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Chapter 17 — CD-ROM Content

Software

- *TubeCalculator* by Bentley Chan and John Stanley, K4ERO, for analysis of operation of popular high power transmitting tubes.
- *PI-EL* by Jim Tonne, W4ENE, for design and analysis of pi and pi-L networks for transmitter output.
- *SVC Filter* by Jim Tonne, W4ENE, for design and analysis of low-pass and high-pass filters using standard value components.
- *MeterBasic* by Jim Tonne, W4ENE, for design and printing of custom analog meter scales.
- *MATCH.EXE* software (for use with Tuned (Resonant) Networks discussion)

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- Tuned (Resonant) Networks (for use with MATCH.EXE)
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- "The Everyham's Amp" by John Stanley, K4ERO
- Everyham's Amp files — construction notes, layouts, modifications for various tube types
- "A 3CX1500D7 RF Linear Amplifier" by Jerry Pittenger, K8RA
- 3CX1500D7 HF amplifier files — PCB layout, Pi-L values spreadsheet
- "A 250 W Broadband Solid-State Linear Amplifier" by Dick Frey, K4XU
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- "The Sunnyvale/Saint Petersburg Kilowatt-Plus," a 4CX1600B HF amplifier project by George Daughters, K6GT

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- "432 MHz 3CX800A7 Amplifier" by Steve Powlishen, K1FO (SK)
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Chapter 17

RF Power Amplifiers

Amateur Radio operators typically use a very wide range of transmitted power — from milliwatts to the full legal power limit of 1.5 kW. 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. The 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. This edition also includes several new amplifier projects in the print edition: The Everyham's Amplifier by K4ERO; an All-mode, 2 Meter, 80 W Linear Amp by W6PQL, and a 2 W 10 GHz Amplifier by KB9MWR. In addition, more amplifier projects and supplemental information are included on the CD-ROM.

17.1 High Power, Who Needs It?

There are certain activities where higher power levels are almost always required for the greatest success — contesting and DXing, for example. While there are some outstanding operators who enjoy being competitive in spite of that disadvantage, the high-power stations usually have the biggest scores and get through the pileups first. On the VHF and higher bands, sometimes high power is the only way to overcome path loss and poor propagation.

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. This can be crucial to effective emergency communications. General operation also benefits from the availability of high power when conditions demand it to maintain contact.

How do you decide that you need amplification? As a rule of thumb, if you have a good antenna and 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 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

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 **Power Sources** and the **Safety** chapters in this *Handbook*.

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 life for the amplifying devices and other amplifier components as well as a cleaner signal result from running an amplifier below its maximum rating. However, just as with automobiles, that extra capability also presents certain temptations. Remember that FCC rules require you to employ an accurate way to determine output power, especially when you are running close to the limit.

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 fully 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.)

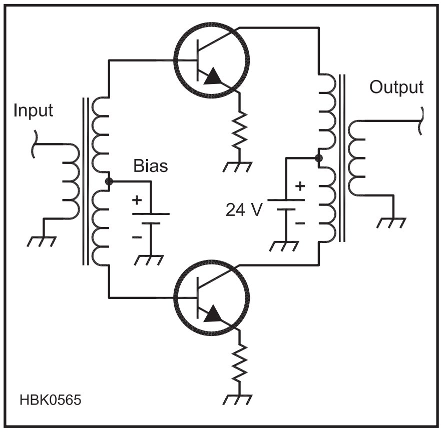


Fig 17.1 — This simple circuit can operate in a linear manner if properly biased.

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.

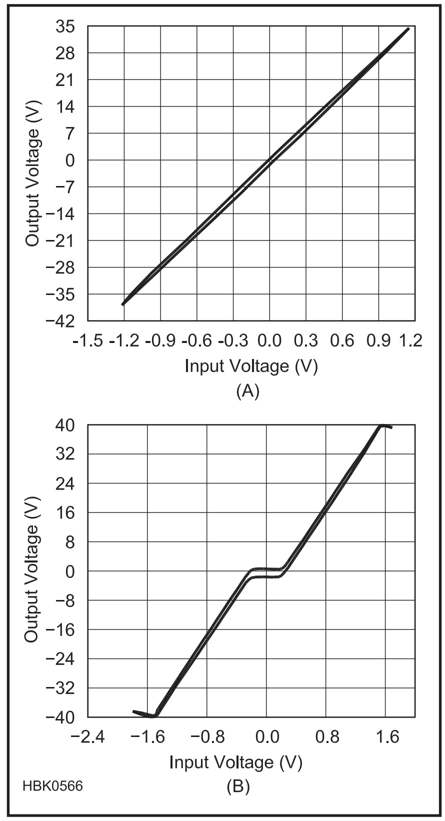


Fig 17.2 — Input versus output signals from an amplifier, as observed with the X-Y display on an oscilloscope. At A, an amplifier with proper bias and input voltage. At B, the same amplifier with improper bias and high input voltage.

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

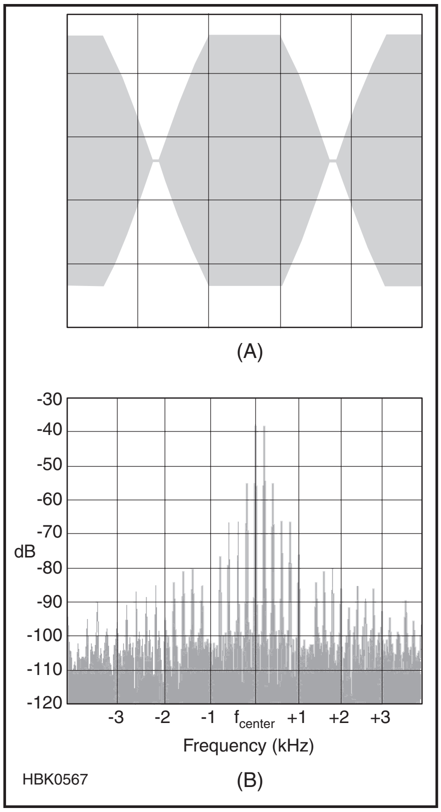


Fig 17.3 — Improper operation of a linear using a two-tone test will show peak clipping on an oscilloscope (A) and the presence of additional frequencies on a spectrum analyzer (B). When these patterns appear, your "linear" has become a "nonlinear" amplifier. Users of adjacent channels will not be happy.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

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. 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 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.

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**.

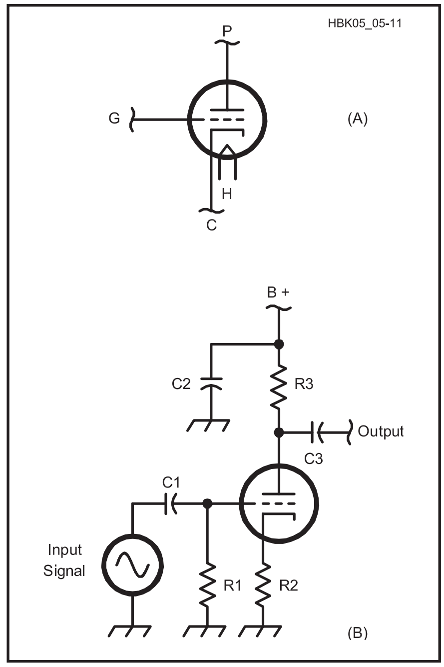


Fig 17.4 — Vacuum tube triode. (A) Schematic symbol detailing heater (H), cathode (C), grid (G) and plate (P). (B) Audio amplifier circuit using a triode. C1 and C3 are dc blocking capacitors for the input and output signals to isolate the grid and plate bias voltages. C2 is a bypass filter capacitor 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.

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.

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 be-

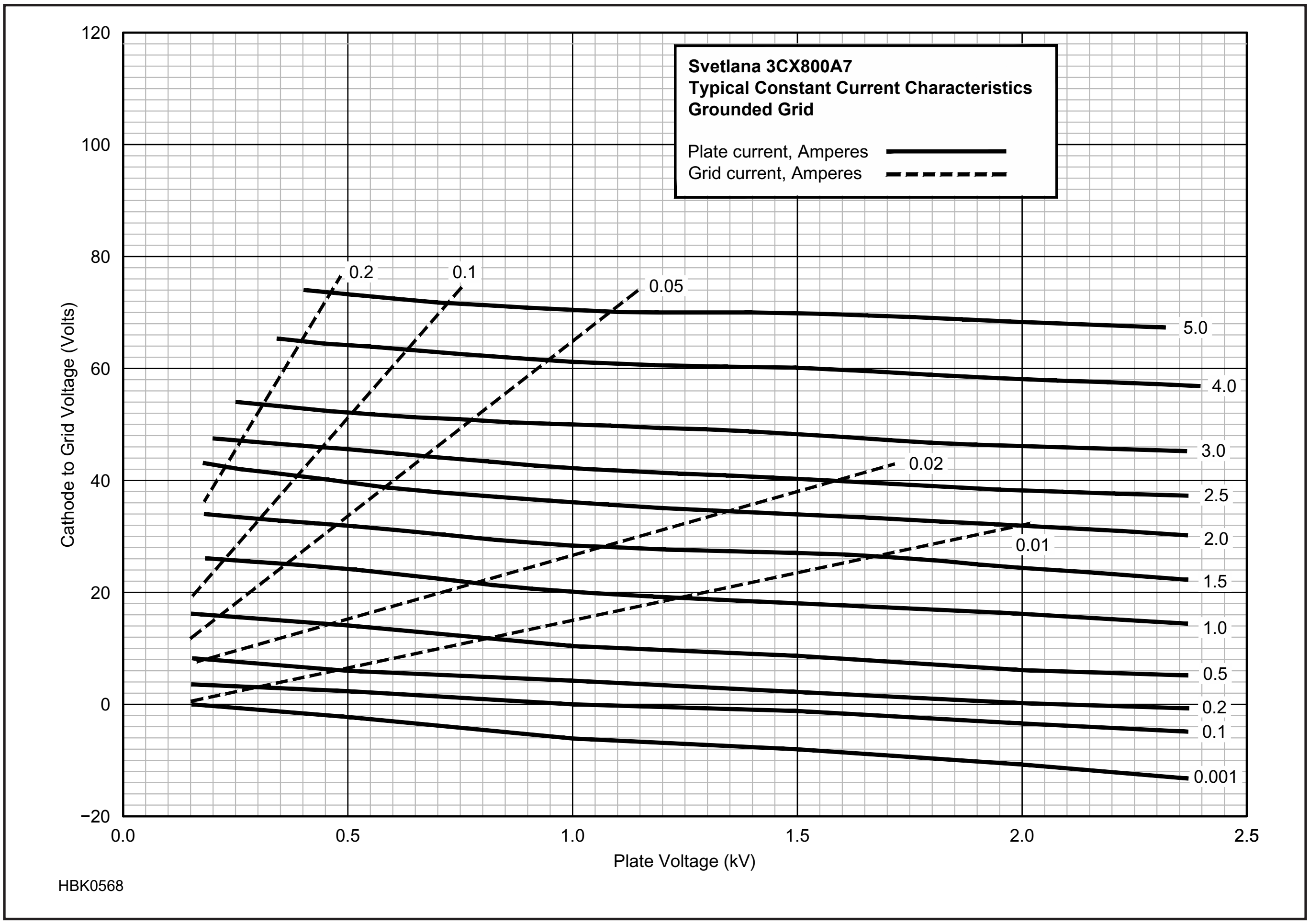


Fig 17.5 — Characteristic curves for a 3CX800A7 triode tube. Grid voltage is plotted on the left, plate voltage along the bottom. The solid lines are plate current, the dashed lines are grid current. This graph is typical of characteristic curves shown in this chapter and used with the *TubeCalculator*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

tween 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

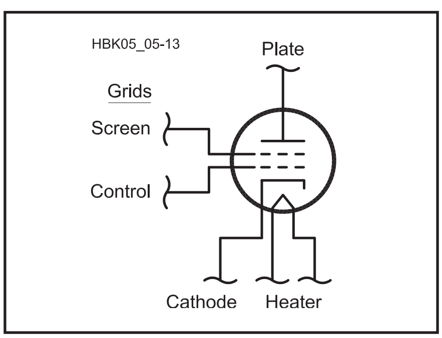


Fig 17.6 — Vacuum tube tetrode. Schematic symbol detailing heater (H), cathode (C), the two grids: control and screen, and the plate (P).

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*,

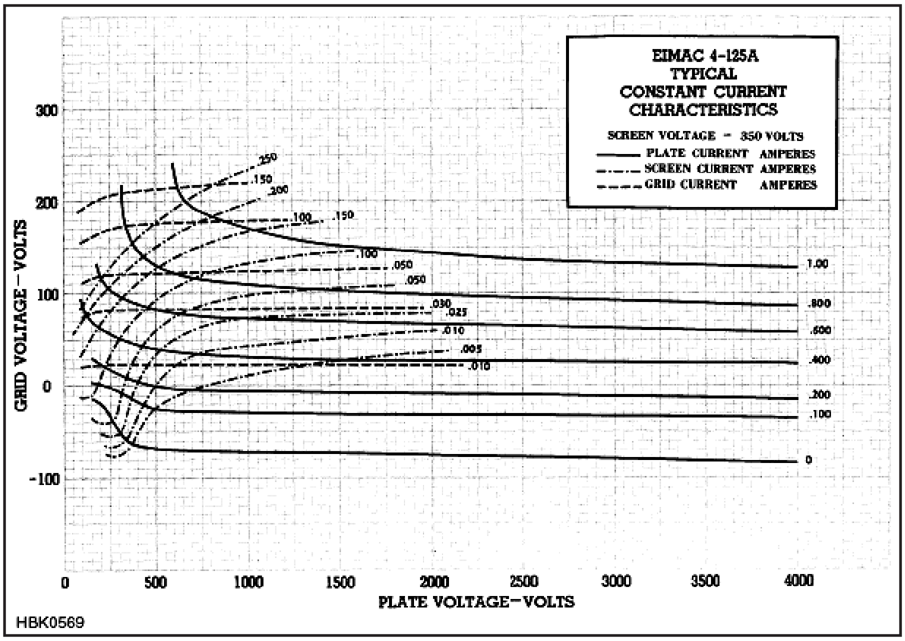


Fig 17.7 — Characteristic curves for a 4-125 tetrode tube.

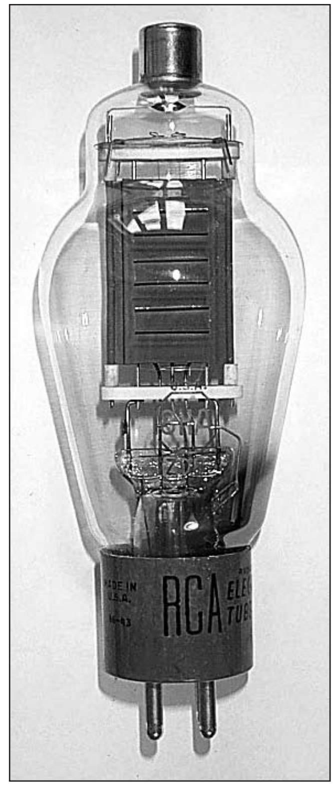


Fig 17.8 — The RCA 811A is an example of a transmitting triode with a glass envelope. [Photo courtesy the Virtual Valve Museum, www.tubecollector.org]

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

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 elements). 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 transmitting 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 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 1/2 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

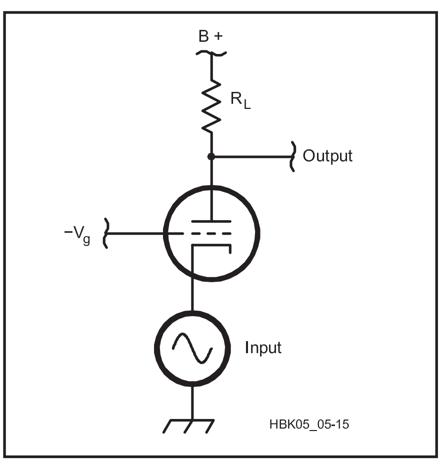


Fig 17.10 — Grounded grid amplifier schematic. The input signal is connected to the cathode, the grid is biased to the appropriate operating point by a dc bias voltage, $-V_G$, and the output voltage is obtained by the voltage drop through R_L that is developed by the plate current, I_P.

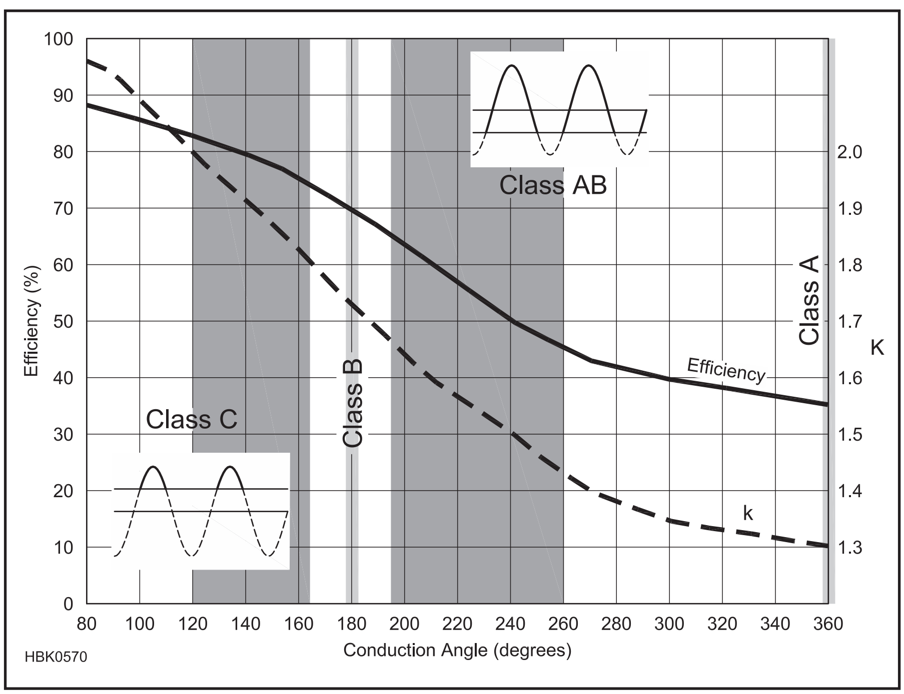


Fig 17.11 — Efficiency and K for various classes of operation. Read the solid line to determine efficiency. Read the dashed line for K, which is a constant used to calculate the plate load required, as explained in the text.

The Flywheel Effect

The operation of a resonant tank circuit (see the **Oscillators and Synthesizers** chapter for a discussion of resonant LC circuits) is sometimes referred to as the "flywheel" effect. A flywheel does illustrate certain functions of a resonant tank, but a flywheel alone is non-resonant; that is, it has no preferred frequency of operation.

A better analogy for a resonant tank — although not exact — is found in the balance wheel used in mechanical watches and clocks in which 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 an 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.

A plot of the rotational (or 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.

The plate (or collector or drain) voltage is a sine wave, even though the plate current is made up of pulses. Just as the escarpment mechanism must apply its kicks at just the right time, the frequency at which current pulses are added to the tank 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.

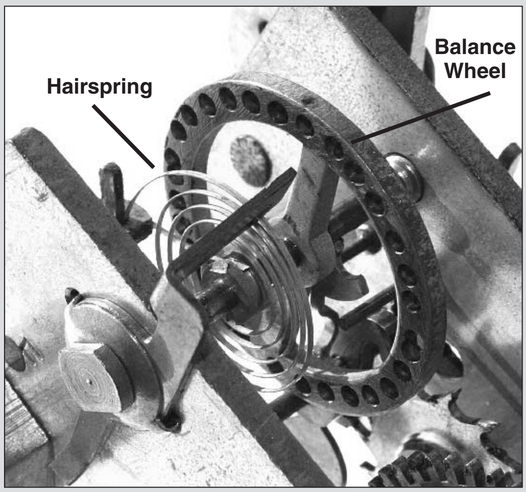


Fig 17.A1 — A balance wheel and hairspring in a mechanical clock illustrate the flywheel effect in tank circuits.

