that snaps the amplifier offline if the grid current exceeds 400 mA. The grid current is drawn through the 10-Ω resistor (R104) connected between the B– line and chassis ground. The current creates a voltage across R104 that is fed to the grid-trip adjustment potentiometer, R106. Q102 is turned on when the base voltage reaches 0.6 V and actuates the grid-trip relay K3. K3 contacts break the +28 V dc input and output relay lines (K3B), locks the relay closed (K3C) and extinguishes the pilot bulb (K3A) of the GRID-TRIP RESET normally closed push-button switch (S3) located on the front panel. Pushing the GRID TRIP RESET switch (S3) breaks the current path for the grid-trip relay K3 and resets the relay. The reason the grid trip was actuated should be determined prior to attempting to use the amplifier again. Usually, this is caused by improper setting of the load capacitor or transmitting into the wrong antenna.

INPUT NETWORK

As mentioned before, this amplifier uses a unique concept for the input-matching network, getting rid of a switched network mechanically ganged to the main band-switch. Not only can such a switching arrangement be awkward mechanically, but obtaining a reasonable network Q and a low SWR over an entire band can be difficult.

Thus the author decided to use a commercial automatic tuner integrated into the RF deck (see Fig 17.60A). The tuner is a kit from W4RT Electronics based on the AT-100 autotuner by LDG Electronics (discontinued as standalone equipment). The kit is supplied without the enclosure and switches. This application is simple but elegant. The unit automatically initiates a retune if the input SWR exceeds approximately 1.5:1. The tuning cycle takes three to five seconds to execute. But retuning does not happen often because the tuner has over 4000 memories and remembers the settings for different frequency ranges. As the amplifier is used on each band, the tuner *learns* and stores settings into the memory. When switching bands, it only takes milliseconds to retrieve the data from memory and actuate the correct tuner relays.

Integration of the tuning network requires connections for RF input and output, +12 V and ground. RF input goes to the center of T1 and ground goes to J2 (clearly marked on the board). RF output goes to J3 and ground goes to J6. The +12-V dc connection is the larger of the three holes at J10 (next to L10). The other two holes are grounds for dc connections.

A momentary contact switch (S4) is mounted on the front panel to provide manual control of the tuner. A normally open contact on S4 is connected to the input pin J9 (next to L12) and ground. (The pin is marked as the ring for the connector that is not installed.) The correct hole is on the C56 side. If the switch is pushed for less than ½ second, the tuner alternates between bypass and in-line modes. If S4 is pressed between 0.5 to 2.5 seconds, it does a memory tune from the stored data tables. If S4 is pressed for more than 2.5 seconds with RF applied, it skips the memory access, retunes and stores the new settings into the memory table. The manual retune function is seldom, if ever, used.

The tuner works perfectly and it really simplified the input-network design and construction. The SWR never exceeds 1.5:1 (typically it is 1.2:1).

PI-L NETWORK

A pi-L network is used to insure good harmonic attenuation. The pi-L circuit is actually a pi-network, followed by an L-network that provides additional harmonic attenuation. The L-section transforms the load of 50 Ω up to an intermediate resistance of 300 Ω. The pi section then transforms 300 Ω up to the desired plate load resistance of 3100 Ω. The plate load is calculated using the following formula:

$$R_L = \frac{E_p}{1.7 \times I_p} = \frac{4000}{1.7 \times 0.750} = 3137\Omega$$

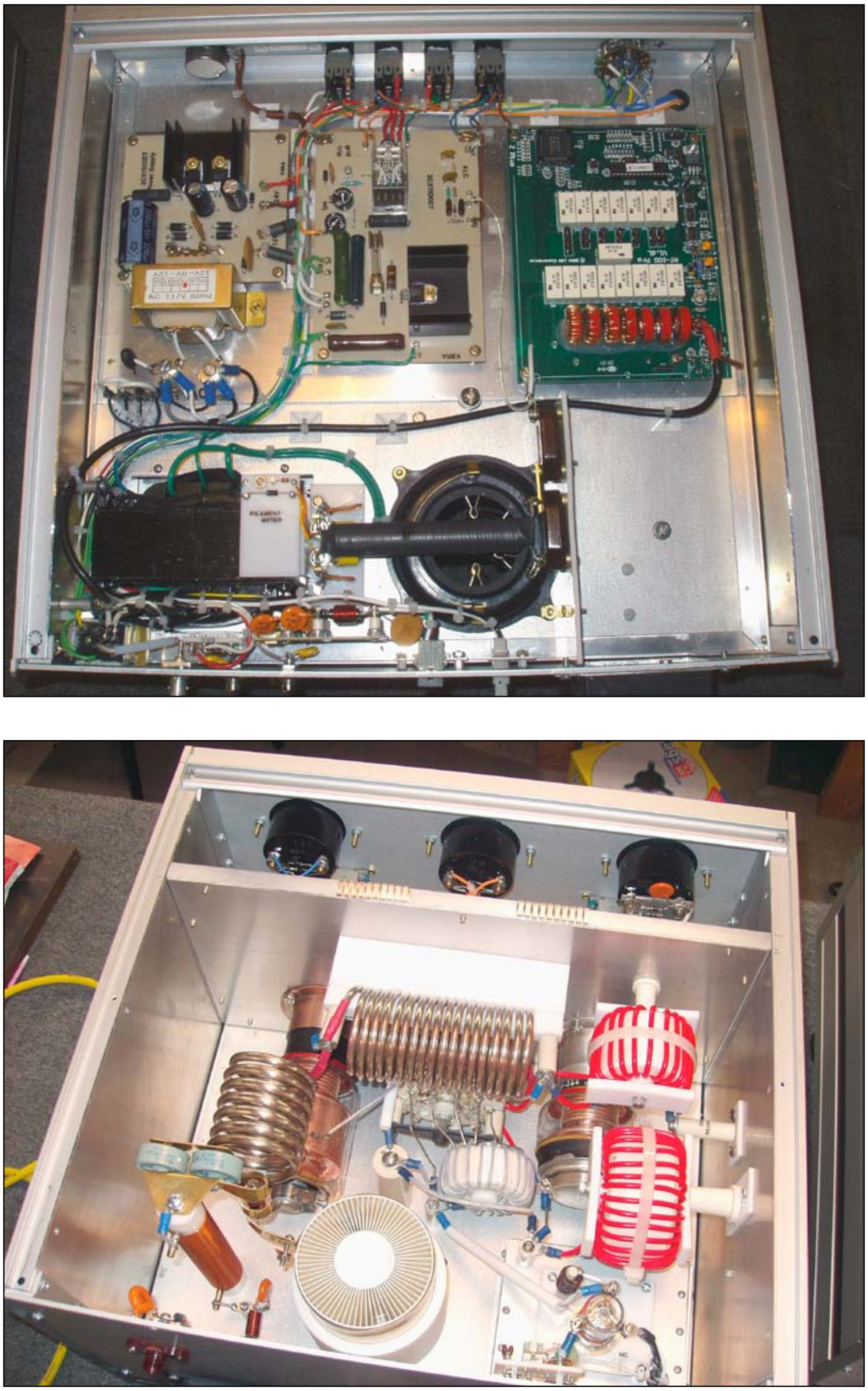


Fig 17.60—At the top, under the chassis of the RF Deck. The autotuner used as the input network for this amplifier is at the upper right. At bottom, view of the Pi-L output network in the RF Deck.

