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Software and Tools
- 135 Degree Pi Network Calculator Spreadsheet by John Stanley, K4ERO
- *MATCH.EXE* software (for use with Tuned (Resonant) Networks discussion)

The following programs are available in the separate Software folder
- *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.
- *MeterBasic* by Jim Tonne, W4ENE, for design and printing of custom analog meter scales.

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, including a sidebar on a design method to avoid interactive tuning, were prepared by John Stanley, K4ERO. The sections on solid-state amplifiers were contributed by Richard Frey, K4XU, along with an overview of the new “pallet” amplifier modules. Roger Halstead, K8RI, contributed material on amplifier tuning and the use of surplus components in amplifier construction that appears with the online content. The section on Pallet Amplifiers was contributed by Jim Klitzing, W6PQL. The section on Solid-State Amplifier IMD was contributed by Rob Sherwood, NCØB; Dick Frey, K4XU; Warren Pratt, NRØV; and Tom Thompson, WØIVJ, who also contributed the material on Adaptive-Predistortion. Amplifier projects using tubes, transistors, and integrated circuits that appeared in previous editions of *The ARRL Handbook* are included with the online content.

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 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.

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*.

High-power amplifiers generate strong RF fields, especially near a tube or transistor and around the output circuits. High levels of RF at frequencies of 50 MHz through microwaves are particularly hazardous. Avoid exposing yourself to intense RF fields when adjusting or measuring energized equipment and follow the RF exposure guidelines for yourself and others. See the **Safety** chapter for more information.

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 ensure that high voltage and current do not occur during switching.

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

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

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

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

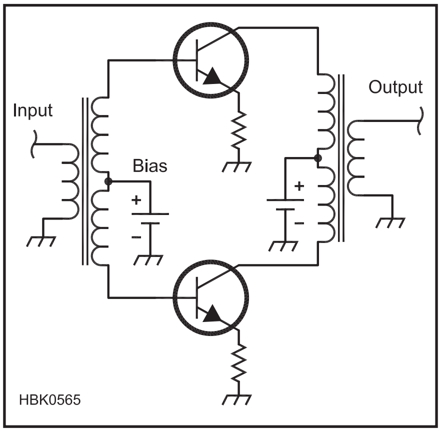


Figure 17.1 — This simple circuit can operate in a linear manner if properly biased.

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

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

ACHIEVING LINEAR AMPLIFICATION

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

Figure 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 **Figure 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 Figure 17.2B, the improperly adjusted amplifier suddenly switches on and produces output when the input signal reaches 0.5 V.

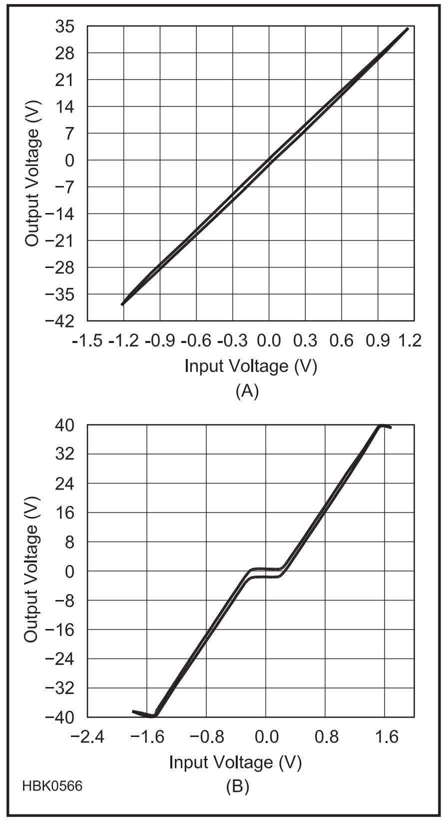


Figure 17.2 — Input versus output signals from an amplifier, as observed with the X-Y display on an oscilloscope. At A, an amplifier with proper bias and input voltage. At B, the same amplifier with improper bias and high input voltage.

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

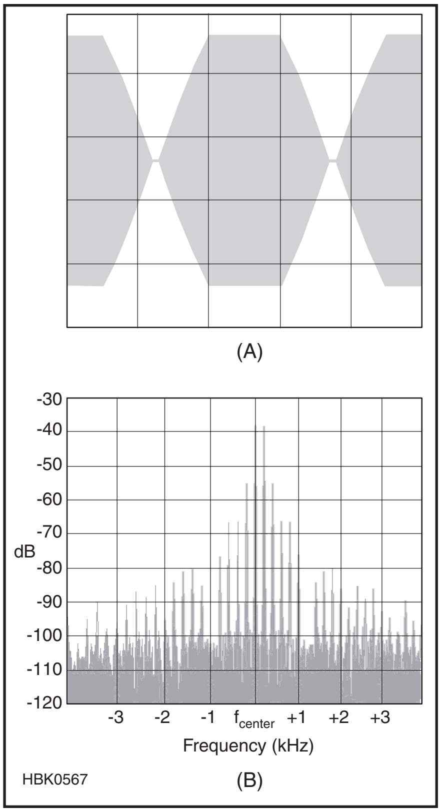


Figure 17.3 — Improper operation of a linear using a two-tone test will show peak clipping on an oscilloscope (A) and the presence of additional frequencies on a spectrum analyzer (B). When these patterns appear, your "linear" has become a "nonlinear" amplifier. Users of adjacent channels will not be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

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 **Figure 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.

Care and Feeding of Power Grid Tubes Available

The classic handbook of power tube design and maintenance from the Eimac Corporation, *Care and Feeding of Power Grid Tubes*, is once again available, courtesy of Communications and Power Industries (**cpii.com** — Eimac is a division of CPI). The book consists of six PDF sections covering all phases of tube operation and design. It can be downloaded from **www.arrl.org/engineering-references** or **www.arrl.org/arrl-handbook-reference**. (A hard copy of the book is available on request from CPI.) Additional references are downloadable from **cpii.com/library.cfm/9**. The ARRL thanks CPI for making this important reference available to radio amateurs!

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%. Tube manufacturers specify an allowable range of filament voltages for proper operation and maximum tube life — follow those recommendations. Tubes should never be operated with filament voltages above the allowable value. Reducing filament voltage below the specified range in hopes of extending tube life is definitely not recommended for tubes with oxide-coated cathodes. In addition, low filament voltage can cause distortion and spurious emissions for any type of tube. Tube failures from low filament emission in amateur service are rare. For more information, see the Reference section entry for T. Rauch, W8JI, on filament voltage management.

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 **Figure 17.4**.

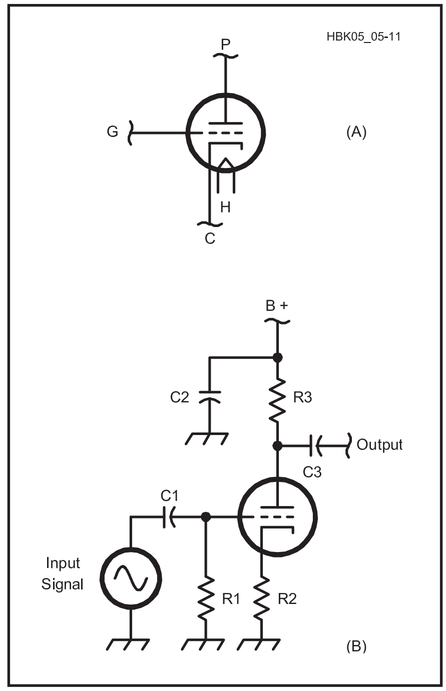


Figure 17.4 — Vacuum tube triode. (A) Schematic symbol detailing heater (H), cathode (C), grid (G) and plate (P). (B) Audio amplifier circuit using a triode. C1 and C3 are dc blocking capacitors for the input and output signals to isolate the grid and plate bias voltages. C2 is a bypass filter capacitor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

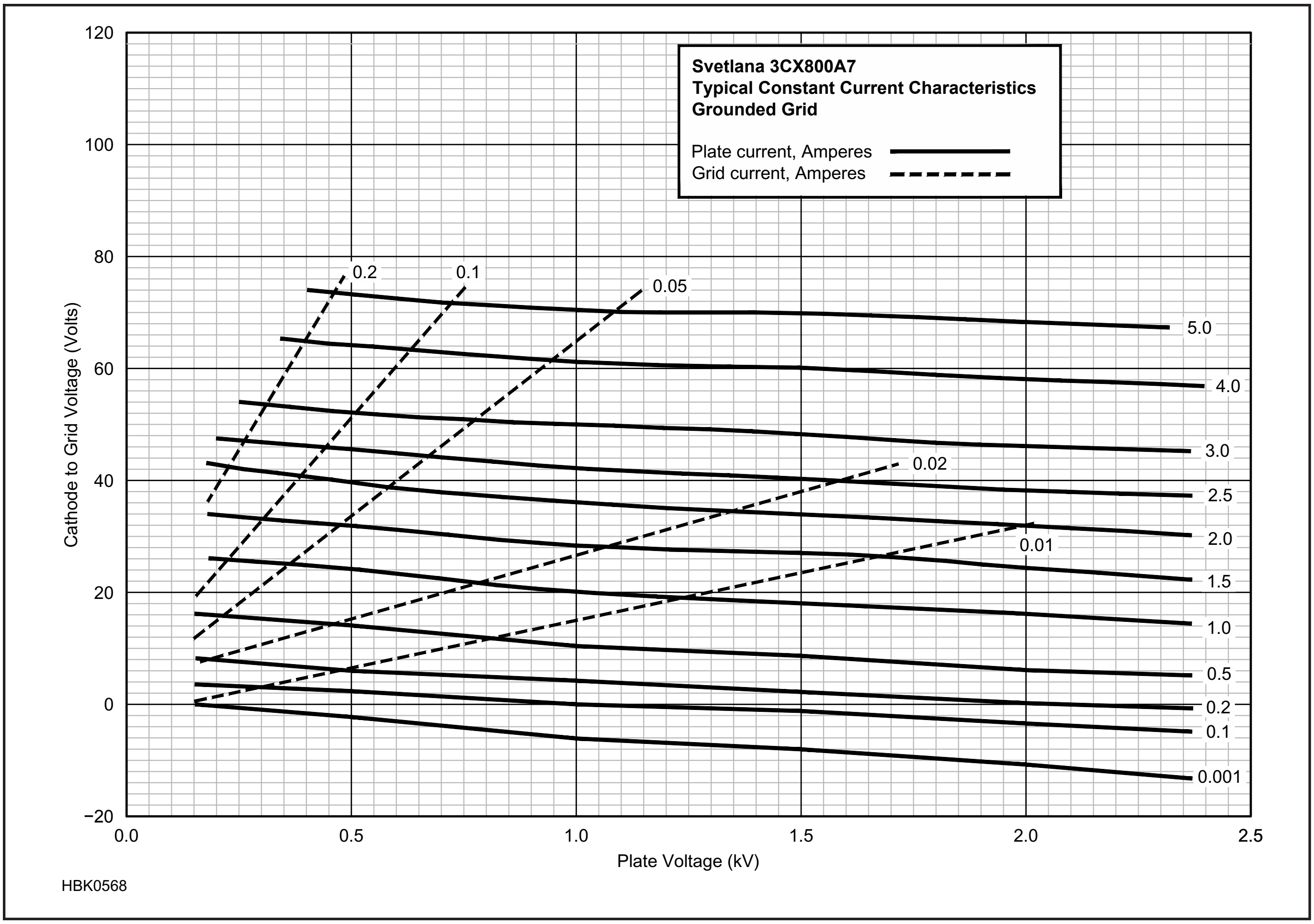


Figure 17.5 — Characteristic curves for a 3CX800A7 triode tube. Grid voltage is plotted on the left, plate voltage along the bottom. The solid lines are plate current, the dashed lines are grid current. This graph is typical of characteristic curves shown in this chapter and used with the *TubeCalculator*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

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

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

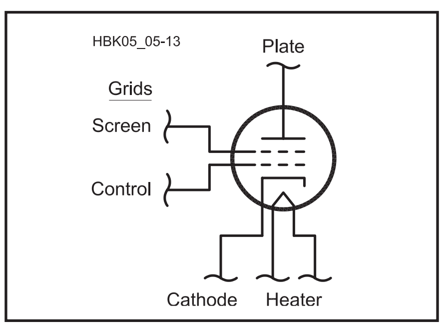


Figure 17.6 — Vacuum tube tetrode. Schematic symbol detailing heater (H), cathode (C), the two grids: control and screen, and the plate (P).

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.

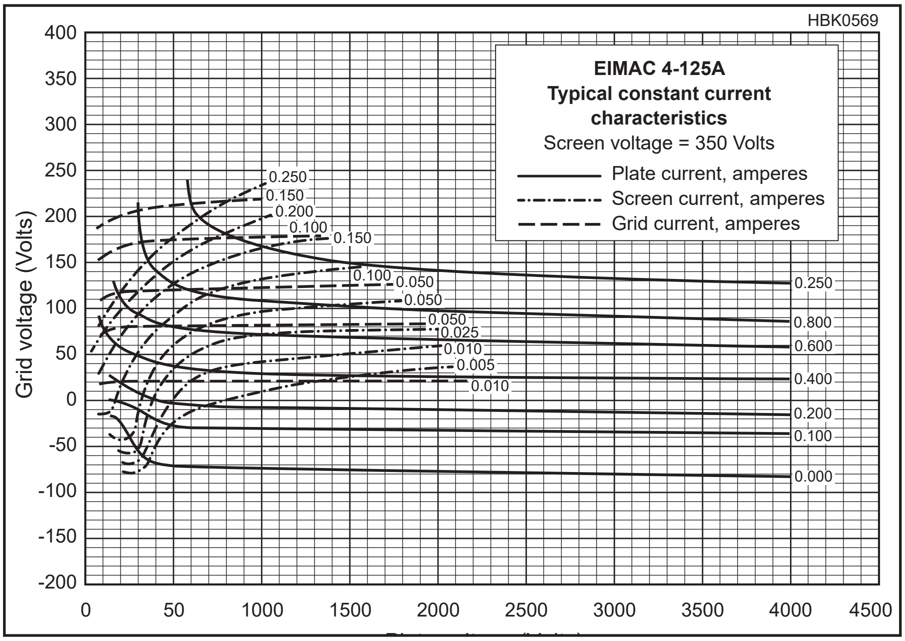


Figure 17.7 — Characteristic curves for a 4-125 tetrode tube.

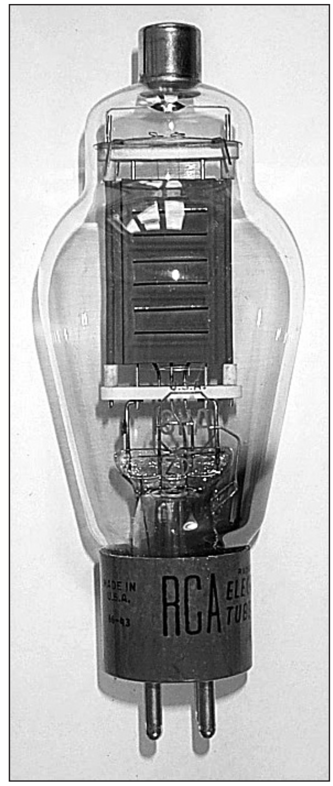


Figure 17.8 — The RCA 811A is an example of a transmitting triode with a glass envelope. [Photo courtesy the Virtual Valve Museum, www.tubecollector.org]

A third grid, called the *suppressor grid*, can be added between the screen grid and the plate. This overcomes the effects of secondary emission in tetrodes. A vacuum tube with three grids is called a *pentode* (penta- for five 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 **Figure 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 (**Figure 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 (Figure 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. Base diagrams of some popular transmitting tubes are provided in the online content.

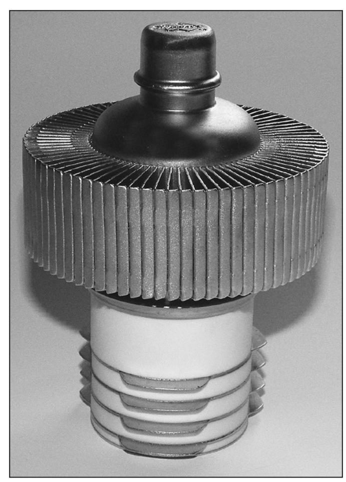


Figure 17.9 — Modern power tubes, such as this 4CX1000A tetrode, tend to use metal and ceramic construction.

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 Figures 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 Figure 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 **Circuits and Components** 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. Figure 17.4B shows the common cathode connection, which gives both current and voltage gain. The common grid (often called *grounded grid*) connection shown in **Figure 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 **Figure 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 to the left in Figure 17.11. The tube is biased well

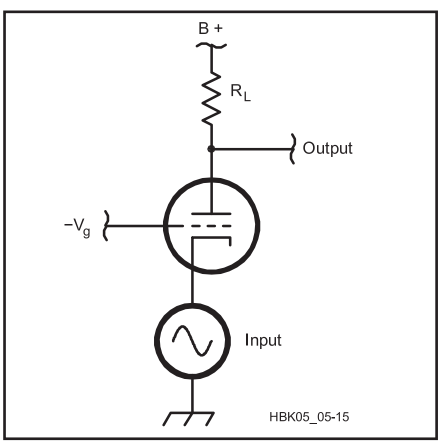


Figure 17.10 — Grounded grid amplifier schematic. The input signal is connected to the cathode, the grid is biased to the appropriate operating point by a dc bias voltage, $-V_G$, and the output voltage is obtained by the voltage drop through R_L that is developed by the plate current, I_P.