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-ROM, 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 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

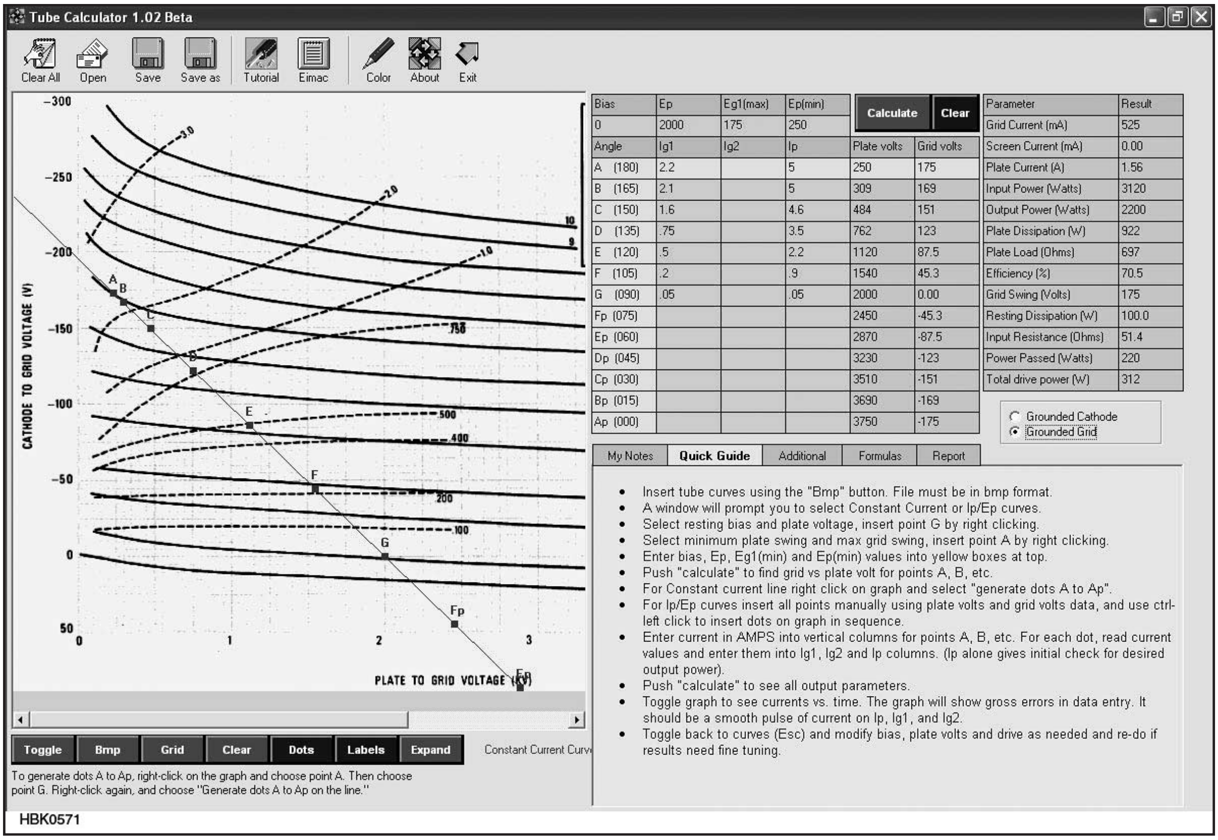


Fig 17.12 — Main screen of *TubeCalculator* program (included on the *Handbook* CD). A characteristic curve plot is loaded in the window on the left side of the screen.

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.

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."

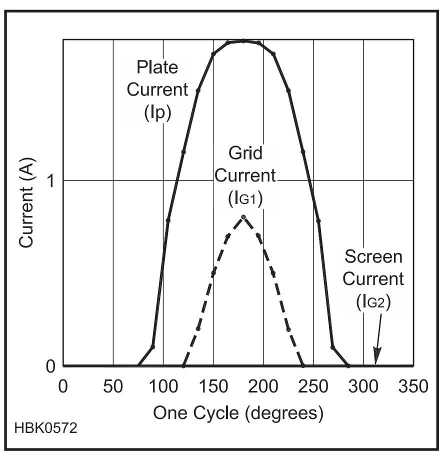


Fig 17.13 — Class AB2 plate and grid current over one cycle as plotted by the *TubeCalculator* program. This plot is for a triode, so there is no curve for screen (G2) current.

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

$$R_L = \frac{V_p}{K \times I_p} \qquad (1)$$

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, $K \approx 1.3$; Class AB, $K \approx 1.5$-1.7; Class B, $K \approx 1.57$-1.8; Class C, $K \approx 2$. 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

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad (2)$$

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.

$$Q = 2\pi \frac{W_S}{W_L} \qquad (3)$$

where

W_S = is the energy stored

W_L = the energy lost to heat and the load

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:

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

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

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

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

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

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

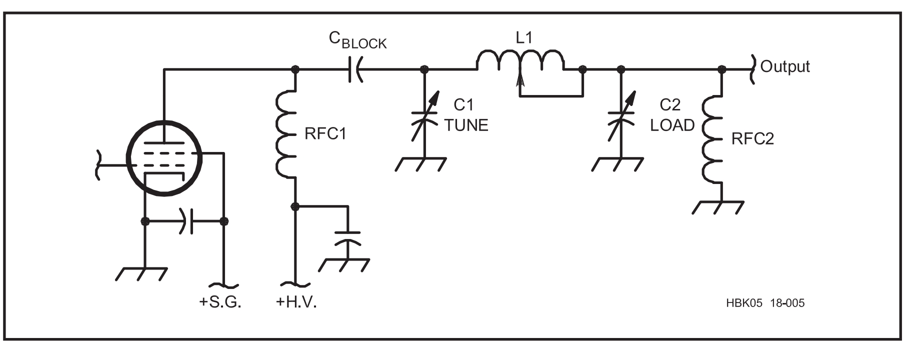


Fig 17.14 — A pi matching network used at the output of a tetrode power amplifier. RFC2 is used for protective purposes in the event C_{BLOCK} fails.

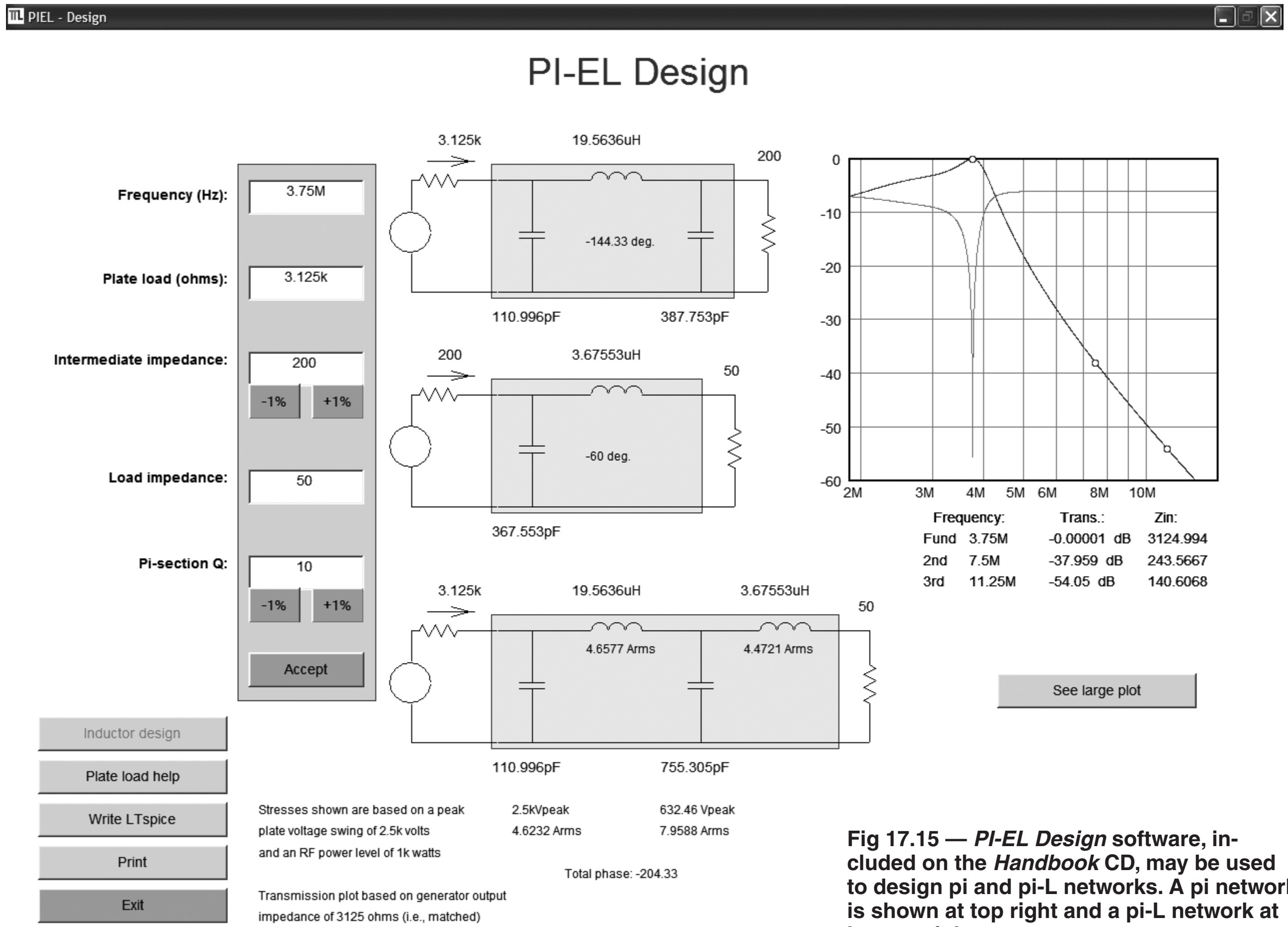


Fig 17.15 — *PI-EL Design* software, included on the *Handbook* CD, may be used to design pi and pi-L networks. A pi network is shown at top right and a pi-L network at bottom right.

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.

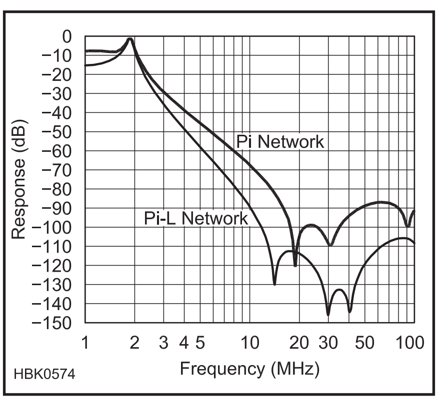


Fig 17.16 — Relative harmonic rejection of pi and pi-L circuits

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

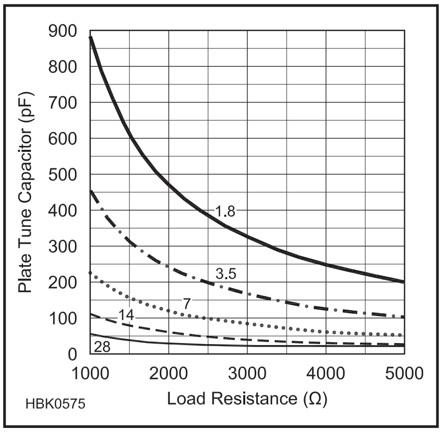


Fig 17.17 — Plate tuning capacitor values for various bands and values of load resistance. Figs 17.17 through 17.19 may be used for manual design of a tank circuit, as explained in the text.

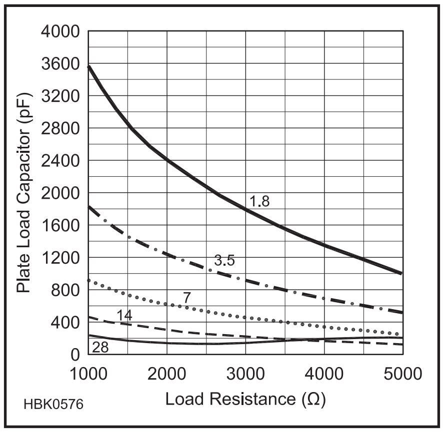


Fig 17.18 — Plate loading capacitor values for various bands and values of load resistance.

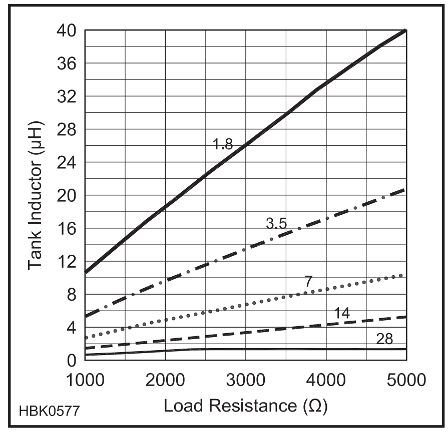


Fig 17.19 — Tank inductor values for various bands and values of load resistance.

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 website, **www.linear.com**. The *PI-EL Design* software mentioned above generates files for *LTspice* automatically. *PI-EL Design* assumes that the 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 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

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

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 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 non-inductive. 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.

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, al-

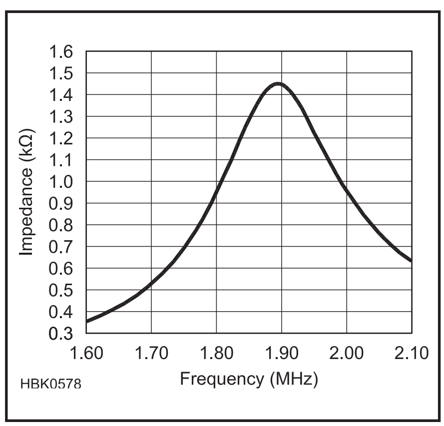


Fig 17.20 — Impedance as would be measured at the plate of an amplifier tube with the output network tuned to 1.9 MHz and the output terminated in its proper load.

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 following procedures may not be exactly the same as the manufacturer's directions 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*

though 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 non-inductive 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 Tube 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

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

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 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.

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 1/8"
	18-28	6 or 3/16"
500	1.8-3.5	12
	7-14	10
	18-28	8 or 1/8"
150	1.8-3.5	16
	7-14	12
	18-28	10

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

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

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 ⅜ 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 foot 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!

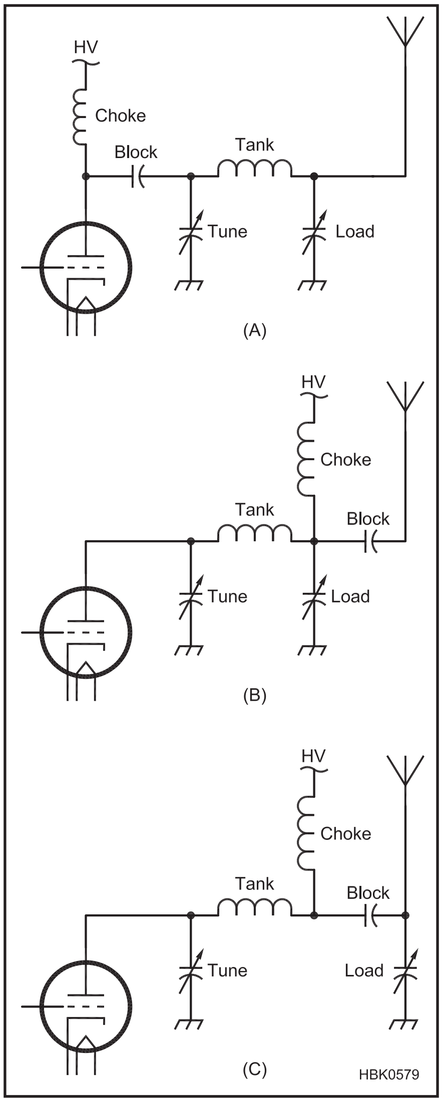


Fig 17.21 — Three ways of feeding dc to a tube via an RF choke. See text for a discussion of the tradeoffs.

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 ¾ 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, 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 de-rated 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 Sources** 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.

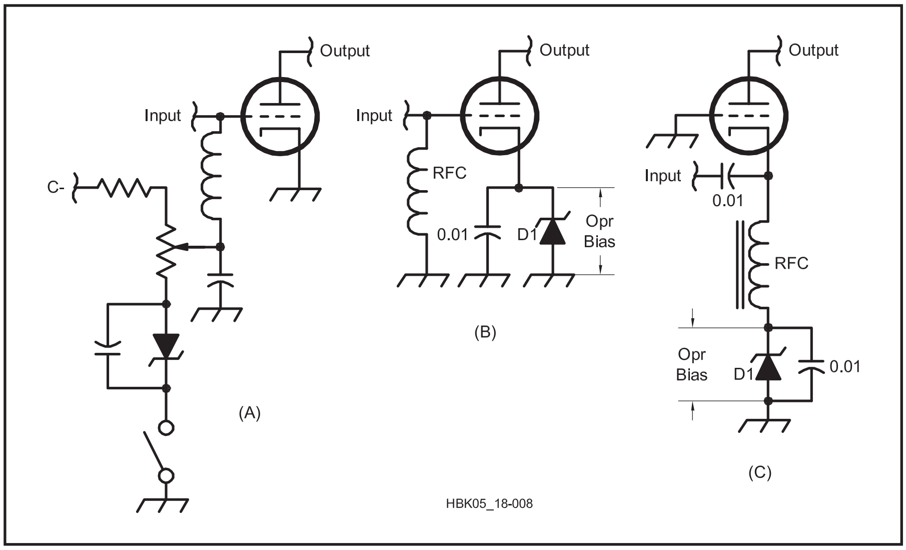


Fig 17.22 — Various techniques for providing operating bias with tube amplifiers.

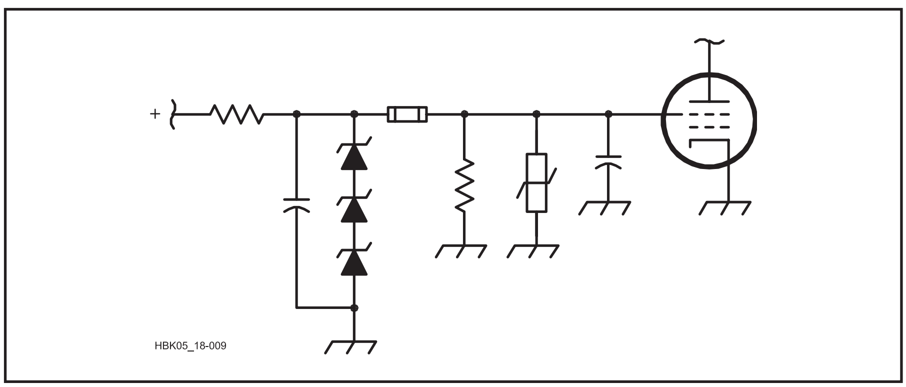


Fig 17.23 — A Zener-regulated screen supply for use with a tetrode. Protection is provided by a fuse and a varistor.

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 bleeder 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.

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.

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.