A nominal Q of 12 is used for the network. But as with most RF amplifiers, the capacitance needed for the higher-frequency bands is less than is physically possible using variable capacitors. For Q = 12, the tune capacitance (C101) for 10 meters is 14 pF. Using a vacuum variable capacitor for C1 helps because the minimum achievable capacitance (12 pF) is substantially less than with an air variable. But the tune capacitance is the sum of variable C101, 7.1 pF for the output capacitance of the 3CX1500D7 tube and any stray capacitance resulting from the physical layout of the amplifier.

The minimum obtainable capacitance is thus on the order of 30 pF, which yields a higher value of loaded Q than optimum. The solution is two fold. First, connect the plate-tune capacitor (C101) one turn into the 10-meter coil. This actually forms an L-pi-L circuit. Second, accept a higher value of loaded Q so that the variable capacitor can still be tuned. **Table 17.8** shows the loaded Q finally used for each band setting. The disadvantages of higher loaded Qs are high circulating currents in the tank circuit and the need to retune during excursions across the higher-frequency bands. This amplifier works fine on all bands, delivering a solid 1500 W output even on 10 meters.

Another pi-L tank circuit design constraint in this amplifier is the bandswitch. Many amplifiers use a single-pole, 12-position, non-shorting switch. Although this type of switch is easier to find, it can be problematic because high voltages are generated that could result in arcing in the bandswitch—usually from the wiper to the high frequency taps. You should use a switch with a multiple-finger wiper (see Fig 17.57) that shorts out lower-frequency coil taps not being used. For example, when the amplifier is used on 20 meters, the 40-, 80- and 160-meter taps are shorted to the wiper.

However, shorting switches only allow for six connections with 30° indexing. The common shorting wiper consumes 180° of switch deck on 160 meters. This results in having to design the 10/12-meter and 15/17-meter bands to use single taps for each frequency pair. Again, this is accomplished by adjusting the loaded Q for each band so that shared bands so they require nearly the same inductance. From Table 17.8, the same band switch position is shared on the 10/12-meter bands (1.4 µH) and the 15/18-meter bands (2.2 µH).

In actual construction of the tank circuit, it is very useful to have access to both a capacitance meter and an inductance meter. The author used an Elenco LCM-1950 meter that measures both capacitance and inductance and is available for under $100 (**www.elenco.com**). With the tune and load capacitors mounted and connected to calibrated knobs or turns counters, make a table of capacitance verses knob settings. This is useful to estimate the initial setting for each band during setup and test. Also, measure the inductance of each coil turn to determine initial coil taps for each band. On this amplifier, only the 10-meter tap had to be adjusted from the predetermined settings.

As mentioned above, the pi-L tank circuit was designed for 3100-Ω plate-load resistance. Such a high plate resistance demands higher inductance values to obtain reasonable tank circuit Qs. Table 17.8 shows that 160 meters requires 42 µH. If air-wound coils were used exclusively, the coils would require many turns and would take up a lot of cabinet space. To maintain a reasonable physical coil size, therefore, toroidal coils were used for

Table 17.8
Pi-L Component Values

Frequency (MHz)	*C1 (pF)*	*C2 pF*	*L1 µH*	*L2 µH*	*Q*
1.850	211	1262	44.3	9.6	12
3.700	105	631	22.2	4.8	12
7.150	65	364	9.7	2.5	14
14.150	33	184	4.9	1.26	14
18.100	45	208	2.23	0.98	23
21.200	33	159	2.21	0.84	20
24.900	36	161	1.48	0.71	25
28.250	29	133	1.43	0.63	23

Tank Circuit Coils

Coil	*Band*	*Inductance*	*Construction*
L1	10/12-15/17 m	2.3 µH	7½ t, ¼-in. copper tube, 2-in. ID silver-plated 10/12-m tap @ 3½ t 15/17-m tap @ 7½ t
L2	20-40 m	7.4 µH	19 t, ³⁄₁₆-in. copper tube, 2-in. ID silver plated 20 tap@ 8 t 40 tap @ 19 t
L3	80 m	12.4 µH	17 t on 3×T225-2 cores, #10 Teflon silver wire
L4	160 m	22.0 µH	23 t on 3×T300-2 cores, #10 Teflon silver wire
L5	L-Coil	9.6 µH	19 t on 2×T225-2 cores, #12 tinned wire w/Teflon sleeve 10/12-m tap @ 2 t 15/17-m tap @ 4 t 20-m tap @ 5 t 40-m tap @ 7 t 80-m tap @ 12 t 160-m tap @ 19 t

80 (L3) and 160 (L4) meters in addition to the output coil (L5) (see Fig 17.60B). You should use substantial core material for high-power operation to avoid core heating. Core sizes were increased by using multiple cores taped together. Each ferrite core is wrapped with three layers of high temperature fiberglass tape, available from RF Parts (**www.rfparts.com**). Teflon-insulated #10 wire was wound to obtain the desired inductance in L3 and L4. Both coils are mounted on ceramic standoffs and held in place with Teflon blocks.

The output coil is wound on a pair of T225-2 cores using #12 tinned wire covered with a Teflon sleeve. Taps onto the coil are made by carefully trimming a small ⅛-inch space from the Teflon sleeve on the inner edge of the core facing the bandswitch. Taps are then made from the back section of the 2-pole bandswitch using #12 tinned wire. The proper placement of each tap is determined by first winding #12 insulated wire around the core. A small slit is carved into each turn and the inductance was measured. The copper wire is removed and the final Teflon-covered #12 tinned wire is wound onto the core. Using the output L-coil (L102) design values in Table 17.8, permanent taps were made.

Note that the taps for the output coil are not extremely critical. Select the closest turn to the value needed. The output coil is mounted on the back of the bandswitch on one of the switch wafer screws using a threaded 1-inch diameter Teflon rod. The Teflon rod holds the position of the coil. The weight is carried by the wire taps from the coil to the bandswitch contacts. Table 17.8 also gives the inductance and construction instructions for each coil.