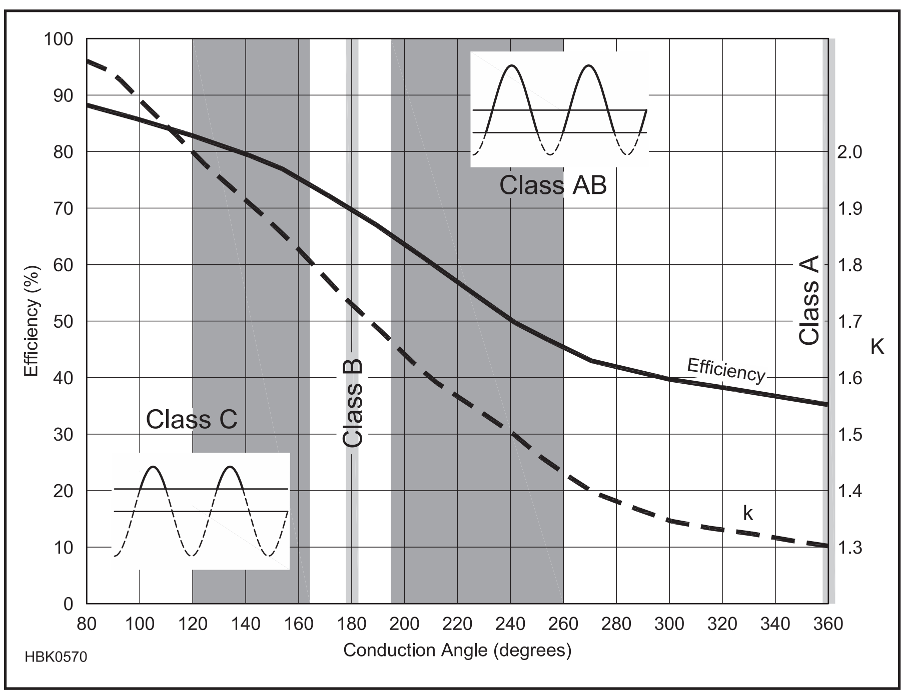


Figure 17.11 — Efficiency and K for various classes of operation. Read the solid line to determine efficiency. Read the dashed line for K, which is a constant used to calculate the plate load required, as explained in the text.

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.

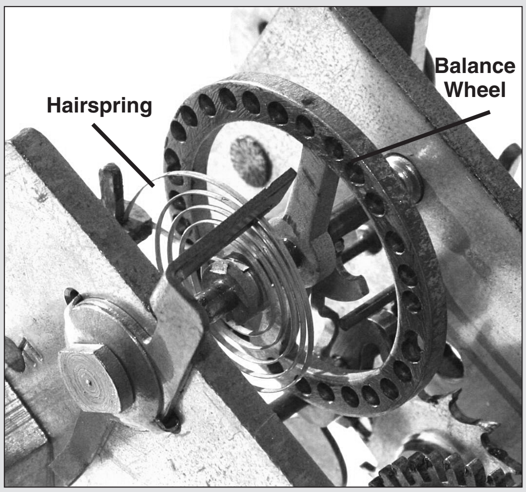


Figure 17.A1 — A balance wheel and hairspring in a mechanical clock illustrate the flywheel effect in tank circuits.

beyond cutoff when no drive signal is applied. Output current flows only during positive crests in the drive cycle, so it con-sists 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 Figures 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 available with this book's online content, 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. **Figure 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 plot shown in **Figure 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 apply-

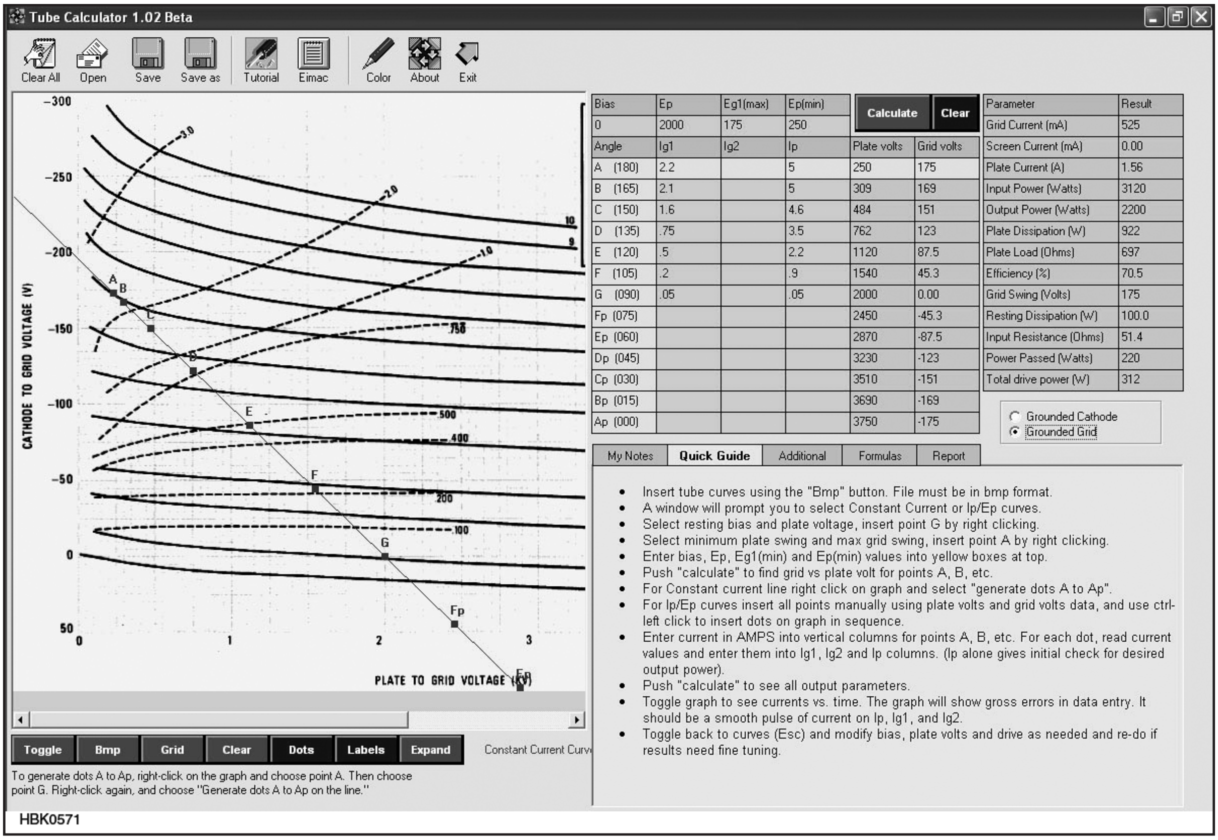


Figure 17.12 — Main screen of *TubeCalculator* program (available with this book's online content). A characteristic curve plot is loaded in the window on the left side of the screen.

ing 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. (See the base diagrams and operating values material in the online content.) 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."

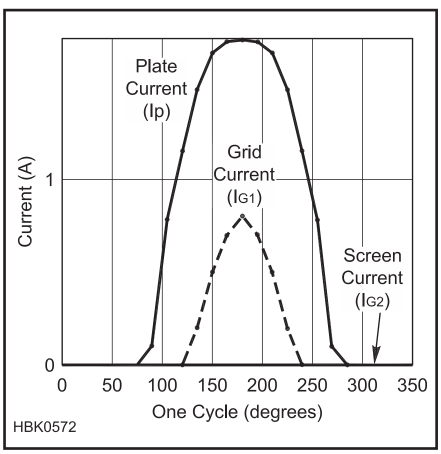


Figure 17.13 — Class AB2 plate and grid current over one cycle as plotted by the *TubeCalculator* program. This plot is for a triode, so there is no curve for screen (G2) current.

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, and

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

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 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

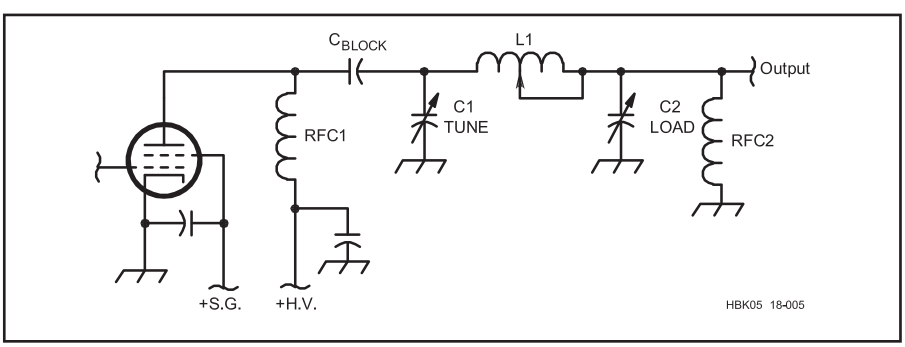


Figure 17.14 — A pi matching network used at the output of a tetrode power amplifier. RFC2 is used for protective purposes in the event C_{BLOCK} fails.

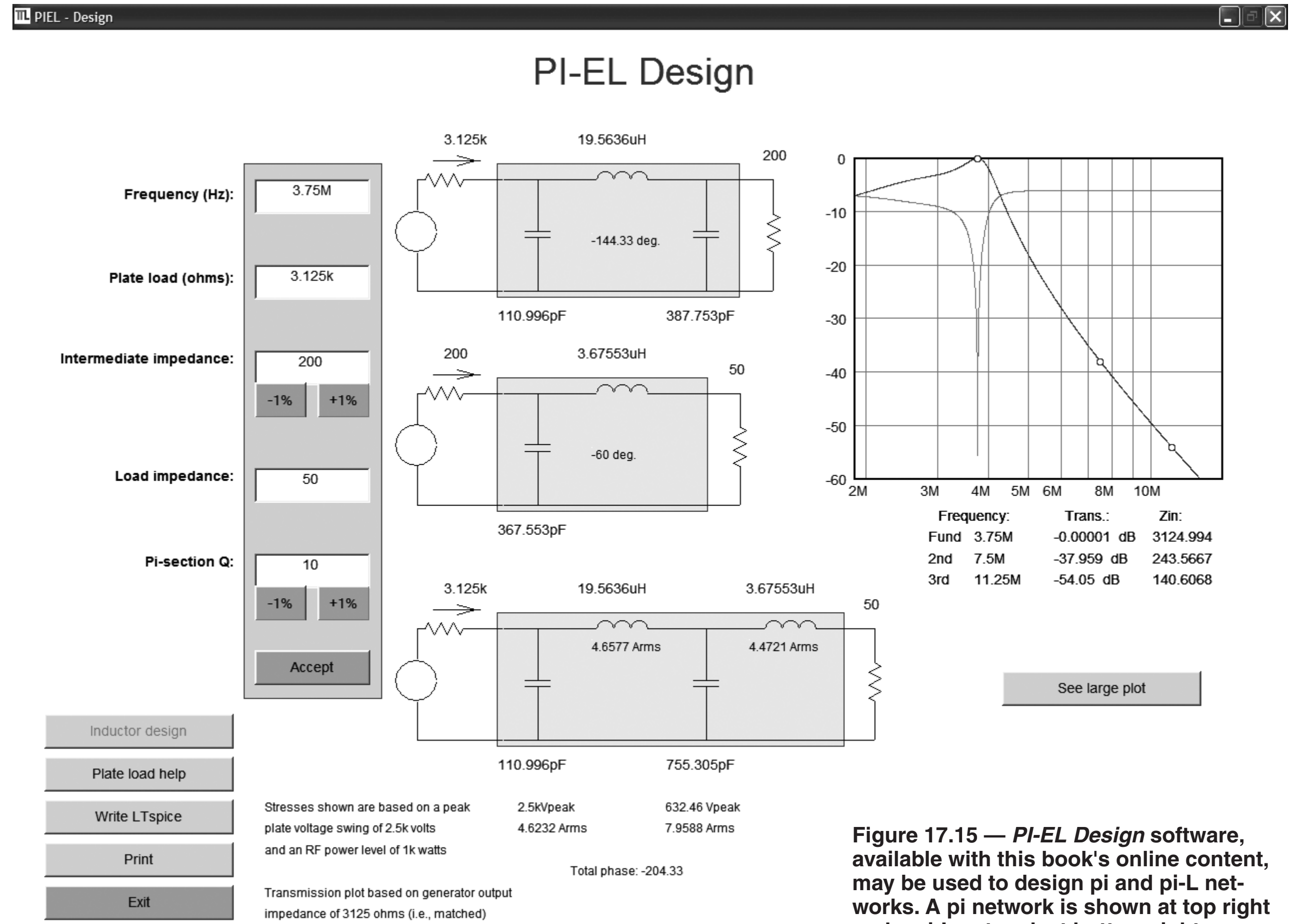


Figure 17.15 — *PI-EL Design* software, available with this book's online content, may be used to design pi and pi-L networks. A pi network is shown at top right and a pi-L network at bottom right.

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 **Figure 17.14**. The **Analog and Digital Filtering** 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 with this book's online content, along with tabular data for finished designs. The input variables are desired plate load resistance, output impedance to be matched (usually 50 Ω), the desired loaded Q (typically 10 or 12) and the frequency. With these inputs one can calculate the values of the components C1, L1 and C2. These components are usually referred to as the plate tune capacitor, the tank inductor and the loading capacitor. In a multi-frequency amplifier, the coil inductance is changed with a band switch and the capacitors are adjusted to the correct value for the band in question.

Tank circuit component values are most easily found using computer software. The program *PI-EL Design* by Jim Tonne, W4ENE, is available with this book's online content 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.

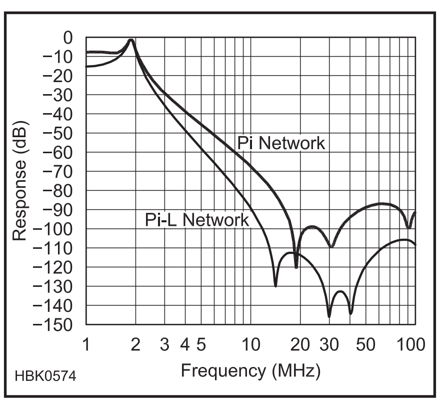


Figure 17.16 — Relative harmonic rejection of pi and pi-L circuits

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 **Figure 17.15**. The harmonic rejection is increased, as shown in **Figure 17.16**, and the 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

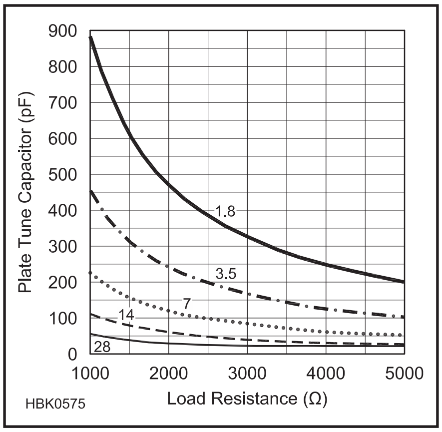


Figure 17.17 — Plate tuning capacitor values for various bands and values of load resistance. Figures17.17 through 17.19 may be used for manual design of a tank circuit, as explained in the text.

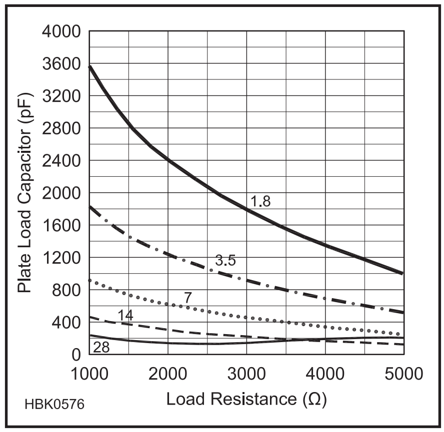


Figure 17.18 — Plate loading capacitor values for various bands and values of load resistance.

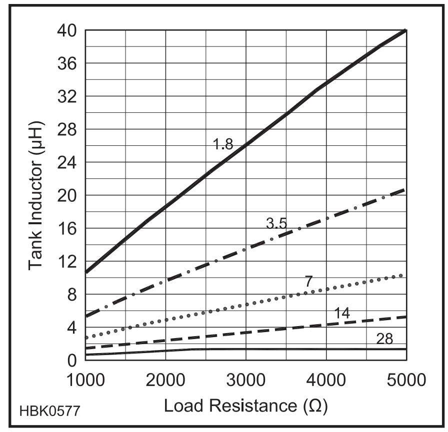


Figure 17.19 — Tank inductor values for various bands and values of load resistance.

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 Figure 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 (**Figure 17.17** to **Figure 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 with this book's online content 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 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.

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

Designing to Avoid Interactive Tune and Load Adjustments

It can be annoying when tuning a tube amplifier if adjusting the loading control requires that we also re-dip the plate current. This happens when the loading capacitor changes the reactance seen by the final tube, rather than just the resistive part of its load. Setting the output power to a chosen value may require numerous readjustments of both controls.

The plate tuning capacitor is directly across the load seen by the tube, hence it only tunes for resonance. The loading capacitor is seen by the tube through the pi network and the network's phase shift will determine how the plate load impedance changes when the loading control is adjusted. For the loading capacitor to change only the resistive part of the tube load, choose a phase shift value for the pi network that is an odd multiple of 45°. Since a pi network with 45° phase shift will have very low Q, the best choice is a network with 135° phase shift.

The Smith chart in **Figure 17.A2** illustrates this effect. The solid line with dots shows the effect of adjusting the plate tuning capacitor. When close to resonance, it does not change the resistive part of the load seen by the tube, but does tune out the reactive part as shown by the fact that it crosses the horizontal resistance axis at a right angle. When the load is purely resistive at resonance, we get a dip in the plate current.

The thick dashed curve shows what happens when we adjust the loading capacitor with a 135° pi network. The resistive part of the tube load changes as shown by the curve moving along the horizontal axis, but the resonance is largely unaffected. The thin dashed line shows an example of what happens when we change the load capacitor with a network with a phase shift that is not 135°. Both the resistive part and the reactive part of the tube load change, making it necessary to retune the plate capacitor to resonance.

The phase shift of a pi network is determined by the Q of the network and the impedance transformation ratio. The value of Q that will give us the "magic" 135° phase shift will thus depend on the plate load to output impedance ratio. For a 50 Ω load at the output of the pi network, the desired Q value can be determined from the upper curve of **Figure 17.A3**. Putting this value of Q into the pi-L design program will generate the pi network component values desired. For example, for 2500 Ω plate load we see that the desired Q of the tank will be 12.4, a reasonable Q to choose that is close to the commonly used value of 10. Thus, with a slight adjustment of the design Q we can get to a "non interactive loading control" design.

As we can see on the upper curve of Figure 17.A3, with very high or low plate load values, the excursion from the normally used Q value of 10 may be excessive and a fully non-interactive tuning solution not practical with the simple pi network. For a plate load of 4000 Ω we see that a Q of 16.4 would be required. This is somewhat high and would lead to excessive losses in the tank. An alternate approach would be to use the pi-L network, and allow the intermediate value to be 100 Ω. The lower curve shows this allows the pi-L network to be designed using a Q value of 11.2, which is more acceptable.

A further discussion of this method, with formulas, a spreadsheet calculator, and more Smith chart graphics illustrating how it works is available with this book's online content.