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. 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

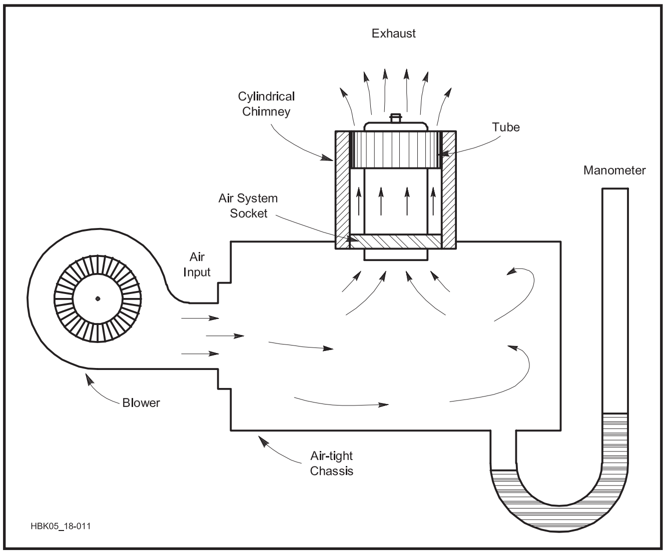


Fig 17.24 — Air is forced into the chassis by the blower and exits through the tube socket. The manometer is used to measure system back pressure, which is an important factor in determining the proper size blower.

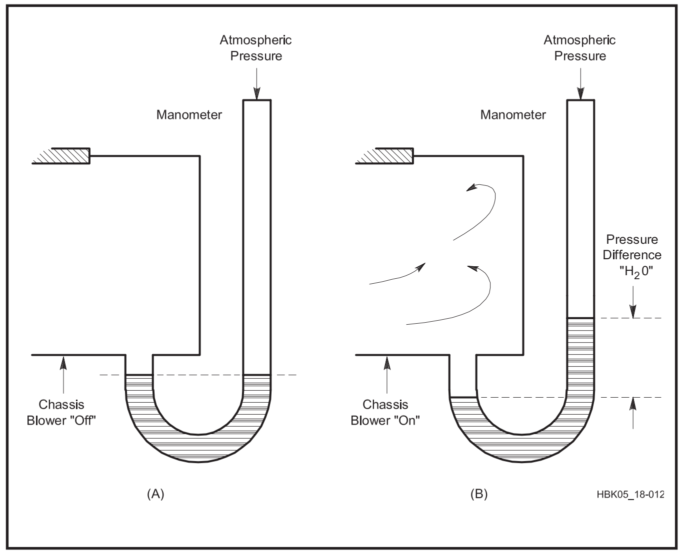


Fig 17.25 — At A the blower is "off" and the water will seek its own level in the manometer. At B the blower is "on" and the amount of back pressure in terms of inches of water can be measured as indicated.

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
2 15⁄16"	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
3 15⁄16"	1 15⁄16"	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**).

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.

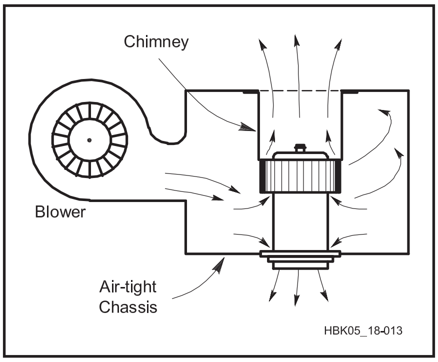


Fig 17.26 — Anode compartment pressurization may be more efficient than grid compartment pressurization. Hot air exits upwards through the tube anode and through the chimney. Cool air also goes down through the tube socket to cool tube's pins and the socket itself.

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 Vacuum Tube 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 its 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**.

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

$$\frac{C1}{C3} = \frac{C_{gp}}{C_{IN}} \qquad (8)$$

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.

Neutralization adjustment is accomplished

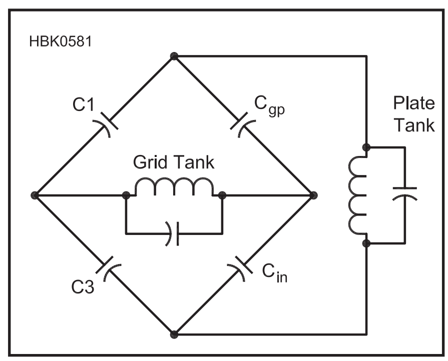


Fig 17.29 — The "bridge neutralization" circuit of Fig 17.28 redrawn to show the capacitance values.

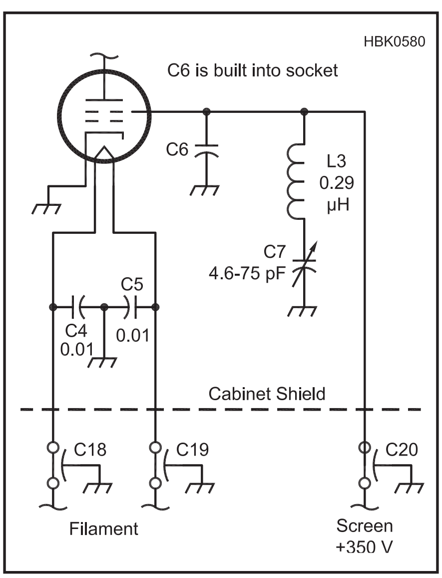


Fig 17.27 — A series-resonant circuit can be used to provide nearly perfect screen or grid bypassing to ground. This example is from a single-band 50 MHz amplifier.

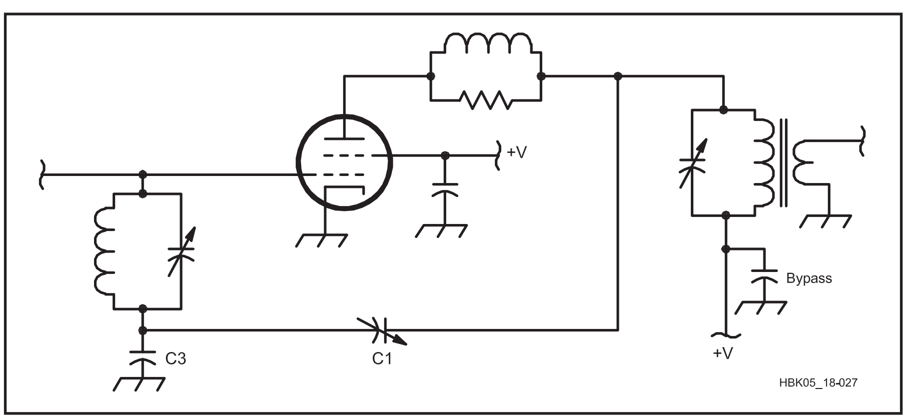


Fig 17.28 — A neutralization circuit uses C1 to cancel the effect of the tube internal capacitance.

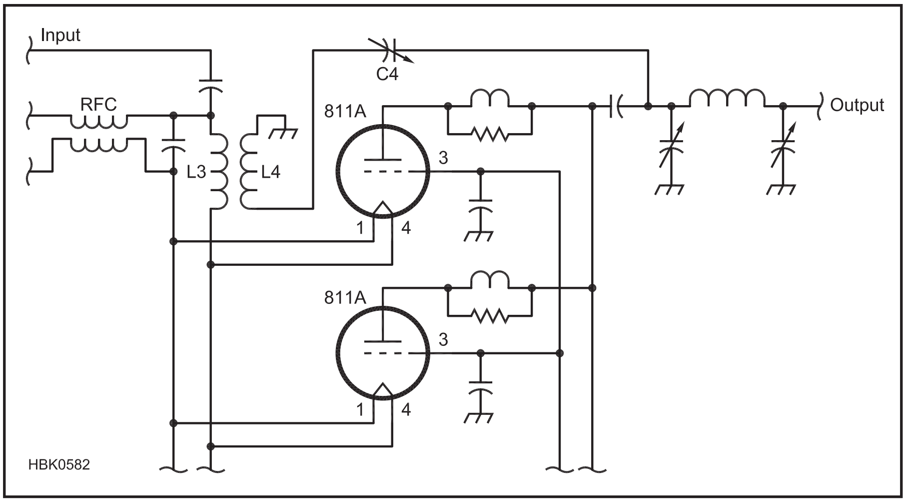


Fig 17.30 — In this neutralizing method, a broadband transformer (L3, L4) provides the needed out-of-phase signal. L3 is 6 turns of #14 wire close wound, ½ inch diameter. L4 is 5 turns of insulated wire over L3. C4 is 6 pF with 0.06-inch spacing. This circuit was originally featured in June 1961 *QST*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

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 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

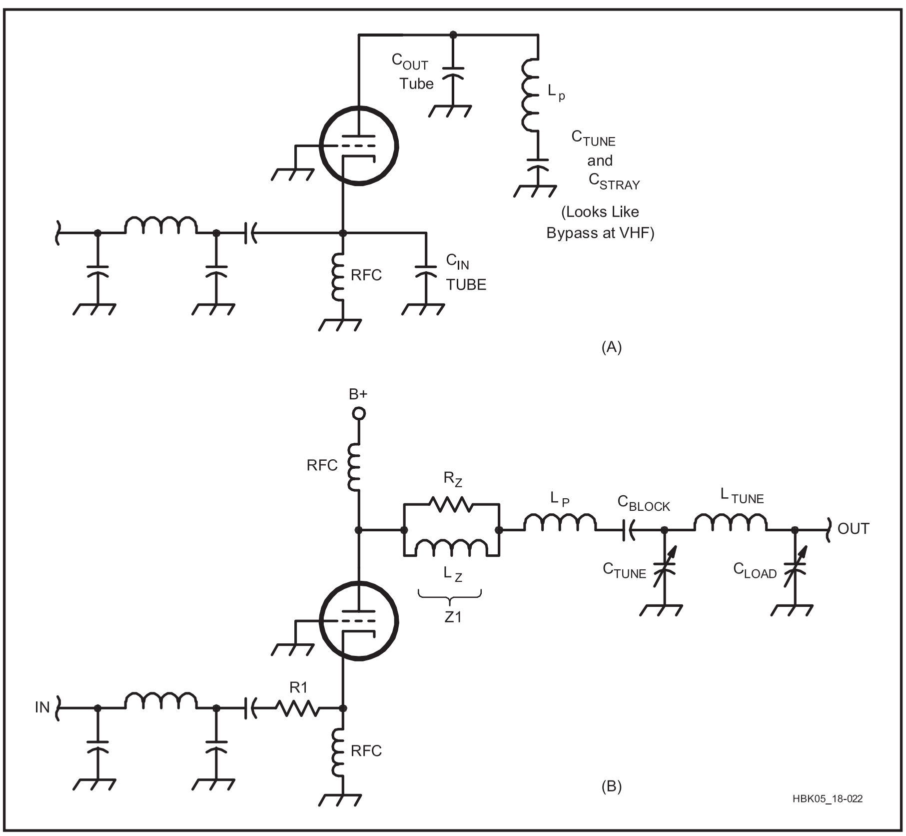


Fig 17.31 — At A, typical VHF/UHF parasitic resonance in plate circuit. The HF tuning inductor in the pi network looks like an RF choke at VHF/UHF. The tube's output capacitance and series stray inductance combine with the pi-network tuning capacitance and stray circuit capacitance to create a VHF/UHF 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.

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 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.

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 net-

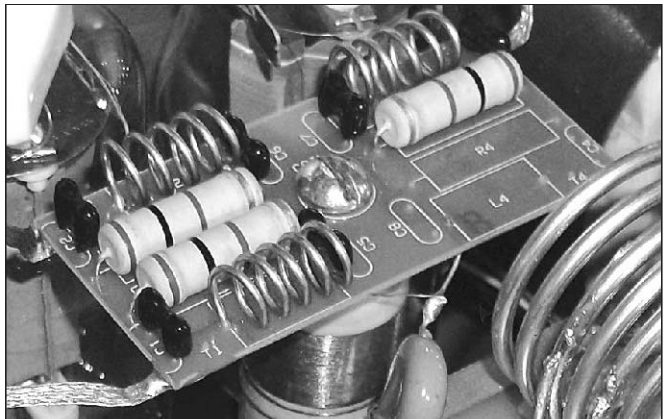


Fig 17.32 — Typical parasitic suppressor.

work 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.

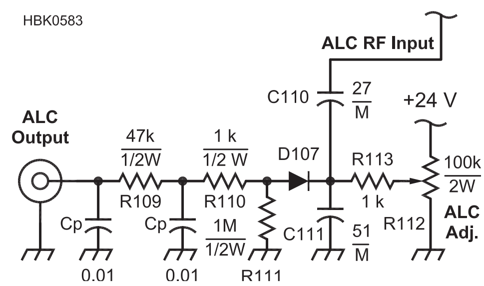


Fig 17.33 — An amplifier ALC circuit can be used to automatically limit the drive power from the transceiver to a safe level.

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 than 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.

17.9 MOSFET Design for RF Amplifiers

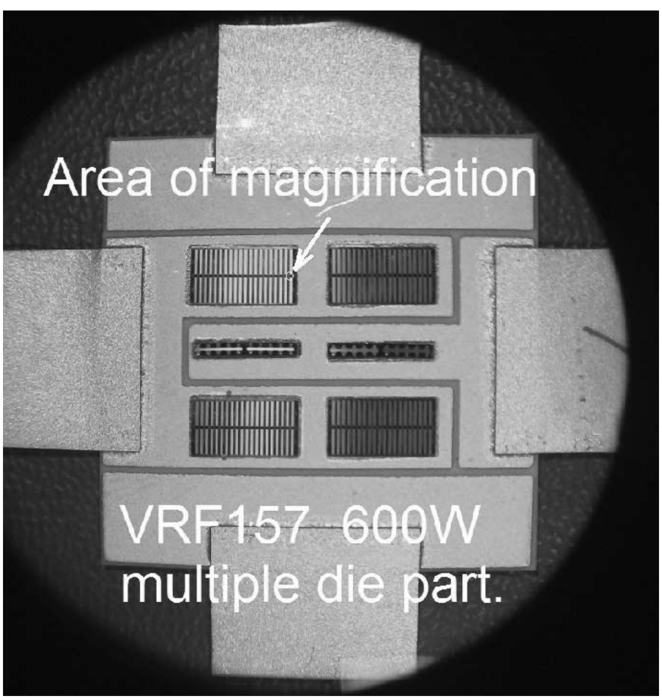
There are two general classes of MOSFETs: high and low frequency designs. (See the **Analog Basics** chapter for MOSFET basics.) The low frequency types are generally optimized for high volume commercial switching applications: computer power supplies, motor controllers, inverters, and so on. They have molded plastic packages, the die are made with aluminum top side metallization, they have maximum junction temperature ratings of 150 to 175°C. Most have polysilicon gate conductors. Polysilicon is easy to manufacture consistently and it's cheap. This works very well for applications up to 200 kHz, but the gate losses start to increase dramatically when they are used at higher frequencies.

A MOSFET gate is essentially a capacitor, but its folded structure is long and skinny. The gate in a 500 W device may be more than a meter long! ("Meter" is not a misprint.) If its conductor material is a lossy material like polysilicon, the gate becomes a long distributed RC network. If an RF signal does make it all the way to the end it will be attenuated and no longer be in phase with the start. This effectively reduces the useful area of the device as frequency increases. It takes a lot of RF current to feed the gate capacitance: I = CVf, where C is the gate capacitance, V is the peak gate voltage, and f is the operating frequency. If the gate capacitance is 500 pF and is being driven to 10 V at 30 MHz, the gate current is 150 mA RMS. While the current is directly proportional to the frequency, the power loss is I^2R. If the gate's top conductor is not low loss, it will fail due to I^2R losses at the gate bond pad (the metallization melts) when used at frequencies much higher than it was designed for.

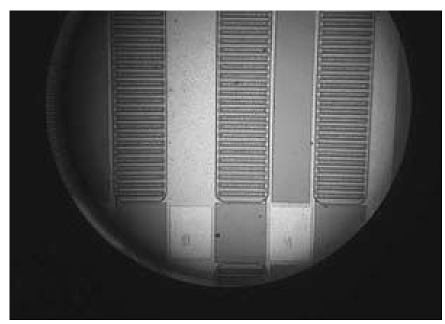
There are two MOSFET manufacturers that use a metal gate instead of polysilicon for their switchmode devices, IXYS and Microsemi. While the die of these devices are quite capable of HF operation, their packaging (usually TO-247 or TO-264) does not provide an optimum HF layout. Because these are aimed at the switchmode market, their drain terminal is on the mounting surface and the source bond-wire length adds gain-killing degeneration at higher frequencies. But on the other hand, they are cheap in terms of cost per watt of dissipation and are acceptable for single-band designs through 20 meters.

When these same metal-gate MOSFET die are placed in packages that are specifically for RF use, the source is often connected to the mounting surface of the package. The source bond wires are thus short, which improves the available gain at all frequencies. This is very convenient because the source is grounded in most RF power amplifier circuits. It also eliminates the need for a mounting insulator, which in turn improves the power dissipation capability.

In MOSFETs specifically designed for RF, the main distinguishing feature is the gate structure. The channels are "shorter" (there is less distance between the gate and source) which reduces the transit time for electrons. As the active area of a device is increased by making the channel "wider," its power dissipation capability is increased. At the same time, the intrinsic (inter-electrode) capacitances also get bigger. A larger device is more difficult to use at higher frequencies because the input impedance (mostly gate capacitance) becomes ever smaller, which makes it harder to drive. In order to solve the gate loss problem mentioned earlier, the long skinny gate is folded into a comb shape. (See **Fig 17.34**.) The gate signal now only has to travel to the end of each finger. The highest frequency designs use multiple combs with shorter fingers. Several of these comb struc-



(A)



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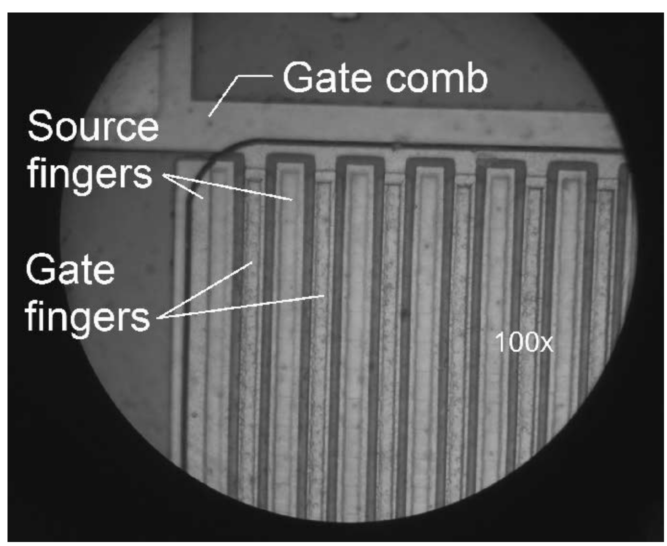


(C)

Fig 17.34 — A shows the layout of a multiple-die RF MOSFET (VRF157). B illustrates the comb structure of the gate. C is a closeup of the gate showing the interleaved source and gate finger structure. [Dick Frey, K4XU, photos]

tures are arrayed on the die and are connected in parallel when the die is wire-bonded in the package.

The top metallization for RF parts is either aluminum or gold. Aluminum is cheaper but gold is best because it has a higher operating temperature rating, up to 225 °C, and it is immune to power cycling failures due to its excellent ductility. The downside is that it is more expensive and the devices are much harder to manufacture because gold likes to dissolve into silicon. Its higher temperature rating means you can get more power from a small device, which offsets their higher cost somewhat.

17.9.1 LDMOS versus VDMOS

So far all the devices discussed are vertical MOSFETs, or VDMOS. Their gate and source electrodes are on the top surface of the die and the drain is on the bottom. For RF applications there is another type, LDMOS. This is a lateral device. Here the MOS structure is laid on edge and all the electrodes are on the top side of the die. Vertical p+ source connections are made through the die to make the bottom side of the die a source contact in order to get the optimum "common source" configuration. The channel area is low so the capacitances are smaller, especially the feedback capacitance, C_{GD}. However, the operating voltage capability is also low. There are none rated for more than 50 V operation. The gates are particularly sensitive to ESD and overdrive. However, they have spectacular high frequency capability and gain, and reasonable ruggedness. Your mobile phone would not work without LDMOS technology.

