Whitaker, Jerry C. "Chapter 4 – Designing Vacuum Tube Circuits"
Power Vacuum Tubes Handbook 2nd Edition.
Ed. Jerry C. Whitaker
Boca Raton: CRC Press LLC, 2000

Chapter

4

Designing Vacuum Tube Circuits

4.1 Introduction

Any number of configurations may be used to generate RF signals using vacuum tubes. Circuit design is dictated primarily by the operating frequency, output power, type of modulation, duty cycle, and available power supply. Tube circuits can be divided generally by their operating class and type of modulation employed. As discussed in Section 2.1.4, the angle of plate current flow determines the class of operation:

- Class A = 360° conduction angle
- Class B = 180° conduction angle
- Class C = conduction angle less than 180°
- Class AB = conduction angle between 180 and 360°

The class of operation has nothing to do with whether the amplifier is grid-driven or cathode-driven. A cathode-driven amplifier, for example, can be operated in any desired class. The class of operation is only a function of the plate current conduction angle. The efficiency of an amplifier is also a function of the plate current conduction angle.

The efficiency of conversion of dc to RF power is one of the more important characteristics of a vacuum tube amplifier circuit. The dc power that is not converted into useful output energy is, for the most part, converted to heat. This heat represents wasted power; the result of low efficiency is increased operating cost for energy. Low efficiency also compounds itself. This wasted power must be dissipated, requiring increased cooling capacity. The efficiency of the amplifier must, therefore, be carefully considered, consistent with the other requirements of the system. Figure 4.1 shows the theoretical efficiency attainable with a tuned or resistive load assuming that the peak ac plate voltage is equal to the plate supply voltage.

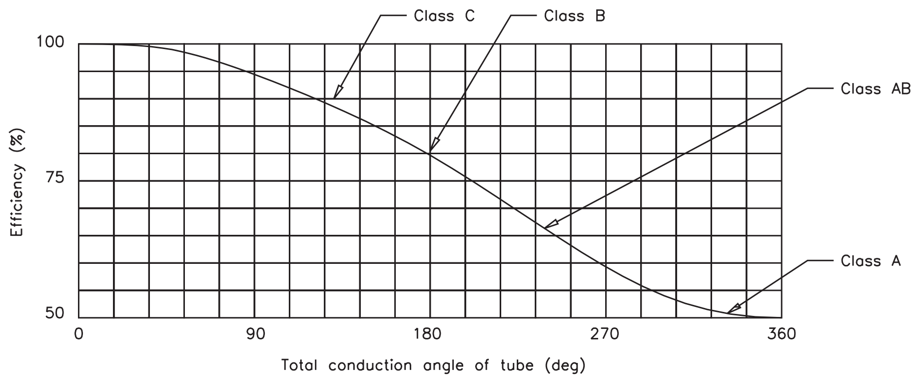


Figure 4.1 Plate efficiency as a function of conduction angle for an amplifier with a tuned load.

4.1.1 Class A Amplifier

A class A amplifier is used in applications requiring low harmonic distortion in the output signal. A class A amplifier can be operated with low intermodulation distortion in linear RF amplifier service. Typical plate efficiency for a class A amplifier is about 30 percent. Power gain is high because of the low drive power required. Gains as high as 30 dB are typical.

4.1.2 Class B and AB Amplifiers

A class AB power amplifier is capable of generating more power—using the same tube—than the class A amplifier, but more intermodulation distortion also will be generated. A class B RF linear amplifier will generate still more intermodulation distortion, but is acceptable in certain applications. The plate efficiency is typically 66 percent, and stage gain is about 20 to 25 dB.

4.1.3 Class C Amplifier

A class C power amplifier is used where large amounts of RF energy need to be generated with high efficiency. Class C RF amplifiers must be used in conjunction with tuned circuits or cavities, which restore the amplified waveform through the *flywheel effect*.

The grounded-cathode class C amplifier is the building block of RF technology. It is the simplest method of amplifying CW, pulsed, and FM signals. The basic configuration is shown in Figure 4.2. Tuned input and output circuits are used for impedance

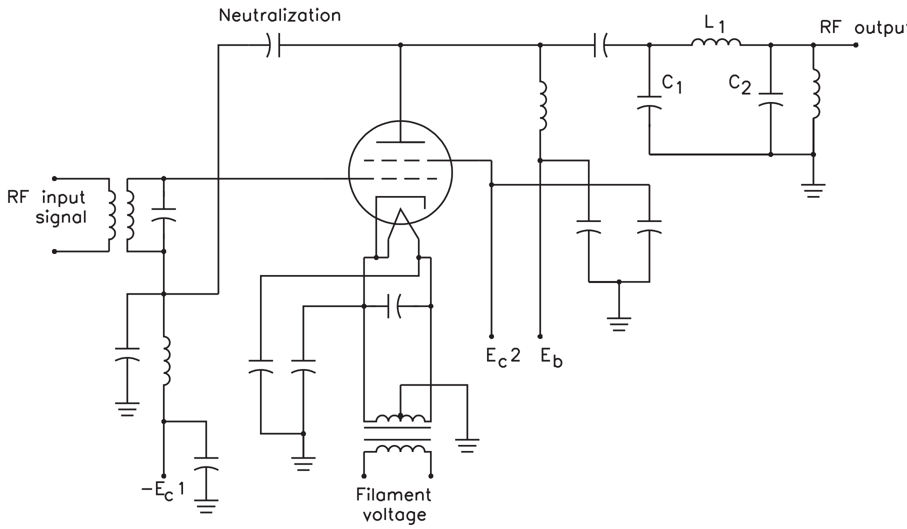


Figure 4.2 Basic grounded-cathode RF amplifier circuit.

matching and to resonate the stage at the desired operating frequency. The cathode is bypassed to ground using low-value capacitors. Bias is applied to the grid as shown. The bias power supply may be eliminated if a self-bias configuration is used. The typical operating efficiency of a class C stage ranges from 65 to 85 percent.

Figure 4.3 illustrates the application of a zero-bias triode in a grounded-grid arrangement. Because the grid operates at RF ground potential, this circuit offers stable performance without the need for neutralization (at MF and below). The input signal is coupled to the cathode through a matching network. The output of the triode feeds a *pi* network through a blocking capacitor.

4.2 Principles of RF Power Amplification

In an RF power amplifier, a varying voltage is applied to the control grid (or cathode, in the case of a grounded-grid circuit) from a driver stage whose output is usually one of the following:

- Carrier-frequency signal only
- Modulation (intelligence) signal only
- Modulated carrier signal

Simultaneous with the varying control grid signal, the plate voltage will vary in a similar manner, resulting from the action of the amplified current flowing in the plate circuit. In RF applications with resonant circuits, these voltage changes are smooth sine

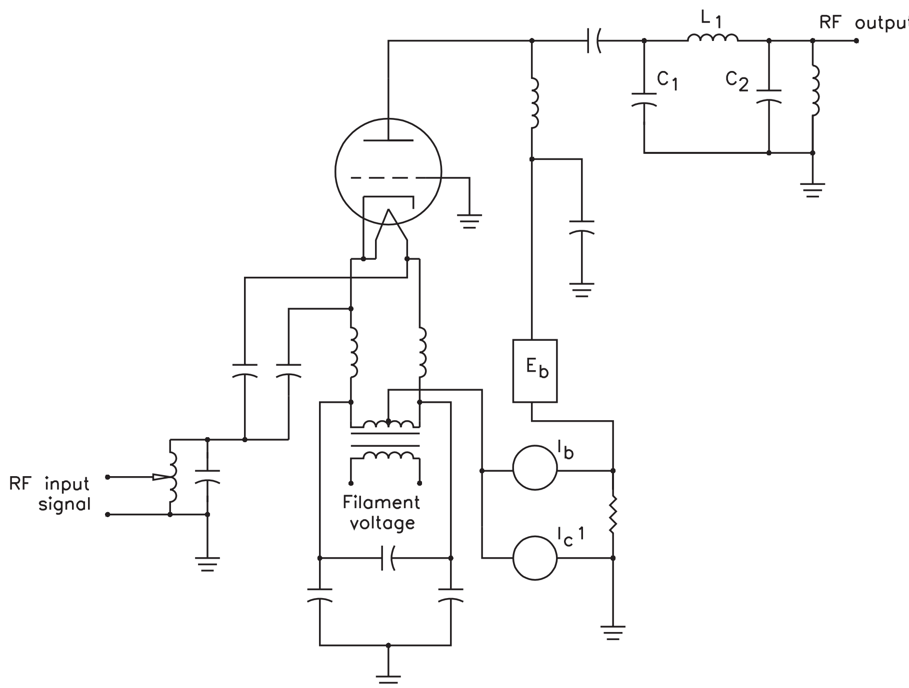


Figure 4.3 Typical amplifier circuit using a zero-bias triode. Grid current is measured in the return lead from ground to the filament.

wave variations, 180° out of phase with the input. The relationship is illustrated in Figure 4.4. Note how these variations center about the dc plate voltage and the dc control grid bias. In Figure 4.5, the variations have been indicated next to the plate voltage and grid voltage scales of a typical constant-current curve. At some instant in time, shown as t on the time scales, the grid voltage has a value denoted e_g on the grid voltage sine wave.

Any point on the operating line (when drawn on constant-current curves as illustrated in Figure 4.5) tells the instantaneous values of plate current, screen current, and grid current that must flow when these particular values of grid and plate voltage are applied to the tube. Thus, by plotting the values of plate and grid current as a function of time t, it is possible to produce a curve of instantaneous values of plate and grid current. Such plots are shown in Figure 4.6.

By analyzing the plate and grid current values, it is possible to predict with accuracy the effect on the plate circuit of a change at the grid. It follows that if a properly loaded resonant circuit is connected to the plate, a certain amount of RF power will be delivered to that circuit. If the resonant circuit is tuned to the fundamental frequency (the same frequency as the RF grid voltage), the power delivered will be that of the fundamental or principal RF component of plate current. If the circuit is tuned to a harmonic

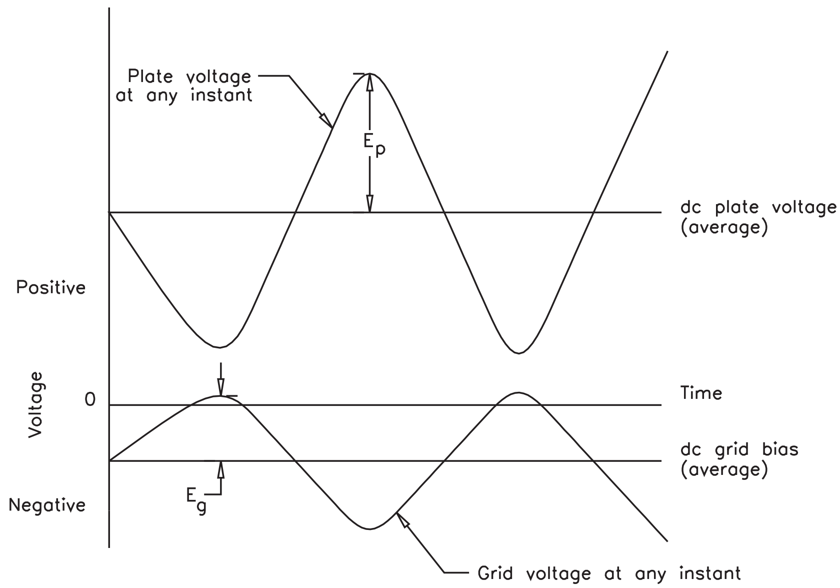


Figure 4.4 Variation of plate voltage as a function of grid voltage.