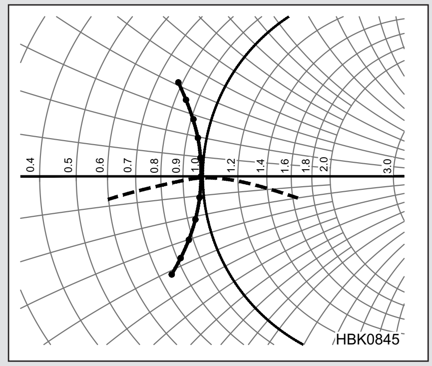


Figure 17.A2 — Effect of plate and load capacitors for various phase shift values through the pi network.

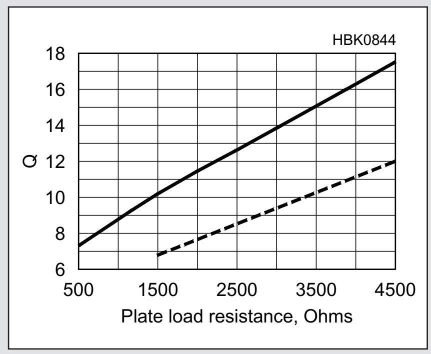


Figure 17.A3 — Pi network Q versus plate load resistance for non-interactive tuning. Use the upper curve for a 50 Ω simple pi network and the lower curve for a pi-L network with a 100 Ω intermediate resistance valu

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 with this book's online content 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

Amplifier Tuning

For commercial amplifiers, "Read the Manual" should be your guide as the following procedures may not be exactly the same as the manufacturer's directions and can be quite different in some cases. 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.

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.

Pre-tuning Preparation

1. 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. If you don't have instructions or pre-set values, set the TUNE control for maximum capacitance.
2. 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.

Grounded-Grid Triode Amplifiers

Triodes used in currently manufactured amplifiers include the 811A, 572B, 3-500Z, and 3CX-type tubes. Tuning a triode-based, grounded-grid amplifier is the simplest: Tune for maximum or legal-limit output power without exceeding the tube ratings, particularly grid current. The typical procedure is:

1. While monitoring grid current, adjust TUNE for a dip in the plate current which should correspond to a peak in output power.
2. 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.
3. Adjust the TUNE control then advance the LOAD control for maximum output.
4. 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 is reached.
5. If necessary, increase drive and re-peak both the TUNE and LOAD controls.

Tetrode Amplifiers

Amplifiers that use tetrodes such as the 4CX800A/GU-74B, 4CX1000 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.

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.

1. 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.
2. As with the triode amplifier, if drive needs to be increased, readjust the TUNE and LOAD controls.

After Tune-Up

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.

What Does Tuning Really Do?

What is actually happening when the TUNE and LOAD controls are operated this way? First, remember that the pi network acts as an adjustable impedance matching circuit between the tube's high impedance (from a few hundred to several kiloohms) to the antenna system's low impedance (typically under 100 ohms). It acts like two L networks back-to-back, changing the impedance in two steps. The TUNE capacitor and a portion of the inductor form one L network while the LOAD capacitor and the remainder of the inductor form the other L network.

By adjusting the two controls, you adjust impedance presented to the tube. Setting the LOAD control to maximum capacitance and adjusting the TUNE control for minimum plate current presents the maximum impedance or lightest load to the tube.

The next steps of decreasing LOAD capacitance and re-adjusting the TUNE control for minimum plate current ("dipping the plate") steadily reduce the impedance presented to the tube (increasing the load), which increases power output. The goal is to increase the LOAD capacitance while keeping plate current minimized, until the desired output power is reached, and the tube is operated at "full load."

The math behind how the pi network changes the impedance and the tube current with power output is explained on the web page **www.edn.com/pi-network-and-dipping-the-final**.

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.

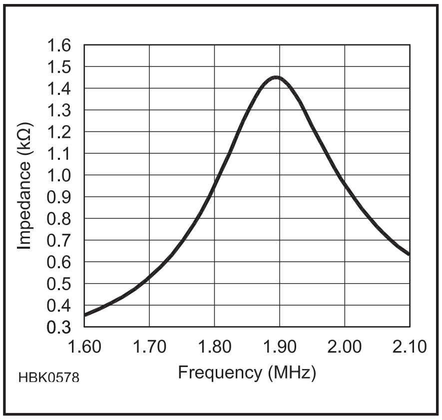


Figure 17.20 — Impedance as would be measured at the plate of an amplifier tube with the output network tuned to 1.9 MHz and the output terminated in its proper load.

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 **Figure 17.20**.

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

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

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.

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.

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

Larger conductors than required for current handling are often used to maximize unloaded Q, particularly at higher frequencies. Where skin depth effects increase losses, the greater surface area of large diameter conductors can be beneficial. Small-diameter copper tubing, up to 3/8 inch outer diameter, can be used successfully for tank coils up through the lower VHF range. Copper tubing in sizes suitable for constructing high-power coils is generally available in 50 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 **Figure 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!

An RF choke performs best well below its self-resonant frequency (SRF) but it can be used throughout the range for which it has an acceptably high impedance. The choke's impedance will be inductive below the SRF and capacitive above the SRF. (See the discussion of RF Chokes in the **RF Techniques** chapter.) In the plate circuit, resonances may produce very high voltages at one or more points along the coil. This can cause arcing and damage to the choke. These chokes are often specially wound to minimize the effect of resonances but this must be done to suit the exact application since distributed capacitance to the enclosure and other components affects the SRF.

W8JI has written an informative paper about SRF and plate chokes which can be viewed at **www.w8ji.com/rf_plate_choke.htm**. His companion web page on inductors used in high-power RF circuits (**www.w8ji.com/loading_inductors.htm**) is also very informative.

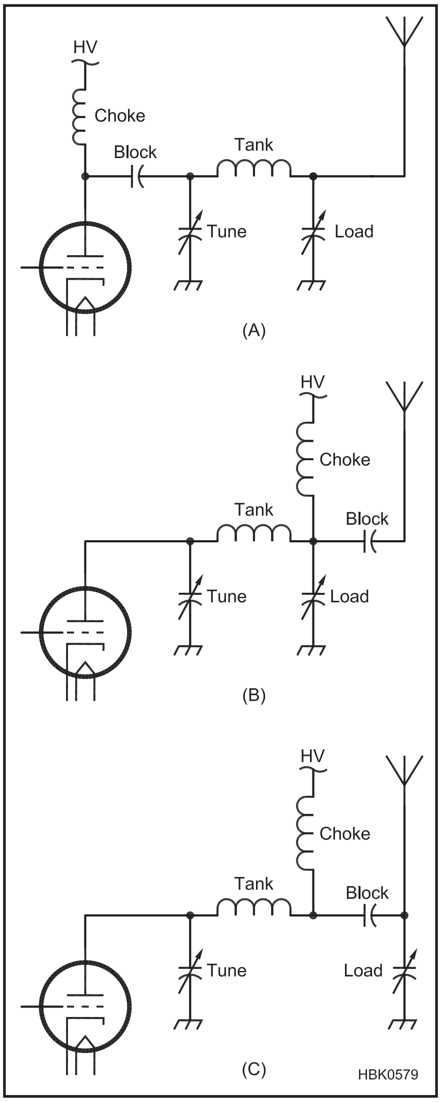


Figure 17.21 — Three ways of feeding dc to a tube via an RF choke. See text for a discussion of the tradeoffs.

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 Figure 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 Figure 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 Figure 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.

Finally, it is also possible to place the plate at ground potential and the filament or cathode at negative HV or B-. This has the benefit of placing the RF choke at a low-impedance part of the circuit, greatly reducing the effects of self-resonance. Blocking capacitors are still required on the input and grid-bias circuits but they carry much less RF current than when in the output plate circuit. The filament transformer will have to have sufficient insulation to withstand the HV across it, as well. While rarely seen in amateur amplifiers, this technique was used in AM broadcast transmitters and is discussed more at **forums.qrz.com/index.php?threads/the-case-for-a-grounded-b-other-innovations-in-the-design-of-a-vacuum-tube-rf-power-amplifier.771103/**.

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.

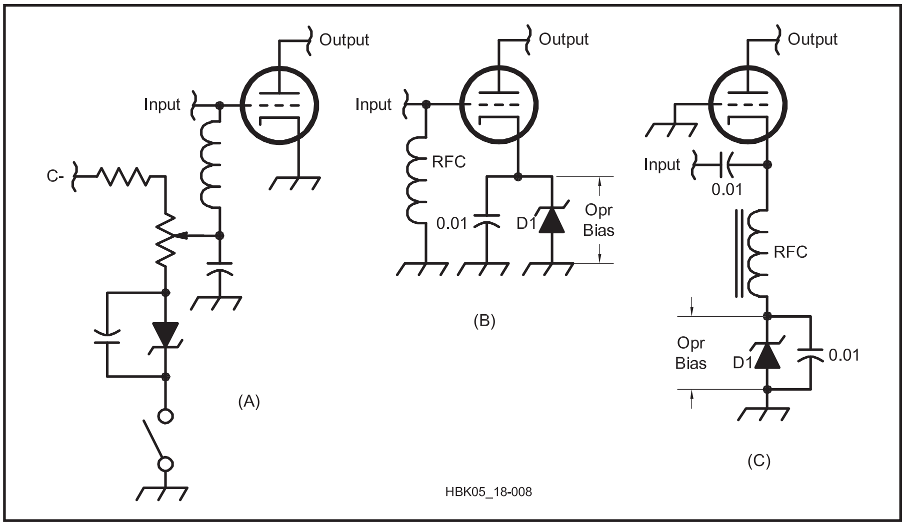


Figure 17.22 — Various techniques for providing operating bias with tube amplifiers.

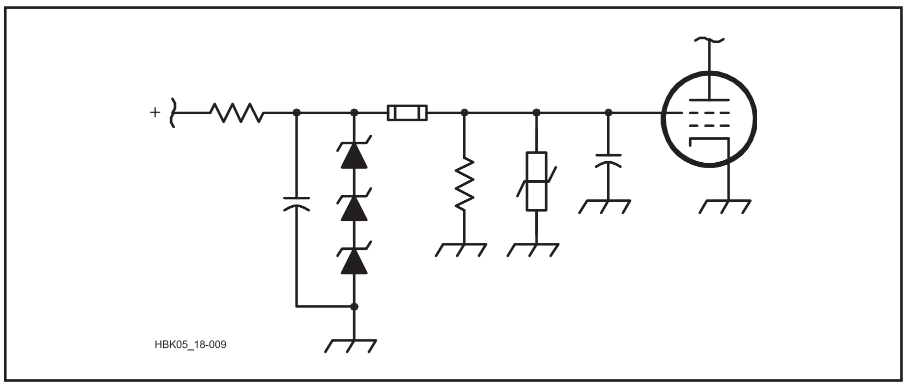


Figure 17.23 — A Zener-regulated screen supply for use with a tetrode. Protection is provided by a fuse and a varistor.

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.

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. **Figure 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 Figure 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 Figure 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.

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

17.7 Tube Amplifier Cooling

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

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

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

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

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.

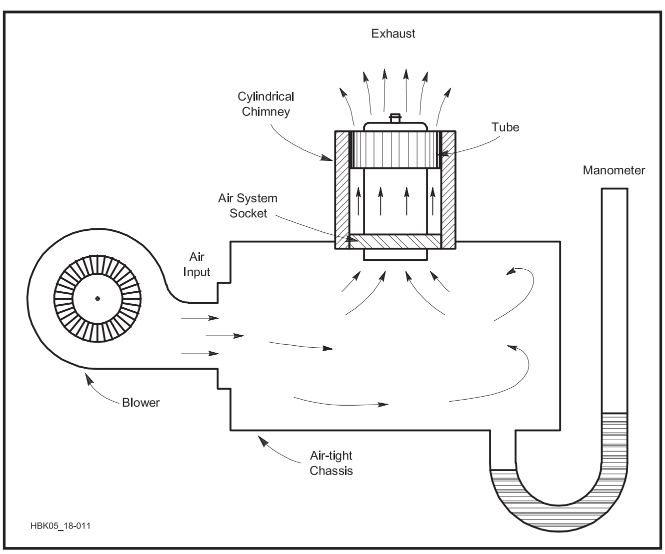


Figure 17.24 — Air is forced into the chassis by the blower and exits through the tube socket. The manometer is used to measure system back pressure, which is an important factor in determining the proper size blower.

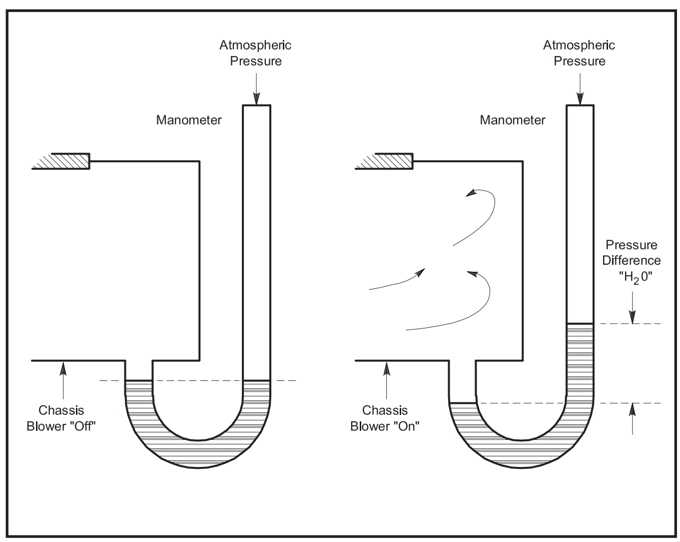


Figure 17.25 — At A the blower is "off" and the water will seek its own level in the manometer. At B the blower is "on" and the amount of back pressure in terms of inches of water can be measured as indicated.

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**).

measured easily in an operational air system as indicated in **Figures 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 Figure 17.25A, the blower is "off" and the water seeks its own level, because the air pressure (ordinary atmospheric pressure) is the same at both ends of the manometer tube. At B, the blower is "on" (socket, tube and chimney in place) and the pressure difference, in terms of inches of water, is measured. For most applications, a standard ruler used for measurement will yield sufficiently accurate results.

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

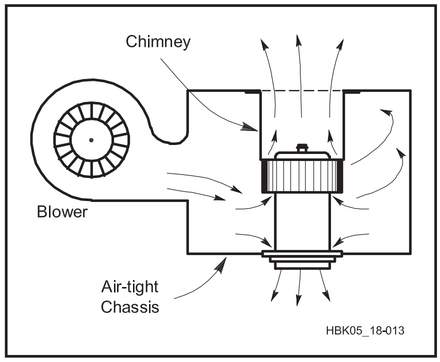


Figure 17.26 — Anode compartment pressurization may be more efficient than grid compartment pressurization. Hot air exits upwards through the tube anode and through the chimney. Cool air also goes down through the tube socket to cool tube's pins and the socket itself.

• 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 Figure 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 **Figure 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.

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), and
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 **Radio Fundamentals** 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 **Figure 17.27**.

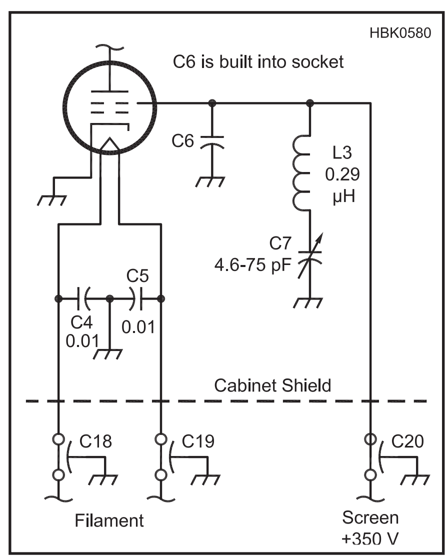


Figure 17.27 — A series-resonant circuit can be used to provide nearly perfect screen or grid bypassing to ground. This example is from a single-band 50 MHz

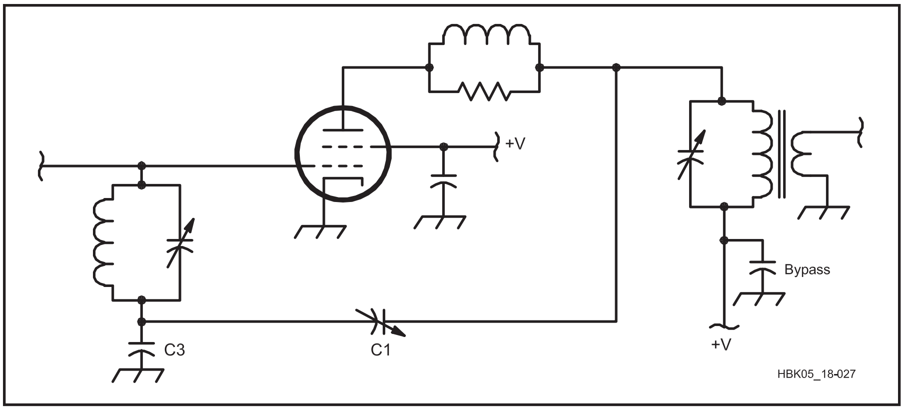


Figure 17.28 — A neutralization circuit uses C1 to cancel the effect of the tube internal capacitance.

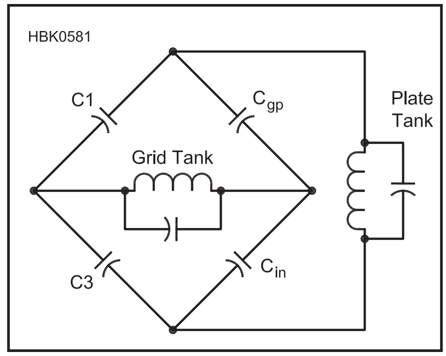


Figure 17.29 — The "bridge neutralization" circuit of Figure 17.28 redrawn to show the capacitance values.