New RF amplifier designs are using LDMOS to replace bipolar transistors, which manufacturers are no longer making. The ability of LDMOS to operate at lower voltages is well suited for 12 V operation and, with its high gain and frequency response, providing 6 meter capability is simple. The downside is that these devices are not as linear as the bipolar transistor they replace.

Bipolar transistors need emitter ballasting resistors in each tiny bipolar cell so they can be paralleled in the die. The resistors also provide negative feedback, which improves linearity. MOSFETs do not require source resistors: paralleled cells will naturally share the load because their ON resistance has a positive temperature coefficient that prevents thermal runaway. As a result, LDMOS amplifiers require more negative feedback to provide comparable IMD performance, which offsets their higher gain advantage.

Regardless of the device type, the packaging is particularly important to an RF device. It must have low parasitics (see the **RF Techniques** chapter) and superior thermal qualities. The package insulator is made of ceramics, beryllia BeO (which is toxic), and/or alumina Al_2O_3, for high temperature capability. The conductors are gold-plated copper or Kovar, and the base flanges are often copper-tungsten or copper-molybdenum. The package is the major determining factor in the cost of an RF part. Parasitic inductance introduced by gate and source bond wires limits the ultimate frequency capability of a VDMOS part. LDMOS parts have gate and drain bond wires. LDMOS devices have an advantage in terms of frequency and package cost because they are free of the gain degeneration caused by source wire inductance and their die may be soldered directly to a metal mounting flange.

17.9.2 Designing Amplifiers with MOSFETs

Designing an amplifier requires a systems approach. You will need to consider how much power supply is required, as well as the cooling and control systems needed to keep it happy. If you have the transistors and want to build them into an amplifier, the design procedure is a little different. The place to start in any case is with its transistor's data sheet. This will show the voltage and power capabilities, and from these the circuit requirements can be calculated and the cooling system defined.

VOLTAGE RATINGS

Designing an amplifier with MOSFETs requires knowledge of the part being used. Generally, the cheap plastic-packaged switchmode parts will be best for single-band operation. Switchmode parts are rated by their drain breakdown voltage (BV_{DSS}), ON resistance (R_{DS}), and power dissipation. They are available in voltage ratings from as little as 5 V to over 1200 V. RF parts are sold by operating voltage, V_{DD}, power dissipation and frequency capability.

Select the part to suit your power supply requirements. A 500 V MOSFET will not work on a 12 V supply and will barely work at 50 V. This is because the MOSFET's intrinsic capacitances are higher at low voltage. Between the drain and source of every MOSFET is a parasitic "body diode" as shown in **Fig 17.35**. It's too slow to rectify RF, but like any diode, its capacitance changes with reverse bias voltage. This relationship is always given in the device's C-V curves on its data sheet. (See **Fig 17.36**.) Data sheets can be found on manufacturer or distributor websites or perform an Internet search for the part number and "data sheet."

In class AB operation, a MOSFET works best when operated at a little less than one-half of its rated breakdown voltage, BV_{DSS}. The drain voltage will swing up to 2 or even 3.562 times (for class E) the supply voltage. Under high VSWR, the drain voltage can be somewhat higher still. The RF voltage breakdown of a MOSFET is typically 20% higher than its data sheet value but is hard to specify reliability, so RF devices are rated by their

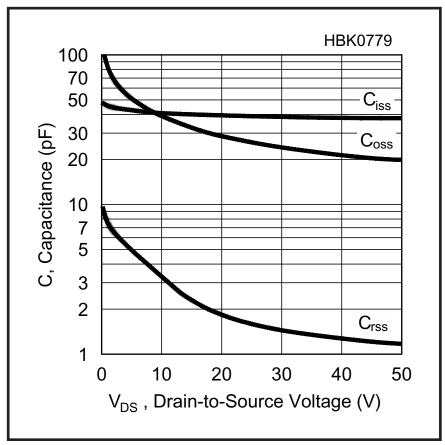


Fig 17.36 — The capacitance versus voltage (C-V) curves for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

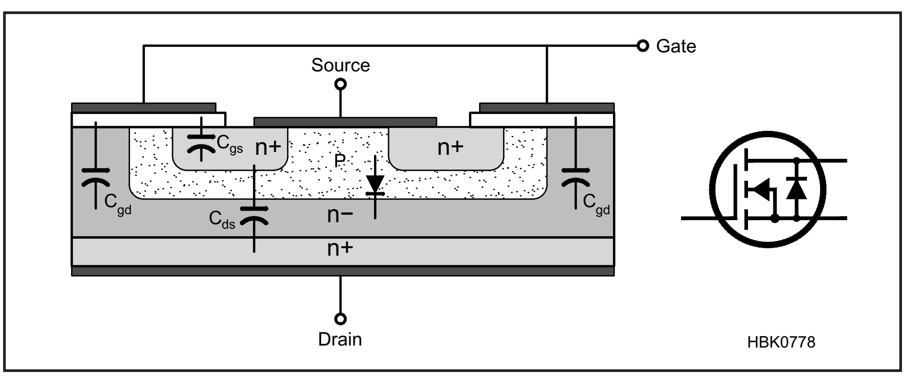


Fig 17.35 — All MOSFETs have parasitic capacitances as shown and a body diode between the drain and source in the cross-section and schematic symbol. The diode is shown for an N-channel device. It is reversed for P-channel devices.

dc operating voltages rather than BV_{DSS}. This takes into account the requirement for operating overhead.

RF parts are usually rated at a specific operating voltage such as 13.5 V, 28 V or 50 V. Originally these were common battery voltages for civilian and military vehicles and the tradition persists. The devices are optimized for their operating voltage. Choosing the operating voltage is a matter of considering many different parameters, not just the breakdown voltage.

THERMAL DESIGN

Suffice it to say that the thermal design of a high power transistor PA is often as challenging as the electrical design. It can be done "by the numbers" but the tricky part is making it fit into the available space. A thermal design example and an Advanced Power Technology application note by the author are provided as a supplemental article on the CD-ROM accompanying this book.

A word of caution is in order: tubes used in power amplifiers have a great deal of "headroom" in their specifications and are quite forgiving of momentary operator errors. RF power transistors, because they are more expensive in terms of cost per watt, are specified much closer to their limits. These limits must be observed at all times. Even though the data sheet says the device can do X watts, the designer must observe the requirements for proper cooling in order to reach this level in practice. In addition, manufacturers rate the power dissipation in theoretical terms. You will be lucky to achieve half of it.

As with tubes, there are CW ratings and SSB ratings. For transistors, the ratings are based on average power. The difference is simply the size of the heat sink required, as the peak power is the same for each. Each transistor has a thermal rating expressed as $R_{\theta JC}$, the thermal resistance from the transistor junction to the bottom of its case. Since the device has an upper junction temperature limit, somewhere between 150 and 200 °C, the power dissipation is determined by the difference between the junction and the case temperature:

$$P_d = (T_J - T_C) / R_{\theta JC}$$

where P_d is the available power dissipation, T_J is junction and T_C is case temperature. What this says is that without any cooling, the transistor's case will be almost the same as the junction temperature so its power dissipation capability is nearly zero. When placed on a heat sink, the case will be cooler and it then has power dissipation capability. It follows that the better the heat sink, the more power can be dissipated by the transistor.

There is another thermal consideration: the thermal resistance between the case of the transistor and the heat sink, $R_{\theta CS}$. Even if the base of the transistor and the heat sink are flat and smooth, microscopic air gaps still exist. These do not conduct heat and so reduce the net effectiveness of the heat sink. The solution is to use a thermal interface compound or *thermal grease*. The simplest and best is silicone oil loaded with zinc oxide powder. The oil does most of the work: the powder thickens it to a paste so it doesn't run out of the joint. It is applied as a very thin coat between the heat sink and device. When using thermal grease, always wiggle the transistor around on the sink to insure a minimum of grease between the part and the sink. Remember, thermal grease is not a good thermal conductor, it's just much better than air. Use it sparingly!

Most commercial high power broadcast amplifiers are cooled with circulated water, or a water-glycol mix if the minimum ambient temperature will be below 0 °C. While it has yet to be introduced to the amateur market, liquid cooling has great potential. The advances in plastic fittings, small pumps and heat exchangers driven by the high-performance computer market have great potential benefits to the amateur high power amplifier. Water is more than four times better than air for absorbing and moving heat. **Fig 17.37** illustrates both open- and closed-loop cooling systems. But regardless of where it is moved to, the heat must still be dissipated. It can warm the air in the shack, heat the rest of the house, or warm the septic tank.

17.9.3 The Transistor Data Sheet

Regardless of manufacturer, all data sheets contain the same basic information. The following should help make sense of what can be very confusing to a first time user. Transistor specifications rely heavily on several ideal conditions that cannot happen in practice but since it is a common practice by all manufacturers, the numbers are very useful for comparing different devices. As long as you have the part number (and it is not a custom part), the corresponding data sheet can be found easily by searching for the part number and "data sheet."

The Microsemi VRF151 N-channel enhancement-mode VDMOS transistor will be used as an example. It is used in the 250 W broadband amplifier project presented in this chapter. The following discussion assumes the reader has obtained a data sheet for this part (see the company website at **www.microsemi.com**) and can refer to it.

MINIMUM, TYPICAL AND MAXIMUM

All of the specification parameters which the manufacturer guarantees are subject to either a minimum value or a maximum value. This is the worst case. As in an automobile, there is a maximum safe stopping distance and a minimum gas tank capacity. Most parameters also have a *typical* value that is generally representative of typical performance than the specified minimum or maximum. Some quantities have both upper and lower limits. Every parameter has specific test conditions under which it is measured.

ABSOLUTE MAXIMUM RATINGS

These are all dc ratings, easily verified with a variable power supply and a multimeter. If the manufacturer finds that any of these have been exceeded, any warranty claims are voided.

V_{DSS} is the maximum drain to source voltage rating, with the gate is shorted to the source. Think of the device as a high voltage Zener diode. As soon as it draws any current, power is dissipated, and temperature rises

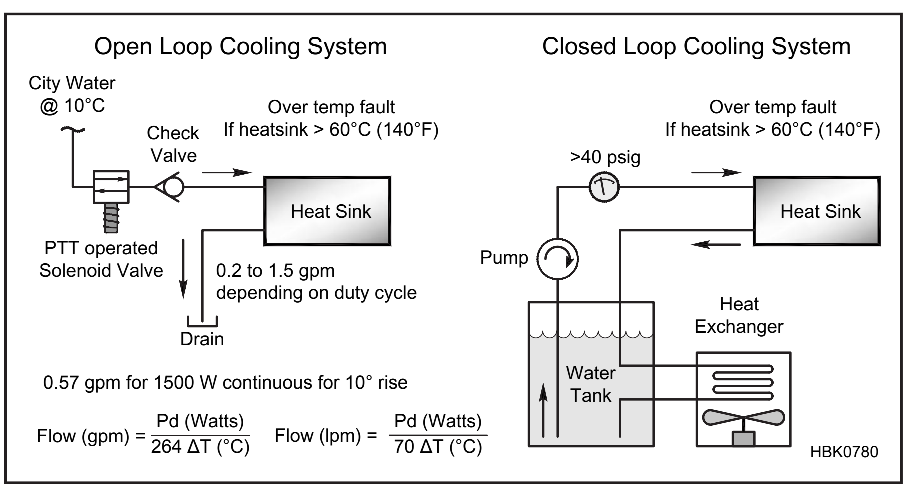


Fig 17.37 — Open- and closed-loop water cooling systems for solid-state amplifiers.

very quickly. Damage occurs either from puncturing through the junction or by arcing over the edge of the die.

Maximum drain current, I_D, is the current that will cause the device to dissipate its maximum rated power when fully turned on. Every device has an ON resistance called $R_{DS(ON)}$. The power dissipated is due to $I_D^2 \times 2.5\ R_{DS(ON)}$. The factor of 2.5 accounts for $R_{DS(ON)}$ having a positive temperature coefficient which causes it to roughly double by the point at which the junction temperature is 200 °C.

V_{GS} is the maximum gate to source voltage. The gate is essentially a capacitor, with a SiO_2 dielectric perhaps 400 to 1000 Angstroms (10^{-10} m) thick. The limit is lower on LDMOS than VDMOS, and cannot be exceeded without destroying the device. Because LDMOS devices have much lower V_{GS} ratings and thinner dielectrics, some LDMOS manufacturers build in diode protection to make them less susceptible to electrostatic discharge, ESD.

P_D is the maximum power dissipation of the device under theoretical conditions: The bottom of the case at 25 °C and the junction at its maximum temperature, T_{jmax}. If not stated directly, the thermal resistance $R_{\theta JC}$ is equal to P_D / (T_{jmax} – 25°C).

Storage temperature is straightforward. Maximum T_J is the junction temperature above which things start to come unsoldered, or the reliability seriously impaired, or smoke emitted.

STATIC ELECTRICAL CHARACTERISTICS

V_{DSS} is specified again, this time showing the measurement conditions, the guaranteed minimum, and the typical production values.

$V_{DS(ON)}$ or sometimes $R_{DS(ON)}$ is the minimum resistance between drain and source that is obtained when the device is fully ON and carrying half the rated current. It is more commonly specified on switchmode parts, but it is of particular importance in high-efficiency saturated modes like class D and E because it limits the maximum obtainable efficiency.

I_{DSS} is the maximum drain current flowing with the gate shorted to the source. In an N-channel enhancement device one must apply a positive voltage to the gate to turn it on. The VRF151 will not conduct more than 1 mA at 100 V by itself without gate bias. This is a leakage current. In a perfect device it is zero.

Similarly, I_{GSS} is gate leakage current with the V_{DS} = 0. It represents a resistor in parallel with the gate capacitor. In this case, 1 μA at 20 V is 20 MΩ. This does not sound like much but if the drain has voltage on it and there is no resistor across the gate, this leakage current will cause the gate to charge from the drain, eventually turning the device fully ON with bad consequences.

Forward transconductance, g_{fs} , is the dc gain of the device expressed in terms of change in drain current per change in gate volts measured at a particular drain current. It has only a mild relationship with RF gain and too high a g_{fs} can cause bias instability and/or parasitics.

V_{TH} is the gate threshold specification. When V_{DS} is10 V, V_{GS} of no less than 2.9 V and no more than 4.4 V will cause 100 mA of drain current to flow. A typical Class AB quiescent bias condition is 100 mA. Of all the parameters, this one has the widest window. Enough variation exists so that manufacturers sort parts into “bins” of values across the range, and assign letter codes to each which are marked on the package. This allows one to make matched pairs within a device type.

THERMAL CHARACTERISTICS

$R_{\theta JC}$ is equal to P_{Dmax} / ($T_{j\ max}$ – 25°C). Specifying the first two parameters generates the third. In this sense it is redundant but is often reiterated for those who are looking for the particular parameter. Because thermal performance is as important as RF performance, much of the information in the data sheet is concerned with it. In addition to the static thermal dissipation, $R_{\theta JC}$, there is a dynamic thermal impedance $Z_{\theta JC}$. Transistor packages have thermal mass. You cannot change its die temperature immediately For pulse operation, the effective $R_{\theta JC}$ is lower than it is for steady-state operation.

The dynamic thermal impedance is shown in **Fig 17.38** (Figure 5 of the VRF151 data sheet). It shows how the device can be used for pulse operation at higher power than it can on CW because the effective thermal impedance is lower for short pulses. This has some application to SSB, but there are other constraints on peak power that boil down to just allowing a smaller heat sink when used only for intermittent service — a bad practice for amateur amplifiers!

DYNAMIC CHARACTERISTICS

Somewhat of a misnomer, these are the parasitic capacitances between the gate, drain and gate. Except for the gate capacitance which is a fixed value based on device dimensions, the other two are a function of the drain voltage, just like varactors (voltage-variable capacitors). Because it is rather difficult to

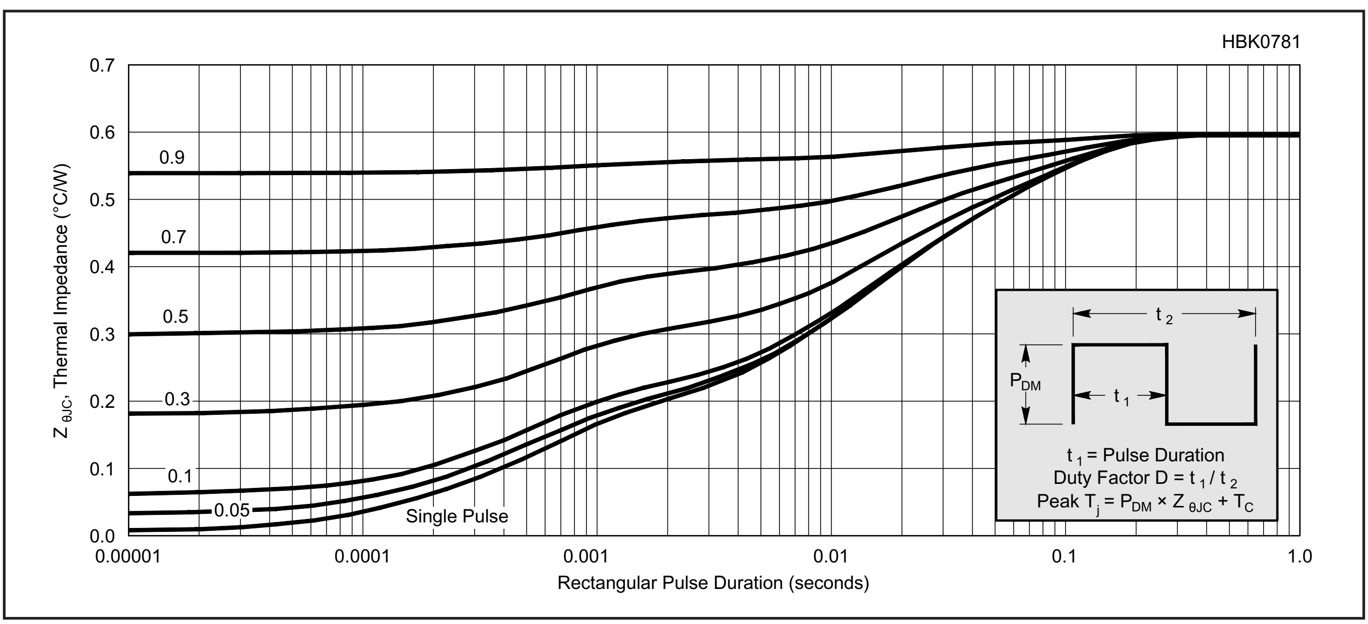


Fig 17.38 — Dynamic thermal impedance for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

measure these parameters, the parameters are defined in a common-source configuration. C_{ISS} is the gate-to-source capacitance with the drain ac-shorted to the source. It is actually C_{GS} in parallel with C_{GD}. The three parameters are usually measured at the specified drain supply voltage with V_{GS} at zero. They are also usually displayed as in Fig 17.36 (Figure 3 of the data sheet), a graph of C vs drain voltage.

This voltage-varying capacitance is one of the causes of IMD in a transistor. Its effect is to impart some phase modulation (PM) on the signal. It can be observed as unequal IMD products on each side of the carrier pair. The PM distortion has pairs of odd-order carriers like AM distortion but the high side carriers are –180° out of phase with the AM products. This causes them to reduce the level of amplitude products on the high side of the carrier pair. (See the reference list entry for Sabin and Schoenike, *Single Sideband Systems and Circuits*.)