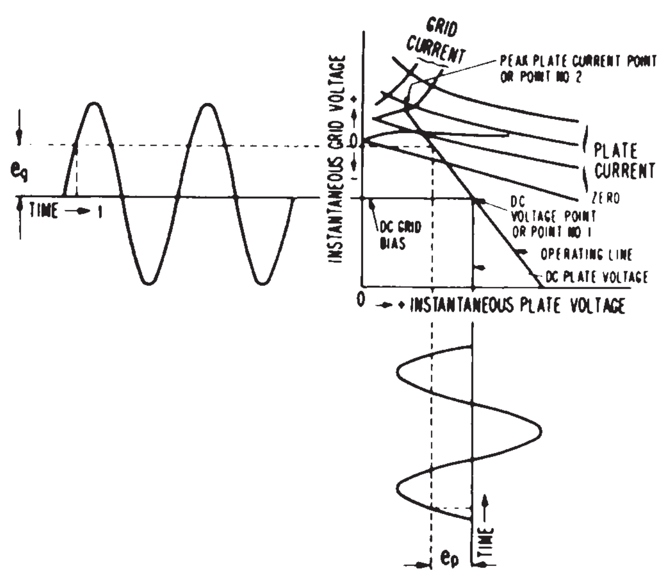


Figure 4.5 Relationship between grid voltage and plate voltage plotted on a constant-current curve.

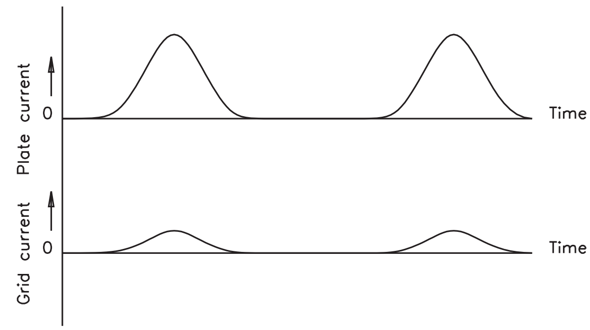


Figure 4.6 Instantaneous values of plate and grid current as a function of time.

of the grid voltage frequency, the power delivered will be the result of a harmonic component of the plate current.

4.2.1 Drive Power Requirements

The technical data sheet for a given tube type lists the approximate drive power required for various operating modes. As the frequency of operation increases and the physical size of the tube structure becomes large with respect to this frequency, the drive power requirements also will increase.

The drive power requirements of a typical grounded-cathode amplifier consist of six major elements:

- The power consumed by the bias source, given by:

$$P_1 = I_{c1} \times E_{c1} \tag{4.1}$$

- The power dissipated in the grid as a result of rectified grid current:

$$P_2 = I_{c1} \times e_{cmp} \tag{4.2}$$

- The power consumed in the tuned grid circuit:

$$P_3 = I_{c,rms}{}^2 \times R_{rf} \tag{4.3}$$

- The power lost as a result of transit-time effects:

$$P_4 = \left\{ \frac{e_{c,rms}}{R_t} \right\}^2 \tag{4.4}$$

R_t is that part of the resistive component of the tube input impedance resulting from transit-time effects, and is given by:

$$R_t = \frac{1}{Kg_m f^2 T^2} \tag{4.5}$$

- The power consumed in that part of the resistive component of the input impedance resulting from cathode lead inductance:

$$P_5 = \frac{e_g^{\,2}}{R_s} \tag{4.6}$$

Input resistance resulting from the inductance of the cathode leads is found from the following:

$$R_s = \frac{1}{\omega^2 g_m L_k C_{gk}} \tag{4.7}$$

- The power dissipated in the tube envelope because of dielectric loss:

$$P_6 = 1.41 f E_1^{\,2} \varepsilon \tag{4.8}$$

Where:
I_{c1} = dc grid current
E_{c1} = dc grid voltage
E_{cmp} = maximum positive grid voltage
$I_{c,rms}$ = rms value of RF grid current
R_{rf} = RF resistance of grid circuit
$e_{c,rms}$ = rms value of RF grid voltage
R_t = resistance resulting from transit-time loading
K = a constant (function of the tube geometry)
g_m = transconductance
f = frequency (Hz)
T = transit-time, cathode to grid
R_s = cathode lead inductance input resistance loading
$\omega = 2\pi f$
L_k = cathode lead inductance (Henrys)
C_{gk} = grid-to-cathode capacitance (Farads)
E_1 = voltage gradient (kilovolt per inch, rms)
ε = loss factor of dielectric materials

The total drive power P_t is, then, equal to:

$$P_t = P_1 + P_2 + P_3 + P_4 + P_5 + P_6 \tag{4.9}$$

Particular attention must be given to grid dissipation when a tube is operated in the VHF and UHF regions. The total driving power required for a given output may be greater than the grid dissipation capability of the device.

Operational Considerations for VHF and UHF

For operation of a tube in the VHF and UHF regions, several techniques may be applied to minimize the driving power without appreciably affecting plate conversion efficiency. The most common techniques are:

- Use the minimum dc control bias. Frequently, it is advisable to bring the bias down to approximately cutoff.
- Maintain a high value of dc screen voltage, even though it appears to increase the fraction of the cycle during which plate current flows.
- Use the minimum RF excitation voltage necessary to obtain the desired plate circuit performance, even though the dc grid current is considerably lower than would be expected at lower frequencies.
- Keep the cathode lead inductance to the output and input circuits as low as possible. This can be accomplished by 1) using short and wide straps, 2) using two separate return paths for the input and output circuits, or 3) properly choosing a cathode bypass capacitor(s).

These techniques do not necessarily decrease the plate efficiency significantly when the circuit is operated at VHF and UHF. The steps should be tried experimentally to determine whether the plate circuit efficiency is appreciably affected. It is usually acceptable, and even preferable, to sacrifice some plate efficiency for improved tube life when operating at VHF and UHF.

Optimum power output at these frequencies is obtained when the loading is greater than would be used at lower frequencies. Fortunately, the same condition reduces driving power and screen current (for the tetrode and pentode), and improves tube life expectancy in the process.

4.2.2 Mechanical and Electrical Considerations

To maintain proper isolation of the output and input circuits, careful consideration must be given to the location of the component parts of the amplifier. All elements of the grid or input circuit and any earlier stages must be kept out of the plate circuit compartment. Similarly, plate circuit elements must be kept out of the input compartment. Note, however, that in the case of the tetrode and pentode, the screen lead of the tube and connections via the socket are common to both the output and input resonant circuits. Because of the plate-to-screen capacitance of a tetrode or pentode, the RF plate voltage (developed in the output circuit) causes an RF current to flow out of the screen lead to the chassis. In the case of a push-pull stage, this current may flow from the screen terminal of one tube to the screen terminal of the other tube. Similarly, because of the grid-to-screen capacitance of the tube, the RF voltage in the input circuit will cause an RF current to flow in this same screen lead to the chassis, or to the opposite tube of the push-pull circuit.

The inductance of the lead common to both the output and input circuits has the desirable feature of providing voltage of opposite polarity to neutralize the feedback voltage of the residual plate to control grid capacitance of the tube. (The properties of neutralization are discussed in Section 3.4.) Note, however, that the mutual coupling from the screen lead to the input resonant circuit may be a possible source of trouble, depending on the design.

Lead Lengths

The interconnecting lead wires close to the tube should be designed with low inductance to minimize the generation of VHF parasitic oscillations. If two or more tubes are used in a given circuit, they should be placed reasonably close together to facilitate short interconnecting leads. The lead lengths of HF circuits usually can be much longer; the length depends, to a large extent, upon the frequency of the fundamental. All of the dc, keying, modulation, and control circuit wires can be relatively long if properly filtered and arranged physically apart from the active RF circuits. The following interconnecting lead wires in a tetrode or pentode power amplifier should have low inductance:

- Filament and screen bypass leads
- Suppressor bypass leads
- Leads from the grid and the plate to the tuning capacitor of the RF circuit (and return)
- Interconnections from tube to tube in push-pull or parallel arrangements (except for parasitic suppressors in the plate)

For a lead to have low inductance, it must have a large surface and be short in length, such as a strap or a ribbon. This consideration also applies to that portion of a lead inside a bypass capacitor.

Power Supply Considerations

The power supply requirements for a triode are straightforward. The degree of regulation and ripple depends upon the requirements of the system. In the case of a linear RF amplifier, it is important to have good plate power supply regulation. Without tight regulation, the plate voltage will drop during the time the plate is conducting current heavily. This drop will cause *flat topping* and will appear as distortion in the output. In push-pull applications where grid current flows, it is important to keep the grid circuit resistance to a minimum. If this is not done, positive peak clipping will occur.

In the case of the tetrode and pentode, the need for screen voltage introduces some new considerations and provides some new operating possibilities. Voltage for the screen grid of a low-power tetrode or pentode can readily be taken from the power supply used for the plate of the tube. In this case, a series resistor or potential-dividing resistor is chosen so that, with the intended screen current flowing, the voltage drop through the resistor is adequate to give the desired screen voltage. The potential-dividing resistor is the preferred technique for those tubes with significant secondary screen emission.

For high-power tubes, screen voltage commonly is taken from a separate power supply. A combination scheme also may be employed, where a dropping resistor is used in conjunction with a low-voltage or intermediate-voltage supply. Often, a combination of series resistor and voltage source can be chosen so that the rated screen dissipation will not be exceeded, regardless of variations in screen current. With a fixed screen supply, there are advantages in using an appreciable amount of fixed grid bias so as to provide protection against the loss of excitation, or for cases where the driver stage is being keyed.

If the screen voltage is taken through a dropping resistor from the plate supply, there is usually little point in using a fixed grid bias because an unreasonable amount of bias would be required to protect the tube if the excitation failed. Under operating conditions with normal screen voltage, the cutoff bias is low (screen voltage divided by the screen μ). When a stage loses excitation and runs statically, the screen current falls to nearly zero. (See the static curves of the tube in question.) If the screen voltage is obtained through a simple dropping resistor from the plate supply, the screen voltage will then rise close to full plate voltage. Because the cutoff bias required is proportional to screen voltage, the grid bias required will be much greater than the amount of bias desired under normal operating conditions. When a screen dropping resistor is used, most of the bias normally is supplied through a grid resistor, and other means are used for tube protection.

The power output from a tetrode or pentode is sensitive to screen voltage. For this reason, any application requiring a high degree of linearity through the amplifier requires a well-regulated screen power supply. A screen dropping resistor from the plate supply is not recommended in such cases.

The suppressor grid power supply requirements are similar to those of the control grid power supply. The suppressor grid intercepts little current, so a low-power supply may be used. Any variation in suppressor voltage as a result of ripple or lack of regulation will appear at the output of the amplifier because of suppressor grid modulation of the plate current.