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 **Figure 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 **Figure 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, and
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 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 **Figure 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

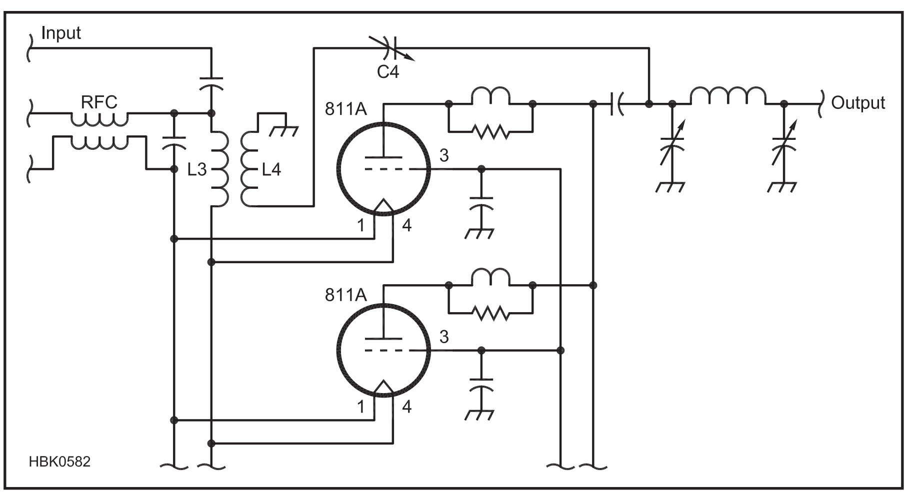


Figure 17.30 — In this neutralizing method, a broadband transformer (L3, L4) provides the needed out-of-phase signal. L3 is 6 turns of #14 wire close wound, ½ inch diameter. L4 is 5 turns of insulated wire over L3. C4 is 6 pF with 0.06-inch spacing. This circuit was originally featured in June 1961 *QST*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

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 **Figure 17.31**. Stray inductance between the tube plate and the output tuning capacitor forms a high-Q resonant circuit with the tube's C_{OUT}. C_{OUT} normally is much smaller (higher X_C) than any of the other circuit capacitances shown. The tube's C_{IN} and the tuning capacitor C_{TUNE} essentially act as bypass capacitors, while the various chokes and tank inductances shown have high reactances at VHF. Thus, the values of these components have little influence on the parasitic resonant frequency.

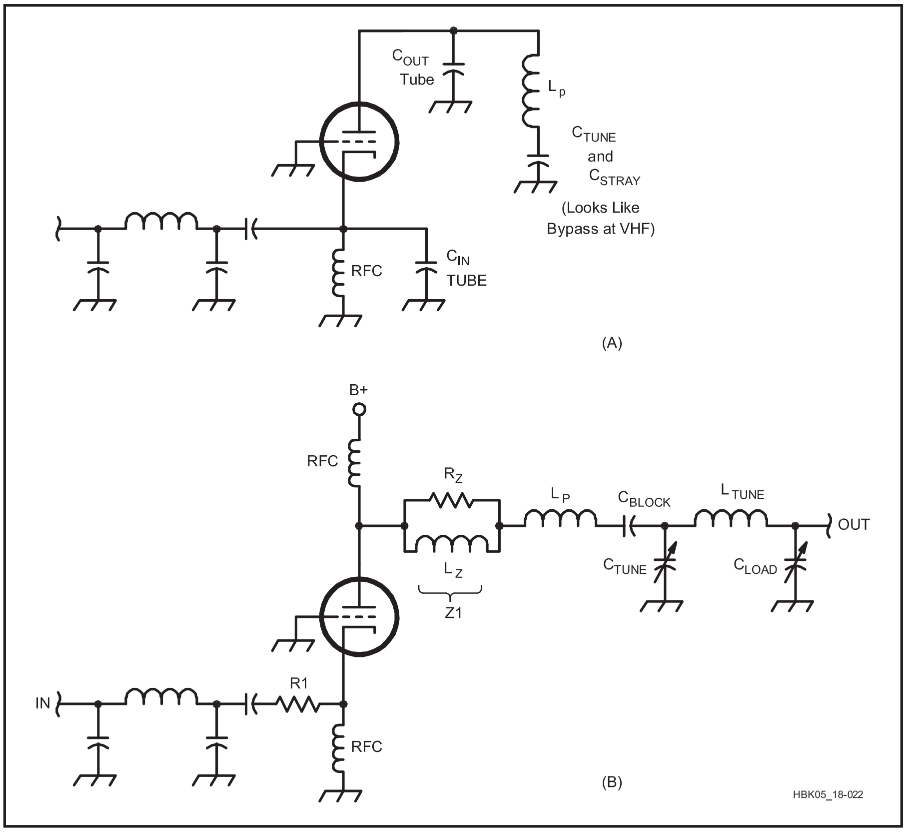


Figure 17.31 — At A, typical VHF/UHF parasitic resonance in plate circuit. The HF tuning inductor in the pi network looks like an RF choke at VHF/UHF. The tube's output capacitance and series stray inductance combine with the pi-network tuning capacitance and stray circuit capacitance to create a VHF/UHF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

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:

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

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, and
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 Figure 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. **Figure 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

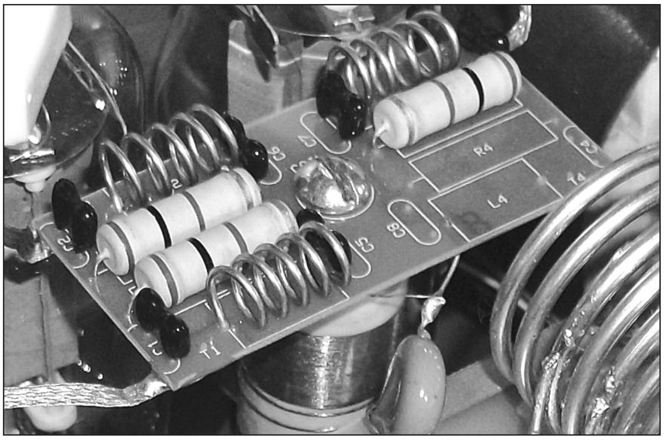


Figure 17.32 — Typical parasitic suppressor.

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 Figure 17.31B. If the tube has a relatively low input impedance, as is typical of grounded-grid amplifiers and some grounded-cathode tubes with large C_{IN}, R1 may dissipate a significant portion of the total drive power.

TESTING TUBE AMPLIFIERS FOR VHF-UHF PARASITIC OSCILLATIONS

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

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

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

LOW-FREQUENCY PARASITIC OSCILLATIONS

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

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

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

17.8.3 Reduction of Distortion

As mentioned previously, a common cause of distortion in amplifiers is over drive (flat topping). The use of automatic level control (ALC) is a practical way of reducing the ill effects of flat topping while still being assured of having a strong signal. This circuit detects the voltage applied to the input of the amplifier. Other circuits are based on detecting the

Overshoot and Overdrive Protection

The ALC and power control circuits of numerous transceivers allow short excess power transients at the beginning of transmission. While tube amplifiers are fairly tolerant of short overloads, solid-state amplifiers are not. To avoid damaging your amplifier input, some kind of protection is necessary. The article "Amplifier Overshoot — Drive Protection" by Phil Salas, AD5X, shows how to use a gas-discharge tube to limit short over-power pulses. It is available as part of the online information for this book.

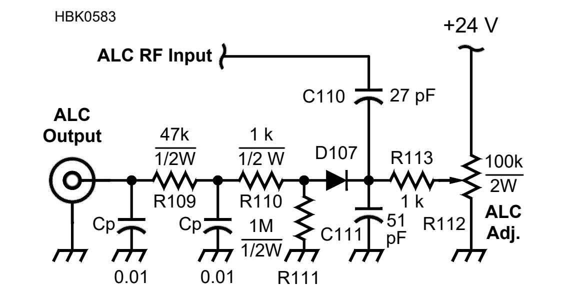


Figure 17.33 — An amplifier ALC circuit can be used to automatically limit the drive power from the transceiver to a safe level.

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 **Figure 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 **Circuits and Components** 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 **Figure 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 structures 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.

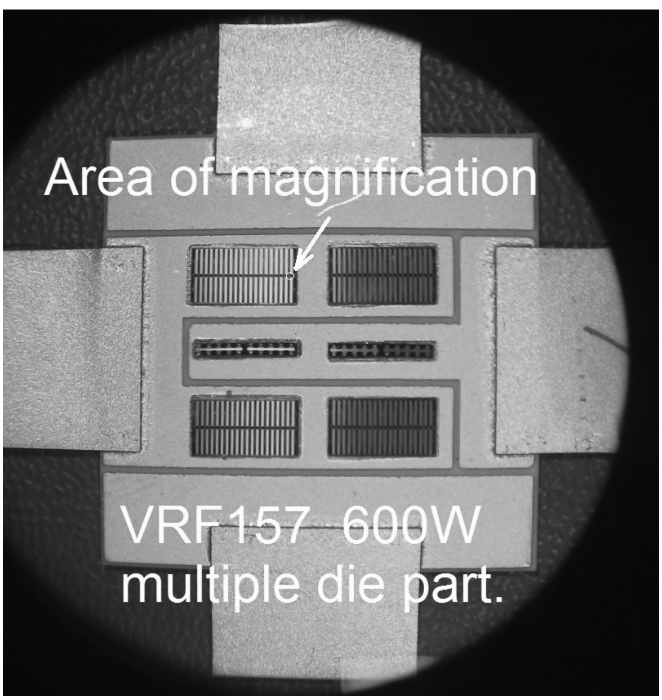
17.9.1 LDMOS versus VDMOS

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.

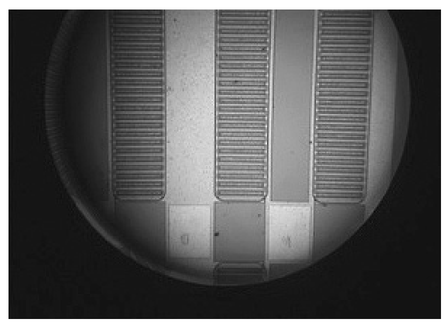
New RF amplifier designs are using LDMOS to replace bipolar transistors, which manufacturers are no longer making. The ability of LDMOS to operate at lower voltages is well suited for 12 V operation and, with its high gain and frequency response, providing 6 meter capability is simple. The downside is 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 degen-



(A)



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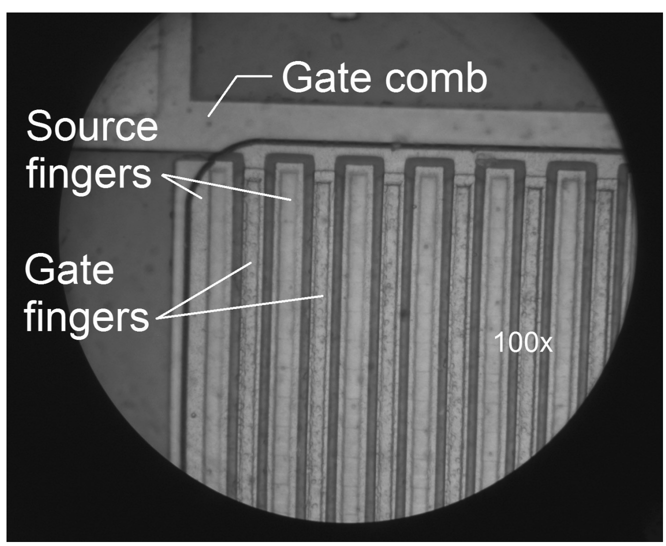


(C)

Figure 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]

eration 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 switch-mode 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 **Figure 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 **Figure 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 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 with this book's online content.

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,

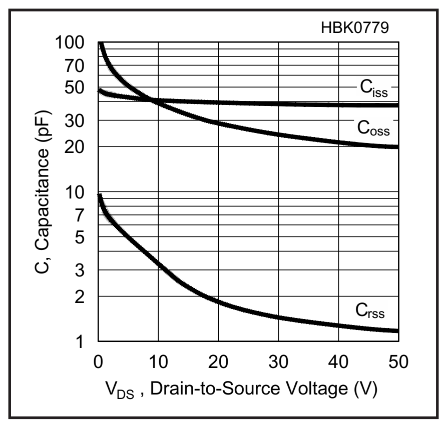


Figure 17.36 — The capacitance versus voltage (C-V) curves for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

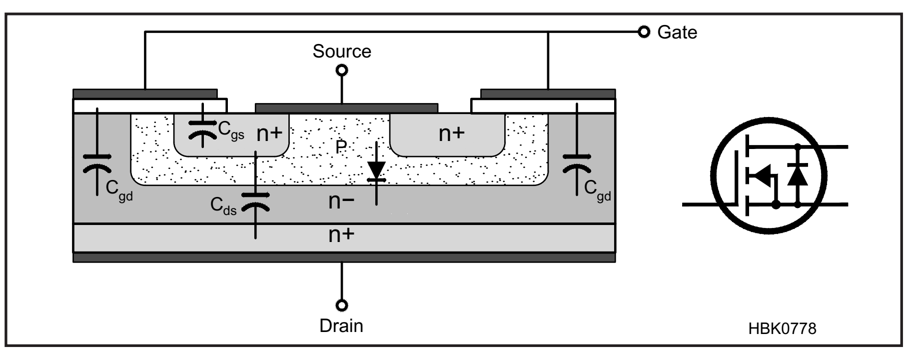


Figure 17.35 — All MOSFETs have parasitic capacitances as shown and a body diode between the drain and source in the cross-section and schematic symbol. The diode is shown for an N-channel device. It is reversed for P-channel devices.

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. **Figure 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 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 ($R_{\Theta JC}$ and P_{Dmax}) generates the third ($T_{j\,max}$). In this

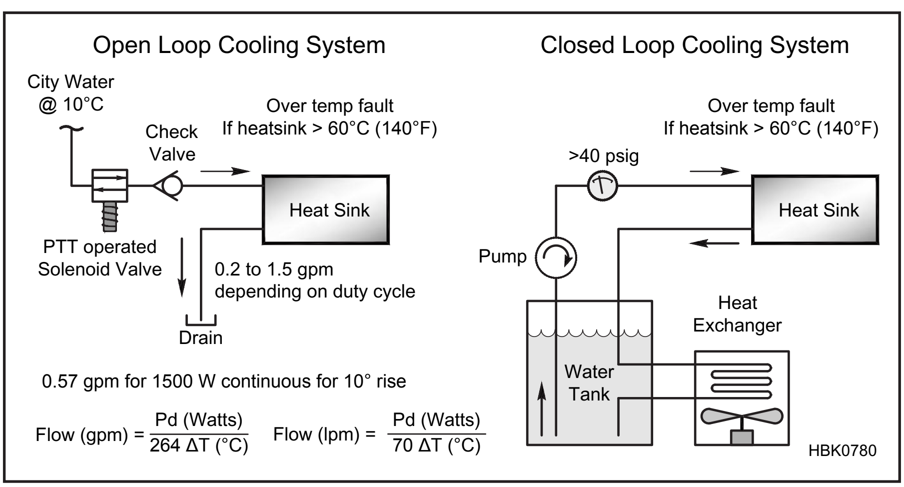


Figure 17.37 — Open- and closed-loop water cooling systems for solid-state amplifiers.

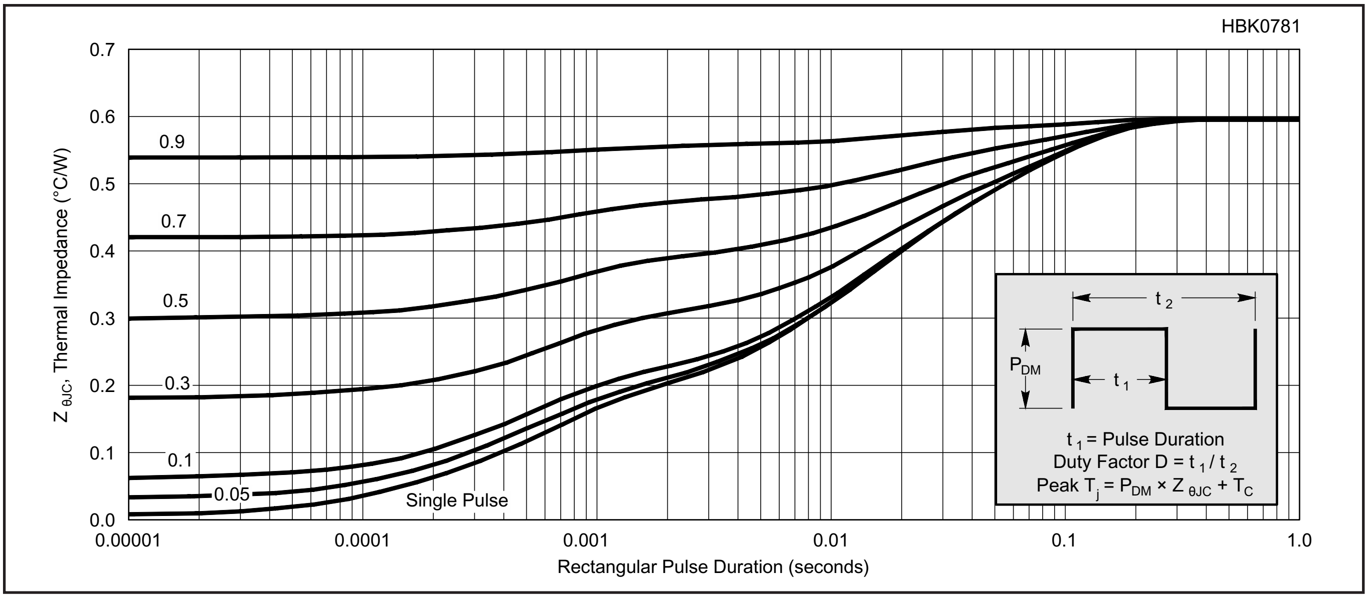


Figure 17.38 — Dynamic thermal impedance for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

sense specifying $T_{j\,max}$ is redundant but it is often reiterated for those who are looking for the particular parameter.

Because thermal performance is just as important as RF performance, much of the information in the data sheet is concerned with it. In addition to the static thermal impedance, $R_{\Theta JC}$, there is a dynamic thermal impedance $Z_{\Theta JC}$. Transistor packages have thermal mass so the die temperature does not change instantaneously. For pulsed operation, the effective $R_{\Theta JC}$, can be much lower than it is for steady-state operation.

The dynamic thermal impedance is shown in **Figure 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 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 Figure 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. PM can be observed as unequal IMD products on each side of the carrier pair in a two-tone test. PM distortion generates 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 and increases them on the lower side. (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 **Figure 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.

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

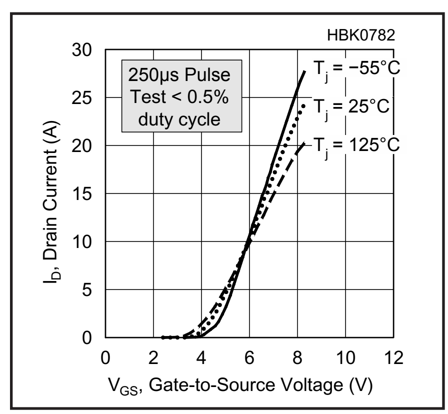


Figure 17.39 — Transfer characteristics for the Microsemi VRF151 RF MOSFET. (Illustration courtesy Microsemi Corp.)

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 **Electronic Design Automation (EDA)** 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 **Electronic Design Automation (EDA)** 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, available with this book's online content, 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.