L1 and L2 (10-40 meters) are silver plated. They were wound using a 2-inch aluminum pipe as a form. Clean the copper tubing with #0000 steel wool prior to winding. Wind the copper tubing close spaced on the pipe. Leave plenty of pigtail on each end of the coil. The ends can be trimmed to fit the mounting positions precisely. After winding the desired number of turns, plug the ends by closing the tube ends with a hammer, spread the coil windings and rinse the coil in acetone to remove any oil. Allow a few minutes to dry. The coil is now ready to plate.

Go to any photo shop and beg/buy a gallon of used photographic fixer solution. Note that used fixer solution has silver remnant. The more the solution has been used, the more the silver content. The coils can be silver plated by dipping the clean coil into the solution. Do not leave the coil in the solution too long or it will turn black. A thin but bright silver coat will be deposited on the copper tube. This is called *flash plating*. After dipping the coil into the solution, immediately rinse in a bath of clean water and blow dry under pressure with an air compressor, heat gun or hair dryer. If a thicker silver coat is desired, electroplating is necessary, a subject beyond the scope of this article.

A #10 lug is crimped and soldered onto the end of each coil and used to mount the coil. The L2 coil is mounted using a Teflon block that is held in place to the front sub-panel with small screws. The block is carefully drilled with ³⁄₁₆-inch holes the desired spacing of the coil about ¾-inch from one edge. The block is sawed down through the holes creating two matching blocks. At each end of the block a hole is drilled and tapped (6-32 tap). The silver plated coil is sandwiched between the two blocks for secure support. The tapped screws serve as the connecting points for the ends of the coil.

METERING

The amplifier uses three separate meters to simultaneously monitor plate current, grid current and a choice of plate voltage, power output or filament voltage. Each meter is identical with a 1-mA full-scale movement. As mentioned previously, the custom scales for each movement were designed using the *Meter Basic* software from Tonne Software. This allows up to three scales on each meter. Scales can be designed as either linear or log and the number of major and minor tick marks can be specified. Each scale can be labeled using different font size and color. The author printed the scales using a color inkjet printer onto glossy photo paper. The scales were carefully cut to match the meter faceplate and glued into place using a thin coat of adhesive.

Plate current is measured by M1 in series in the B– line using a current divider (R114 and R115) as shown in Fig 17.58. Adjust R114 to obtain full scale with 1-A of plate current. The meter was calibrated prior to installation using a low-voltage power supply with adjustable current limiting in series with an accurate digital meter.

M2 monitors grid current by measuring the voltage drop created by grid current flow through the 10-Ω resistor, R104. Connecting a voltage source (ie, small variable power supply) across R104 and measuring the actual current flow with an external meter provides a way to set the calibration pot, R105.

M3 is a multimeter that reads HV, RF power or filament voltage. The metering circuit is selected using a 2P3T rotary switch (S5). The HV metering circuit is in the HV power supply and fed to the RF deck through the control cable. The filament-voltage detect circuit is shown on the control circuit diagram (Fig 17.57: D207, D201 and R202). Adjust R202 for the proper reading on Meter M3. The 3.1-V zener (D201) expands the meter scale for more precise reading.

The RF wattmeter circuit is also shown in Fig 17.57. Only forward power is measured and potentiometer R403 is used for calibration. The wattmeter is not a precise instrument but gives a relative output reading. It is adequate for peaking power output when tuning. The meter provides good accuracy through 40 meters and then begins to read lower on the higher-frequency bands. This is due to the simplicity of the circuit and the toroid used. Quite honestly, don't expect much accuracy from this wattmeter.

HV POWER SUPPLY

The matching HV Power Supply (Fig 17.58) provides approximately 4000 V under load. It uses a full-wave bridge rectifier and is filtered using a single 53 µF/5000-V oil filled capacitor (C301). Whenever the HV supply is plugged into the 240-V line, live 120 V ac is routed to the RF deck through the control cable. The 120-V ac line is obtained from L1 and neutral of the 240-V ac line. The neutral line is isolated from ground for safety.

Actuating the on/off switch S1 on the front panel of the RF deck provides ac power to the low-voltage power supply. In turn, +12 V is returned to the power supply through the control cable and routed to a pair of solid state power relays (K1, K2). Also, +12 V is routed to a timer relay that provides a two-second delay in applying +12 V to the second pair of solid-state relays (K3, K4). During the two-second delay, each leg of the 240-V ac primary voltage is routed through a 25-Ω resistor (R301, R302) to reduce the current surge when charging the filter capacitor, C301.

HV is metered at the bottom of the two series 100,000-Ω bleeder resistors (R303, R304). A current divider is created using a small potentiometer (R306) in parallel with a 25-Ω/5-W resistor (R305). The current divider is in series with the bleeder resistors and tied to the B− line. R306 is set to allow 1 mA of current to flow to the HV meter located on the front panel of the RF deck with 5000 V HV dc. The potentiometer R306 and the paralleled fixed resistor R305 need handle only a small amount of power, since the voltage and current flow is quite small at this point in the circuit.

The HV cable between the RF Deck and power supply is made from a length of automotive-ignition cable that has a #20 wire and 60,000-V insulation. Be sure to get a solid-wire center conductor and not the resistive carbon material. Also use high-quality HV connectors that are intended for such an application. Millen HV connectors (50001) were used in this amplifier. Coax boots intended for coaxial cable are used on each connector for added insulation and physical strength. The mounting holes for the Millen connectors are oversized and plastic screws were used for safety.

TUNING AND OPERATION

The amplifier is very easy to tune after the initial settings of the tune (C101) and load (C102) controls are determined. The correct settings are determined with a plate current of 700 mA with a corresponding grid current of 200 mA. The turn counters provide excellent resetability once the proper settings have been found initially. Required drive power is about 75 W for 1500 W output.

Thanks to CPI Eimac Division, LDG Electronics, and MTI Inc (Radio Switch) for their support in this project.

Project: A 6-Meter Kilowatt Amplifier

The Svetlana 4CX1600B tube has attracted a lot of attention because of its potent capabilities and relatively low cost. Because of its high gain and its large anode dissipation capabilities, the tube has relatively large input and output capacitances—85 pF at the input and 12 pF at the output. Stray capacitance of about 10 pF must be added in as well. On bands lower than 50 MHz, these capacitances can be dealt with satisfactorily with a broadband 50-Ω input resistor and conventional output tuning circuitry.

See the article by George Daughters, K6GT, "The Sunnyvale/Saint Petersburg Kilowatt-Plus" in 2005 and earlier Handbooks and included in the Templates section of the *Handbook CD* for details on suitable control and power-supply circuitry. This 6-meter amplifier uses the same basic design as K6GT's, except for modified input and output circuits in the RF deck. See **Fig 17.61**, a photograph of the front panel of the 6-meter amplifier built by the late Dick Stevens, W1QWJ.