4.2.3 Bypassing Tube Elements

Operation at high frequencies requires attention to bypassing of the tube elements. Areas of concern include:

- Filament circuit
- Screen grid circuit
- Suppressor grid circuit

Filament Bypassing

Low-inductance bypass capacitors should be used at the filament in an RF amplifier. Good engineering practice calls for placement of the capacitor directly between the filament socket terminals. If the circuit design allows it, strap one filament directly to the chassis; if not, use a second bypass capacitor from one terminal to the chassis.

If two or more tubes are used in a push-pull or parallel circuit, a short strap interconnecting one of the filament terminals of each socket can be used. The midpoint of the interconnecting strap is then bypassed or grounded directly.

For tubes equipped with an isolating screen cone terminal, the general circuit arrangement is usually different. The filament or cathode should go directly, or through bypass capacitors, to the cavity wall or chassis to which the screen terminal is bypassed.

Screen and Suppressor Grid Bypassing

Low-inductance leads are generally advisable for screen and suppressor grid terminal connections. For all frequencies, good engineering practice calls for routing of the screen bypass capacitors directly from the screen to one filament terminal. The suppressor grid is bypassed in the same manner when the suppressor is operated with a potential other than cathode potential. With the suppressor operating at cathode potential, the suppressor should be grounded to the chassis directly in a circuit where the cathode is at chassis potential. This applies to tubes in push-pull as well as tubes in single-ended stages.

In the VHF region, the connection to the screen terminals—for those tubes with two screen pins—should be made to the midpoint of a strap placed between the two screen terminals of the socket. This arrangement provides for equal division of the RF currents in the screen leads and minimizes heating effects.

At operating frequencies above the self-neutralizing frequency of the tetrode or pentode being used, the screen bypass capacitors are sometimes variable. By proper adjustment of this variable capacitor, the amount and phase of the screen RF voltage can be made to cancel the effects of the feedback capacitance within the tube. Thus, neutralization is accomplished. The screen lead inductance and the variable capacitor are not series resonant. The variable capacitor is adjusted so that a net inductive reactance remains to provide the proper voltage and phase for neutralization.

The preceding guidelines apply directly to tubes having the screen and suppressor grids mounted on internal supporting lead rods. Tube types having isolating screen cone terminals tend to work best when the screen or suppressor bypass capacitor is a flat sandwich type built directly onto the peripheral screen-contacting collet of the socket. The size of the bypass capacitor is a function of the operating frequency. The dielectric material may be of Teflon, mica, isomica, or a similar material. (Teflon is a trademark of DuPont.)

Application Examples

The use of bypassing capacitors for a grounded-cathode RF amplifier is illustrated in Figure 4.7. Bypassing for a grounded-grid amplifier is shown in Figure 4.8. The con-

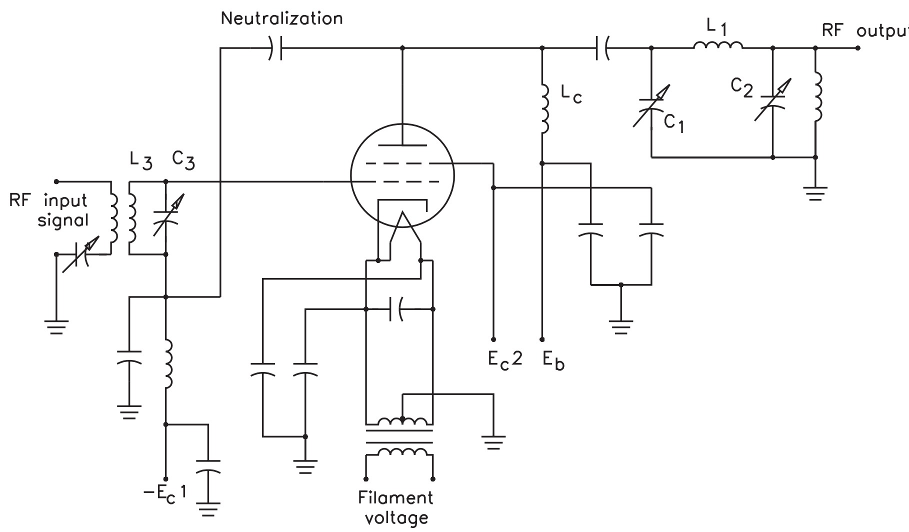


Figure 4.7 Schematic of a tetrode RF power amplifier showing tube element bypass capacitors.

trol grid may be operated at dc ground potential, shown in Figure 4.8*a*, or bypassed directly at the socket, shown in Figure 4.8*b*. For the sake of simplicity, operating the grid at dc ground potential is preferred. Functionally, these two circuits are identical. They vary only in the method used to measure grid current. In Figure 4.8*a*, grid current is measured in the return lead from ground to filament. In Figure 4.8*b*, the grid is raised 1 Ω above dc ground to allow grid current to be measured.

4.2.4 Parasitic Oscillations

Most self-oscillations in RF power amplifiers using gridded tubes have been found to fall within the following three classes:

- Oscillation at VHF from about 40 to 200 MHz, regardless of the operating frequency of the amplifier
- Self-oscillation on the fundamental frequency of the amplifier
- Oscillation at a low radio frequency below the normal frequency of the amplifier

Low-frequency oscillation in an amplifier usually involves the RF chokes, especially when chokes are used in both the output and input circuits. Oscillation near the fundamental frequency involves the normal resonant circuits and brings up the question of neutralization. (Neutralization is discussed in Section 3.4.) When a parasitic self-os-

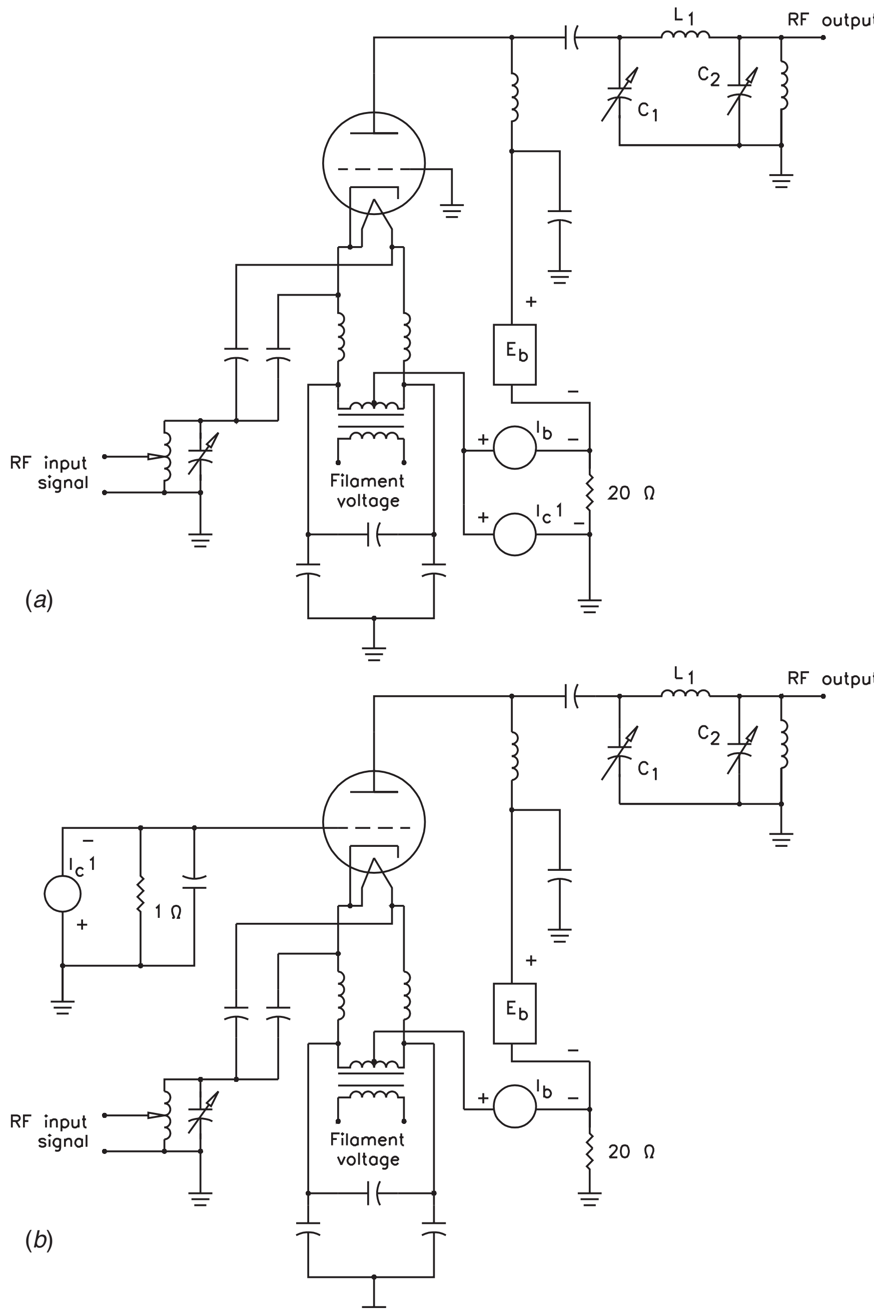


Figure 4.8 Schematic of a triode RF power amplifier showing tube element bypass capacitors: (*a*) control grid at dc ground potential, (*b*) control grid bypassed to ground.

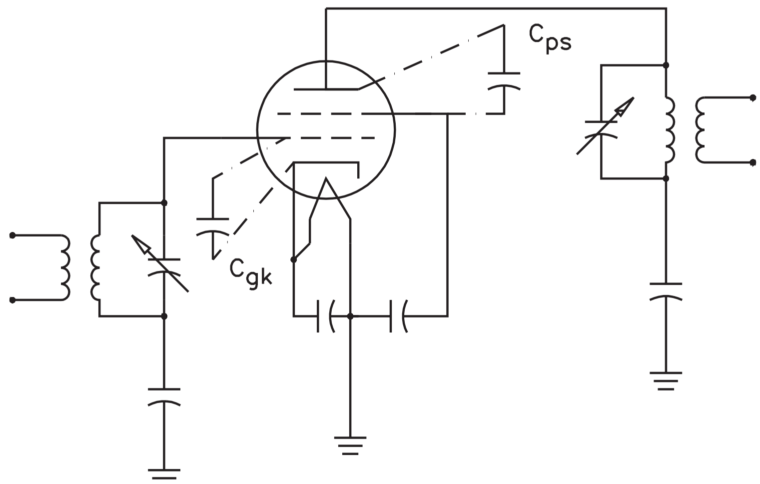


Figure 4.9 Interelectrode capacitances supporting VHF parasitic oscillation in a high-frequency RF amplfier.

cillation is found on a very high frequency, the interconnecting leads of the tube, tuning capacitor, and bypass capacitors typically are involved.

VHF oscillation occurs commonly in amplifiers where the RF circuits consist of coils and capacitors, as opposed to cavity designs. As illustrated in Figure 4.9, the tube capacitances effectively form a tuned-plate, tuned-grid oscillator.

The frequency of a VHF parasitic typically is well above the self-neutralizing frequency of the tube. However, if the self-neutralizing frequency of the device can be increased and the frequency of the parasitic lowered, complete suppression of the parasitic may result, or its suppression by resistor-inductor parasitic suppressors may be made easier.

It is possible to predict, with the use of a grid dip wavemeter, the parasitic frequency to be expected in a given circuit. The circuit should be complete and have no voltages applied to the tube. Couple the meter to the plate or screen lead, and determine the resonant frequency.