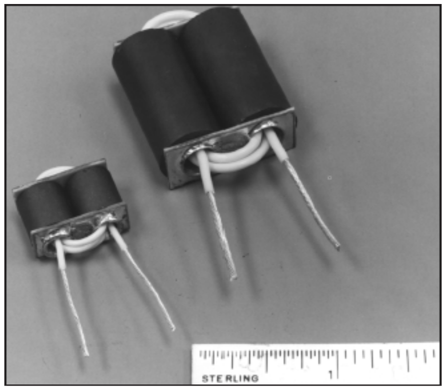
A single-transistor amplifier can be made in the same way, and in fact most single-band VHF amplifier "bricks" are. A tuned matching network provides the proper load impedance for the transistor and transforms it up to 50 Ω. The major difference is that the proper load impedance for a transistor, at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

BROADBAND TRANSFORMERS

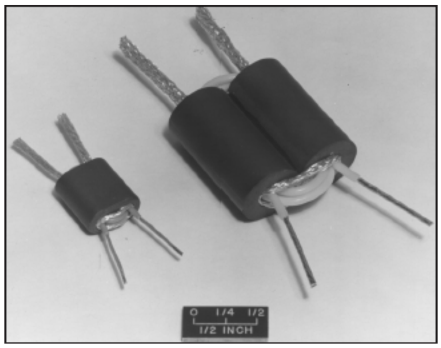
Broadband transformers are often used in matching to the input impedance or optimum load impedance in a power amplifier. Unlike the tuned matching circuits, transformers can provide constant impedance transformation over a wide range of frequency without tuning. Multi-octave power amplifier performance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

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.



(A)



(B)

Figure 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.

Typical examples of conventional transformers are shown in **Figure 17.40**. In Figure 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 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 Figure 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

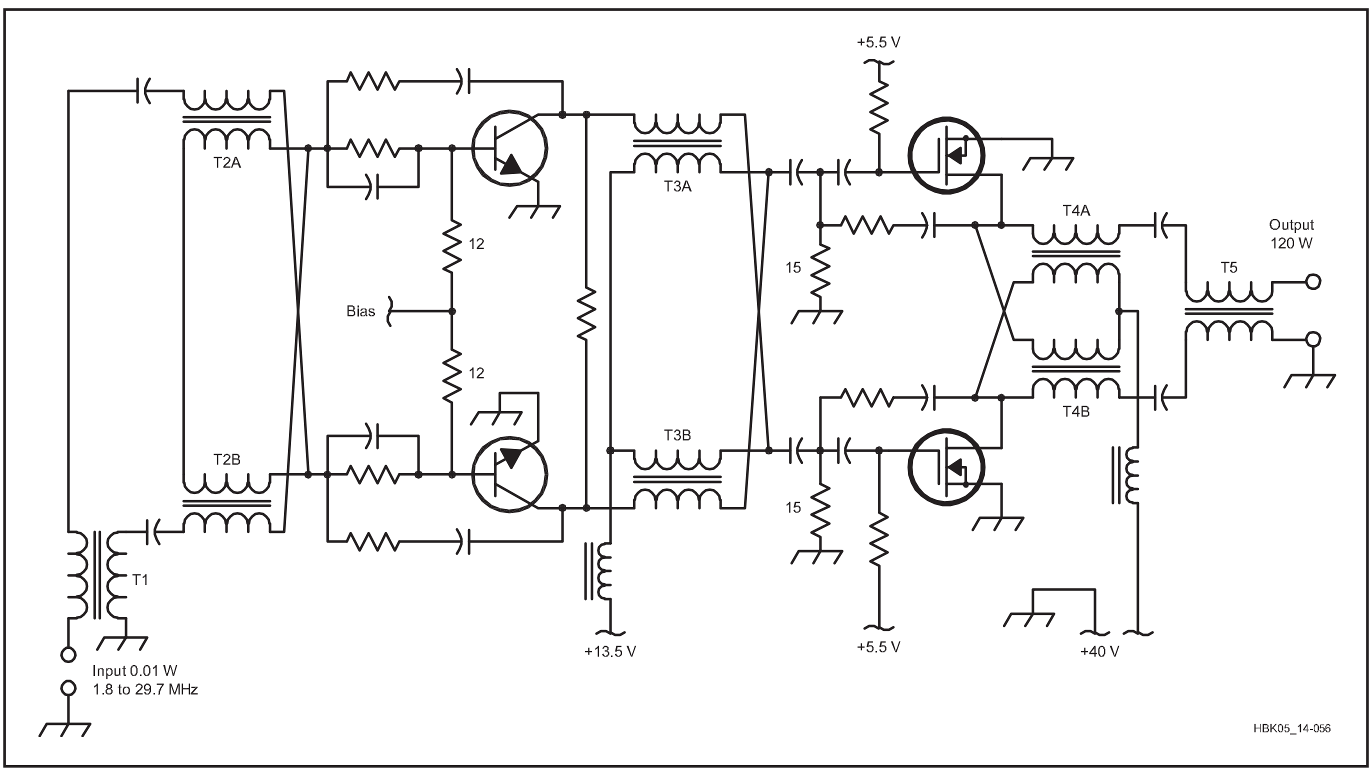


Figure 17.41 — Typical use of transmission line transformers as baluns and combiners in solid state power amplifiers.

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

Figure 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.

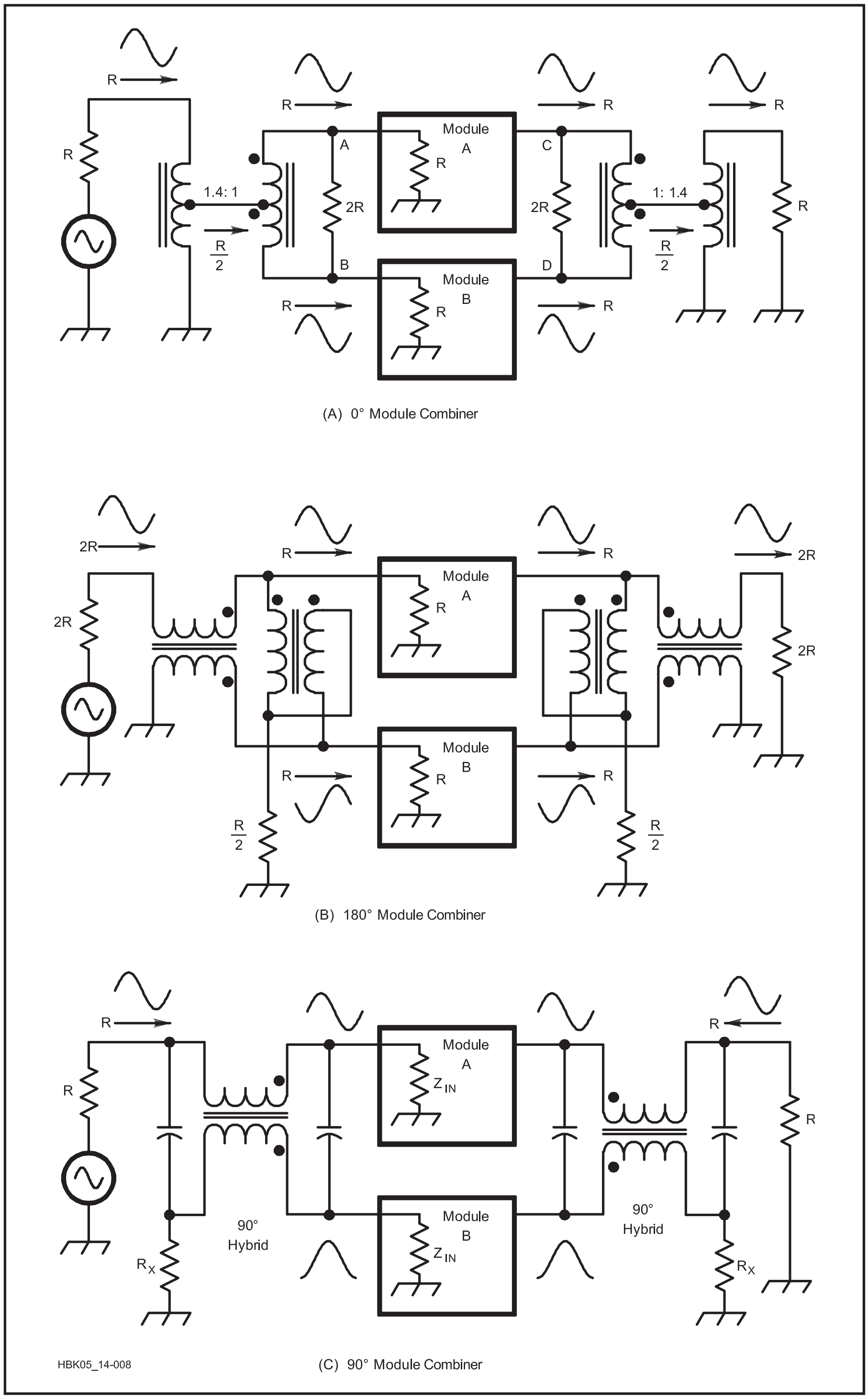


Figure 17.42 — Three basic techniques for combining modules.

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.

Figure 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

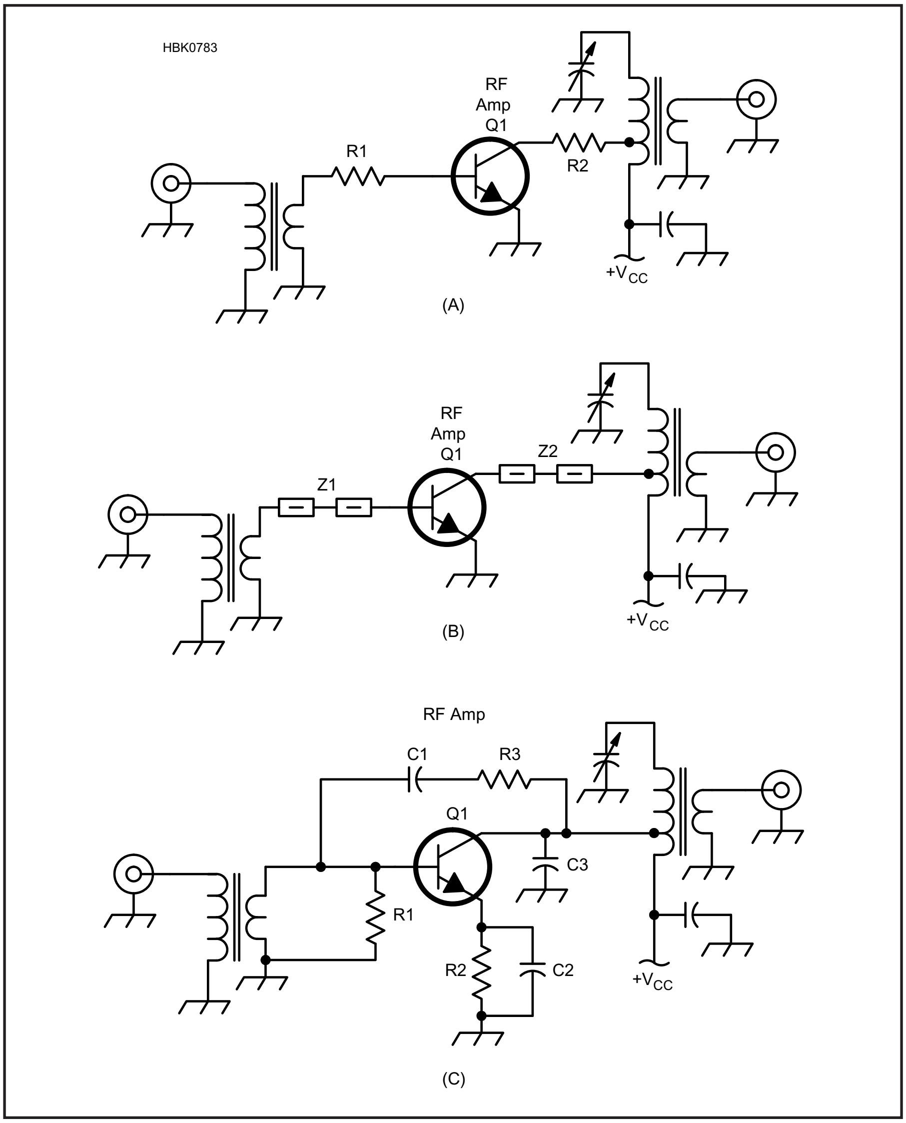


Figure 17.43 — Suppression methods for VHF and UHF parasitics in solid state amplifiers. At A, small base and collector resistors are used to reduce circuit Q. B shows the use of ferrite beads to increase circuit impedance. In circuit C, C1 and R3 make up a high-pass network to apply negative feedback.

if the generator resistance is closely equal to R. The output combiner separates the two outputs 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 **Figure 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 Figure 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 Figure 17.43C can be added to some power amplifiers to dampen VHF/UHF parasitic oscillations. The capacitor should be low in reactance at VHF and UHF, but must present a high reactance at the operating frequency. The exact value selected will depend upon the collector impedance. A reasonable estimate is to use an X_C of 10 times the collector impedance at the operating frequency. Silver-mica or ceramic chip capacitors are suggested for this application. An additional advantage is the resultant bypassing action for VHF and UHF harmonic energy in the collector circuit. C3 should be placed as close to the collector terminal as possible, using short leads.

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

LOW-FREQUENCY PARASITIC OSCILLATIONS

Bipolar transistors 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 Figure 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 Figure 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 Figure 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.

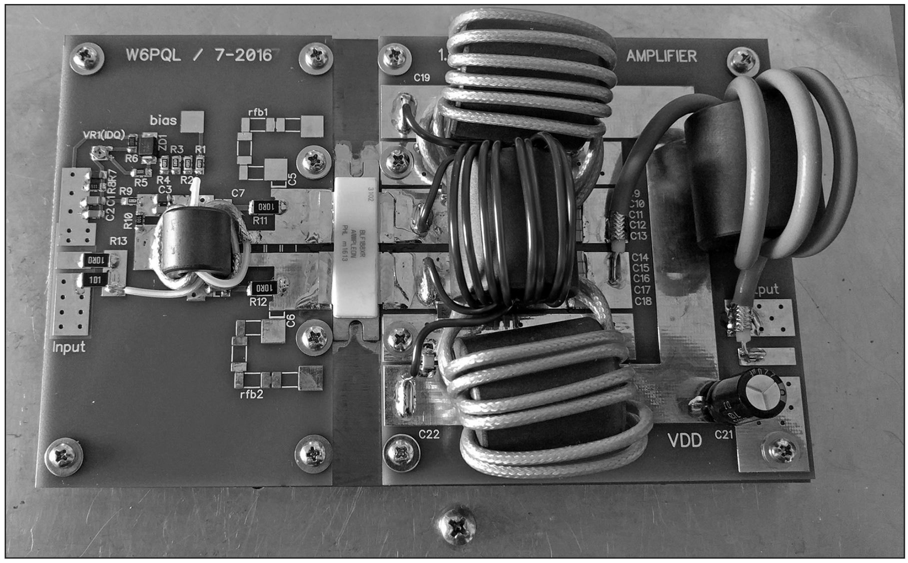


Figure 17.44 — Typical HF amplifier "pallet" designed by Jim Klitzing, W6PQL (www.w6pql.com/index.htm). This design is capable of 1 kW output power and uses the NXP BLF188XR LDMOS amplifier transistor module. [Jim Klitzing, W6PQL, photo]

17.10.7 "Pallet" Amplifiers

[*This section, written by Jim Klitzing, W6PQL, explains pallet construction and how pallets are used in an amplifier system. More information is available at the author's website:* ***www.w6pql.com***.]

In the world of RF power, a "pallet" is a fully functional solid-state power amplifier (SSPA) module intended to be used as a sub-assembly in a high-power amplifier system. A "pallet amplifier" is a complete RF amplifier based on one or more pallet modules. **Figure 17.44** shows a pallet designed by W6PQL for HF amplifiers. Modules and amplifiers are available for the MF/HF, VHF, UHF, and microwave amateur bands.

THE PALLET MODULE

A typical pallet amplifier module consists of at least the following components:

• The RF transistor itself — The most common type in use today for high-power applications is the LDMOS transistor, usually two identical devices inside the same ceramic or plastic package and intended to be used as a push-pull amplifier. An amplifier so configured has the advantage of suppressing even harmonics, making the design of an output filter less complex. Typical device part numbers are BLF188XR, ART1K6, MRF1K50, MRFX1K80, and many others. The pallet in Figure 17.44 is based on the NXP BLF188XR.

• A copper *heat spreader* base — A heat spreader distributes heat from the device transistor(s) over a larger area. The base of the spreader is then bolted to an even larger aluminum heat sink where the heat is drawn away with airflow provided by fans.

• Input and output matching circuits — The transistor module's input is normally fed with a broadband center-tapped transformer that will need to handle only a few watts of drive power. However, the output circuit must be able to handle 300 W to 2 kW as of early 2022. This level of power involves many amps of RF current and high RF voltages. Capacitors, coaxial or ferrite transformers, and coupling capacitors must be adequately rated to survive these conditions.

• External stabilization components — Most pallet modules have a gain of 30 dB at 2 meters, and much higher gain at HF and below. If this gain is not limited with stabilizing components, there will be spurious signals in the output stream. In the worst case, a module can oscillate at high power levels that ensure the transistors will be quickly destroyed along with other nearby components.

• An adjustable bias supply — An LDMOS transistor is usually an enhancement mode device. Such a transistor designed for 50 V operation may only require a positive bias voltage of 2 – 3 V applied to the gates to conduct. Without bias the device is cut off and will not conduct, though RF drive itself can provide some stimulus.

Transistor modules are either bolted or soldered to the heat spreader. The direct bolt-down method requires heat sink compound to be used on the module where it mates to the copper spreader. This allows for easier device replacement but is less robust than soldering the device directly to the spreader. A properly managed flow-solder process ensures the best possible RF connection to the circuit common ground plane and maximum heat flow into the spreader. Soldering can be performed in a home workshop using a hotplate.

TRANSISTOR BIAS

The pallet's bias supply must be well-regulated and adjustable; once the drain is conducting, an upward change of just a few dozen millivolts on the gates can cause the LDMOS idling current to rise from of a couple of amps to full saturation.

Bias voltage should also be temperature compensated. As the LDMOS transistors heats up during operation, the idling current will drift up as well. If uncompensated, drain current will rise quickly depending on how hot the transistor junctions are. This effect can cascade into thermal runaway.