On the 50-MHz band the tube's high input capacitance must be tuned out. The author used a T network so that the input impedance looks like a nonreactive 50 Ω to the transceiver. To keep the output tuning network's loaded Q low enough for efficient power generation, he used a 1.5 to 46 pF Jennings CHV1-45-5S vacuum-variable capacitor, in a Pi-L configuration to keep harmonics low. You should use a quarter-wave shorted coaxial stub in parallel with the output RF connector to make absolutely sure that the second harmonic is reduced well below the FCC specification limits.

To guarantee stability, the author had to make sure the screen grid was kept as close as possible to RF ground. This allows the screen to do its job "screening"—this minimizes the capacitance between the control grid and the anode. He used the Svetlana SK-3A socket, which includes a built-in screen bypass capacitor, and augmented that with a 50-MHz series-tuned circuit to ground. In addition, to prevent VHF parasitics, he used a parasitic suppressor in the anode circuit.

Unlike the K6GT HF amplifier, this 6-meter amplifier uses no cathode degeneration. The author wanted maximum stable power gain, with less drive power needed on 6 meters. He left the SK-3A socket in stock form, with the cathode directly grounded. This amplifier requires about 25 W of drive power to produce full output.

Fig 17.62 is a schematic of the RF deck. The control and power supply circuitry are basically the same as that used in the K6GT HF amplifier, except that plate current is monitored with a meter in series with the B− lead, since the cathode in this amplifier is grounded directly. The K6GT power supply is modified by inserting a 250-Ω, 25-W power resistor to ground in place of the direct ground connection. See **Fig 17.63**. In Fig 17.62, C1 blocks grid-bias dc voltage from appearing at the transceiver, while L1, L2 and C2 make up the T-network that tunes out the input capacitance of V1. R1 is a non-reactive 50-Ω 50-W resistor.

Fig 17.61—Photo of the front panel of the 6-meter 4CX1600B amplifier.

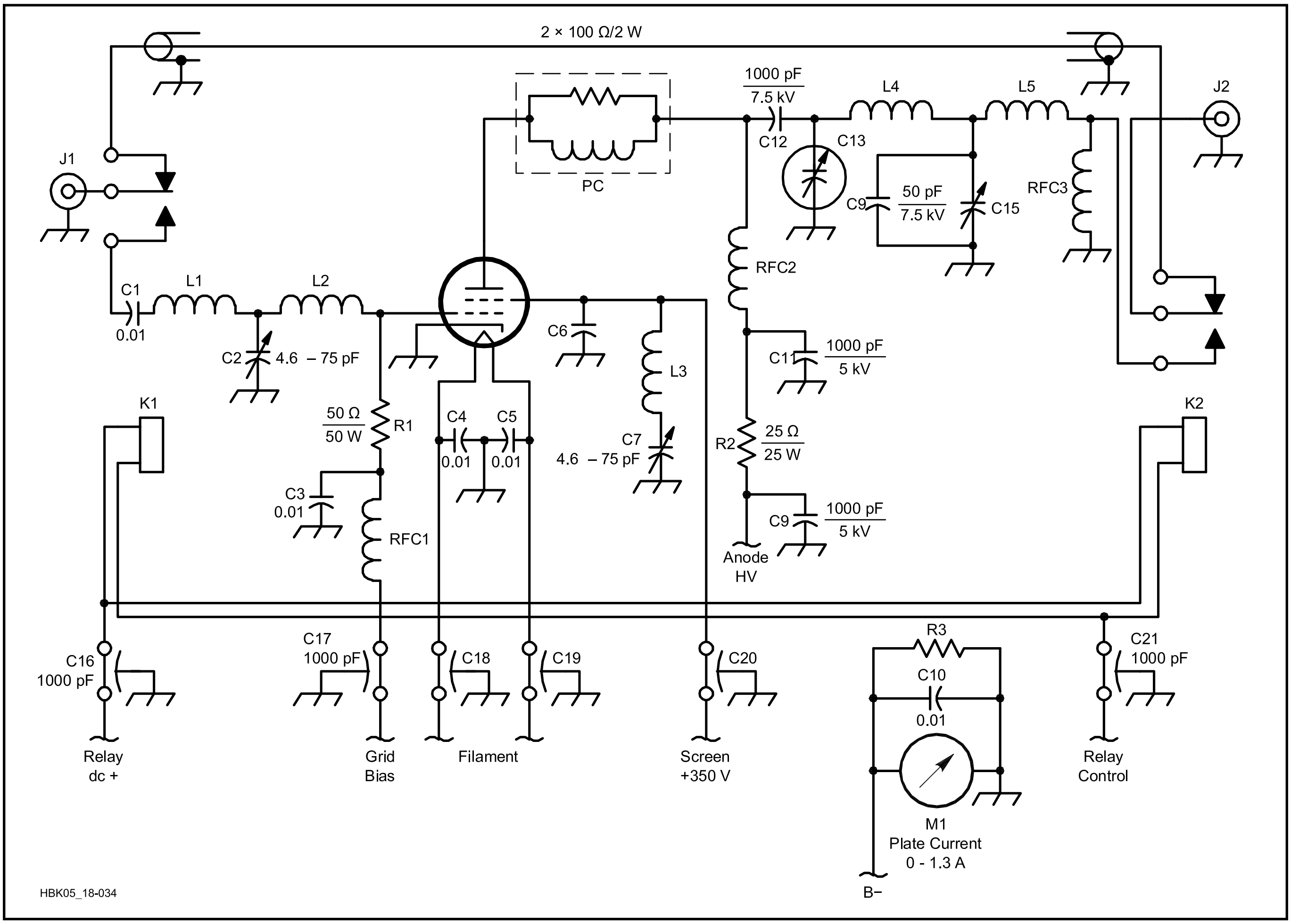


Fig 17.62—Schematic for the RF deck for the 6-meter 4CX1600B amplifier. Capacitors are disc ceramic unless noted.

C2, C7—4.6-75 pF, 500-V air-variable trimmer capacitor, APC style.
C6—Screen bypass capacitor, built into SK-3A socket.
C13—1-45 pF, 5 kV, Jennings CHV1-45-5S vacuum-variable capacitor.
C14—50 pF, 7.5 kV, NP0 ceramic doorknob capacitor.
C15—4-102 pF, 1100V, HFA-100A type air-variable capacitor.
C16, 17, 18, 19, 20, 21—1000 pF, 1 kV feedthrough capacitors.
L1—11 turns, #16, ⅜-inch diameter, 1-inch long.
L2—9 turns #16, ⅜-inch diameter, close-wound.
L3—8 turns #16, ⅜-inch diameter, ⅞-inch long.
L4—¼-inch copper tubing, 4½ turns, 1¼ inches diameter, 4½ inches long.
L5—5 turns #14, ½-inch diameter, 1⅜ inches long.
M1—0-1.3 A meter, with homemade shunt resistor, R3, across 0-10 mA movement meter.
PC—Parasitic suppressor, 2 turns #14, ½-inch diameter, shunted by two 100-Ω, 2-W carbon composition resistors in parallel.
RFC1—10 μH, grid-bias choke.
RFC2—Plate choke, 40 turns #20, ½-inch diameter, close-wound.
RFC3—Safety choke, 20 turns #20, ⅜ inch diameter.