Elimination of the VHF parasitic oscillation may be accomplished using one of the following techniques:

- Place a small coil and resistor combination in the plate lead between the plate of the tube and the tank circuit, as shown in Figure 4.10*a*. The resistor-coil combination usually is made up of a noninductive resistor of about 25 to 100 Ω, shunted by three or four turns of approximately 1/2 in diameter and frequently wound around the resistor. In some cases, it may be necessary to use such a suppressor in both the plate and grid leads. The resistor-coil combination operates on the principle that the resistor loads the VHF circuit but is shunted by the coil for the lower fundamental frequency. In the process of adjusting the resistor-coil combination, it is

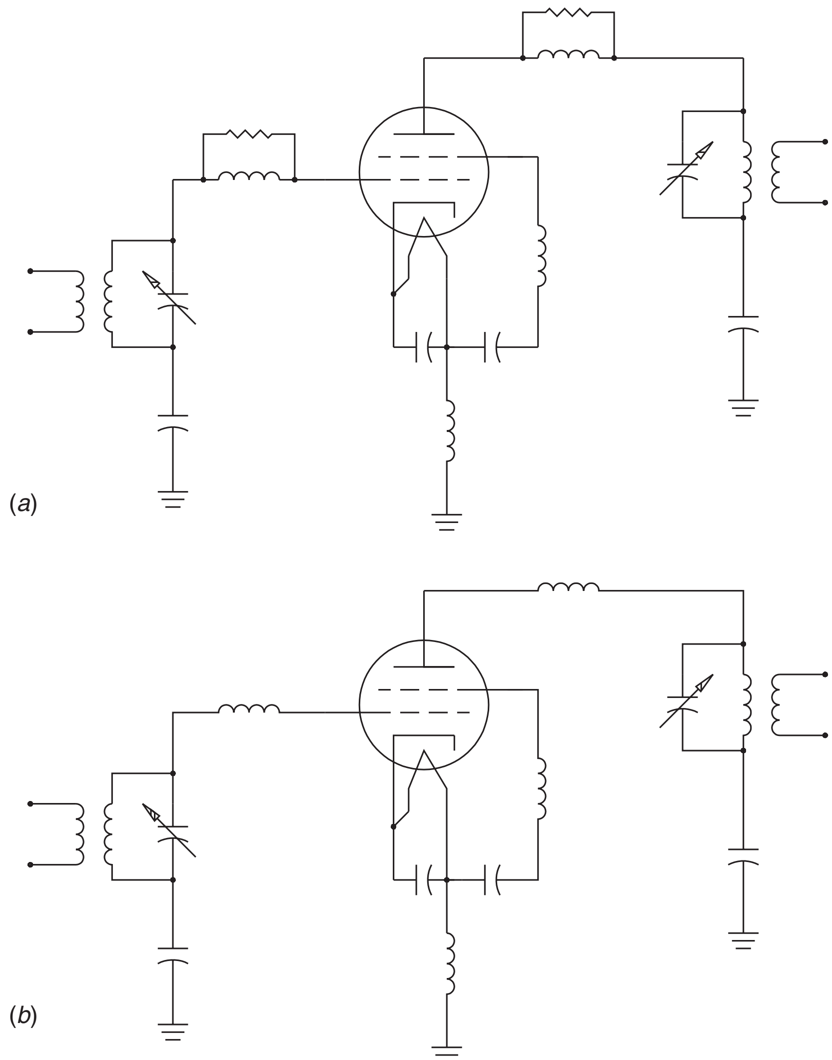


Figure 4.10 Placement of parasitic suppressors to eliminate VHF parasitic oscillations in a high-frequency RF amplifier: (*a*) resistor-coil combination, (*b*) parasitic choke.

often found that the resistor runs hot. This heat usually is caused by the dissipation of fundamental power in the resistor, and it is an indication of too many turns in the suppressor coil. Use only enough turns to suppress the parasitic, and no more. Once the parasitic has been suppressed, there will be no parasitic voltage or current present. Therefore, there is no parasitic power to be dissipated.

- Use small parasitic chokes in the plate lead, as shown in Figure 4.10*b*. The size of the coil will vary considerably depending upon the tube and circuit layout. A coil of four to ten turns with a diameter of approximately 1/2 in is typical. The presence of the choke in the frequency-determining part of the circuit lowers the frequency of a possible VHF parasitic so that it falls near the self-neutralizing frequency of the tube and bypass leads. In addition to variation in the size of the suppressor choke, the amount of inductance common to the screen and filament in the filament grounding strap may be a factor. This parameter can be varied simultaneously with the suppressor choke.

Of the two methods outlined for suppressing VHF parasitic oscillations, the first is probably the simpler and is widely employed.

Dynatron Oscillation

Another form of commonly encountered self-oscillation is known as *dynatron oscillation*, caused when an electrode in a vacuum tube has *negative resistance*. At times, there may be more electrons leaving the screen grid than arriving. If the screen voltage is allowed to increase under these conditions, even more electrons will leave the grid. This phenomenon implies a negative resistance characteristic. If there is high alternating current impedance in the circuit from the screen grid through the screen grid power supply, and from the plate power supply to the plate, dynatron oscillation may be sustained.

Dynatron oscillation typically occurs in the region of 1 to 20 Hz. This low-frequency oscillation usually is accompanied by another oscillation in the 1 to 2 kHz region. Suppression of these oscillations can be accomplished by placing a large bypass capacitor (1000 μF) across the output of the screen grid power supply. The circuit supporting the oscillation also can be detuned by a large inductor. Increasing the circuit losses at the frequency of oscillation is also effective.

Harmonic Energy

It is generally not appreciated that the pulse of grid current contains energy on harmonic frequencies and that control of these harmonic energies may be important. The ability of the tetrode and pentode to isolate the output circuit from the input over a wide range of frequencies is significant in avoiding feedthrough of harmonic voltages from the grid circuit. Properly designed tetrode and pentode amplifiers provide for complete shielding in the amplifier layout so that coupling external to the tube is prevented.

In RF amplifiers operating either on the fundamental or on a desired harmonic frequency, the control of unwanted harmonics is important. The following steps permit reduction of the unwanted harmonic energies present in the output circuit:

- Keep the circuit impedance between the plate and cathode low for the high harmonic frequencies. This requirement may be achieved by having some or all of the tuning capacitance of the resonant circuit close to the tube.

- Completely shield the input and output compartments.
- Use inductive output coupling from the resonant plate circuit and possibly a capacitive or Faraday shield between the coupling coil and the tank coil, or a high-frequency attenuating circuit such as a *pi* or *pi-L* network.
- Use low-pass filters on all supply leads and wires coming into the output and input compartments.
- Use resonant traps for particular frequencies.
- Use a low-pass filter in series with the output transmission line.

4.2.5 Shielding

In an RF amplifier, shielding between the input and output circuits must be considered. Triode amplifiers are more tolerant of poor shielding because power gain is relatively low. If the circuit layout is reasonable and no inductive coupling is allowed to exist, a triode amplifier usually can be built without extensive shielding. Even if shielding is not necessary to prevent fundamental-frequency oscillation, it will aid in eliminating any tendency toward parasitic oscillation. The higher the gain of an amplifier, the more important the shielding.

Pierced Shields

Tetrode and pentode power amplifiers require comprehensive shielding to prevent input-to-output circuit coupling. It is advisable to use nonmagnetic materials such as copper, aluminum, or brass in the RF fields to provide the shielding. Quite often, a shield must have holes through it to allow the passage of cooling air. In the LF and part of the HF range, the presence of small holes will not impair shield effectiveness. As the frequency is increased, however, the RF currents flowing around the hole in one compartment cause fields to pass through the hole into another compartment. Currents, therefore, are induced on the shield in the second compartment. This problem can be eliminated by using holes that have significant length. A piece of pipe with a favorable length-to-diameter ratio as compared to the frequency of operation will act as a *waveguide-beyond-cutoff* attenuator.

If more than one hole is required to pass air, a material resembling a honeycomb may be used. This material is available commercially and provides excellent isolation with a minimum of air pressure drop. A section of honeycomb shielding is shown in Figure 4.11. Some tube sockets incorporate a waveguide-beyond-cutoff air path. These sockets allow the tube in the amplifier to operate at high gain and up through VHF.

Metal Base Shells and Submounted Sockets

Some tetrodes and pentodes are supplied with metal base shells. The shell is typically grounded by the clips provided with the socket. This completes the shielding between

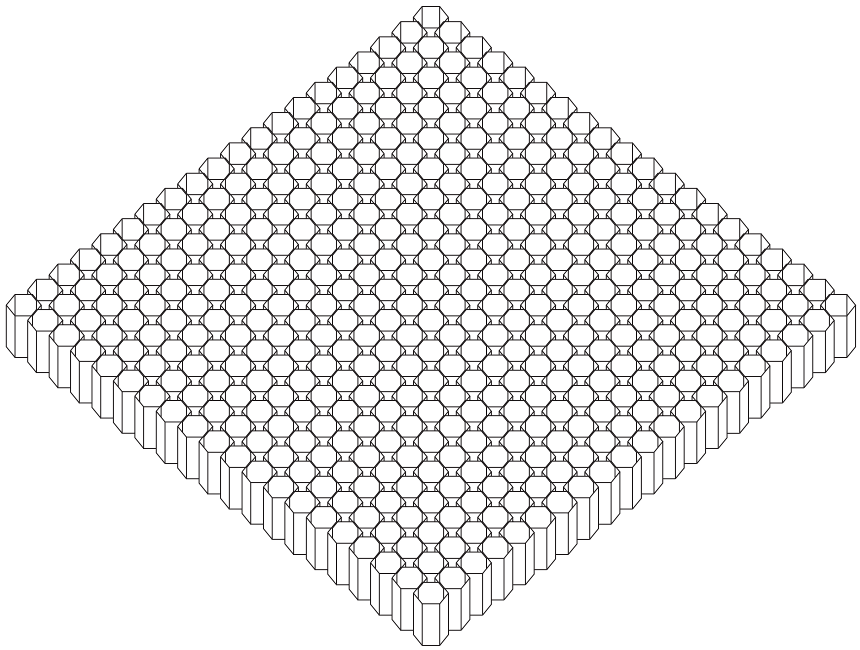


Figure 4.11 A section of honeycomb shielding material used in an RF amplifier.

the output and input circuits because the base shell of the tube comes up opposite the screen shield within the tube itself.

Some pentodes use this metal base shell as the terminal for the suppressor grid. If the suppressor is to be at some potential other than ground, then the base shell must not be dc grounded. The base shell is bypassed to ground for RF and insulated from ground for dc.

There is a family of tetrodes and pentodes without the metal base shell. For this type of tube structure, it is good practice to submount the socket so that the internal screen shield is at the same level as the chassis deck. This technique will improve the input-to-output circuit shielding. In submounting a tube, it is important that adequate clearance be provided around the base of the device for the passage of cooling air.