Bias supply control can be performed by active or passive circuits. The active method uses a voltage regulator and a variable resistor. The passive method uses a Zener diode instead of the active regulator and has the advantages of being simpler to implement and less susceptible to RFI. With either method, a semiconductor PN junction or a thermistor can be employed as the temperature-sensing element.

RF DECOUPLING AND DC POWER

Most high-power LDMOS devices are designed for 50 V dc power. Newer transistors are designed for 55 – 65 V. In order to get to the kW output level or higher at 50 V, the required drain current can be 30 amps or more depending on the efficiency of the design. Higher drain voltages allow full power output at lower current levels.

To route high current to the transistor drains, some designers route dc power through the RF transformers at a properly bypassed RF-neutral winding tap. An alternate method is to route dc power separately to the drain connections through rf chokes bypassed at the dc connection. This second method reduces the additional IR losses in the RF transformers caused by the high currents. The first method is simpler (fewer parts) and the second lets the RF transformers run cooler. Both methods work well and are illustrated in **Figure 17.45**.

POWER LIMITATIONS

LDMOS transistors are available at up to 2 kW rated output in a single push-pull package. This makes a legal-limit HF pallet amplifier using a single device a real possibility.

At frequencies above HF, single-band pallets are the norm. IR losses mount with upward frequency changes and currently available components for a single-device pallet limit output to about 1200 W continuous at 2 meters. Some high-end pallets can easily develop 1500 W in SSB and CW modes where the duty cycle is low.

For FM or digital modes where the duty cycle can be very high, above 1200 W the transformers can over-heat with temperatures high enough to melt the solder holding them in place. One can mitigate this a bit with additional airflow across the pallet, but 1200 W is about the limit. To get to 1500 W in digital modes, two separate pallets or a dual-device pallet and combiners are often used.

At 1296 MHz and above, the maximum single-device pallet output is usually lower and you may need as many as 4 devices (properly combined) for a 1500 W EME amplifier.

There are also limitations on some of the other components in the design and these must be considered. RF and dc voltages and currents are higher, so two possible responses are:

• Increasing the size of the transformer core and wires, and

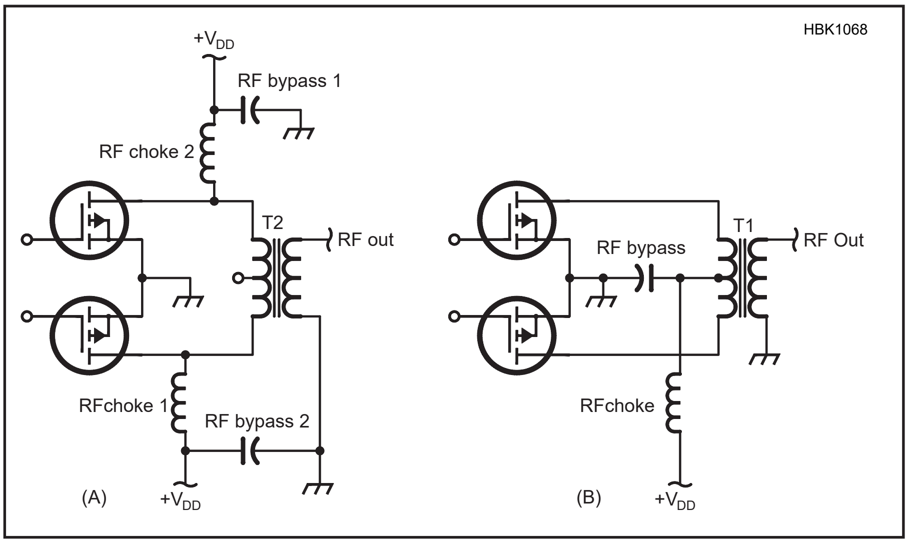


Figure 17.45 — Two methods of feeding dc power to the RF transistors. At (A), power is supplied separately to each transistor through an RF choke. At (B), power is supplied through the shared primary winding center-tap. See text for a comparison of the two methods.

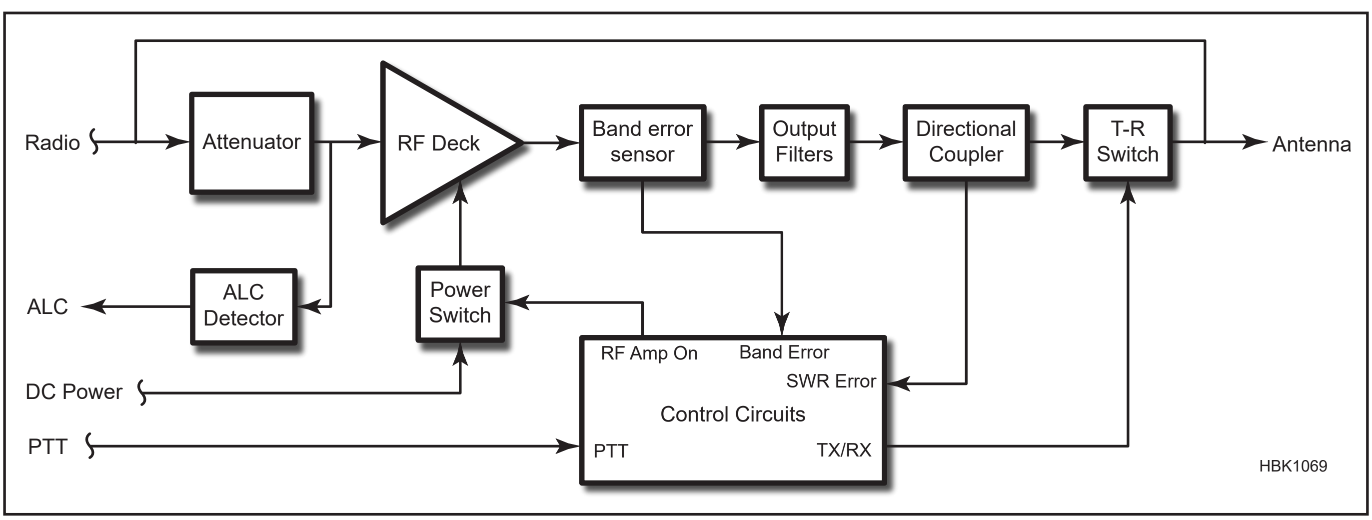


Figure 17.46 — A block diagram of a complete pallet-based RF power amplifier.

•Using matching capacitors rated at higher voltages and currents.

Capacitor current limits can be overcome by using two or more capacitors of the same type in parallel. For example, if one needs a 300 pF capacitor, using two 150 pF parts in parallel will do, as will using three 100 pF parts, etc. This method is currently employed in many designs.

BANDWIDTH AND DISTORTION

Pallet amplifiers designed for use on HF frequencies are normally very broadband, requiring no tuning into a 50 Ω load, and can cover several octaves if the transformers are designed properly. The no-tune feature does require well-matched antennas to facilitate the best power transfer — we give up the "plate" and "load" tuning adjustments associated with tube-type amplifiers for this convenience.

To achieve no-tune broadband performance some compromises are usually made. One of the most notable is efficiency; if one were to optimize for efficiency only, the result would be more narrow band. At VHF and above the narrower bandwidth approach is the most common, and resulting efficiencies are often at least 10% higher than at HF.

If the pallet amplifier is properly biased and not over-driven, typical IMD3 is –30 to –35 dBc. When not driven higher than the 1dB compression point it will be difficult for listeners to tell whether the amplifier is a solid-state pallet design or some other type. Linearity of solid-state RF power amplifiers is discussed the following section.

PALLET AMPLIFIER SYSTEM DESIGN

A useable pallet amplifier involves a complete design, including associated protective and matching circuits, the output filter, power supply, and indicator/monitoring/control circuits. **Figure 17.46** is a block diagram showing the different functions and their relationship. The complete system consists of at least the following parts:

- RF deck (pallet)
- Antenna transmit-receive (T-R or changeover) switch
- Output filtering for harmonic suppression
- Cooling system
- Protective circuits for band selection errors (for multi-band systems), high SWR, and excessive temperature
- Band-select system (for multi-band systems)
- High current switches for dc power
- Control system for the system components
- Power monitors for forward and reflected power
- Metering for voltage and current
- RF attenuator to limit input power
- Power supply
- A suitable cabinet to house all components

Antenna T-R Switch

The antenna T-R switch has two important jobs to do; in its default (unpowered) state, it connects the antenna to the radio. When switched to the transmit state, it will place the amplifier in the transmit path between the driving radio and the antenna.

The most common antenna switch is usually an electro-mechanical device (a relay or relays) but could also be a solid-state switch of some type (typically a PIN diode switch). The mechanical types can be general-pur-pose relays or coaxial types; general-purpose relays are often used at HF and coaxial relays are more commonly found at VHF and above.

For systems with a separate input and output relay, the input relay needs to handle no more power than the driving radio can produce, but the output relay must be more robust and able to handle as much as 1.5 kW. Some relays have two poles and are capable of handling both receive and transmit switching if the isolation between the poles is adequate. A good rule of thumb is to have at least 10 dB more isolation than the amplifier has gain; this is necessary to avoid oscillation caused by positive feedback from leakage of the amplifier output to the input.

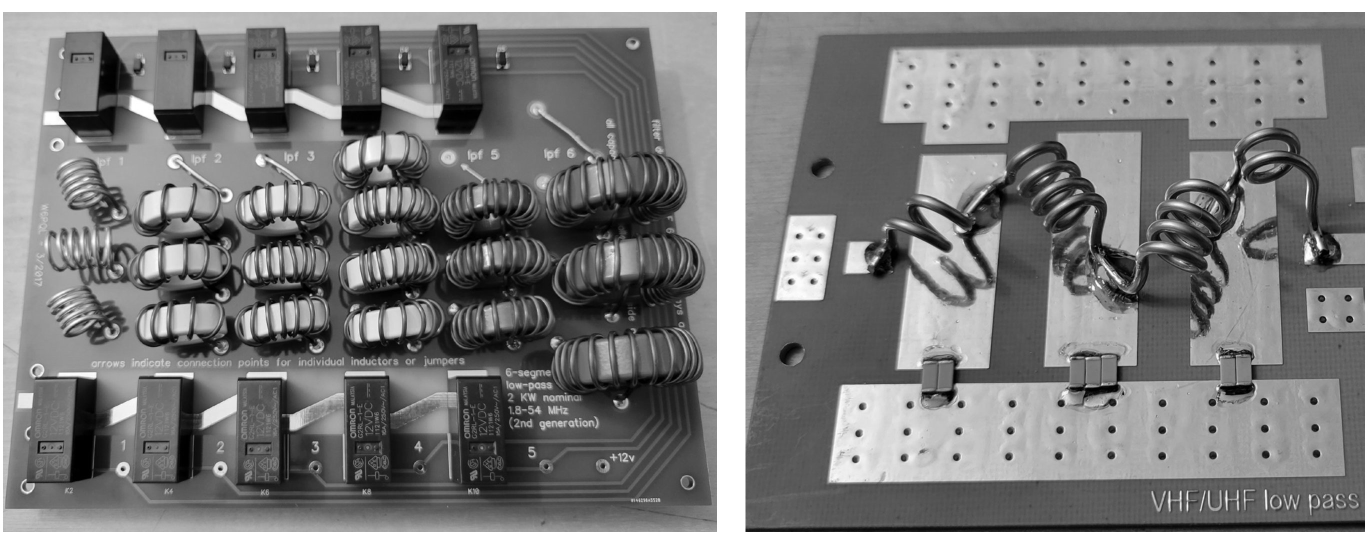
Output Filtering

Output filtering for single-band systems can consist of a single bandpass filter or a low-pass filter (LPF), the latter being the most common. **Figure 7.47** shows two examples of amplifier output filters. Figure 7.47A is a multi-band HF filter assembly and Figure 7.47B is a single-band low-pass filter for a 220 MHz amplifier.

For multi-band HF amplifier systems, the filter usually consists of several separate LPFs which are selected according to the band in use. For example, one LPF section with adequate harmonic suppression above 24 MHz can be used for the 14, 18 and 21 MHz bands. It is typical for the LPF to have 5 or 6 different selectable sections with different cut-off frequencies to cover the full range of 160 through 6 meters. These can be assembled on one PC board and selected using general-purpose relays capable of handling the high RF voltages and currents.

Cooling System

The RF deck on its copper heat spreader should be mounted to a heat sink (aluminum extrusion is widely used) with a thin coating



(A) (B)

Figure 17.47 — Output LPF filters for a multi-band HF amplifier (A) and a VHF single-band amplifier (B).

of heat sink compound covering the surface of the copper base where it mates to the heat sink. A good rule of thumb for cooling is six pounds of aluminum heatsink with a 0.375-inch base under the fins and 80 – 120 cfm airflow per kilowatt of power generated.

Some builders have fitted the copper base with water cooling. While this method is quiet and very efficient it also has some disadvantages for convenience and adds additional complexity. It is seldom necessary unless the unit will be operated in an extremely hot environment.

Thermal dynamics is a science all its own and some experimentation may be in order; for example, if the distance between the output of the fans and the heat sink fins is too close, the back pressure to the fans will be high and restrict the airflow. You can evaluate this before you build, changing the distance and monitoring the airflow on the other side of the heat sink to see this effect. About 3/4 inch seems to be the correct distance in most cases. The warm exhaust air must also be directed out of the cabinet in such a manner that it does not flow back over the top or sides of the heat sink and dump heat back into the system. One method which works well is to pull cool air in from the bottom rear of the cabinet and then expel the warm air out the top of the front.

Protective Circuits

Human error is a certainty as is the occasional unexpected act of nature. We get distracted and forget to reconnect our antenna to the radio or amplifier after doing some sort of test, moving equipment around, or perhaps an animal chews on the feed line. It happens and whatever the cause we don't want all that RF energy coming back to overload circuits and cause components to fail.

A control system relies on an SWR sensor to lock out the amplifier when reflected power gets close to damage levels. A simple lockout for a kW amplifier should activate around 100 W of reflected power. Some sensors are even more sophisticated and can sense 2:1 SWR or higher with just a dozen watts of forward power but the simple lockout sensor is adequate, provided it is fast enough to prevent damage. Such a sensor depends on sampling reflected power and is usually part of a dual directional detector monitoring both forward and reflected power levels. See **Figure 17.48** for an example circuit.

A high-temperature lockout sensor should be able to trigger the control system if the amplifier cooling system cannot maintain safe operating temperatures. Unsafe conditions can be caused by hot weather, blocked exhaust ports, or something else causing the amplifier to overheat. Today's modern LDMOS devices are tough but letting things cool down until safe operating temperatures are reached again will prolong the operating life of those and many other components. A small thermistor attached to the main heat sink makes a good sensor. Thermistors are not rapid responders, and this is a good match for the gradual changes in heat sink temperature. Fast sensors tend to make the fans 'stutter' (turn on and off rapidly) and this can be an unnecessary annoyance.

Another protective circuit which is good to have with multi-band (HF) systems is a band-select error sensor. If you transmit on 20 meters with the band switch on 40 meters, the output LPF will reflect all the power to the RF deck. Since the SWR sensor is on the output side of the LPF it won't see this error, so you'll need a separate reverse power sensor between the RF deck and LPF to catch this condition. This sensor would not be needed if the amplifier is fitted with a band-select decoder allowing the radio to control the amplifier band. The sensor threshold is set to a high enough level not to trip on harmonic energy reflected from the LPF.

High-current Switch

A switch for disconnecting the RF deck from the 50 V power connection during receive is another safety feature. This disconnect absolutely shuts off the transistor and helps to ensure stability. For example, during receive the antenna switch will leave the pallet with no load and if left active in that state it could be unstable.

The switch can be a relay, but with power MOSFETs the power can safely be switched on or off in less than 1 msec. In the ON state, the device forward resistance is just a few milliohms so there is very little IR loss (½ V or less). These devices can also handle 35 amps or more of current at voltages higher than 50 V and can be paralleled for additional current capacity if necessary. **Figure 17.49** shows a sample current switch circuit.

Control System

With so many components making up the amplifier system, an effective method of coordinated control is necessary. For the system to work properly some functions need to happen in the proper order. This part of the control system is called a *sequencer*. Other things (like fans) can operate independently. The control system circuits are usually consolidated onto one PC board with the appropriate sensor inputs and control outputs. (A sample control system schematic and PC board layout can be seen on the author's website.)

Everything requiring coordination begins with the radio signaling the amplifier it is going into the transmit state. The channel for this is the PTT connection between the radio and the amplifier; this is usually a control line being connected to ground. When this happens, the following should take place in this order:

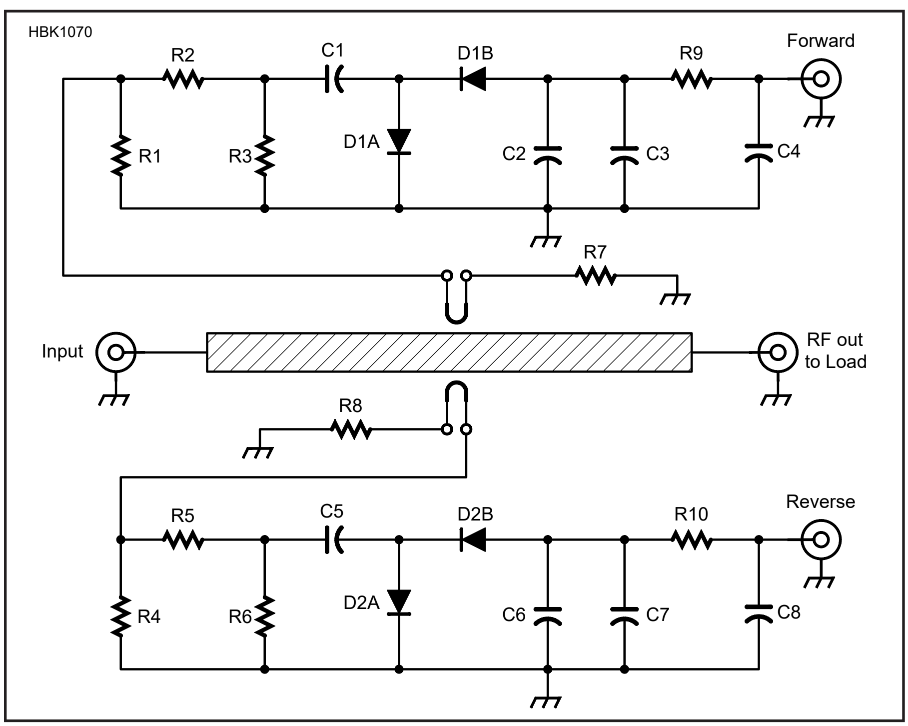


Figure 17.48 — Basic schematic for a dual directional RF coupler, used to monitor SWR in an RF amplifier.