TRANSFER CHARACTERISTICS

As discussed above, g_{fs} describes the gain of a MOSFET as the change in drain current per change in gate voltage: $\Delta I_D / \Delta V_{GS}$. This means that g_{fs} is the slope of the transfer curve. The transfer curve for the VRF151 is shown in **Fig 17.39** (the data sheet's Figure 2). There are three curves in this graph, depicting the transfer characteristic at three different temperatures.

The three curves cross each other at the "thermal neutral point." This is where the temperature coefficient of V_{TH} changes from negative to positive and it is usually at a current much higher than the part normally operates. This explains why MOSFETs must have thermally compensated gate bias. Below the crossover point, where the part would be biased for class AB, the temperature coefficient of V_{TH} is negative. This means the gate voltage required for a given current goes down as the part heats up. Without thermal bias compensation we can have thermal runaway. The part will usually melt before it reaches the crossover point.

The transfer curve also shows that the gain of the device is not constant. It is quite low at low current and increases to a nominal value over its linear range and then the curves flatten out at higher current as the part saturates. This demonstrates why very low distortion amplifiers are usually operated in class A.

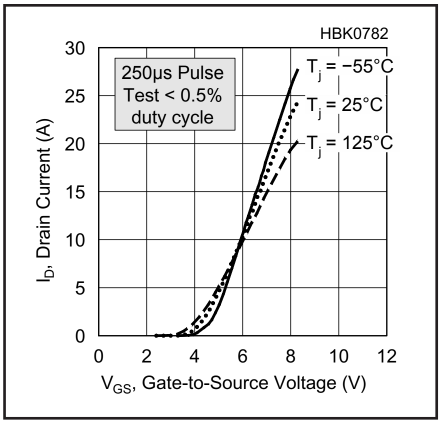


Fig 17.39 — Transfer characteristics for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

FUNCTIONAL CHARACTERISTICS

This section is where the RF performance is specified — how much gain at what frequency, IMD performance, and ruggedness. While gain and IMD are easily understood, ruggedness is more difficult and it is very poorly defined by most manufacturers. The VRF151 has a ruggedness specification at all phase angles of a 30:1 VSWR when putting out 150 W PEP at 30 MHz. The test circuit is shown on page 4 in the data sheet. Nothing is said about how long the test takes or how well the device is cooled. If the test time and cooling conditions are not given, the specification is inadequate to guarantee that a design will remain within the device power limits. In the author's experience, 90% of all VSWR-related failures are due to over-dissipation. This says that limiting the amplifier's drain current is a simple and effective means for VSWR protection.

DATA SHEET EXTRAS

Test circuits are usually provided so customers can duplicate the test conditions for gain, IMD and output power. Sometimes circuit layouts and complete part lists are also provided. Note however, that while most high power parts are usually used in push-pull circuits, parts in data sheets are always tested one at a time in single-ended circuits. The last page of the data sheet gives the mechanical outline and sometimes mounting information. The VRF151 is "binned" (sorted into similarly performing groups) for V_{TH} as it exits the final testing and the bin letter code is marked on the package. This allows very close matching of gate threshold which is important when used in a push-pull circuit with a common bias supply for both parts. Most designers use separate bias adjustments regardless of matching, but it insures a measure of gain matching also, especially if the parts are from the same lot date code.

17.9.4 Summary Observations

MOSFETs are not perfect devices. g_{fs}, V_{TH}, BV and R_{DS} are all affected by die temperature. g_{fs} goes down with increasing die temperature. While this might cause the output power to sag a bit as the amplifier gets hotter, it is generally a benefit. MOSFETs can be paralleled without requiring source resistors because as the one carrying more current heats up, it loses gain and its resistance goes up, allowing the others to share the load. R_{DS} rises with temperature, a useful trait in hard-switching applications using paralleled devices. Breakdown voltage, BV_{DSS}, increases with temperature and so is usually not a concern in a part that heats up as it is being used. It has implications in cases of extreme cold, as in satellites.

One caveat when paralleling MOSFETs: always be mindful that they will easily oscillate at UHF. The device's inter-electrode capacitances and bonding wire inductances conspire to form a cross-coupled multivibrator. The solution is to place a small resistor in series with the gate leads, 3.3 Ω will usually suffice. This lowers the circuit Q enough to prevent oscillation from starting. [ref: Motorola *Engineering Bulletin 104*, et al]

17.10 Solid State RF Amplifiers

17.10.1 Solid State vs Vacuum Tubes

Solid state amplifiers have become the norm in transceivers, 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 as the broadcast and industrial RF industry converts to solid state, making RF power transistors available for amateur use at lower prices. A number of legal-limit solid state Amateur Radio amplifiers are available.

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 RF amplifiers increased linearity, mainly in 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 common 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 in order to prevent key clicks.

Class D and E are most efficient, up to 95% in practical circuit applications. But both of these both require a narrow band tuned tank circuit to achieve this efficiency. They are not linear; their output is essentially either on or off. They can be used quite effectively 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 transistor models they use. 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 also be designed before the circuit can be modeled. 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 particularly useful for looking at parts of the whole.

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. *SVC Filter Designer* by Jim Tonne, W4ENE, 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 up 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

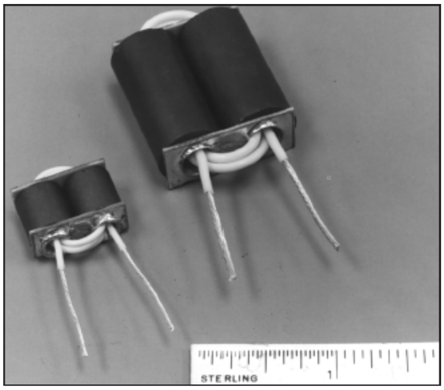
Broadband transformers are often used in matching to the input impedance or optimum load impedance in a power amplifier. Unlike the tuned matching circuits, transformers can provide constant impedance transformation over a wide range of frequency without tuning. Multi-octave power amplifier performance 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.

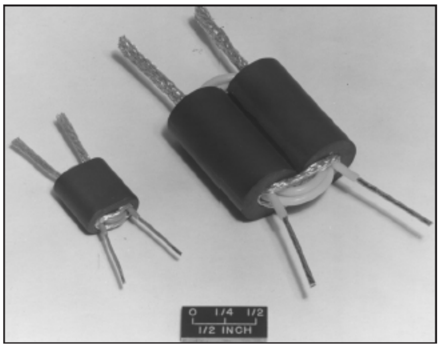
The conventional transformer is wound much the same way as a power transformer. Primary and secondary windings are wound around a high-permeability core, usually made from a ferrite or powdered-iron material. Coupling between the secondary and primary is made as tight as possible to minimize leakage 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 impedance 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



(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.

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. Because of the increased coupling between the primary and secondary of the transformer made with multiple pieces of coax, leakage reactance is reduced and bandwidth performance is increased.

The cores used must be large enough so the 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, the reactance of 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 caused 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 the square of their turns ratios. Those with single turn primaries are limited to integer values of 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 impedance ratio is 50/3.125 = 16, which is provided by 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. There are higher voltage devices developed for the industrial markets and they are gradually finding their way into amateur designs. Being able to adjust the impedance to a convenient value for a common transformer turns ratio by adjusting the operating voltage is a powerful design option. These device and transformer limitations on output power can also 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 higher voltage approach quite attractive.

TRANSMISSION LINE TRANSFORMERS AND COMBINERS

Fig 17.41 illustrates, in skeleton form, how transmission-line transformers can be used in a push-pull solid state power amplifier. The idea is to maintain highly balanced stages so that each transistor shares equally in the amplification in each stage. The balance also minimizes even-order harmonics so that low-pass filtering of the output is made much easier. In the diagram, T1 and T5 are current (choke) baluns that convert a grounded connection at one end to a balanced (floating) connection at the other end, with a high impedance to ground at both wires. T2 transforms the 50 Ω generator to the 12.5 Ω (4:1 impedance) input impedance of the first stage. T3 performs a similar step-down transformation from the collectors of the first stage to the gates of the second stage. The MOSFETs require a low impedance from gate to ground. The drains of the output stage require an impedance step up from 12.5 Ω to 50 Ω, performed by T4. Note how the choke baluns and the transformers collaborate to maintain a high degree of balance throughout the amplifier. Note also the various feedback and loading networks that help keep the amplifier frequency response flat.

Three methods are commonly used to combine modules: parallel (0°), push-pull (180°) and quadrature (90°). In RF circuit design, the combining is often done with special types of "hybrid" transformers called splitters and combiners. These are both the same type of transformer that can perform either function. The splitter is at the input, the combiner at the output.

Fig 17.42 illustrates one example of each of the three basic types of combiners. In a 0° hybrid splitter at the input the tight coupling between the two windings forces the voltages at A and B to be equal in amplitude and also equal in phase if the two modules are identical. The 2R resistor between points A and B greatly reduces the transfer of power between A and B via the transformer, but only if the generator resistance is closely equal to R. The output combiner separates the two outputs

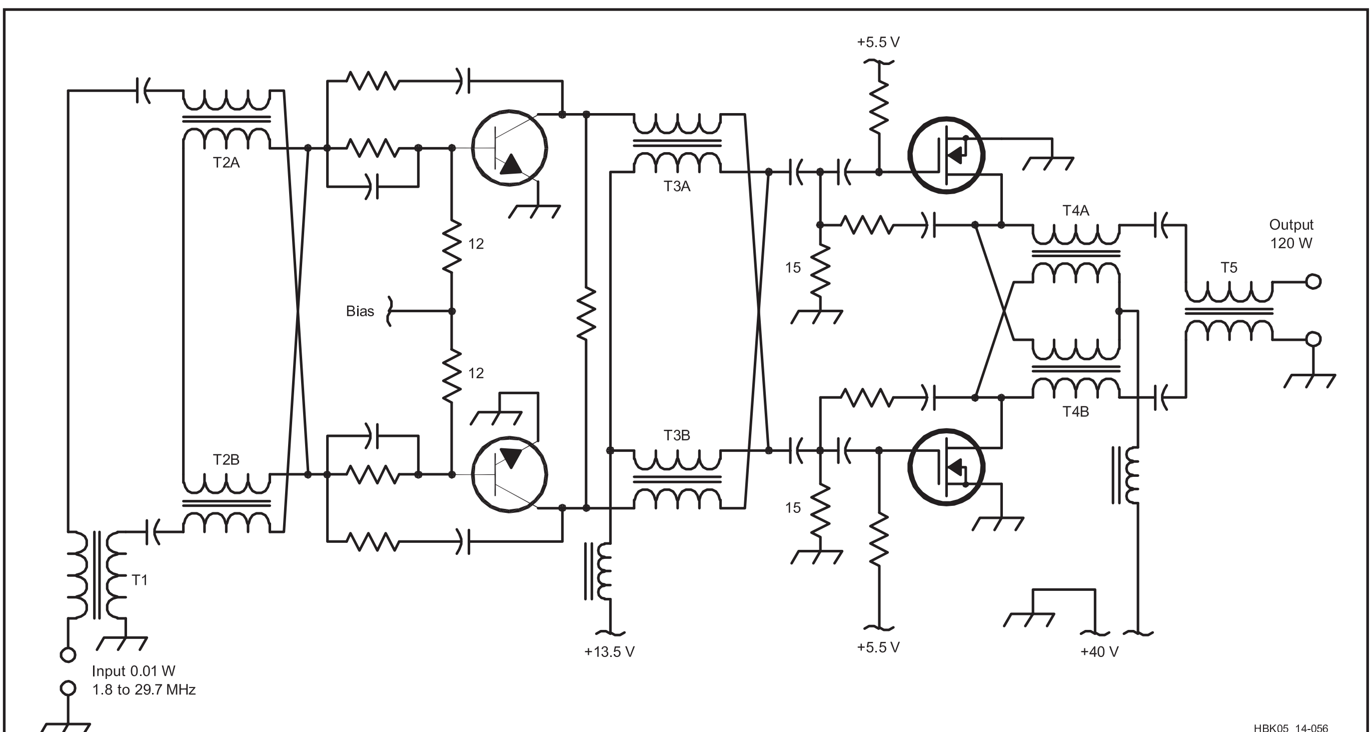


Fig 17.41 — Typical use of transmission line transformers as baluns and combiners in solid state power amplifiers.

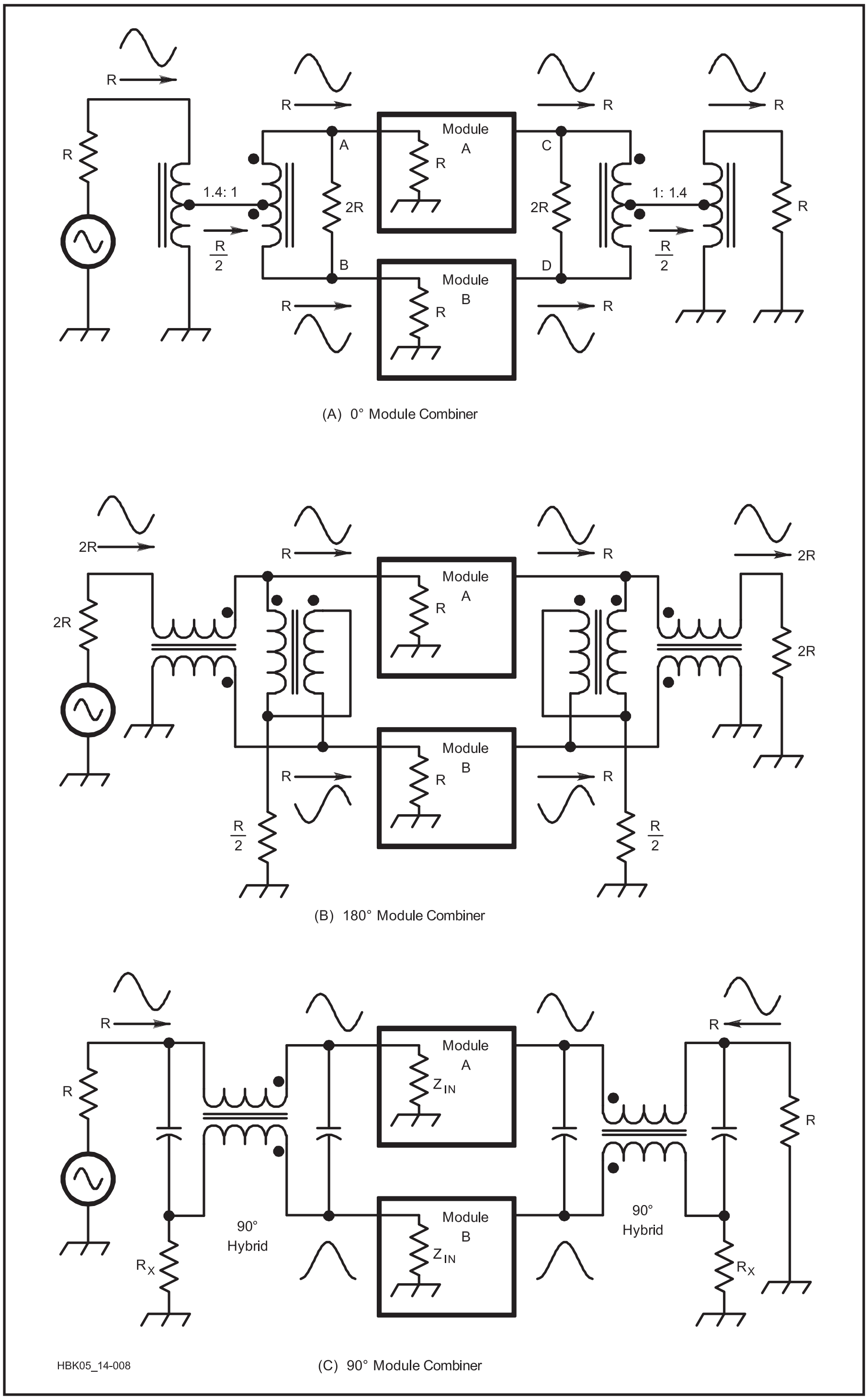


Fig 17.42 — Three basic techniques for combining modules.

C and D from each other in the same manner, if the output load is equal to R, as shown. No power is lost in the 2R resistor if the module output levels are identical. This section covers the subject of combiners very lightly. We suggest that the reader consult the considerable literature for a deeper understanding and for techniques used at different frequency ranges.

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 or away from the operating frequency 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. Unless 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 17.43A**. 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 17.43B 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 17.43C can be added to some

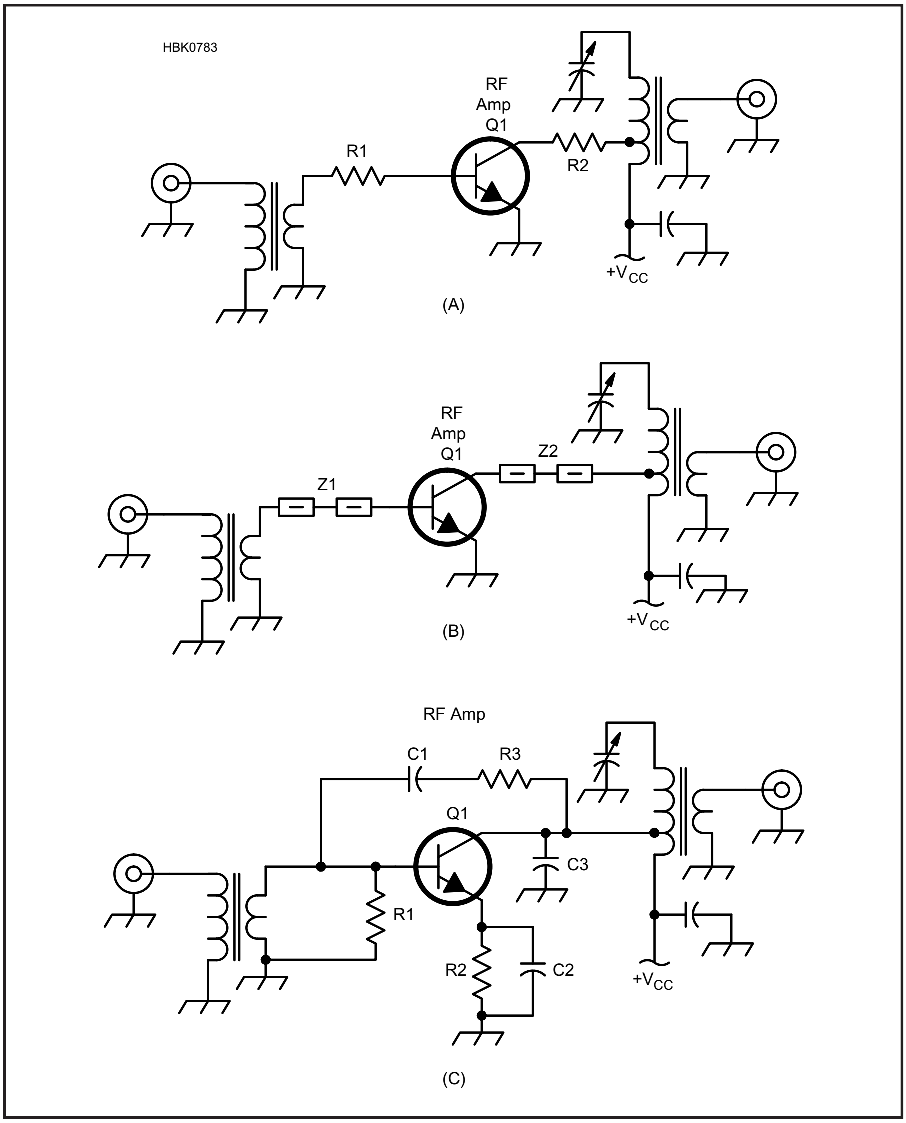


Fig 17.43 — Suppression methods for VHF and UHF parasitics in solid state amplifiers. At A, small base and collector resistors are used to reduce circuit Q. B shows the use of ferrite beads to increase circuit impedance. In circuit C, C1 and R3 make up a high-pass network to apply negative feedback.