Compartments

By placing the tube and related circuits in completely enclosed compartments and properly filtering incoming supply wires, it is possible to prevent coupling of RF energy out of the circuit by means other than the desired output coupling. Such filtering prevents the coupling of energy that may be radiated or fed back to the input section or to earlier stages in the amplifier chain. Energy fed back to the input circuit causes undesirable interaction in tuning and/or self-oscillation. If energy is fed back to ear-

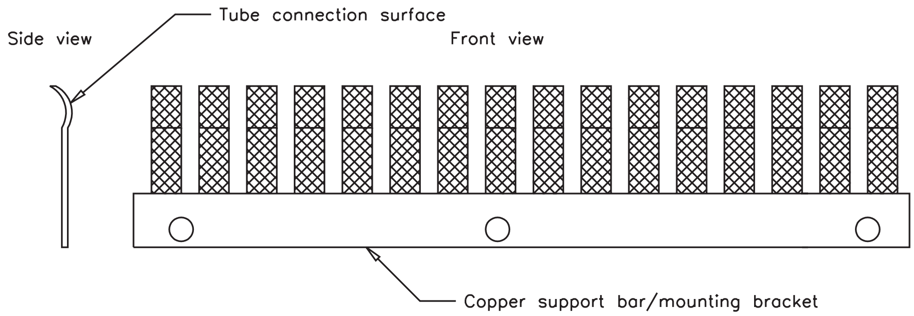


Figure 4.12 A selection of common fingerstock.

lier stages, significant operational problems may result because of the large power gain over several stages.

In the design of an RF amplifier, doors or removable panels typically must be used. The requirement for making a good, low-resistance joint at the discontinuity must be met. There are several materials available commercially for this purpose. Fingerstock, shown in Figure 4.12, has been used for many years. *Teknit*[1] is also a practical solution. Sometimes, after the wiring has been completed, further shielding of a particular conductor is required. Various types of shielding tapes[2] can be wound on a conductor as a temporary or permanent solution.

4.2.6 Protection Measures

Power grid tubes are designed to withstand considerable abuse. The maximum ratings for most devices are conservative. For example, the excess anode dissipation resulting from detuning the plate circuit will have no ill effects on most tubes if it is not applied long enough to overheat the envelope and the seal structure.

Similarly, the control, screen, and suppressor grids will stand some excess dissipation. Typically, the maximum dissipation for each grid indicated on the data sheet should not be exceeded except for time intervals of less than 1 s. The maximum dissipation rating for each grid structure is usually considerably above typical values used for maximum output so that ample operating reserve is provided. The time of duration of overload on a grid structure is necessarily short because of the small heat-storage capacity of the grid wires. Furthermore, grid temperatures cannot be measured or seen, so no warning of accidental overload is apparent.

---

1 Technical Wire Products, Inc., Cranford, NJ
2 Perfection Mica Co., Magnetic Shield Division, Chicago, IL

Table 4.1 Protection Guidelines for Tetrode and Pentode Devices

Circuit Failure Type	Fixed Screen Supply		Screen Voltage Through Dropping Resistor	
	Fixed Grid Bias	Resistor Grid Bias	Fixed Grid Bias	Resistor Grid Bias
Loss of excitation	No protection required	Plate current relay	Plate current relay	Plate current relay or screen control circuit
Loss of antenna loading	Screen current relay	Screen current relay	Grid current relay	No protection required
Excess antenna loading	Screen undercurrent relay	Screen undercurrent relay	Plate current relay	Plate current relay
Failure of plate supply	Screen current relay	Screen current relay	Grid current relay	No protection required
Failure of screen supply	Grid current relay	No protection required	Does not apply	Does not apply
Failure of grid bias supply	Plate current relay or screen current relay	Does not apply	Plate current relay or grid current relay	Does not apply

The type and degree of protection required in an RF amplifier against circuit failure varies with the type of screen and grid voltage supply. Table 4.1 lists protection criteria for tetrode and pentode devices. The table provides guidelines on the location of a suitable relay that should act to remove the principal supply voltage from the stage or transmitter to prevent damage to the tube.

For designs where screen voltage is taken through a dropping resistor from the plate supply, a plate relay provides almost universal protection. For a fixed screen supply, a relay provides protection in most cases. For protection against excessive antenna loading and subsequent high plate dissipation, a screen undercurrent relay also may be used in some services.

The plate, screen, and bias voltages may be applied simultaneously to a tetrode. The same holds true for a pentode, plus the application of the suppressor voltage. In a grid-driven amplifier, grid bias and excitation usually can be applied alone to the tube, especially if a grid leak resistor is used. Plate voltage can be applied to the tetrode and pentode before the screen voltage with or without excitation to the control grid. Never apply screen voltage before the plate voltage. The only exceptions are when the tube is cut off so that no space current (screen or plate current) will flow, or when the excitation and screen voltage are low. If screen voltage is applied before the plate voltage and screen current can flow, the maximum allowable screen dissipation will almost always be exceeded, and tube damage will result.

Table 4.2 lists protection guidelines for a triode. The table covers the grid-driven triode amplifier and the high-μ (zero-bias) cathode-driven triode amplifier. Drive voltage must never be applied to a zero-bias triode amplifier without plate voltage. The table indicates the recommended location of a suitable relay that should act to remove the prin-

Table 4.2 Protection Guidelines for a Triode

Circuit Failure Type	Triode		Zero-Bias Triode
	Fixed Grid Bias	Resistor Grid Bias	
Loss of excitation	No protection required	Plate overcurrent relay	No protection required
Loss of antenna loading	RF output detector and relay	RF output detector and relay	Grid overcurrent relay
Excess antenna loading	RF output detector and relay	RF output detector and relay	RF output detector and relay
Failure of plate supply	No protection required	No protection required	Grid overcurrent relay
Failure of grid bias supply	Plate overcurrent relay	Does not apply	Does not apply

cipal supply voltage from the stage or transmitter to prevent damage to the tube or transmitter.

4.3 Cavity Amplifier Systems

Power grid tubes are ideally suited for use as the power generating element in a cavity amplifier. Because of the physical dimensions involved, cavity designs typically are limited to VHF and above. Lower-frequency RF generators utilize discrete *L* and *C* devices to create the needed tank circuit. Two types of cavity amplifiers commonly are used: 1/4-wavelength and 1/2-wavelength systems.

In a cavity amplifier, the tube becomes part of a resonant transmission line. The stray interelectrode and distributed capacity and inductance of the tube are used to advantage as part of the resonant line. This resonant line is physically larger than the equivalent lumped constant *LRC* resonant circuit operating at the same frequency, and this larger physical size aids in solving the challenges of high-power operation, skin effect losses, and high-voltage-standoff concerns.

A shorted 1/4-wavelength transmission line has a high, purely resistive input impedance. Electrically, it appears as a parallel resonant circuit, as shown in Figure 4.13. When the physical length of the line is less than wavelength, the impedance will be lower and the line will appear inductive, as illustrated in Figure 4.14. This inductance is used to resonate with the capacitive reactance in the tube and the surrounding circuit.

4.3.1 Bandwidth and Efficiency

Power amplifier bandwidth has a significant effect on modulation performance. Available bandwidth determines the amplitude response, phase response, and group-delay response. Performance tradeoffs often must be considered in the design of a cavity amplifier, including bandwidth, gain, and efficiency.

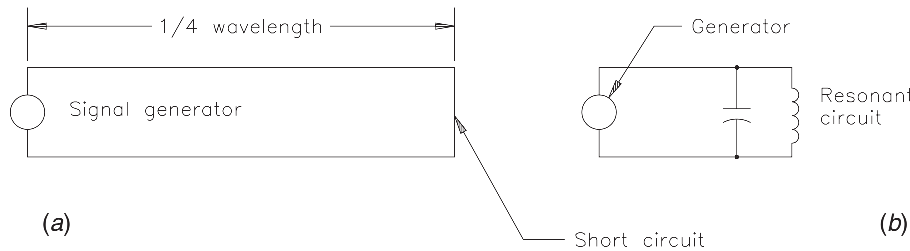


Figure 4.13 Shorted 1/4-wavelength line: (*a*) physical circuit, (*b*) electrical equivalent.

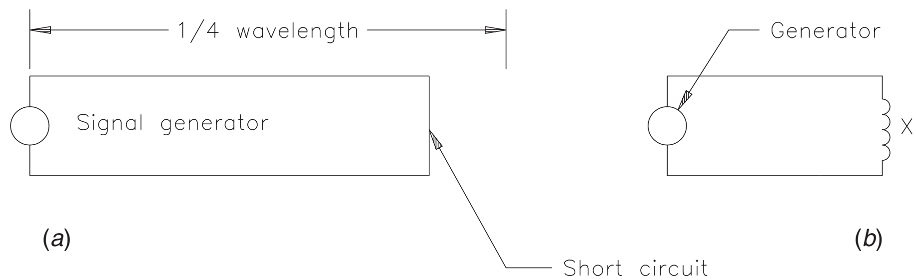


Figure 4.14 Shorted line less than 1/4-wavelength: (*a*) physical circuit, (*b*) electrical equivalent.

Power amplifier bandwidth is restricted by the equivalent load resistance across the parallel tuned circuits in the stage. Tuned circuits are necessary to cancel low reactive impedance presented by the relatively high input and output capacitances of the amplifying device. The bandwidth for a single tuned circuit is proportional to the ratio of capacitive reactance X_c to load resistance R_l appearing across the tuned circuit:

$$BW \approx \frac{K}{2\pi f_c C \; R_l} \approx \frac{K(X_c)}{R_l} \tag{4.10}$$

Where:
BW = bandwidth between half-power (- 3 dB) points
K = proportionality constant
R_l = load resistance appearing across tuned circuit
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
X_c = capacitive reactance of C
f_c = carrier frequency

The RF voltage swing across the tuned circuit also depends on the load resistance. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

The efficiency of a cavity PA depends primarily on the RF plate voltage swing, the plate current conduction angle, and the cavity efficiency. Cavity efficiency is related to the ratio of loaded to unloaded Q as follows:

$$N = 1 - \frac{Q_l}{Q_u} \times 100 \tag{4.11}$$

Where:
N = cavity efficiency in percent
Q_l = cavity loaded Q
Q_u = cavity unloaded Q

The loaded Q is determined by the plate load impedance and output circuit capacitance. Unloaded Q is determined by the cavity volume and the RF resistivity of the conductors resulting from the skin effect. (Skin effect is discussed in Chapter 7.) For best cavity efficiency, the following conditions are desirable:

- High unloaded Q
- Low loaded Q

As the loaded Q increases, the bandwidth decreases. For a given tube output capacitance and power level, loaded Q decreases with decreasing plate voltage or with increasing plate current. The increase in bandwidth at reduced plate voltage occurs because the load resistance is directly related to the RF voltage swing on the tube anode. For the same power and efficiency, the bandwidth also can be increased if the output capacitance is reduced. Power tube selection and minimization of stray capacitance are important considerations when designing for maximum bandwidth.

4.3.2 Current Paths

The operation of a cavity amplifier is an extension of the current paths inside the tube. Two elements must be examined in this discussion:

- The input circulating currents
- The output circulating currents

Input Circuit

The grid/cathode assembly resembles a transmission line whose termination is the RF resistance of the electron stream within the tube. Figure 4.15 shows the current path of an RF generator (the RF driver stage output) feeding a signal into the grid/cathode circuit.