C6 is the built-in screen bypass capacitor in the SK-3A socket, while L3 and C7 make up the series-tuned screen bypass circuit. RFC3 is a safety choke, in case blocking capacitor C12 should break down and short, which would otherwise place high voltage at the output connector.

CONSTRUCTION

Like the K6GT amplifier, this amplifier is constructed in two parts: an RF deck and a power supply. Two aluminum chassis boxes bolted together and mounted to a front panel are used to make the RF deck. **Fig 17.64** shows the 4CX1600B tube and the 6-meter output tank circuit.

Fig 17.65 shows the underside of the RF deck, with the input circuitry shown in more detail in **Fig 17.66**. The 50-Ω, 50-W noninductive power resistor is shown in Fig 17.66. Note that the tuning adjustment for the input circuit is accessed from the rear of the RF deck.

AMPLIFIER ADJUSTMENT

The tune-up adjustments can be done without power applied to the amplifier and with the top and bottom covers removed. You can use readily available test instruments: an MFJ-259 SWR Analyzer and a VTVM with RF probe.

1. Activate the antenna changeover relay, either mechanically or by applying control voltage to it. Connect a 2700-Ω, ½-W carbon composition resistor from anode to ground using short leads. Connect the SWR analyzer, tuned to 50 MHz, to the output connector. Adjust plate tuning and loading controls for a 1:1 SWR. You are using the Pi-L network in reverse this way.
2. Now, connect the MFJ-259 to the input connector and adjust the input

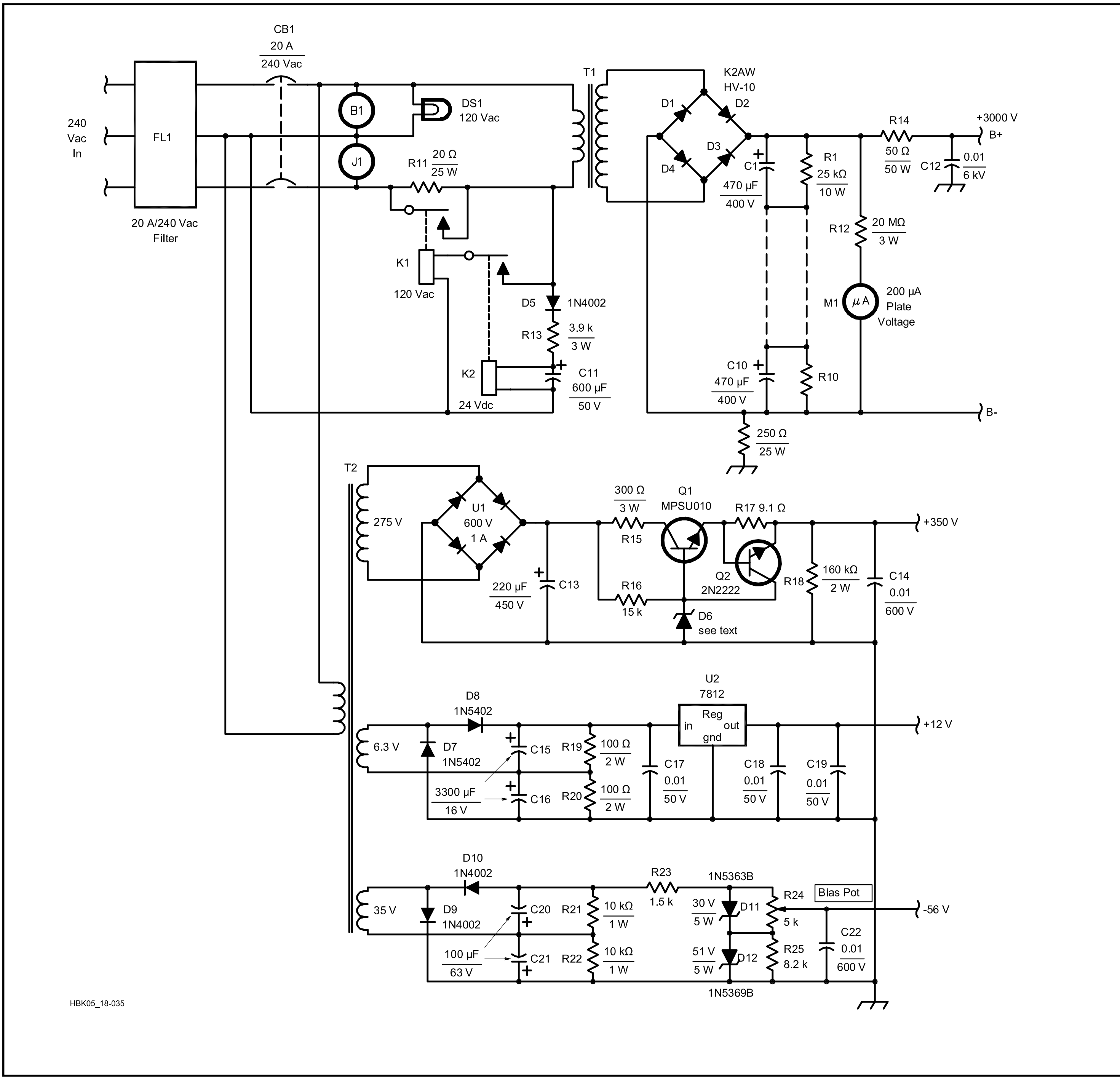


Fig 17.63— Schematic diagram of the K6GT high-voltage plate and regulated screen supply for the 6 meter 4CX1600B linear amplifier. K1 , K2 and associated circuitry provide a "step-start" characteristic to limit the power-on surge of charging current for the filter capacitors. Resistors are ½ W unless noted. Capacitors are disc ceramic unless noted and those marked with a + are electrolytic.

T-network for a 1:1 SWR. Some spreading of the turns of the inductor may be required.

3. Disconnect the Pi-L output network from the tube's anode, leaving the 2700-Ω carbon composition resistor from the anode still connected. Connect the RF probe of the VTVM to the anode and run your exciter at low power into the amplifier's input connector. Tune the screen series-tuned bypass circuit for a distinct dip on the VTVM. The dip will be sharp and the VTVM reading should go to zero.
4. Now, disconnect the 2700-Ω carbon resistor from the anode and replace the covers. Connect the power supply and control circuitry. When you apply power to the amplifier, you should find that only a slight tweaking of the output controls will be needed for final adjustment.