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 and MOSFETs 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 17.43C, 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.

R2 of Fig 17.43C 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 17.43C 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 Solid State Amplifier Projects

This section presents three solid state amplifier designs that operate on a very wide range of frequencies. The first is an HF+6 meter design by Dick Frey, K4XU that will amplify the output of QRP radios in the 10-15 W range to 250 W. It provides an excellent design example of current solid state amplifier design practices described in the previous sections, especially the control and protection circuitry. A second amplifier by James Klitzing, W6PQL, covers 2 meters. It is a linear design suitable for both SSB/CW and FM and has an output power of around 80 W. The final design by Steve Lampereur, KB9MWR, operates on the microwave band of 10 GHz. While the 2 W output may not sound like a lot of power, at X band it is considered QRO especially when combined with a dish antenna that provides tens of dB of gain!

Complete design and construction information for all of these amplifier designs is included on the CD-ROM accompanying this book. In addition, several other amplifier designs are presented for both HF and VHF/UHF operation.

17.11.1 *Project:* A 250 W Broadband Linear Amplifier

The amplifier described here and shown in **Fig 17.44** 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 design project was undertaken for the 2010 edition of the *ARRL Handbook* as a detailed design example as well as a practical solid state amplifier you can build. It is a continuing project. The full description of the amplifier including construction and operating details is provided on the CD-ROM accompanying this book.

A block diagram for the project is given in **Fig 17.45**. 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* since the 2010 edition. 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 51 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.

THE 1.8 TO 55 MHz PA

Fig 17.46 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 even 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 kinds of feedback. Shunt feedback (from drain to gate) is provided by the link on T2

Fig 17.44 — 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.

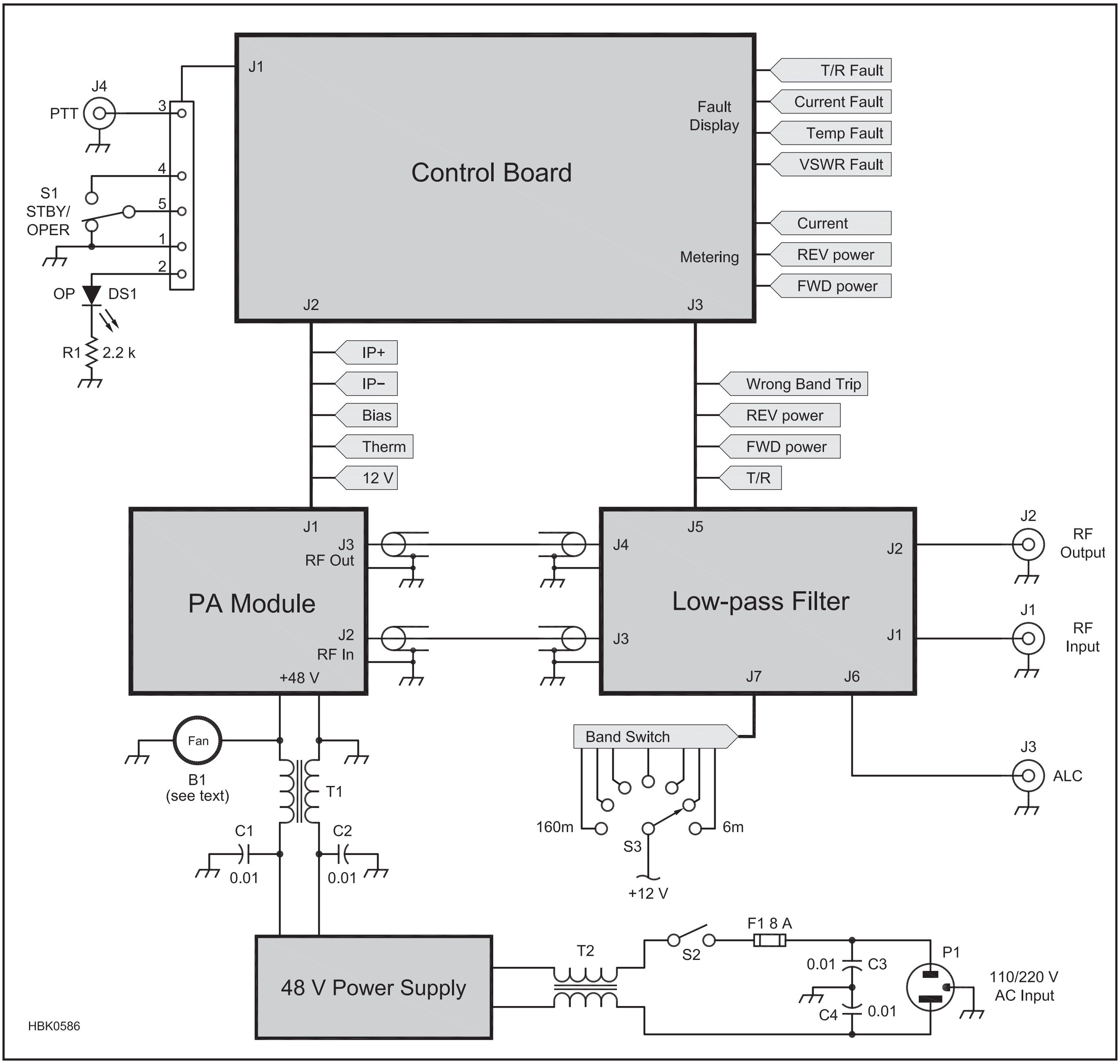


Fig 17.45 — Block diagram of the 250 W solid state amplifier. It is built on three PC boards, which are described in the text and accompanying diagrams. T1 and T2 are 2 turns through a Fair-Rite 2643540002 ferrite bead if needed to suppress RFI from the switching power supply.

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 about 30 dB of gain (×1000) at some frequencies, tending to make it unstable — prone to parasitic oscillation. And the linearity would be terrible, –25 dBc 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.

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. 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. High 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 forcing the amplifier into the standby position and

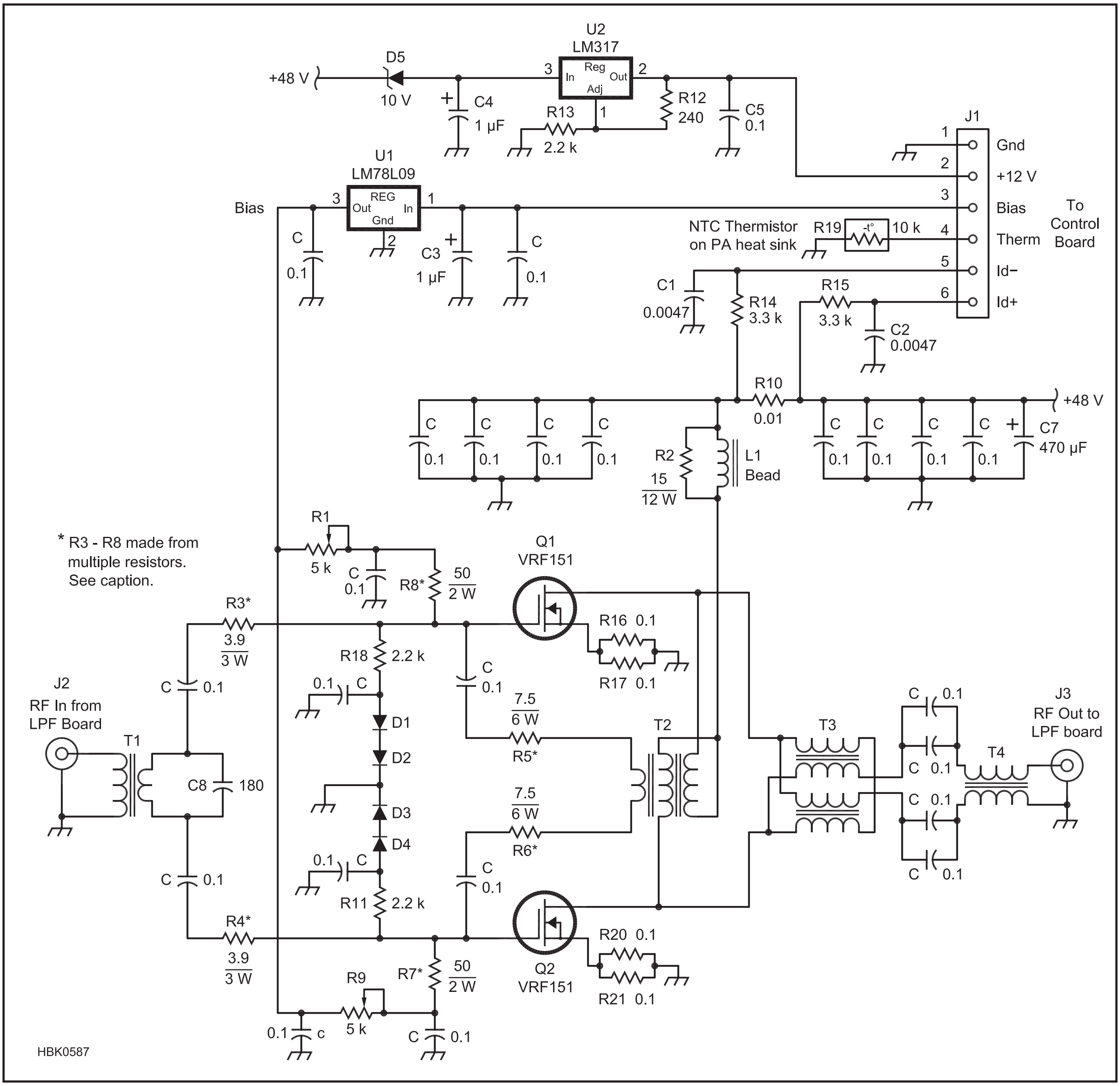


Fig 17.46 — Schematic diagram of the 250 W amplifier PA module. A complete parts list may be found on the *Handbook* CD-ROM.

out of the RF path, and lighting an error LED. 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.

PERFORMANCE

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 17.47**.

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.

17.11.2 *Project:* All-Mode, 2 Meter, 80 W Linear Amp

This solid state amplifier by James Klitzing, W6PQL, is designed for the many

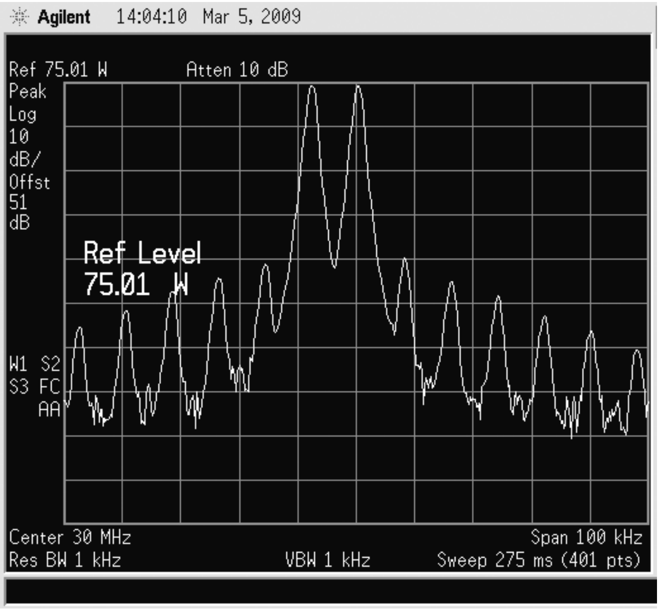


Fig 17.47 — Output of the 250-W amplifier during two-tone IMD testing. All IMD products are better than 38 dB down from either tone.

Fig 17.48 — The 80 W 2 meter linear amplifier is suitable for operation with any mode.

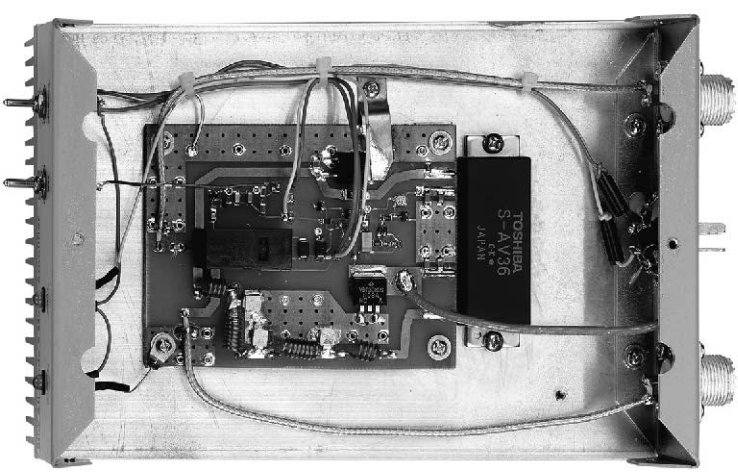


Fig 17.49 — Inside of the compact amplifier showing the simplicity of the final design. The amplifier module is the black rectangle connected to the right edge of the PC board.

low-power 2 meter rigs, ranging from handheld transceivers for FM to older multimode transceivers and even the newer all purpose types such as the Yaesu FT-817 or the Elecraft 2 meter transverters.

The project was featured in the May 2013 issue of *QST* and that article is available on the CD-ROM that accompanies this *Handbook*. The article explains more about the amplifier design and supplies additional construction details. Artwork for the PC board is provided on the *QST* in Depth web page (**www.arrl.org/qst-in-depth**), along with fabrication drawings for sheet metal parts.

The amplifier shown in **Fig 17.48** is low in cost and simple (no preamp or power meters), yet capable of fixed station or mobile operation in any mode and operation from the usual nominal 12 V dc power supply. The same supply that powers a 100 W HF transceiver can likely power the amplifier. The amplifier includes:

- An output low-pass filter to comply with FCC regulations for harmonic and spurious suppression.
- A low-loss antenna relay.
- An RF-sensing TR switch for remote operation, as well as a hard key option.
- TR sequencing to protect the S-AV36 module and prevent hot switching of the antenna relay.
- Indicator LEDs and control switches.
- Reverse polarity protection.

The inside construction is shown in **Fig 17.49**. The small PC board can be made at home (see the **Construction Techniques** chapter) or ordered from the author (see **www.arrl.org/qst-in-depth**). The schematic is provided in **Fig 17.50** and a complete parts list is included in the full article on the CD-ROM. The input and output power of the amplifier with the built in attenuator for a 10 W exciter is provided in **Table 17.6**. This also shows the current required at 13.8 V dc.

Table 17.6
2 Meter Amplifier Operating Parameters
Output power and current required with resting current at 8 A

Drive Power (W)	*Output Power (W)*	*Current (A) at 13.5 V*
1	12	8.2
2	29	9.0
3	44	9.5
4	53	10.0
5	66	11.0
6	74	11.5
7	80	12.0
8	85	12.5
9	89	12.8
10	92	13.0

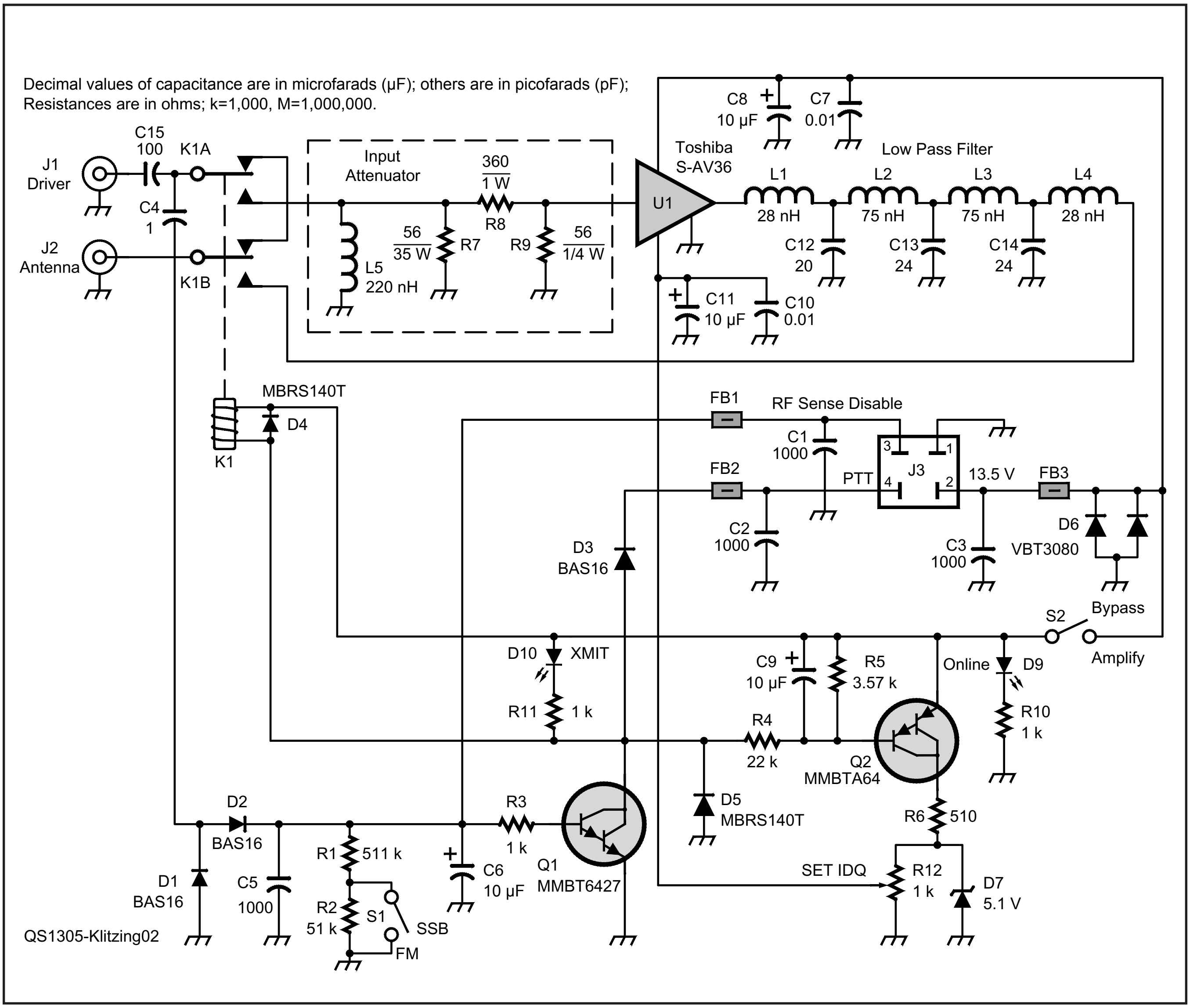


Fig 17.50 — Schematic of the 80 W, 2 meter linear amplifier. A complete parts list is included in the full article on the *Handbook* CD-ROM.

The amplifier is designed around the Toshiba S-AV36 module. The module provides 50 Ω input and output impedances and supplies enough gain that less than 50 mW can drive it to full output in any mode. This design will work with any exciter providing 1 to 10 W of drive, through the use of a built-in attenuator (R7, R8, and R9 in the schematic — see the full article for a table of resistor values that create different levels of attenuation.)

The low-pass filter (L1-L4, C12-C14) is a standard pi network, seven-pole Chebyshev filter. The design provides the required additional harmonic suppression to meet the FCC requirements with very little insertion loss at the operating frequency.

A PCB-mount DPDT general purpose relay was chosen for TR and bypass switching. The contacts are rated at 8 A. At 2 meters, a bit of reactance is introduced by this part, but compensated for by a small capacitor (C15) in series with its input.