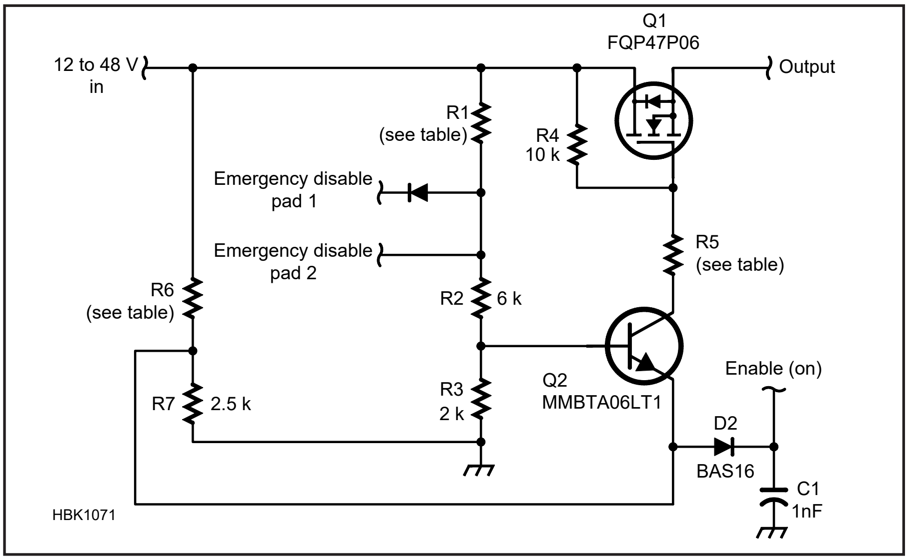


Figure 17.49 — Schematic for a high-current dc switch using a power MOSFET.

1. The fans should be switched on and the change-over relays in the amplifier should switch into the transmit state. Failure to do so would cause them to be hot-switched which shortens the life of the relay contacts and often triggers a high-SWR lockout during the time the relays are settling. The duration of settling is about 20 – 30 msec with general-purpose relays.

2. 30 – 50 msec after Step 1 the high-current dc switch feeding power to the RF deck should be turned on.

3. 30 – 50 msec after Step 2 the bias should be turned on, allowing the amplifier to be fully activated. At this time, transmit inhibit (if used on the driving radio) should be released.

The three events listed above should be switched off in reverse order when the PTT signal from the radio is deactivated, allowing the receiver to function again.

One of the independent safety functions is the heat sensor. It should monitor the temperature of the heat sink continuously and signal the fans to turn on if the temperature rises to moderate levels. If the temperature rises to exceed safe operating levels, the sensor should signal the system to drop out of the transmit path and keep the fans on. This can be done by disabling the PTT function, allowing the sequencer to shut the system down gracefully. Resetting this lockout can be allowed to happen automatically once safe temperatures are reached. Some hysteresis should be built into the mechanism to keep this control from chattering on and off at the temperature thresholds.

A reflected power sensor must trigger an immediate lockout if reflected power exceeds safe levels. Damage can happen quickly with this type of fault so there should be no sequencing delays for this lockout; the control mechanism should immediately command the high current switch to remove power from the RF deck and at the same time tell the sequencer to begin removing the system from the transmit path.

Lastly, some mechanism should be provided to place the amplifier online or offline. An easy way to accomplish this is to use a front panel switch to interrupt power to the sequencer mechanism, or just use the switch to disconnect the PTT input.

Overdrive Protection and ALC

Overdriving an amplifier can cause problems, the most serious of which is damaging the power transistors. One way to provide protection is to use ALC feedback to the driving radio, but this must be done properly to avoid distortion. Whether using a tube or solid-state amplifier, if ALC is only used to control output levels it will cause an amplifier to be non-linear in its response and thus create distortion.

ALC can be employed to prevent overdrive. For example, if an amplifier reaches full linear output with 50 W of drive, setting the ALC to begin reducing drive at 50 to 55 W will prevent damage to the amplifier. In this case, reducing the radio's maximum drive power to just below the point where ALC begins to assert control will get the job done correctly. This is difficult to do on a multi-band amplifier because of different drive levels needed for each band, but if the operator is using a radio capable of controlling power output on each band it will work very well.

An even better way of protecting the amplifier from overdrive damage is to use an input attenuator. Should the radio be capable of a maximum output of 100 W and the amplifier only requires 50 W to reach full output, then an input attenuator reducing the radio's full output power to 80 W or less will provide enough protection to avoid damage. Some accidental overdrive will not damage the transistors, but one must still be aware an overdriven amplifier will cause distortion and unnecessary splatter no matter what method of damage protection is used.

Monitoring Power Levels

Monitoring forward and reflected power is useful for helping the operator make proper adjustments to drive levels and watch for trouble in the antenna system. A good example might be operating in bad weather when high winds, water leakage, or an ice storm might be affecting the antenna; if so, reflected power levels will alert the operator to problems. Adjusting a manual antenna tuner for a good match is also made easy with power monitoring.

The monitoring displays or meters can get their signal information from power detectors; these detectors are usually placed at the output of the LPF with the detectors feeding both a power monitor and the control system at the same time.

While meters are suitable if properly interpreted, peak power is best monitored using peak-reading displays. SSB has lower average power than some of the other modes, but peak power levels are quite a bit higher and should be monitored for compliance with power limit regulations. An LED bar graph display, common on transceivers and amplifiers responds nearly instantly to changing levels and shows peak readings clearly. (Peak and average power measurements are also discussed in the **Test Equipment and Measurements** chapter.)

Power Supplies

Both linear and switching power supplies have advantages and disadvantages. A linear power supply is much heavier due to its iron transformer making the voltage conversion from the AC mains at 50 – 60 Hz. A power transformer for a legal-limit amplifier can weigh 50 pounds or more. The main advantage of linear supplies is the absence of RF noise.

The switching power supply is now the most widely used. A supply equivalent to the linear type might weigh 10 pounds or less and be as efficient as 90%. (See the **Power Sources** chapter for more about switchmode power conversion.) Most supplies are properly shielded and filtered to minimize interference generated by the switching waveforms. Still, it is always good practice to use ferrite chokes on all the power leads going into and out of the power supply enclosure. If this is done properly, even sensitive receivers will be unaffected.

17.11 Solid-State Amplifiers and Intermodulation Distortion

Trade-offs are a fact of life, and we certainly see them in amateur transceivers and linear RF amplifiers (referred to just as "amplifiers" elsewhere in this section). Solid-state transmitters and amplifiers are more convenient to operate with no tuning and automatic band-changing. However, tube transmitters with 6146A tube finals and negative feedback produced much "cleaner" signals with less IMD (intermodulation products — often called "splatter"). The difference is significant since a SSB signal today can be at least twice as wide as for the top tube transmitters such as a Collins 32S-3 or Yaesu FT-102 (when measured where the signal's sideband amplitudes have decreased to 60 dB below their peak value).

The first step, of course, must be to adjust the transceiver so the output signal is clean, with a minimum of IMD products. However, an excessively "wide" signal cannot be blamed solely on transceiver-generated IMD. An amplifier will contribute its own IMD products, and today's solid-state amps produce IMD that can be 10 to 20 dB greater than a tube amp of equivalent power. Why does this happen, and what are the remedies?

17.11.1 Solid-State Device Linearity

QST Product Reviews publish input/output linearity graphs for amplifiers. **Figure 17.50** is an example of a modern LDMOS amp with a relatively straight linearity curve. **Figure 17.51** is an example of an LDMOS amp with a curve that begins to flatten out just above half power. For a clean SSB signal with a minimum of IMD products, this amp should be run at half power. We have to share our crowded bands, and driving an amplifier into its non-linear regions isn't being a good neighbor.

With all the advances in solid-state technology, why are today's MOSFETs worse, from the perspective of amplifier linearity, than the previous generations of bipolar transistors? (See this chapter's section "MOSFET Design for RF Power Amplifiers".)

Bipolar RF power transistor circuits are inherently more linear than MOSFETS because of their thermal characteristics. Bipolar transistors cannot be directly paralleled because the hottest one — even on a perfectly designed IC die — will have increased gain, consume more of the total current, and eventually overheat and destroy itself. The technique, called *emitter ballasting*, deposits a low-value metal resistor in series with each emitter where it connects to the emitter buss of the device.

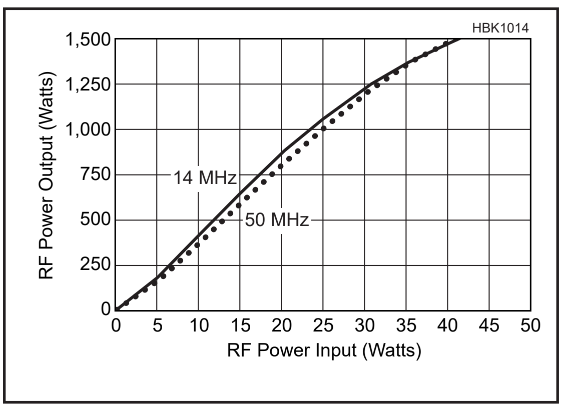


Figure 17.50 — A relatively straight linearity curve for a legal-limit LDMOS solid-state power amplifier.

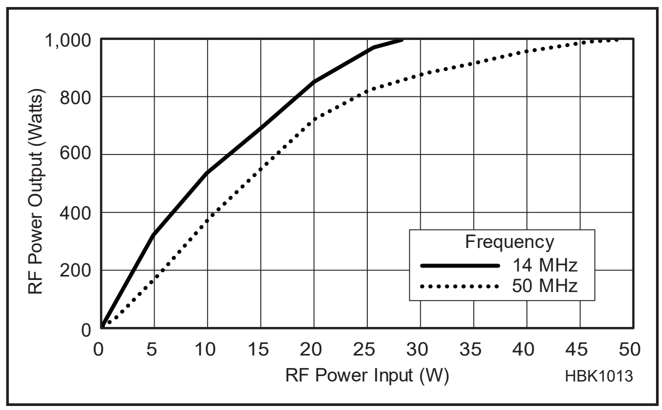


Figure 17.51 — Linearity curve for a 1 kW LDMOS solid-state amplifier that begins to flatten out (becomes nonlinear) at around half of its rated power.

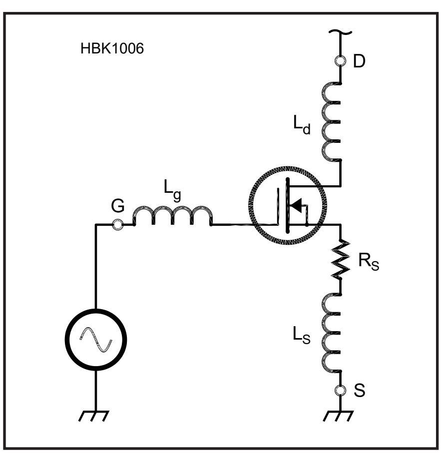


Figure 17.52 — Schematic of a MOSFET connected as a common-source amplifier showing parasitic lead inductance and the intrinsic source resistance.

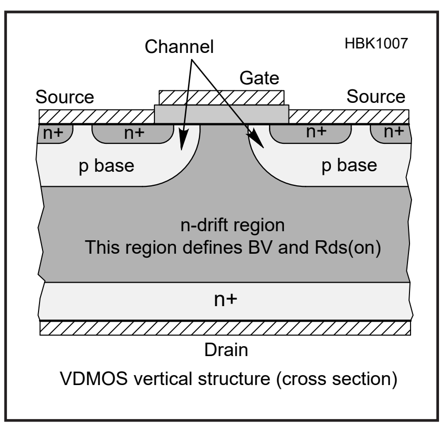


Figure 17.53 — Cross-section of a VDMOS transistor showing the vertical structure with the drain at the bottom which is connected to the device package.

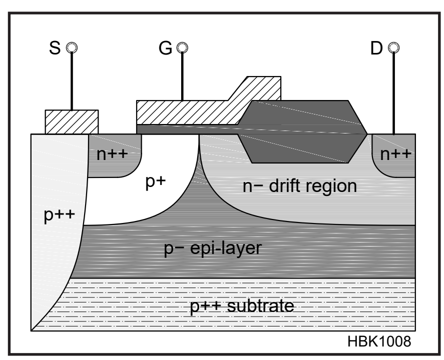


Figure 17.54 — Cross-section of an LDMOS transistor showing the lateral (horizontal) structure with a vertical connection ("sinker") between the source terminal and device substrate.

This solves the thermal problem by adding negative resistive feedback but reduces the available gain by about 3 dB.

Conversely, MOSFET (LDMOS and VDMOS) cells can easily be paralleled (if their size is chip-scale) because the hottest one becomes more resistive via the intrinsic positive temperature coefficient of its drain resistance. This reduces gain and shifts current to the other MOSFETs, stabilizing the transistors thermally.

COMMON-SOURCE AMPLIFIERS

In a common-source amplifier (see **Figure 17.52**), the source is grounded, the drive is applied to the gate, and the output is taken at the drain. Parasitic elements inside the device, such as bond wires connecting the die elements to the package terminals, cannot be ignored. Any impedance in series with the source causes a voltage that subtracts from the drive signal producing negative feedback. Resistance gives feedback that is constant with frequency, and inductance gives feedback that increases with frequency. Parasitic inductance in series with the gate and/or drain can usually be managed by the matching circuits. In general, negative feedback will improve device linearity at the expense of circuit gain.

VDMOS TRANSISTORS

In V(vertical)DMOS parts, the back side of the die is the drain as shown in **Figure 17.53**. The reason LDMOS transistors are more powerful (i.e. much lower thermal resistance) per die size than their VDMOS cousins is that the die does not need to be electrically isolated from its package mounting flange. It can be soldered directly to a thermally and electrically conductive header, which is the source that we wanted to run at ground potential anyway in the common-source configuration. However, it is now impossible to isolate the source enough to introduce some negative feedback there.

The VDMOS gate and source are wirebonded to the package terminals. The die is attached to an electrically insulating BeO (beryllium oxide) substrate that is brazed to the package flange. The thermal resistance of this substrate can easily double the device's junction-to-case thermal resistance ($R_{\theta c}$), reducing its available power dissipation by half.

LDMOS TRANSISTORS

In L(lateral)DMOS devices, all the electrodes are laid out flat rather than stacked vertically, so the gates and drains are edge-on, giving very low gate-to-drain feedback and very high gain. (See **Figure 17.54**.) The sources are connected to the back side of the die using p-well "sinkers" (at left in the figure), so bond wires for the source connections are no longer needed. The die is only 2 or 3 mils thick, so there is virtually no negative feedback ("source degeneration").

A diffused p++ sinker through the p– epitaxial layer is used to connect the source to the back of the substrate. Each source is thus reliably connected to the back side of the wafer, which is then attached directly to the package flange. The gate and drain are wirebonded to the package terminals.

LDMOS parts are relatively low voltage because the breakdown voltage of the drain-source junction is a function of doping resistivity and the distance between electrodes. In a lateral device, more distance requires more expensive die area, reduces gain, and lowers efficiency. If the p-well sinkers connecting the individual cell sources to the back side of the die could be made more resistive, there would be an improvement in linearity, just as in the bipolar devices.

In today's LDMOS devices there is a patented high-pass network on the gates that keeps the gain from being almost infinite on lower frequencies, such as 160 meters. It uses a high-value but somewhat-nonlinear doped silicon resistor paralleled by a capacitor, which acts as a divider against the shunt capacitance of the gate. The resistor is needed for biasing the gate. This explains why LDMOS gates are rather intolerant of overdrive in addition to being another source of nonlinearity that is built-in to the device itself.

LDMOS devices are being used for new amateur amplifier designs because they are the dominant commercial technology and thus, the most cost-effective. Amateurs should be thankful there are plenty of high-power RF applications to keep the price of transistors down and availability high.

17.12 Adaptive Predistortion

17.12.1 Predistortion Overview

Predistortion is applied today in many commercial applications to reduce IMD and thereby reduce the bandwidth consumed by transmissions and conserve frequency spectrum. Some current applications include cell phone transmissions, digital TV transmissions, and digital FM broadcasts. While one offshore amateur transceiver manufacturer has provided predistortion since 2014, several newer amplifiers have built-in samplers for predistortion, and other manufacturers have discussed providing predistortion, but this capability has not yet been widely deployed in amateur radio. This is at least partially because the challenges of applying this technology in an amateur radio environment are different than those of commercial environments and new and innovative solutions are often required.

When configuring predistortion for these commercial applications, the antenna and its load impedance, the amplifier gain, phase, and other characteristics, as well as the specific frequency or frequency range of operation are all known in advance. The frequency range is typically a small percentage of the absolute frequency. The situation is very different for amateur radio — amateurs use a wide variety of antennas presenting different load impedances to the amplifier, a wide variety of amplifiers using different technologies and with very different characteristics, and the frequency range of many transceivers and power amplifiers spans from 1.8 MHz to 54 MHz, the highest frequency being thirty times the lowest!

There are certainly other less technical approaches to reducing bandwidth by reducing IMD. These are usually less effective and less convenient compared to predistortion. For example, an amplifier can be operated in its most linear region as much as possible. This usually means operating at power levels well below the maximum rated power. Another approach is to obtain a Class A amplifier; however, these are very inefficient and not common in amateur radio.

The sections that follow present a brief look at predistortion. There are many variations on predistortion, and this material relates approaches that have been demonstrated to work for amateur radio applications.

17.12.2 IMD — The Cause and a Solution

The human voice has two key characteristics that result in the production of IMD in operating modes such as SSB and AM: (1) the amplitude of the signal varies, and (2) the signal contains multiple frequencies. If an amplifier has perfect gain linearity and constant phase shift as the signal amplitude varies, then, the output of the amplifier is a perfect amplified replica of the input. However, if either of these conditions is not met, the amplifier exhibits intermodulation, forming products at new frequencies from the existing frequencies in the input signal. Some of these new frequencies fall within the pass-band of the signal, producing distortion, while others fall outside the pass-band making them spurious emissions, referred to as splatter, which may interfere with communications on nearby frequencies.

Predistortion compensates for this by inserting an additional predistorter stage that modifies the input signal to the amplifier to cancel the distortions the amplifier introduces. **Figure 17.55** illustrates how a predistorter compensates for gain non-linearity.