B1 — Muffin fan (Rotron SU2A1 or similar).
C1-C10 — Filter capacitors; 470 µF, 400 V electrolytic.
C11 — 600 µF, 50 V electrolytic.
C12 — 0.01 µF, 6 kV disc ceramic.
C13 — 220 µF, 450 V electrolytic.
C14, C22 — 0.01 µF, 600 V disc ceramic.
C15, C16 — 3300 µF, 16 V electrolytic.
C17, C18, C19 — 0.01 µF, 50 V disc ceramic.
C20, C21 — 100 µF, 63 V electrolytic.
CB1 — two pole 20 A, 240 V ac circuit breaker.
D1-D4 — K2AW's HV-10 rectifier diodes.
D5, D9, D10 — 1N4002.
D6 — Zener diodes, three 1N4764A and one 1N5369B to total approximately 350 V dc.
D7, D8 — 1N5402.
D11 — Zener diode, 1N5363B (30 V, 5 W).
D12 — Zener diode, 1N5369B (51 V, 5 W).
DS1 — 120 V ac indicator lamp (red).
FL1 — 240 V ac, 20 A EMI filter.
K1 — 120 V ac DPDT relay; both poles of 240 V ac/15 A contacts in parallel.
K2 — 24 V dc relay; 120 V ac, 5 A contacts.
M1 — 200 µA meter movement.
Q1 — MPSU010.
Q2 — 2N2222.
R1-R10 — Bleeder resistors; 25 kΩ, 10 W.
R11 — 20 Ω, 25 W.
R12 — 20 M Ω, 3 W (Caddock MX430).
R13 — 3.9 k Ω, 3 W.
R14 — 50 Ω, 50 W mounted on standoff insulators.
R15 — 300 Ω, 3 W.
R18 — 160 k Ω, 2 W.
R19, R20 — 100 Ω, 2 W.
R21, R22 — 10 k Ω, 1 W.
R24 — 5 kΩ potentiometer; sets control grid bias for desired no-signal cathode current.
T1 — Plate transformer (Peter W. Dahl No. ARRL-002, contact Harbach Electronics, www.harbachelectronics.com, for equivalent parts).
T2 — Power transformer, 120 V / 275 V at 0.06 A, 6.3 V at 2 A, 35 V at 0.15 A.
U1 — 600 V, 1 A rectifier bridge.
U2 — 7812, +12 V IC voltage regulator.

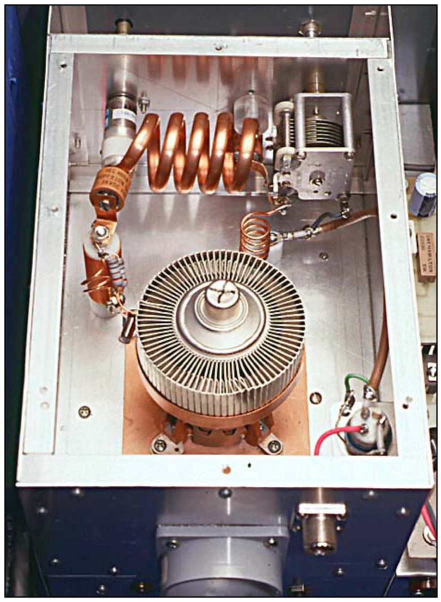


Fig 17.64—Close-up photo of the anode tank circuit for 6-meter kW amplifier. The air-cooling chimney has been removed in this photo.

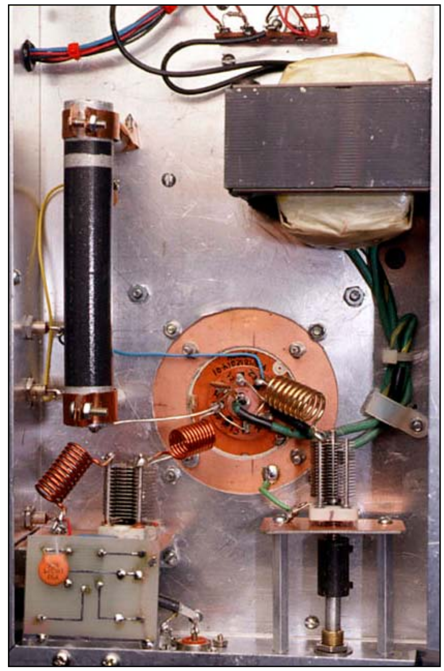


Fig 17.66—Close-up photo of the input circuitry for the 6-meter kW amplifier. Input tuning capacitor C2 is adjusted from the rear panel during operation, if necessary. The series-tuning capacitor C7 used to thoroughly ground the screen for RF is shown at the lower right. It is adjusted through a normally plugged hole in the rear panel during initial adjustment only.

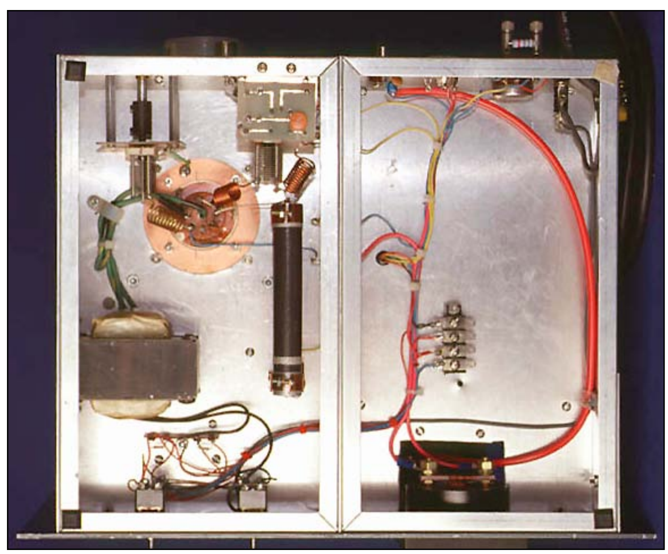


Fig 17.65—Underneath the 6-meter kW amplifier RF deck, showing on the left the tube socket and input circuitry.

Using "Surplus" Parts for Your Amplifier

First-time builders of power amplifiers soon discover that buying all new electronic components from a parts list is costly. Manufacturers realize significant savings by purchasing components in bulk, but the builder of a single unit will pay top dollar for each new part thus negating any savings in the labor costs. While some impressive equipment has been built using all new parts, pride of design and workmanship are usually the goals, not cost savings.

The way to break this economic barrier is to use surplus parts. These are items left over when a project is finished, a design changes, or a war ends. Sometimes, costly parts can be recovered from relatively new but obsolete equipment. Alert dealers locate sources of surplus parts, buy them at auction, often for pennies on the dollar, and make them available in stores, on-line or at hamfests.