The amplifier can be switched ON and OFF by using a control line back to the driving radio (PTT) or an RF sensing circuit is includes that samples the drive from the input connector to provide the transmit trigger.

TR switching is sequenced to prevent the module from attempting to transmit until the relay contacts have switched and settled completely. Hot-switching damages the relay contacts and stressed the amplifier module. The circuitry associated with the base of Q1 does the sampling and controls the sequence timing. Less than ½ W of drive will trigger TR operation.

17.11.3 *Project:* 10 GHz 2 W Amplifier

Generating RF power above a few milliwatts in the 10 GHz band used to be very difficult. Thankfully, Hittite Microwave Corp (**www.hittite.com**) has the HMC487 (the version currently available is the HMC487LP5), which is an easy-to-use X-band amplifier chip that requires no complicated external RF circuitry or special voltage biasing. The HMC487 costs around $60 in single quantities and the evaluation board — which is highly recommended — is a couple hundred dollars. This amplifier project by Steve Lampereur, KB9MWR, is based on the HMC487 evaluation board to help make construction of the final amplifier very easy, even

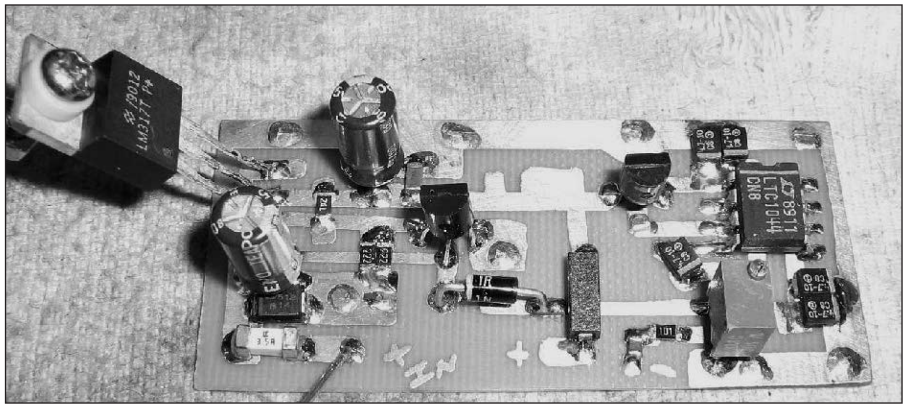


Fig 17.51 — Power supply board that generates the +7 V dc operating and -0.3 V dc bias voltages for the amplifier evaluation board.

for a beginner microwave experimenter.

The complete article is available on the CD-ROM accompanying this book, including construction details and adjustment instructions. The HMC487 data sheet is available from the manufacturer's website.

The HMC487 has around 20 dB of gain from 9 to 12 GHz and is internally matched to 50 Ω on both the RF input and output. It will easily generate 1 W (+30 dBm) of RF output with a 10 mW (+10 dBm) RF input over most of the X-band. It saturates at around 2 W (+33 dBm). The only real drawback to the HMC487 is the large amount of heat it needs to dissipate. Its RF efficiency is only around 20% and the rest of this energy will need to be dissipated as heat.

There are two components to the project — the power supply shown in **Fig 17.51** and the evaluation board. The power supply must generate +7 V at high current to supply the amplifier and –0.3 V for biasing the amplifier gate.

The evaluation board is shown in **Fig 17.52**, where it is mounted on an aluminum plate that acts as a heat sink. A 10 GHz isolator is added to the output to prevent any power reflected from the load from reaching the amplifier IC. Short cables with SMA connectors are used to connect the evaluation board to the chassis-mounted receptacles.

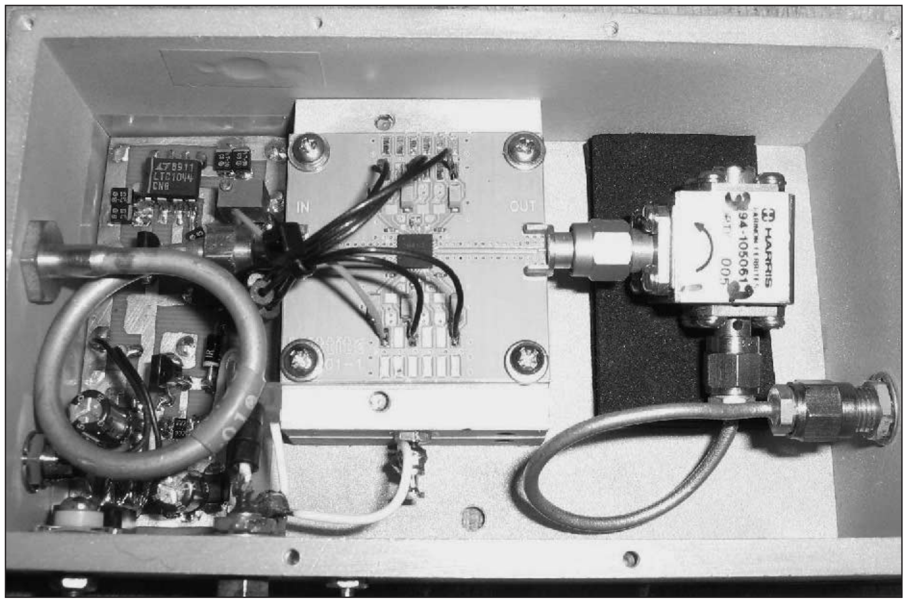


Fig 17.52 — Assembled 10 GHz amplifier. The power supply board is at the left of the enclosure. The evaluation board at center is mounted on an aluminum block to dissipate heat. An isolator at right protects the amplifier output from high SWR.

17.12 Tube Amplifier Projects

Vacuum tube amplifier projects involve several aspects that the builder, particularly beginning amplifier builders, should be aware of and respect. First and foremost are the high voltages associated with these projects — even at the low end of the output power scale. Read the section on High Voltage Techniques in the **Power Sources** chapter and the cautions on measuring high voltages in the **Test Equipment and Measurements** chapter — comply with those recommendations and cautions. Make sure your antenna system and test equipment are suitable for the power and voltages level you will encounter. Consider RF exposure, particularly at 10 meters and higher frequencies at which the Maximum Permissible Exposure limits can be easily reached with a high-power amplifier. (See the **Safety** chapter for more information on RF exposure.) Respect the capabilities of high-power RF!

This section contains overviews of three tube amplifier projects.

- The "Everyham's Amp" is a simple, entry level amplifier design that can develop an output of several hundred watts. The amplifier can be built around several combinations of inexpensive tubes currently available new or surplus. It is intended for the first-time amp builder or for someone who needs a utility amplifier, perhaps for a single band.
- The 3CX1500D7 amplifier is a sophisticated legal-limit design competitive with top-of-the-line commercial models. This model is intended for intermediate builders who have experience with basic metalworking skills and high-power amplifiers.
- A 6 meter kilowatt amplifier design is also suitable for intermediate builders and uses a single 4CX1600 ceramic tube to reach power outputs of about 1 kW, depending on drive level.

Complete design and construction information for all of these amplifier designs is included on the CD-ROM accompanying this book. In addition, several other amplifier designs are presented for both HF and VHF/UHF operation.

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.

Fig 17.A2 — 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.

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

17.12.1 *Project:* The Everyham's Amplifier

Recent editions of the *Handbook* have featured amplifiers that are true works of art, nearly as capable and as elegant as any on the market. Amplifiers such as the HF legal limit amplifier design by Jerry Pittenger, K8RA, in this and recent editions clearly show what advanced amateurs can do.

Many hams who would like to build a basic linear amplifier for their station may not feel ready for advanced amplifier designs — either technically or financially. Contributed by John Stanley, K4ERO, the amplifier project in **Fig 17.53** is a very basic design that will satisfy the need for more power and encourage the reader to experience amplifier building, while providing only the bands and features desired. Once a simple amplifier has been constructed successfully, more advanced designs will be easier to tackle.

The following is an overview and summary of the amplifier's design. A complete description of this amplifier, including additional photographs and drawings, construction details, and additional features is available on the CD-ROM that accompanies this book. An additional three-band design by Leigh Tregellas, VK5TR, is also included on the CD-ROM.

AMPLIFIER CONFIGURATION

The design presented here is based on a modular approach. Each of the three major 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, or you can sometimes bolt the tubes into place with straps and screws.[2]

Using vacuum capacitors really adds class and value to an amplifier. Check the vacuum by setting the capacitor to minimum capacitance 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 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 Sources** 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*

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

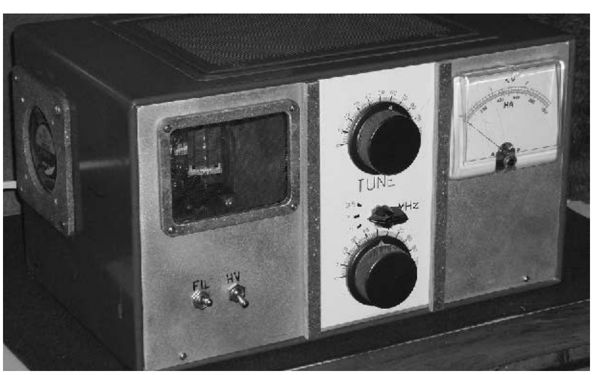


Fig 17.53 — The front panel of the Everyham’s Amplifier. This version is designed to cover three bands (80, 40, and 20 meters). A window is included at left to allow viewing of the tube.

sections (tubes, tuning network, power supply) is somewhat independent and can be changed out separately. In addition, the starting design is “bare bones,” containing nothing that is not absolutely essential for the amplifier to work and provide a minimum level of safety against overloads, abuse, or accidents. TR switching is included in the most basic design because it gives significant advantages at low cost. For the basic design, each part will be described and its purpose explained.

Several options for the major components are presented with the builder being able to choose what best meets his or her needs and best uses the components available in junk boxes, online, or at hamfests. Many small parts such as resistors and diodes are cheap enough that one need not buy them used. Control transformers and tubes are easy to find through online auction sites. Be sure to check shipping costs on heavy items such as transformers.

By shopping carefully, you can avoid the budget being busted by that one essential component costing lots more than expected. Don't overlook acquiring a damaged commercial amplifier or a "basket case" homebrew project — the parts and enclosure hardware available from these are often worth many times the asking price! Refer to the sidebar in this chapter, "Using 'Surplus' Parts for Your Amplifier" for more information about purchasing used and surplus amplifier parts.

Additions and modifications that will improve performance are described in the supplemental article on the CD-ROM and can be added after the initial amplifier is built and tested. Most can be installed in a day, so you need not discontinue using the amp in order to upgrade.

DESIGN OVERVIEW

The basic circuit is shown in **Fig 17.54**. The tubes used in this design are triodes, connected in a grounded-grid configuration. The amplifier can be constructed to use a single 3-500Z or a pair of 811A, 572B, GI-6B or GI-7B tubes. All of these are currently available new or surplus at reasonable cost. The input grid circuit input is driven directly. Tuned input circuits are discussed later as an option.

C3, L1 and C4 form an impedance-matching pi network that transforms the 50 Ω load at the output to the several thousand ohm load that the tube requires. In addition, this "tank" filters out the harmonics from the tube while passing the desired fundamental signal. An optional band switch, S3, is shown. In its simplest form it shorts out sections of L1 as the operating frequency increases. A single-band design doesn't need S3.

When the amplifier is OFF or the transceiver is in receive, the antenna connects directly to the transceiver antenna jack via the two relays as shown in **Fig 17.55**. (K2 and K3 in the main schematic.) When the transceiver is in transmit *and* the amplifier turned on, the transceiver output is routed through the amplifier and amplified.

As part of the TR process, we will want to bias the amplifier off during receive. This saves energy, cools the tube(s) and also removes any RF noise the tube(s) might produce that could get into the receiver. When opened during receive, K1 in Fig 17.54 lifts the center tap of the filament transformer (tube cathode) from ground. K1 can be a separate relay or an additional set of contacts on the TR relays K2 and K3 if they are internal to the amplifier. Usually a resistor (R11), perhaps 47 kΩ or so, is left between the center tap and

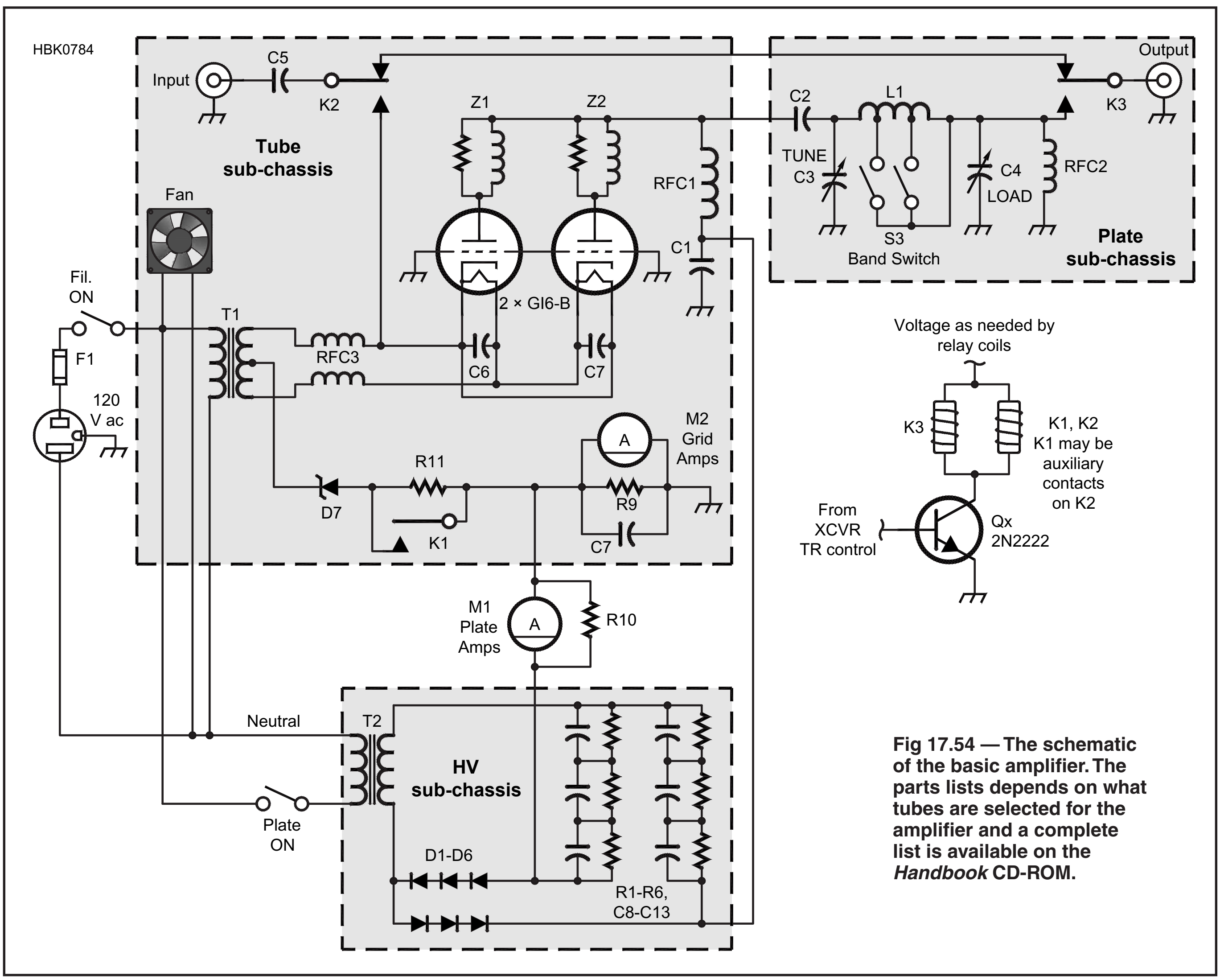


Fig 17.54 — The schematic of the basic amplifier. The parts lists depends on what tubes are selected for the amplifier and a complete list is available on the *Handbook* CD-ROM.

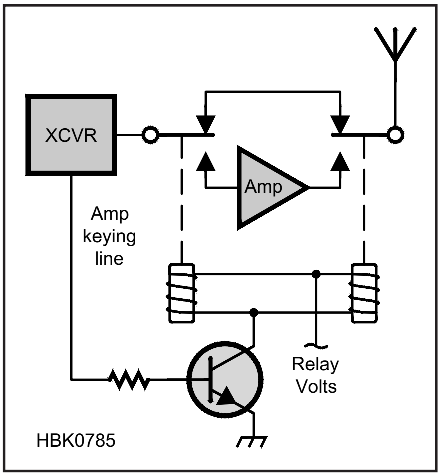


Fig 17.55 — The TR switching of transmit and receive signals for the basic amplifier is controlled by the transceiver's amplifier keying output signal, assumed here to be a positive voltage to key the amplifier with a suitable current-limiting resistor and relay drive transistor installed in the amplifier.

ground to provide a dc current path.

The power supply shown is a basic voltage doubler. (See the **Power Sources** chapter for information on rectifier and voltage multiplier circuits, as well as an alternative power supply design.) If a high-voltage secondary transformer is available, a rating of 500 VA is required for intermittent service up to 1 kW output. A 250 VA rating is sufficient for intermittent service to 500 W with a voltage quadrupler. At one time, TV transformers were commonly used for amplifiers of this size. However, they are becoming hard to find and their age leads to insulation breakdown, thus they are not recommended for this project.

For T2, the main power transformer, the circuit uses two 480/240 to 120 V "control" transformers with the two 480 V windings in series as a secondary. Control transformers are used in industry to reduce the plant's 480 or 240 V equipment wiring to 120 V for the purpose of operating instrumentation and control electronics. These typically have two 240 V windings that can be connected in series for 480 V or run in parallel for higher current. These transformers are commonly available at online auction sites.

The rectifier diodes D1-D6 are rated at 3 A and 1 kV PIV. The filter capacitors C8-C13 are 100 µF, 450 V electrolytics. Modern rectifiers do not need equalizing resistors and capacitors as seen in older articles and designs. Today, it is cheaper to overrate the diode stack instead. C8-C13 do need equalizing resistors, which also serve as bleeder resistors to discharge the capacitors when the supply is turned off.

A meter in the negative lead of the supply monitors the plate current. Several inexpensive options are included in the parts list. Some way to indicate RF output voltage or current is needed, but if you have an external SWR or forward/reflected power meter already, it need not be included in the amp.

CONSTRUCTION NOTES

The type of construction chosen will depend on the enclosure (chassis) which is required for good RF performance and for safety reasons. Having a spacious enclosure and an oversized chassis will make construction easier and upgrades simple. If your situation dictates, you can certainly build everything on a single chassis.

The power supply should be included in the enclosure and not built as a separate piece of equipment. Constructing safe high voltage connections between pieces of equipment requires experience and strict attention to safety-related details that may not be obvious to the beginning amplifier builder. For that reason, this basic design assumes an internal power supply. Rectifiers and filter caps can be mounted on an insulated board (plexiglass or other plastic), or an etched PC board. Install that board on top of the transformer for a neat and compact installation. Board layouts are included on the CD-ROM for both the doubler and quadrupler circuits.

If you find an old piece of equipment you can reclaim, that may dictate the layout. Hamfests usually provide a choice of obsolete tube-type lab or medical instrumentation with high quality cabinets at low cost. They may include desirable items such as a blower or meter in addition to the enclosure. The author chose to use a cabinet from an old piece of Heathkit gear, which had no chassis, and so built a front panel and several sub-chassis sections for the various parts of the amplifier. This allows changing out various parts of the circuit easily and even to maintain several different sub-sections to allow experiments with different tubes, frequency ranges or power supply types.