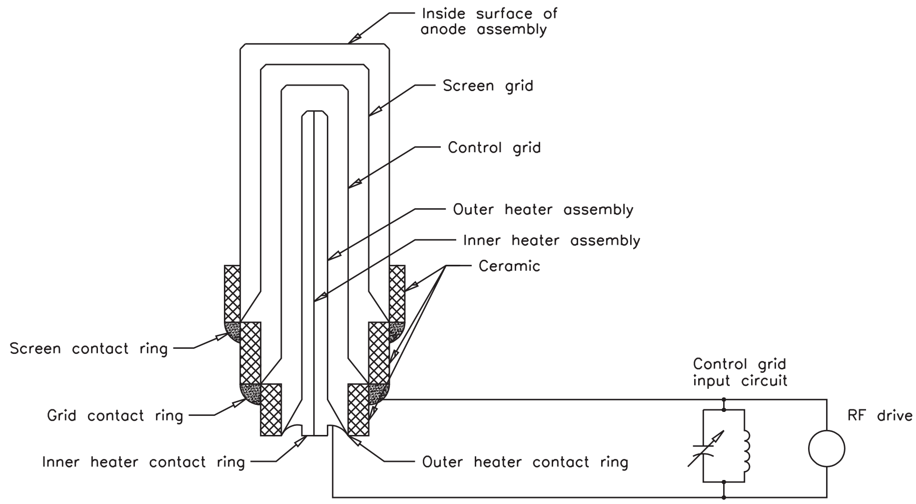


Figure 4.15 A simplified representation of the grid input circuit of a PA tube.

The outer contact ring of the cathode heater assembly makes up the inner conductor of a transmission line formed by the cathode and control grid assemblies. The filament wires are returned down the center of the cathode. For the input circuit to work correctly, the cathode must have a low RF impedance to ground. This cathode bypassing may be accomplished in one of several ways.

Below 30 MHz, the cathode can be grounded to RF voltages by simply bypassing the filament connections with capacitors, as shown in Figure 4.16*a*. Above 30 MHz, this technique does not work well because of the stray inductance of the filament leads. Notice that in (*b*), the filament leads appear as RF chokes, preventing the cathode from being placed at RF ground potential. This causes negative feedback and reduces the efficiency of the input and output circuits.

In Figure 4.16*c*, the cathode circuit is configured to simulate a 1/2-wave transmission line. The line is bypassed to ground with large-value capacitors 1/2 wavelength from the center of the filament (at the filament voltage feed point). This transmission line RF short circuit is repeated 1/2 wavelength away at the cathode (heater assembly), effectively placing it at ground potential.

Because 1/2-wavelength bypassing is usually bulky at VHF (and may be expensive), RF generators often are designed using certain values of inductance and capacitance in the filament/cathode circuit to create an artificial transmission line that will simulate a 1/2-wavelength shorted transmission line. As illustrated in the figure, the inductance and capacitance of the filament circuit can resemble an artificial transmission line of 1/2 wavelength, if the values of *L* and *C* are properly selected.

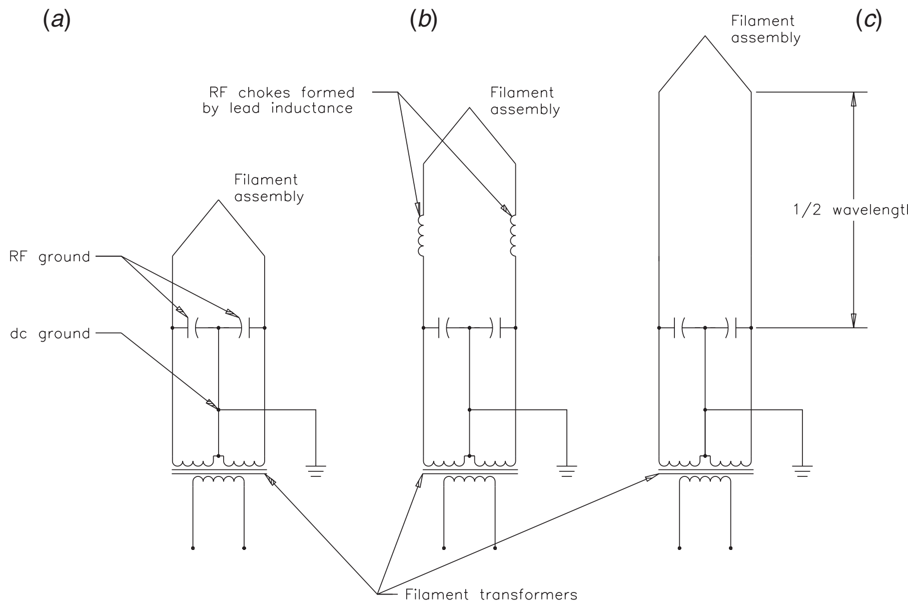


Figure 4.16 Three common methods for RF bypassing of the cathode of a tetrode PA tube: (*a*) grounding the cathode below 30 MHz, (*b*) grounding the cathode above 30 MHz, (*c*) grounding the cathode via a 1/2-wave transmission line.

Output Circuit

The plate-to-screen circulating current of the tetrode is shown in Figure 4.17. For the purposes of example, consider that the output RF current is generated by an imaginary current generator located between the plate and screen grid. The RF current travels along the inside surface of the plate structure (because of the skin effect), through the ceramic at the lower half of the anode contact ring, across the bottom of the fins, and to the band around the outside of the fins. The RF current then flows through the plate bypass capacitor to the RF tuned circuit and load, and returns to the screen grid.

The return current travels through the screen bypass capacitor (if used) and screen contact ring, up the screen base cone to the screen grid, and back to the imaginary generator.

The screen grid has RF current returning to it, but because of the assembly's low impedance, the screen grid is effectively at RF ground potential. The RF current generator, therefore, appears to be feeding an open-ended transmission line consisting of the anode (plate) assembly and the screen assembly. The RF voltage developed by the anode is determined by the plate impedance (Z_p) presented to the anode by the resonant circuit and its load.

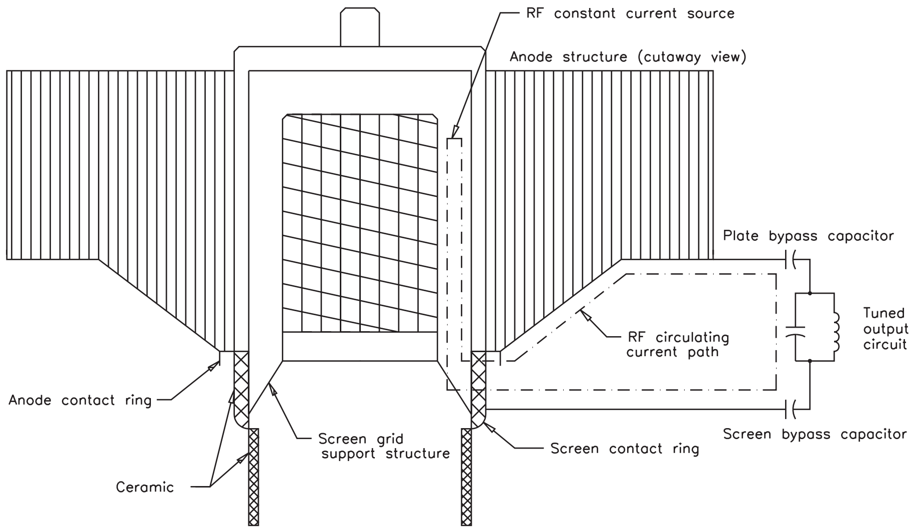


Figure 4.17 RF circulating current path between the plate and screen in a tetrode PA tube.

When current flows on one conductor of a transmission line cavity circuit, an equal-magnitude current flows in the opposite direction on the other conductor. This means that a large value of RF circulating current is flowing in the cavity amplifier outer conductor (the cavity box). All of the outer conductor circulating currents start at and return to the screen grid (in a tetrode-based system). The front or back access panel (door) of the cavity is part of the outer conductor, and large values of circulating current flow into it, through it, and out of it. A mesh contact strap generally is used to electrically connect the access panel to the rest of the cavity.

4.3.3 The 1/4-Wavelength Cavity

The 1/4-wavelength PA cavity is common in transmitting equipment. The design is simple and straightforward. A number of variations can be found in different RF generators, but the underlying theory of operation is the same.

A typical 1/4-wave cavity is shown in Figure 4.18. The plate of the tube connects directly to the inner section (tube) of the plate-blocking capacitor. The exhaust chimney/inner conductor forms the other plate of the blocking capacitor. The cavity walls form the outer conductor of the 1/4-wave transmission line circuit. The dc plate voltage is applied to the PA tube by a cable routed inside the exhaust chimney and inner tube conductor. In the design shown in the figure, the screen-contact fingerstock ring mounts on a metal plate that is insulated from the grounded cavity deck by an insulating material. This hardware makes up the screen-blocking capacitor assembly. The dc

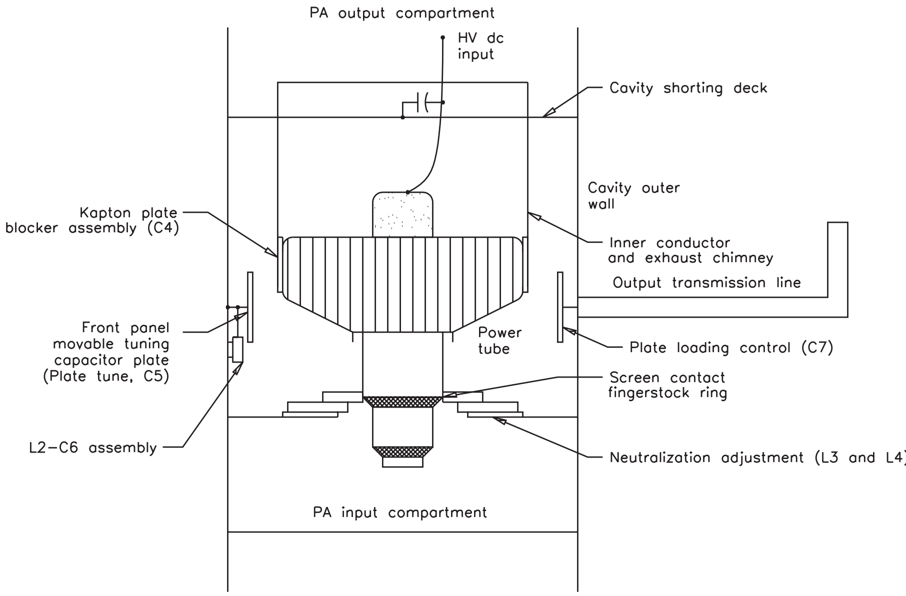


Figure 4.18 The physical layout of a common type of 1/4-wavelength PA cavity designed for VHF operation.

screen voltage feeds to the fingerstock ring from underneath the cavity deck through an insulated feedthrough assembly.

A grounded screen configuration also may be used in this design in which the screen-contact fingerstock ring is connected directly to the grounded cavity deck. The PA cathode then operates at below ground potential (in other words, at a negative voltage), establishing the required screen voltage to the tube.

The cavity design shown in the figure is set up to be slightly shorter than a full 1/4 wavelength at the operating frequency. This makes the load inductive and resonates the tube's output capacity. Thus, the physically foreshortened shorted transmission line is resonated and electrically lengthened to 1/4 wavelength.