Parts can become available when a project is abandoned for which parts have been gathered. Many hams maintain a "junk box" of parts against future needs, only to find it has grown beyond a manageable size. This may contain a mixture of new, used and NOS parts (new old stock, meaning unused but stored for many years).

When buying electronic parts from other than a trusted supplier of new stock, some precautions are in order. Before shopping the surplus shelves or making an on-line purchase, do some research. You are leaving behind the security of buying a specific part that the designer of the amplifier has tested in actual operation. One has to be sure not only that the part is sound, but that it is suited for the intended purpose. This effort is justified by the anticipated savings of up to 80%. Using "odd" parts may make your amplifier a "one of a kind" project, since the next builder won't be able to obtain the same items. Your amplifier may not be reproducible, but it can still be an object of pride and usefulness.

Surplus vacuum tubes may be of military origin. In the 50s and 60s WWII tubes powered many ham amplifiers. Tubes made in Russia in past decades are now available[1]. Many hams have used these tubes for new designs and even as replacements for hard-to-find or more expensive tube types in existing amplifiers. Since tubes have a limited life, one may wish to buy a spare or two before committing to using a surplus type, since future availability is always a concern. Broadcast or medical "pulls" or "pull-outs" provide a source of good used tubes. For highest equipment reliability, these tubes are replaced on a scheduled basis rather than at failure. They will have less life than a new tube, but can be so cheap that they are still a good deal.

When buying used tubes at a hamfest, a quick ohmmeter check is important. Between the filament pins there should be only a few ohms, with high resistance between all other connections. Signs of obvious overheating, such as discoloration or warping, are a concern and call at the least for a reduced price. There is no 100% guarantee that a tube is good apart from trying it in a circuit, since tube testers usually will not test large power tubes. Experience with on-line suppliers of NOS tubes indicates that a small percentage will be duds. If your supplier guarantees them you should be willing to pay a bit more. Some dealers will test or even match the tubes for parallel operation.

Finding sockets for odd tubes may be a challenge. Several enterprising hams supply sockets for popular types, such as the Russian GI7B[2], or you can sometimes bolt the tubes into place with straps and screws.

Using vacuum capacitors really adds class and value to an amplifier. Check the vacuum by setting the capacitor to minimum capacity and watching to see if the vacuum is sufficiently strong to pull the plates all the way in when the screw or hand pressure is released. Weak vacuum means the capacitor will not withstand its rated voltage. With the plates fully meshed, check for a short circuit, which indicates damaged plates. Be sure that the adjusting screw moves freely. Lubrication won't always free up a sticky screw. Check that the voltage and current ratings are suitable for your application. The voltage rating is usually printed on the capacitor. The current rating will be on a factory data sheet or use physical size as an indication.

Vacuum TR (transmit-receive) relays work well in an amplifier and often switch quickly enough to be used for full-QSK (full break-in) operation. If you shop carefully, a surplus relay can cost little more than a much-inferior open-frame type. Visual inspection of the contacts and an ohmmeter check on the coil will generally insure that the relay is good. If the relay's coil voltage is 24 V or some value other than what you wanted, remember that a small power supply can be added cheaply. Be aware that the amplifier keying circuits of most transceivers are not rated to switch the large coil currents of surplus relays, particularly those that operate at 24 V or higher. An outboard switching interface will be needed for these relays. Older non-vacuum relays also have long switching times that, if not accounted for in the amplifier design, may cause "hot-switching" and contact damage.

Inductors can usually be evaluated by inspection since hidden faults are rare. The current rating is determined by conductor size. With formulas found in the **Electrical Fundamentals** chapter, you can determine the inductance. Carry a hand calculator when shopping.

Choose RF switches based on physical size and inspection. Avoid badly arced-over or pitted contacts or insulators.

For power transformers, finding the exact required voltage may be difficult. You can use extra low-voltage windings to buck or boost the primary voltage. External transformers can also be used

17.13 References and Bibliography

TUBE TYPE AMPLIFIER SECTION

ARRL members can download many *QST* articles on amplifiers from: **www.arrl.org/QST**.

All About the GI-7B, K4POZ, **gi7b.com**

Badger, G., W6TC, "The 811A: Grandfather of the Zero Bias Revolution," **www.arrl.org/qst**

Care and Feeding of Power Grid Tubes, **www.cpii.com/library.cfm/9**

EIMAC tube performance computer, **www.cpii.com/library.cfm/9**

Grammer, G., "How to Run Your Linear," *QST,* Nov 1962, pp 11-14.

Knadle, R., "A Strip-line Kilowatt Amplifier 432 MHz," *QST*, Apr 1972, pp 49-55.

LTSPICE, download from Linear Technology , **www.linear.com/designtools/software/ltspice.jsp**

Meade, "A High-Performance 50-MHz Amplifier," *QST*, Sep 1975, pp 34-38.

Meade, "A 2-KW PEP Amplifier for 144 MHz," *QST*, Dec 1973, pp 34-38.

Measures, R., "Improved Parasitic Suppression for Modern Amplifier Tubes," *QST*, Oct 1988, pp 36-38, 66, 89.

Neutralizing Amplifiers, **www.w8ji.com/neutralizing_amplifier.htm**

Orr, *Radio Handbook*, 22nd Ed (Howard W. Sams & Co, Inc, 1981).

in an autotransformer connection for adjustment. Use of a variable autotransformer (Variac or similar) can adapt a transformer to your needs, but the extra winding and core losses may cause voltage sag under load. Amplifiers normally use a separate transformer for supplying the tube filaments as filaments are often turned on before the high-voltage supply. Therefore, filament windings on a plate transformer are not needed. Also, with either center-tapped, full bridge or voltage doubler circuits, you have three options for ac secondary voltage.

Transformers should be rated for 50 or 60 Hz operation, not 400 Hz as used in aircraft. Check that the current and overall VA (volt-amp) ratings are adequate. Use an ohmmeter to detect shorts or open windings, and to be sure that you understand the terminal connections. Check each winding to ground and to each other. Reject transformers with charred insulation. Use your nose to detect burned smells from internal overloads. A rusty core can be wire brushed and spray painted, but might indicate moisture has gotten to the windings. Test the primary with 120 or 240 V ac, if possible. High current or loud buzzing can indicate shorted turns. Be very careful when measuring the secondary windings as the high voltage can easily destroy your multimeter, not to mention endanger your life! (See the **Power Supplies** chapter for a method of testing transformers safely.)

Blowers can be found on otherwise worthless equipment, such as a rack of old tube gear. It may be worth buying the whole rack just to get a nice blower. The pressure and volume ratings of a blower may not be known, but apart from noise, there is no disadvantage in over-sizing your blower. An over-sized blower can be slowed down by reducing the voltage or throttled back with baffles. Remember that if the voltage rating is strange, providing an odd voltage may be cheaper than buying a more expensive blower. Avoid 400 Hz aircraft blowers. For blowers needing an external capacitor, try to obtain the capacitor with the blower.