The tube circuit layout will depend on which tube type is chosen. The 811A/572B option is a great place to start. There is just no other option that can compete in price, counting the tubes and sockets. A pair is recommended for this design although up to four tubes in parallel are common in ham-built amplifiers. A single 3-500Z is another popular choice and, for higher power, a pair of them if the power supply is adequately rated. Ceramic sockets are available for the 3-500Z. Surplus Russian triodes such as the GI-6B and GI-7B are becoming popular and one or a pair is commonly used. The sockets will have to be built as described in the construction details or purchased from ham sources.

External anode tubes such as the Russian GI-6B or GI-7B will need a blower. With the 811A or 3-500Z options, a muffin fan blowing on the tubes will allow them to work much harder. This fan also keeps the other components cooler for extended operation.

Finding suitable capacitors and coils and the band switch will be a significant part of the total procurement process. This is where you can save a lot by using surplus parts. The parts list will specify certain types, mainly so you will know what to look for, but if you simply buy new parts from the list, you will spend your entire budget on these items alone. You don't need exact values for the circuit to work. Limiting the bands to be covered can greatly relax the coil, capacitor and switch requirements.

17.12.2 *Project:* 3CX1500D7 RF Linear Amplifier

This project is a 160 through 10 meter RF linear amplifier that uses the compact Eimac

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.

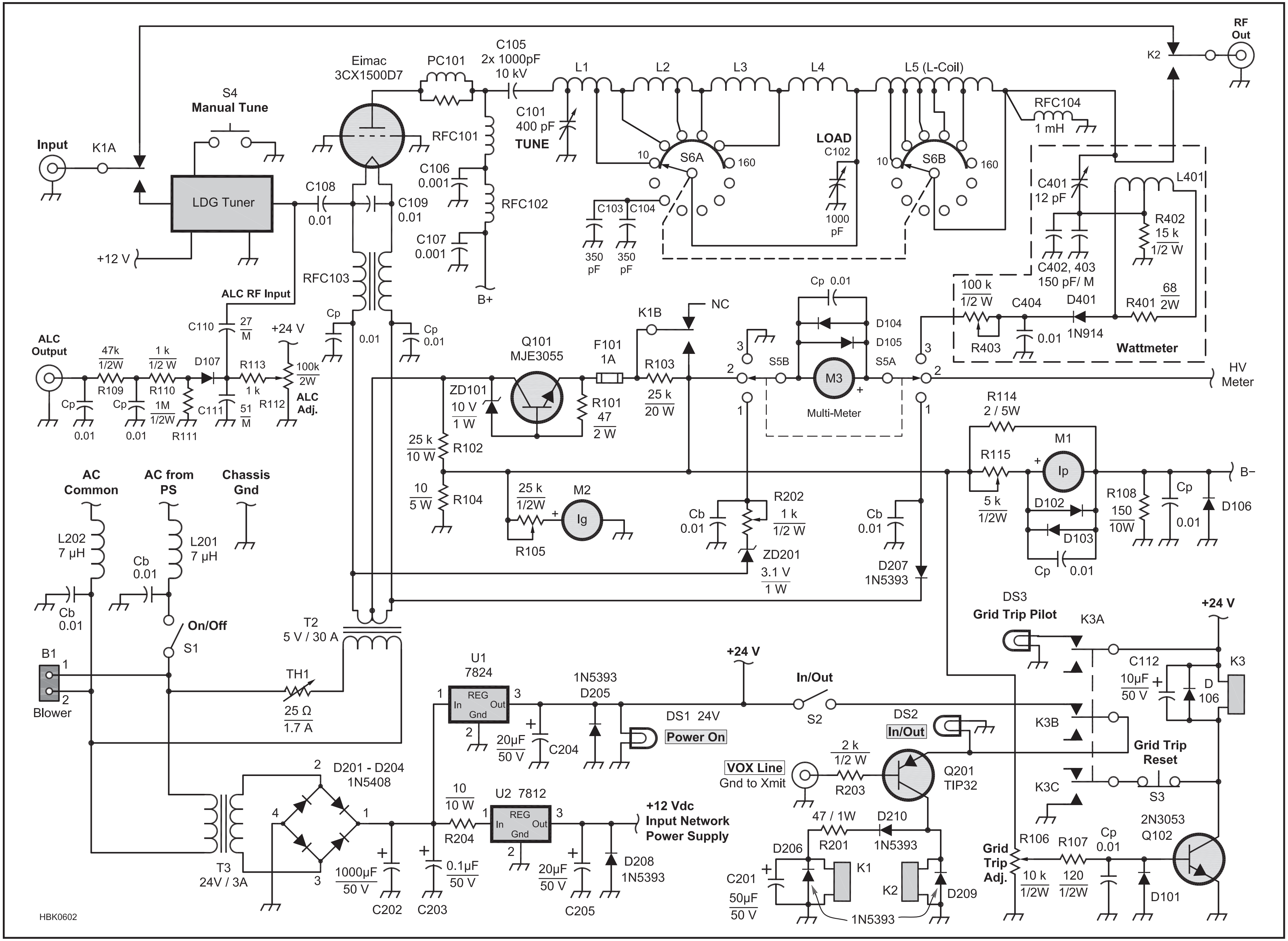


Fig 17.57 — Schematic of the RF deck and control circuitry. A complete parts list is available on the *Handbook* CD-ROM.

3CX1500D7 metal ceramic triode. It was designed and constructed by Jerry Pittenger, K8RA. The complete description of this amplifier including construction and operating details is provided on the CD-ROM accompanying this book. The 3CX1200D7 tube may be more affordable, and the design can be adapted to use it with somewhat less plate dissipation.

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 bandswitch. 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 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

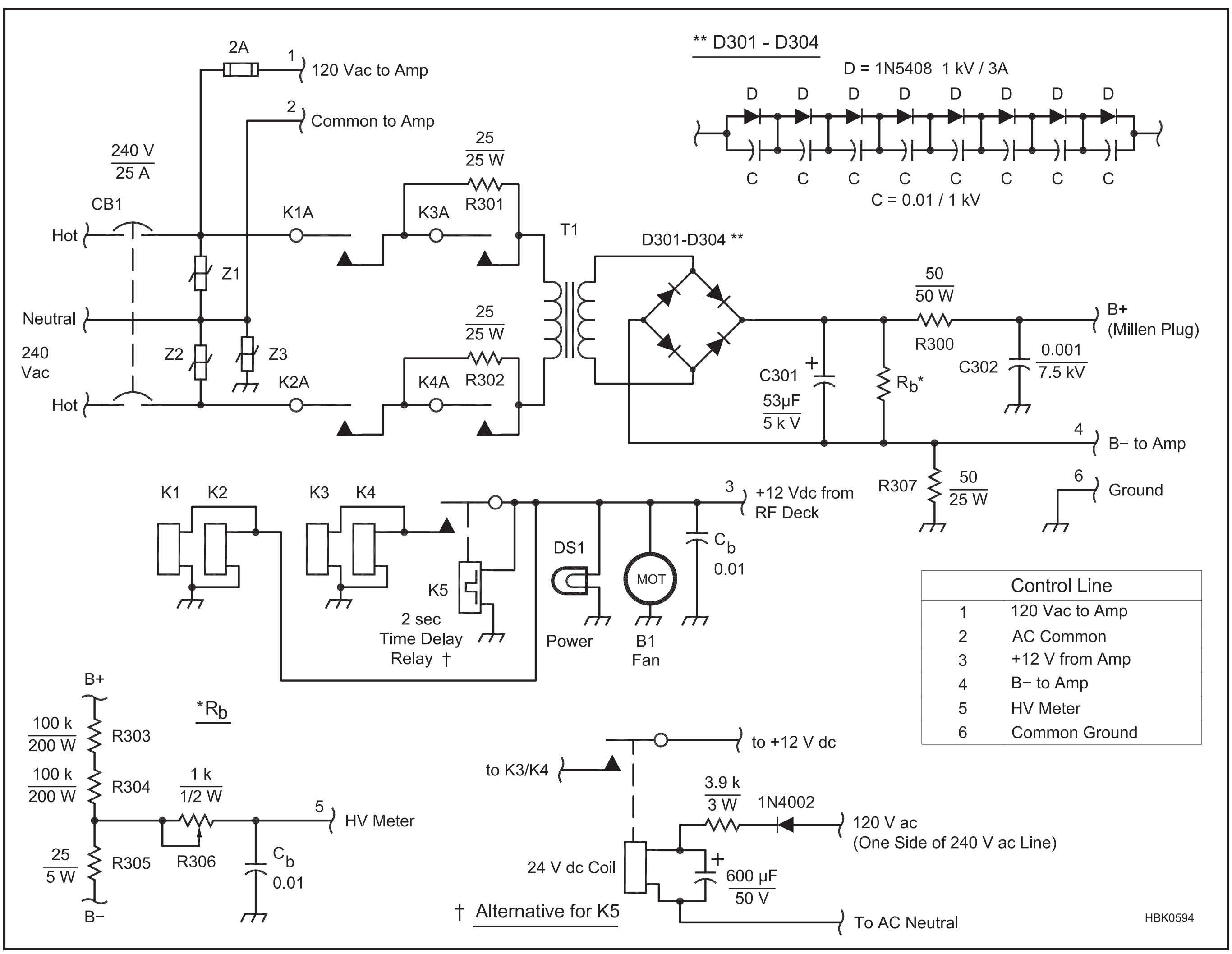


Fig 17.58 — Schematic for power supply for 3CX1500D7 amplifier. A complete parts list is available on the *Handbook* CD-ROM.

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**.

Since this project was first published, Peter W. Dahl has discontinued business and transformers built to the Dahl specifications are now available from Hammond Manufacturing Company, Inc. of Cheektowaga, NY (**www.hammfg.com**). Contact Hammond to cross-reference the Peter Dahl part numbers in the parts list for T1, T2 and RFC103 with current stock or equivalent designs.

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.

GENERAL CONSTRUCTION NOTES

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.

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 ensure safety, but most people can accomplish this project with careful planning and diligence.

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 pack-

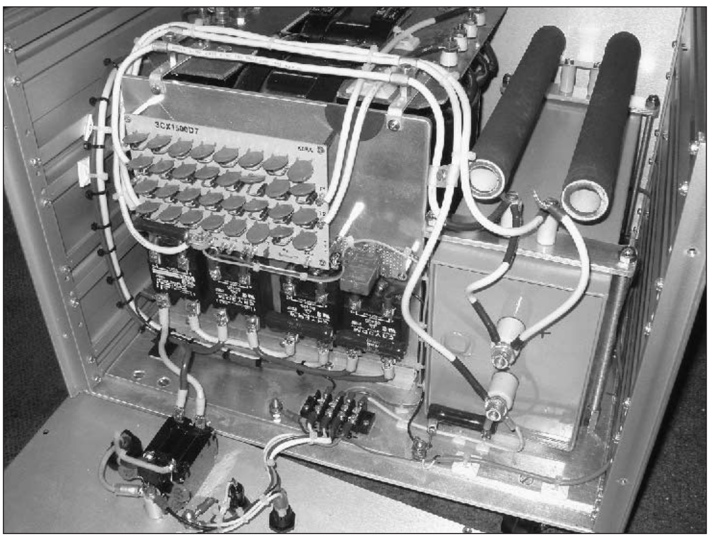


Fig 17.59 — Inside view of the power supply, showing rectifier stack, control relays and HV filter capacitor with bleeder resistors. The heavy-duty high voltage transformer is at the upper left in this photo.

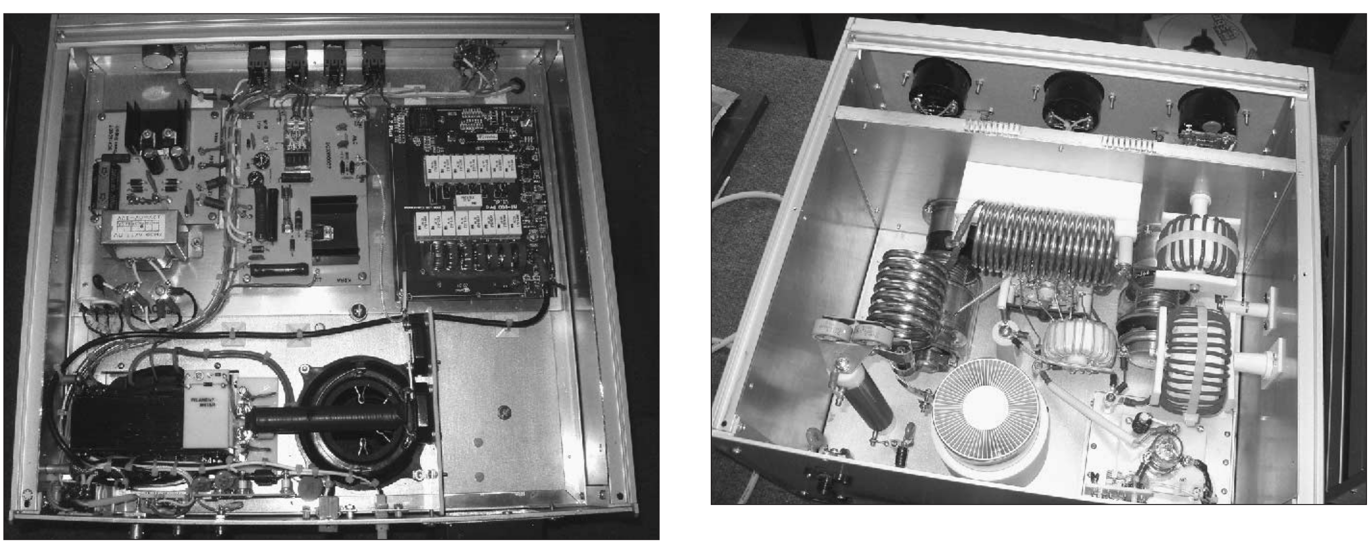


Fig 17.60 — At the left, under the chassis of the RF Deck. The autotuner used as the input network for this amplifier is at the upper right. At right, view of the Pi-L output network in the RF Deck.

age. The printed-circuit boards were designed using a free layout program called *ExpressPCB* (**www.expresspcb. com,** see also the **Computer-Aided Circuit Design** chapter). 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 excellent 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 *ExpressPCB* files for the PC boards are all included on the CD-ROM that accompanies this book.

17.12.3 *Project:* A 6 Meter Kilowatt Amplifier

This amplifier is based on an original design by George Daughters, K6GT, "The Sunnyvale/Saint Petersburg Kilowatt-Plus" in 2005 and earlier editions of this book and that article is included on this book's CD-ROM 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 Dick Stevens, W1QWJ (SK).

The Svetlana 4CX1600B attracted a lot of attention because of its potent capabilities and relatively low cost. (The currently available model of the tube is the 4CX1600U.) **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

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

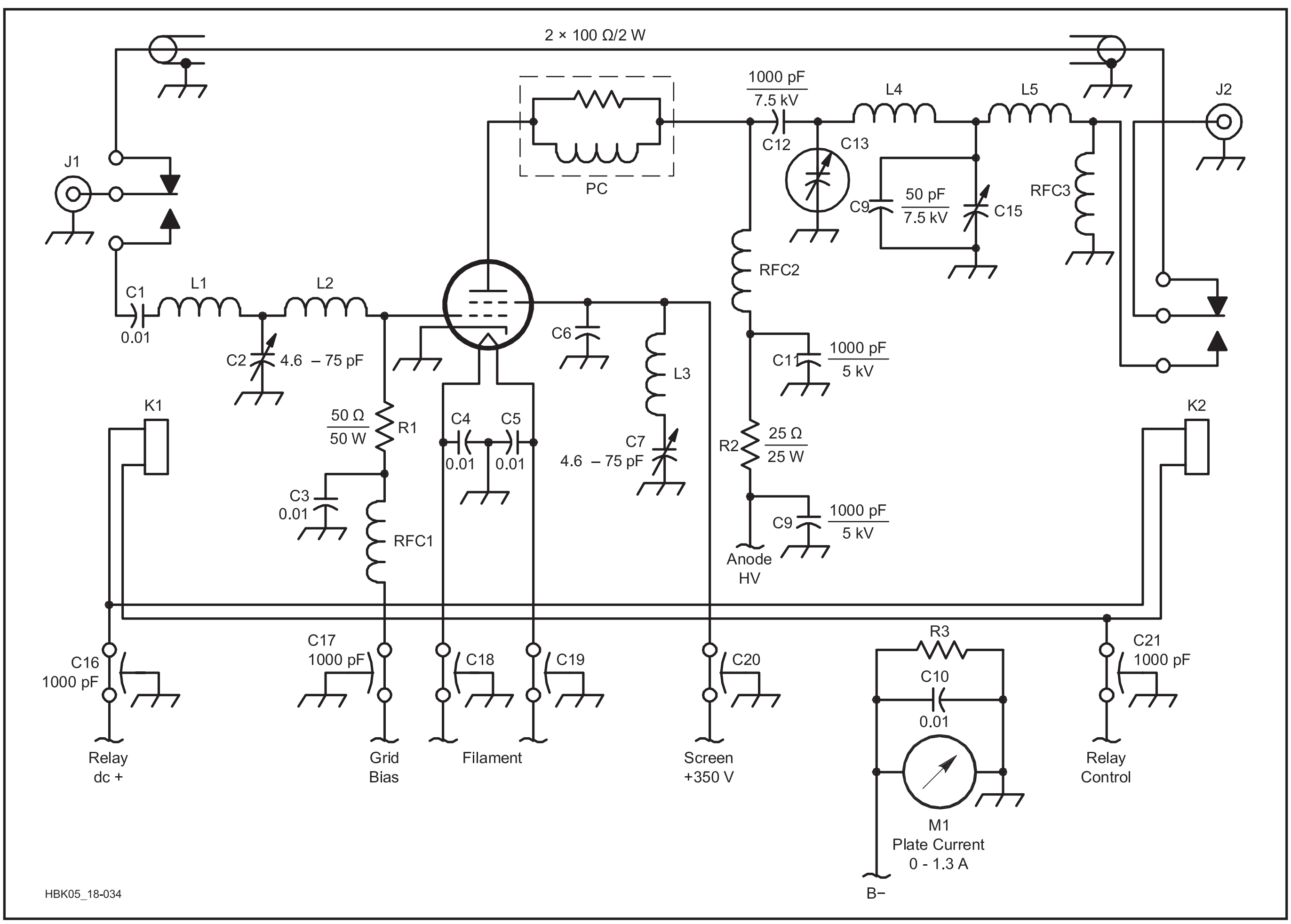


Fig 17.62 — Schematic for the RF deck for the 6 meter 4CX1600B amplifier. The power supply schematic and a complete parts list is available on the *Handbook* CD-ROM.

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 17.63** shows the 4CX1600B tube and the 6 meter output tank circuit.

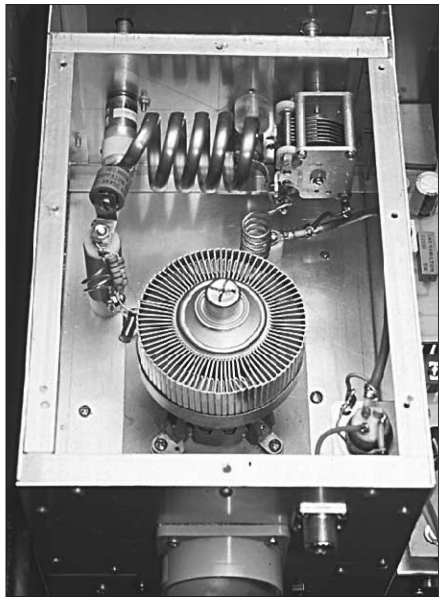


Fig 17.63 — Close-up photo of the anode tank circuit for 6 meter kW amplifier. The air-cooling chimney has been removed in this photo.

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