The predistorter must "know" the gain and phase characteristics of the amplifier to apply compensating predistortion to the signal. The solution is to measure and respond to the characteristics of the amplifier using hardware and software in the transceiver. By doing that repeatedly, the system can automatically adapt to changing bands, temperatures, voltages, and other conditions that may impact the characteristics of the amplifier. This is called *adaptive predistortion*.

Figure 17.56 illustrates the gain characteristics of the same MOSFET amplifier on two different bands. The horizontal axis shows input signal amplitude with a maximum output of 100 W PEP into a 50 Ω load. As can be observed, the required predistortion correction is very different in the two cases.

17.12.3 Predistortion System Design

Figure 17.57 displays a simplified block diagram of a predistortion solution implemented for a direct-sampling radio. The low-level transmitter at the upper left is implemented in software and implements typical functions such as audio processing, band-pass filtering, modulation, and ALC. It produces a fully modulated and filtered baseband signal, the 0 Hz IF. (See the **SDR and DSP Fundamentals** chapter for more information.) Its output goes to the predistorter software that modifies the signal's amplitude and phase. The result is then up-converted to the RF frequency and DAC sample rate and then converted to an analog RF signal by the DAC.

To measure the amplifier's characteristics, a sample of its output is needed. This is obtained with a sampler that can be a resistive or capacitive divider, or perhaps a directional coupler. The sample is digitized by the receiver ADC and down-converted to baseband through the receive path. The predistortion correction required (both gain and phase) is calculated by comparing the down-converted amplifier output with the amplifier input signal just before up-conversion. The result of that calculation — the needed correction information — goes to the predistorter.

17.12.4 Implementing Predistortion

Digital signals are frequently processed in quadrature (*I*/*Q*) form. (See the **Modulation** chapter for more information.) In direct-sampling radios, this is true of the signals down-converted to baseband for receive and of those to be up-converted for transmit. For quadrature digital signals, each signal sample is represented by two numbers, *I*, the In-phase component, and *Q*, the Quadrature component. *Q* differs from *I* in that each frequency component is phase-shifted by 90 degrees. Quadrature signals provide a very convenient way to represent both the magnitude and the phase of a signal. Since *I* and *Q* are in quadrature, the magnitude of each signal sample is

$$M = \sqrt{I^2 + Q^2}$$

and the phase is conveyed by the ratio of *I* and *Q*, specifically $\phi = \tan^{-1}(Q / I)$.

As shown in Figure 17.57, to calculate the required correction, each input sample to the amplifier is compared with the corresponding output sample. For the magnitude

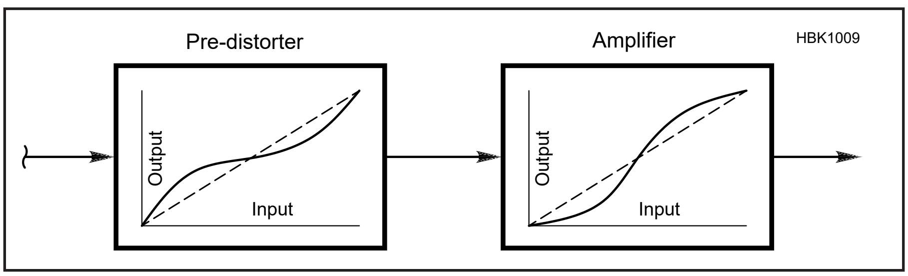


Figure 17.55 — Gain compensation by predistortion.

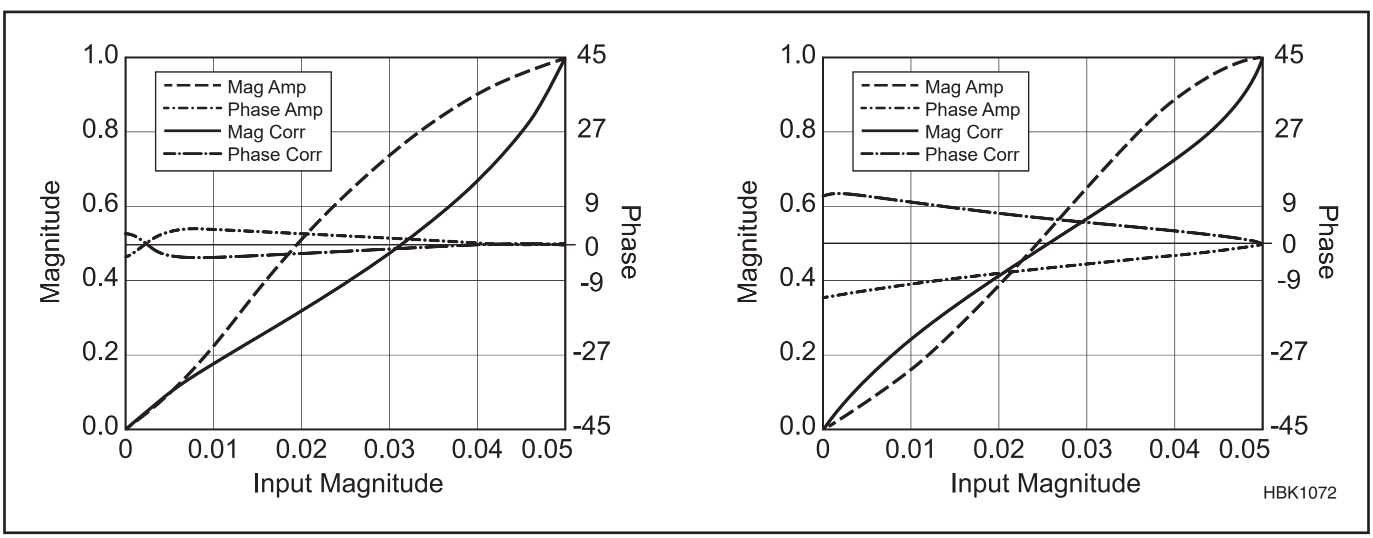


Figure 17.56 — Amplifier characteristics on two bands, 160 meters (left) and 10 meters (right). Dotted lines show the amplifier's magnitude and phase response at different amplitudes. Solid lines show the compensating magnitude and phase predistortion to correct for the amplifier's response. Note the differences between the amplifier characteristics on these two bands.

level of each output sample, the gain correction factor, G, and the phase correction factor, φ, are calculated. After some algebraic manipulation,

$$G = S(\sqrt{I_{in}^{\ 2} + Q_{in}^{\ 2}} \ / \ (\sqrt{I_{out}^{\ 2} + Q_{out}^{\ 2}})$$

$$\varphi = \ tan^{-1}((\text{-I}_{\text{in}} \times Q_{\text{out}} + Q_{\text{in}} \times I_{\text{out}} \ ) \ / \ (I_{\text{in}} \times I_{\text{out}} + Q_{\text{in}} \times Q_{\text{out}}))$$

where I_{in} and Q_{in} are the amplifier input sample, I_{out} and Q_{out} are the amplifier output sample, and S is a scale factor to compensate for the amplifier's gain and for the attenuation in the amplifier output feedback path.

Assuming the correction for all output sample values has been calculated, the correction is then applied. The magnitude of the predistorter input sample is first calculated, the corresponding values of G and φ are found, and these values are applied to the samples according to the equations below. G, the gain correction factor, is a multiplier and φ, the phase correction factor, shifts the phase by changing the ratio of I_{pd} and Q_{pd}, the resulting predistorted values of I and Q.

$$I_{pd} = G(I\cos(\varphi) - Q\sin(\varphi))$$
$$O_{pd} = G(I\sin(\varphi) - Q\cos(\varphi))$$

With regard to the phase correction, only the values of sin (φ) and cos (φ) are used to calculate I_{pd} and Q_{pd}. This means that, unless φ is to be used for some other purpose, e.g., a real-time plot of amplifier characteristics, it need not be explicitly calculated. Only sin (φ) and cos (φ) need to be calculated and stored.

There are several alternatives for the storage and retrieval of the correction factors. Two common approaches are table look-up and polynomial form. In using table look-up, it is normally not practical to store values of G and φ for each possible magnitude value, M. When retrieving values for correction, then the problem is what table values to use for a given M. Possible solutions include (1) choosing those at the nearest available value of M, or (2) interpolating between table values. In using a polynomial form, polynomials that are good fits to the G and φ versus M data are generated. Then for correction, these polynomials are evaluated at the signal values of M.

17.12.5 Practical Amateur Radio Implementation

When designing a robust predistortion solution for amateur radio, the expression "The devil is in the details" is to be revered. Some of these implementation challenges are discussed here.

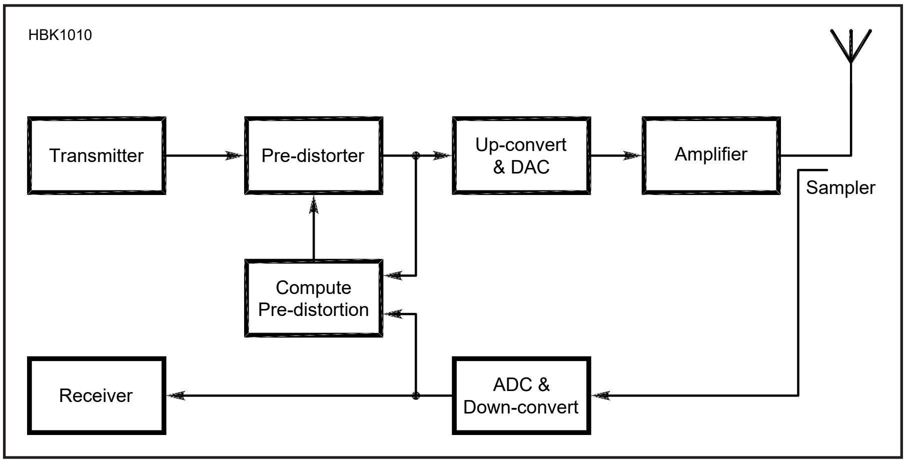


Figure 17.57 — Block diagram of an adaptive predistortion system for a direct-sampling radio.

CORRECTION BANDWIDTH

The preceding section discussed how to modify the gain and phase of the signal to achieve correction. It did not discuss what bandwidth is required for (1) the amplifier output feedback signal, and (2) the corrected upconverted signal driving the DAC. Some technical details are beyond our scope here; however, basically, if the predistorter can't "see" IMD products of a certain order, it can't correct them. This implies that the feedback bandwidth must at least accommodate the highest order products to be corrected. Likewise, if the output of the predistorter cannot convey IMD products of a certain order to the DAC, they cannot be corrected. This implies that the upconversion path must at least accommodate products to be corrected.

In practice, feedback and upconversion bandwidths of 48 kHz are generally adequate for SSB and AM operation. A frequently asked question is "Can a predistorter be implemented in an external box with the correction being applied to the microphone input signal?" The requirement for correction bandwidth is one of the reasons this is not practical since the microphone input signal path is limited by the band-pass filter. (Other reasons include lack of *I-Q* input for phase correction and the fact that other processing, such as compression, is often present in the microphone signal path.)

SAMPLE SYNCHRONIZATION

As discussed above, to calculate the correction factors, samples of the amplifier input and output that correspond in time are compared. I.e., each output sample is compared with the input sample that caused it. However, looking at the simplified diagram in Figure 17.57, it is evident that the feedback path from the amplifier output is much longer, having more delay, than that from the input. Furthermore, the amplifier itself introduces some amount of delay. Also, depending upon the communication channels, the delay from the amplifier output may not always be consistent.

DOWNCONVERSION FILTER RESPONSE

A second problem arises in that the samples from the amplifier output go through a downconversion process which includes the process of decimation to reduce the sample rate. Decimation is implemented in digital filters, which have time-domain overshoot and undershoot which can distort the signal.

A practical solution for both these problems is to provide a second downconversion and communication path, identical to the first. The samples going into the transmit DAC, i.e., the amplifier input samples, are then intercepted and processed through this second downconversion before comparison with the amplifier output samples. This ensures identical filter responses and identical downconversion and communication delays. Still remaining is some added delay in the DAC, amplifier, and ADC. This can be compensated for in software, at baseband, by resampling the feedback signals.

SCALE FACTOR (S)

For a given set of operating conditions and resulting correction solution, S is calculated to maintain precisely the same peak output power as the amplifier has demonstrated prior to the application of the new solution. This calculation can be done from the same data used to calculate the correction. In the event that a peak-power event is not present in the data, the data can be extrapolated to provide a good estimate.

CONTROLLING THE FEEDBACK LEVEL

To obtain optimum IMD reduction and overall performance and considering that the operator may adjust station power output from a few watts to 1.5 kW, an automatic mechanism to adjust the feedback level into the ADC should be provided. To the extent possible, the feedback level should be automatically adjusted to (1) maximize use of the full dynamic range of the ADC, but (2) take into account that passive IMD in the receiver filters and input circuits may increase at very high signal levels. Any passive IMD in the feedback circuit will increase the IMD of the transmitted signal. The typical result would be that the ADC is driven to within just a few dB of its clipping level. This control is easily done by monitoring the ADC peak output level and using that information to adjust a preceding attenuator.

OVER-DRIVEN AMPLIFIERS

Amplifiers that are being over-driven, to the point of having a flat or decreasing output with increasing input, present a special problem in the calculation of magnitude correction. To calculate this correction, the question to be answered is essentially "What is the input signal level that produces each particular output signal level?" However, if there is a range of high input levels for which the output is always the same, the answer to this question is indeterminate. This case can be handled by slightly modifying the measured amplifier response such that a solution is always available. Note that this only happens prior to the application of a predistortion solution; once predistortion is operating, the amplifier characteristic should never have this problem.

17.12.6 Adapting the Solution

In an amateur radio setting, having an adaptive predistortion solution that modifies itself as conditions change is nearly essential. Predistortion requires a high degree of precision since it changes the levels of very small IMD products that, with predistortion applied, may be 40 dB to 70 dB down from the desired signal. Therefore, to obtain best results, it is very important to repeatedly adjust the predistortion solution for even small changes in operating conditions. The solution must adapt to things like equipment temperature changes, small frequency changes, and certainly major events such as band changes and antenna changes.

Fortunately, while applying predistortion, the amplifier's output and input can still be measured and compared. A "next" solution can be calculated while a current solution is in use. Finally, there must be a transition from the current solution to this next solution. This adaptation process continues repeatedly.

Care must be taken in the transition from one solution to another. Considering that predistortion changes the magnitude and phase of the sample stream, an abrupt change in the solution would create a discontinuity in the waveform conveyed by the sample stream. This discontinuity would create a momentary broadening of the frequency spectrum and perhaps an audible "click." The solution is to gradually transition, over a period of a few milliseconds, from one solution to the next.

17.12.7 Memory Effects

Memory effects are present in an amplifier when the amplifier's output at the present time is affected by not only the present input but also by previous inputs. All of our amateur radio amplifiers have memory effects; however, depending upon technologies used and the design, some have more severe memory effects than others. The ability of amplifier components to store energy causes memory effects. For example, if the envelope waveform of the signal is at a high level for a brief time, solid state devices store thermal energy, i.e., they heat up, and this changes the device characteristics such as gain. The amplification of the next time-slice of signal will be affected by these characteristics. The same type of effect happens with voltages and currents. For example, the drain voltage power supply may sag due to a strong burst of signal and the voltage will have some non-zero recovery time. So, the following signal will see a different drain voltage depending upon the strength of the preceding signal. These effects have durations in the microsecond and millisecond range.

The issue for predistortion is that the gain and phase characteristics of the amplifier vary depending upon previous signals. Calculating corrections, when the amplifier characteristics are rapidly changing, presents a challenge. Collecting a set of samples and plotting amplifier output versus amplifier input, in the presence of memory effects, the samples do not line up perfectly along a curve. There is some scattering around the centerline of the curve.

A reasonably simple solution which is generally effective in amateur radio is to do a best fit of the available samples to gain and phase curves and use those curves to calculate correction. Much more advanced solutions are also available. These involve creating a predistortion solution based upon not only the input at the current time, but also based upon previous input values. To apply the solution, both the current input and previous inputs are considered. The Volterra series serves as an excellent mathematical model for such nonlinear time-varying systems. However, many other simplified models that are more practical to implement have been developed. (The Volterra series is used for modeling non-linear systems when including memory effects is important. See Kenington, P., *High-Linearity RF Amplifier Design*, (Artech House, 2000) pp.

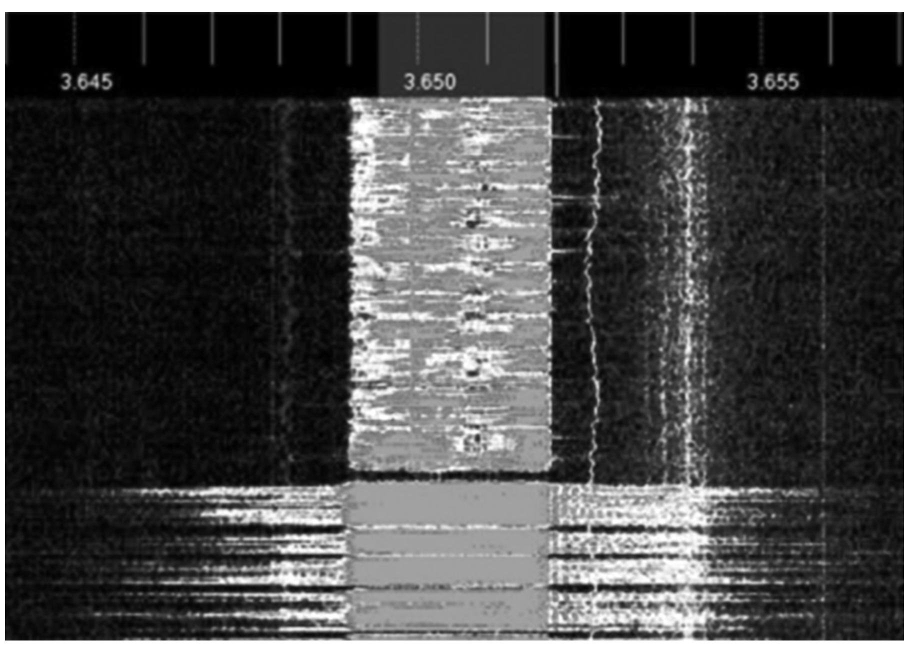


Figure 17.58 — Waterfall display showing two SSB signals without (bottom) and with (top) predistortion.

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17.12.8 On-the-Air Effects of Predistortion

The waterfall display in **Figure 17.58** shows a typical SSB signal (bottom) without predistortion followed by a very clean transmission using predistortion (top). The IMD produced without predistortion (at the bottom in the figure) creates a signal consuming about three times the bandwidth. Both stations were running legal-limit amplifiers.

17.13 References and Bibliography

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