Figure 4.19 illustrates the paths taken by the RF circulating currents in the circuit. RF energy flows from the plate, through the plate-blocking capacitor, along the inside surface of the chimney/inner conductor (because of the skin effect), across the top of the cavity, down the inside surface of the cavity box, across the cavity deck, through the screen-blocking capacitor, over the screen-contact fingerstock, and into the screen grid.

Figure 4.20 shows a graph of RF current, voltage, and impedance for a 1/4-wavelength coaxial transmission line. The plot shows that infinite impedance, zero RF current, and maximum RF voltage occur at the feed point. This would not be suitable for a

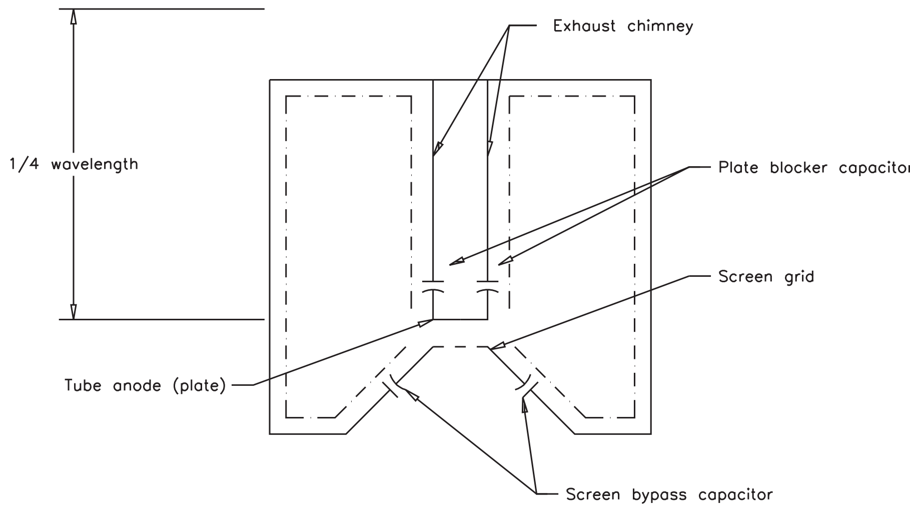


Figure 4.19 The RF circulating current paths for the 1/4-wavelength cavity shown in Figure 4.18.

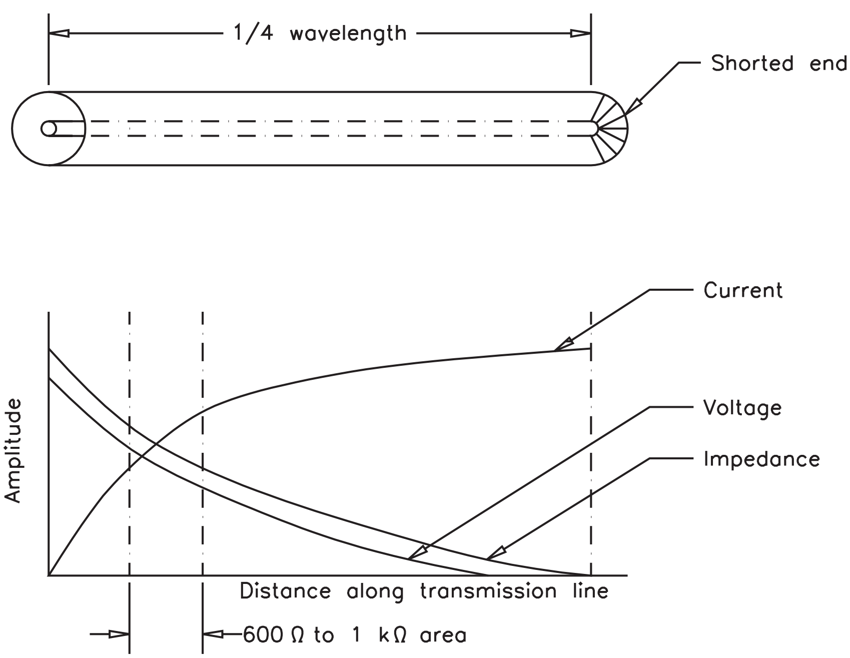


Figure 4.20 Graph of the RF current, RF voltage, and RF impedance for a 1/4-wavelength shorted transmission line. At the feed point, RF current is zero, RF voltage is maximum, and RF impedance is infinite.

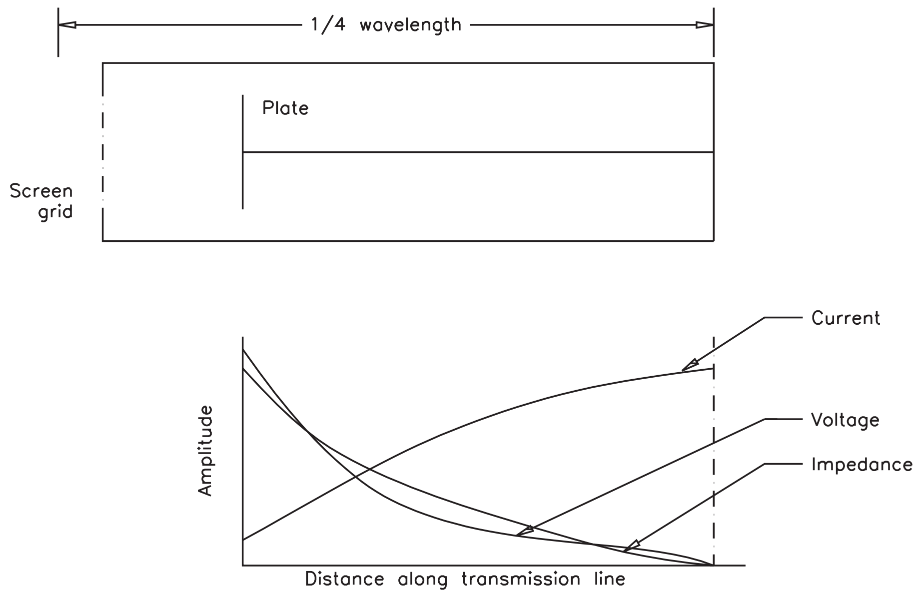


Figure 4.21 Graph of the RF current, RF voltage, and RF impedance produced by the physically foreshortened coaxial transmission line cavity.

practical PA circuit because arcing would result from the high RF voltage, and poor efficiency would be caused by the mismatch between the tube and the load.

Notice, however, the point on the graph marked at slightly less than 1/4 wavelength. This length yields an impedance of 600 to 800 Ω and is ideal for the PA plate circuit. It is necessary, therefore, to physically foreshorten the shorted coaxial transmission line cavity to provide the correct plate impedance. Shortening the line also is a requirement for resonating the tube's output capacity, because the capacity shunts the transmission line and electrically lengthens it.

Figure 4.21 shows a graph of the RF current, voltage, and impedance presented to the plate of the tube as a result of the physically foreshortened line. This plate impedance is now closer to the ideal 600 to 800 Ω value required by the tube anode.

Tuning the Cavity

Coarse-tuning of the cavity is accomplished by adjusting the cavity length. The top of the cavity (the cavity shorting deck) is fastened by screws or clamps and can be raised or lowered to set the length of the cavity for the particular operating frequency. Fine-tuning is accomplished by a variable-capacity plate-tuning control built into the cavity. In a typical design, one plate of the tuning capacitor—the stationary plate—is fastened to the inner conductor just above the plate-blocking capacitor. The movable

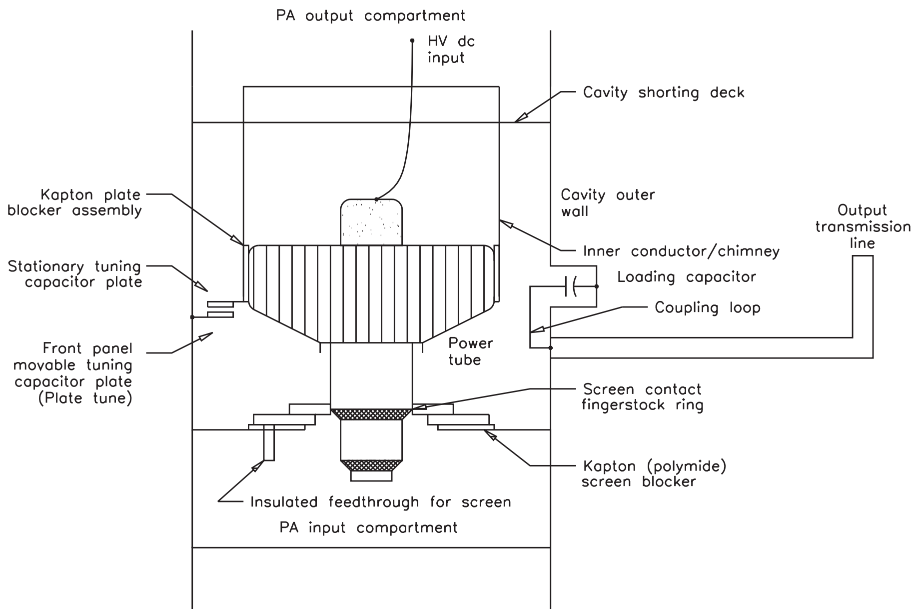


Figure 4.22 Basic loading mechanism for a VHF 1/4-wave cavity.

tuning plate is fastened to the cavity box—the outer conductor—and mechanically linked to the front-panel tuning control. This capacity shunts the inner conductor to the outer conductor and is used to vary the electrical length and resonant frequency of the cavity.

The 1/4-wavelength cavity is inductively coupled to the output port. This coupling is usually on the side opposite the cavity access door. The inductive pickup loop can take one of several forms. In one design it consists of a half-loop of flat copper bar stock that terminates in the loading capacitor at one end and feeds the output transmission line inner conductor at the other end. This configuration is shown in Figure 4.22. The inductive pickup ideally would be placed at the maximum current point in the 1/4-wavelength cavity. However, this point is located at the cavity shorting deck, and, when the deck is moved for coarse-tuning, the magnetic coupling will be changed. A compromise in positioning, therefore, must be made. The use of a broad, flat copper bar for the coupling loop adds some capacitive coupling to augment the reduced magnetic coupling.

Adjustment of the loading capacitor couples the 50 Ω transmission line impedance to the impedance of the cavity. Heavy loading lowers the plate impedance presented to the tube by the cavity. Light loading reflects a higher load impedance to the amplifier plate.

Methods commonly used to increase the operating bandwidth of the cavity include minimizing added tuning capacitance. The ideal case would be to resonate the plate ca-

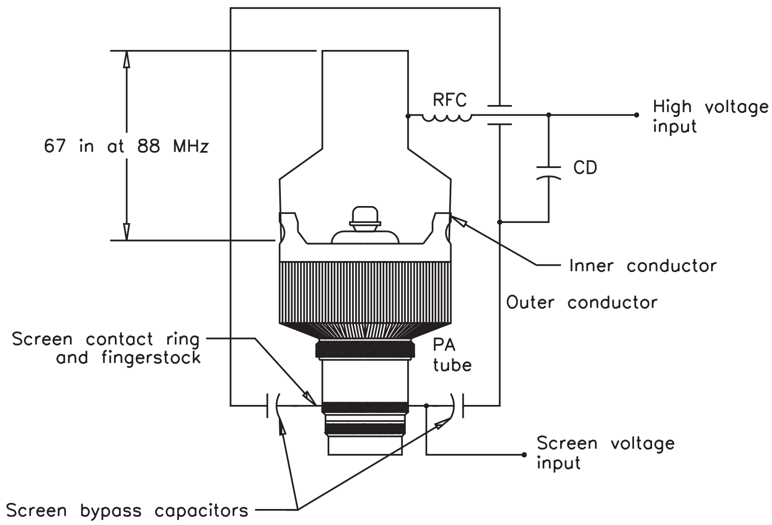


Figure 4.23 The 1/2-wavelength PA cavity in its basic form.

pacitance alone with a "perfect" inductor. Practical cavities, however, require either the addition of a variable capacitor or a variable inductor using sliding contacts for tuning. Other inherent mechanical and electrical compromises include:

- The requirement for a plate dc-blocking capacitor.
- The presence of maximum RF current at the grounded end of the simulated transmission line where the conductor may be nonhomogeneous. This can result in accompanying losses in the contact resistance.