Racks and chassis components can often be refurbished for an amplifier. Standard 19-inch rack-type cabinets are widely available surplus and at hamfests. Even new, they are available at modest cost and have good fit and finish. Filling the many panel holes and painting or adding a new front panel can make them look like new, while saving a lot of expensive and time consuming metal work.[3]

— *Roger Halstead, K8RI*

Photo A — A hamfest is often a good source of used, surplus, and even new parts for building amplifiers and other high-power RF equipment, such as impedance matching units or full-power filters. Follow the cautions in this sidebar when evaluating parts and pieces.

[1]**www.nd2x.net/base-1.html**
[2]**k4poz.com/gi7b/index.html**
[3]W Yoshida, KH6WZ, "Recycling Old Cabinets and Chassis Boxes," *QST*, Jul 2008, p 30.

Pappenfus, Bruene and Schoenike, *Single Sideband Principles and Circuits* (McGraw-Hill, 1964).

Peck, Floyd K., "A Compact High-Power Linear," *QST*, Jun 1961, pp 11-14.

Potter and Fich, *Theory of Networks and Lines* (Prentice-Hall, 1963).

RCA Transmitting Tubes Technical Manual TT-5, Radio Corporation of America, 1962.

Reference Data for Radio Engineers, ITT, Howard & Sams Co, Inc.

Simpson, *Introductory Electronics for Scientists and Engineers* (Allyn and Bacon, 1975).

Terman, *Electronic and Radio Engineering* (McGraw-Hill).

Wingfield, E., "New and Improved Formulas for the Design of Pi and Pi-L Networks," *QST*, Aug 1983, pp 23-29. (Feedback, *QST*, Jan 1984, p 49.).

Wingfield, E., "A Note on Pi-L Networks," *QEX*, Dec 1983, pp 5-9.

SOLID-STATE AMPLIFIER SECTION

Abulet, Mihai, *RF Power Amplifiers* (Noble Publishing, 2001, ISBN 1884932126).

Blinchikoff, Zverev, *Filtering in the Time and Frequency Domains* (Noble Publishing, 2001, ISBN: 1884932177).

Bowick, C. Ajluni, and J. Blyler, *RF Circuit Design,* 2nd ed. (Newnes, 2007, ISBN-10: 0750685182).

Classic Works in RF Engineering : Combiners, Couplers, Transformers and Magnetic Materials Edited by Myer, Walker, Raab, Trask (Artech House, 2005, ISBN-13: 9781580530569).

Cripps, *Advanced Techniques in RF Power Amplifier Design* (Artech House, 2002, ISBN-10: 1580532829).

Davis and Agarwal, *Radio Frequency Circuit Design* (Artech House, 2001, ISBN: 0471350524).

Dye, Norman. and Helge Granberg, *Radio Frequency Transistors — Principles and Practical Applications,* (Butterworth-Heinemann, 1993, ISBN: 0750690593; 2nd ed, 2001, ISBN 0750672811).

Frenzel, Louis E., *RF Power for Industrial Applications*, (Prentice Hall, 2003, ISBN 0-1309-6577-4).

Grebennikov and Sokal, *Switchmode RF Power Amplifiers,* (Newnes, 2007, ISBN: 075067962X).

Grebennikov, *RF and Microwave Power Amplifier Design,* (McGraw-Hill, 2004, ISBN: 0071444939).

Kraus, Bostian and Raab, *Solid State Radio Engineering,* (Wiley, 1980, ISBN 04710301X).

Oxner, E., *Power FETs and Their Applications*, (Prentice-Hall, 1982, ISBN 0136869238).

Raghavan, Srirattana, Laskar *Modeling and Design Techniques for RF Power Amplifiers,* (Wiley-IEEE Press, 2008, ISBN: 0471717460).

Rutledge, David B., *The Electronics of Radio,* (Cambridge University Press, 1999, ISBN-13: 9780521646451).

Sabin, Schoenike, et al, *Single Sideband Systems and Circuits*, (McGraw-Hill, 1993, ISBN 0079120385).

Sevick, J., *Transmission Line Transformers,* 4th ed (Noble Publishing, 2001, ISBN: 1884932185).

Shirvani, A. and Wooley, B.A., *Design and Control of RF Power Amplifiers*, (Springer; 2003, ISBN-10: 1402075626).

Vizmuller, *RF Design Guide: Systems, Circuits, and Equations*, (Artech House, 1995, ISBN: 0890067546).

Zverev, A. I., *Handbook of Filter Synthesis*, (Wiley, 1970, ISBN: 0471986801).

Articles

Davis and Rutledge, "Industrial Class-E Power Amplifiers With Low-cost Power MOSFETs and Sine-wave Drive" Conf. Papers for RF Design '97, Santa Clara, CA, Sep 1997, pp 283-297.

Franke and Noorani, "Lumped-Constant Line Stretcher For Testing Power Amplifier Stability," *RF Design*, Mar/Apr 1983.

Frey, *A 50 MHz, 250W Amplifier using Push-Pull ARF448A/B*, APT Application Note APT9702.

Granberg, H., "Broadband Transformers and Power Combining Techniques for RF," AN-749, Motorola Semiconductor Products Inc.

Hilbers, A.H., *Design of HF Wideband Power Transformers,* Philips Semiconductors, ECO 6907, Mar 1998.

Hilbers, A.H., *Design of HF Wideband Power Transformers Part II,* Philips Semiconductors, ECO 7213, Mar 1998.

Hilbers, A.H., *Power Amplifier Design,* Philips Semiconductors, Application Note, Mar 1998.

Trask, C., "Designing Wide-band Transformers for HF and VHF Power Amplifiers," *QEX,* Mar/Apr 2005, pp 3-15.

Microwave — Cellular Applications

Cripps, *RF Power Amplifiers for Wireless Communication*, (Artech House, 1999, ISBN: 1596930187).

Groe and L. E. Larson, *CDMA Mobile Radio Design*, (Artech House, 2000, ISBN: 1580530591).

Pozar, *Microwave Engineering,* 3rd ed. (Wiley, 2004, ISBN: 0471448788).

Raghavan, Srirattana, Laskar *Modeling and Design Techniques for RF Power Amplifiers,* (Wiley-IEEE Press, 2008, ISBN: 0471717460).

van Nee and Prasad, *OFDM for Wireless Multimedia Communications*, (Artech House, 2000, ISBN: 0890065306).

York and Popovic, *Active and Quasi-Optical Arrays for Solid-State Power Combining*, (Wiley, 1997, ISBN: 0471146145).