4.3.4 The 1/2-Wavelength Cavity

The operation of a 1/2-wavelength cavity follows the same basic idea as the 1/4-wavelength design outlined previously. The actual construction of the system depends upon the power level used and the required bandwidth. An example will help to illustrate how a 1/2-wavelength cavity operates.

The design of a basic 1/2-wavelength PA cavity for operation in the region of 100 MHz is shown in Figure 4.23. The tube anode and a silver-plated brass pipe serve as the inner conductor of the 1/2-wave transmission line. The cavity box serves as the outer conductor. The transmission line is open at the far end and repeats this condition at the plate of the tube. The line is, in effect, a parallel resonant circuit for the PA tube.

The physical height of the circuit shown (67 in) was calculated for operation down to 88 MHz. To allow adequate clearance at the top of the transmission line and space for

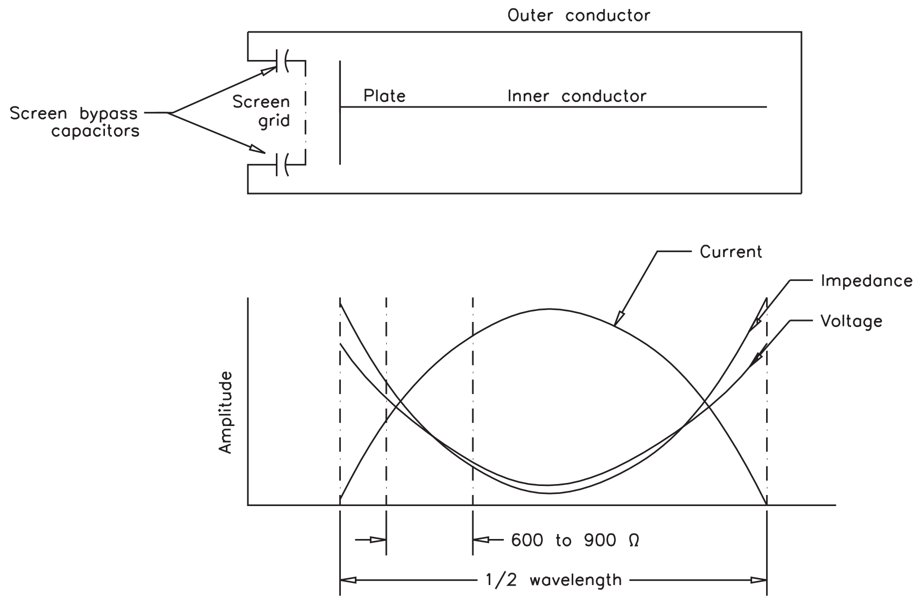


Figure 4.24 The distribution of RF current, voltage, and impedance along the inner conductor of a 1/2-wavelength cavity.

input circuitry at the bottom of the assembly, the complete cavity box would have to be almost 8 ft tall. This is too large for any practical transmitter.

Figure 4.24 shows RF voltage, current, and impedance for the inner conductor of the transmission line and the anode of the tube. The load impedance at the plate is thousands of ohms. Therefore, the RF current is extremely small, and the RF voltage is extremely large. In the application of such a circuit, arcing would become a problem. The high plate impedance also would make amplifier operation inefficient.

The figure also shows an area between the anode and the 1/4-wavelength location where the impedance of the circuit is 600 to 800 Ω. As noted previously, this value is ideal for the anode of the PA tube. To achieve this plate impedance, the inner conductor must be less than a full 1/2 wavelength. The physically foreshortened transmission-line circuit must, however, be electrically resonated (lengthened) to 1/2 wavelength for proper operation.

If the line length were changed to operate at a different frequency, the plate impedance also would change because of the new distribution of RF voltage and current on the new length of line. The problem of frequency change, therefore, is twofold: The length of the line must be adjusted for resonance, and the plate impedance of the tube must be kept constant for good efficiency.

To accommodate operation of this system at different frequencies, while keeping the plate impedance constant, two forms of coarse-tuning and one form of fine-tuning are built into the 1/2-wave PA cavity. Figure 4.25 shows the tube and its plate line (inner

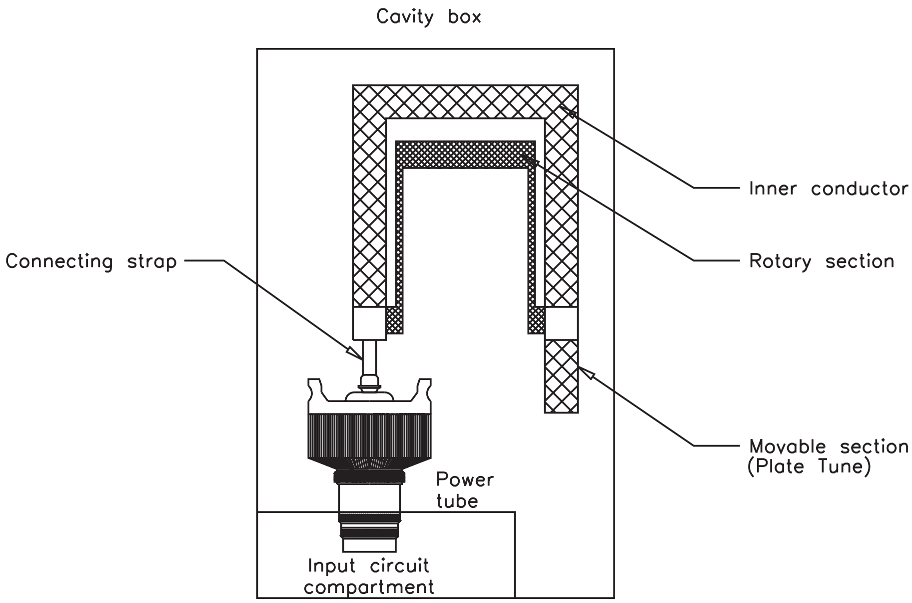


Figure 4.25 The configuration of a practical 1/2-wavelength PA cavity.

conductor). The inner conductor is U-shaped to reduce the cavity height. With the movable section (the plate tune control) fully extended, the inner conductor measures 38 in, and the anode strap measures 7 in. The RF path from the anode strap to the inside of the tube plate (along the surface because of the skin effect) is estimated to be about 8 in. This makes the inner conductor maximum length about 53 in. This is too short to be a 1/2 wavelength at 88 MHz, the target low-end operating frequency in this example. The full length of a 1/2-wave line is 67.1 in at 88 MHz.

The coarse-tuning and fine-tuning provisions of the cavity, coupled with the PA tube's output capacity, resonate the plate line to the exact operating frequency. In effect, they electrically lengthen the physically foreshortened line. This process, along with proper loading, determines the plate impedance and, therefore, the efficiency.

Lengthening the Plate Line

The output capacity of the tube is the first element that electrically lengthens the plate line. A 1/2-wave transmission line that is too short offers a high impedance that is both resistive and inductive. The output capacity of the tube resonates this inductance. The detrimental effects of the tube's output capacity, therefore, are eliminated. The anode strap and the cavity inner conductor rotary section provide two methods of coarse frequency adjustment for resonance.

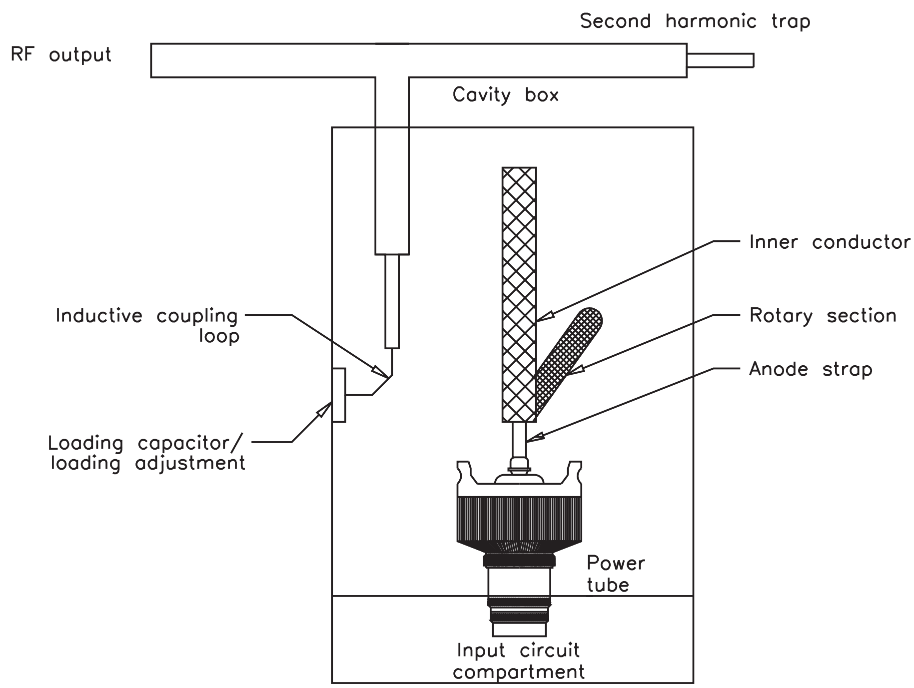


Figure 4.26 Coarse-tuning mechanisms for a 1/2-wave cavity.

The anode strap, shown in Figure 4.26, has a smaller cross-sectional area than the inner conductor of the transmission line. Therefore, it has more inductance than an equal length of the inner conductor. The anode-coupling strap acts as a series inductance and electrically lengthens the plate circuit.

At low frequencies, one narrow anode strap is used. At midband frequencies, one wide strap is used. The wide strap exhibits less inductance than the narrow strap and does not electrically lengthen the plate circuit as much. At the upper end of the operating band, two anode straps are used. The parallel arrangement lowers the total inductance of the strap connection and adds still less electrical length to the plate circuit.

The main section of the plate resonant line, together with the rotary section, functions as a parallel inductance. RF current flows in the same direction through the transmission line and the rotary section. Therefore, the magnetic fields of the two paths add. When the rotary section is at maximum height, the magnetic coupling between the main section of the transmission line and the rotary assembly is maximum. Because of the relatively large mutual inductance provided by this close coupling, the total inductance of these parallel inductors increases. This electrically lengthens the transmission line and lowers the resonant frequency. The concept is illustrated in Figure 4.27*a*.

When the rotary section is at minimum height, the magnetic coupling between the two parts of the inner conductor is minimum. This reduced coupling lowers the mutual inductance, which lowers the total inductance of the parallel combination. The reduced

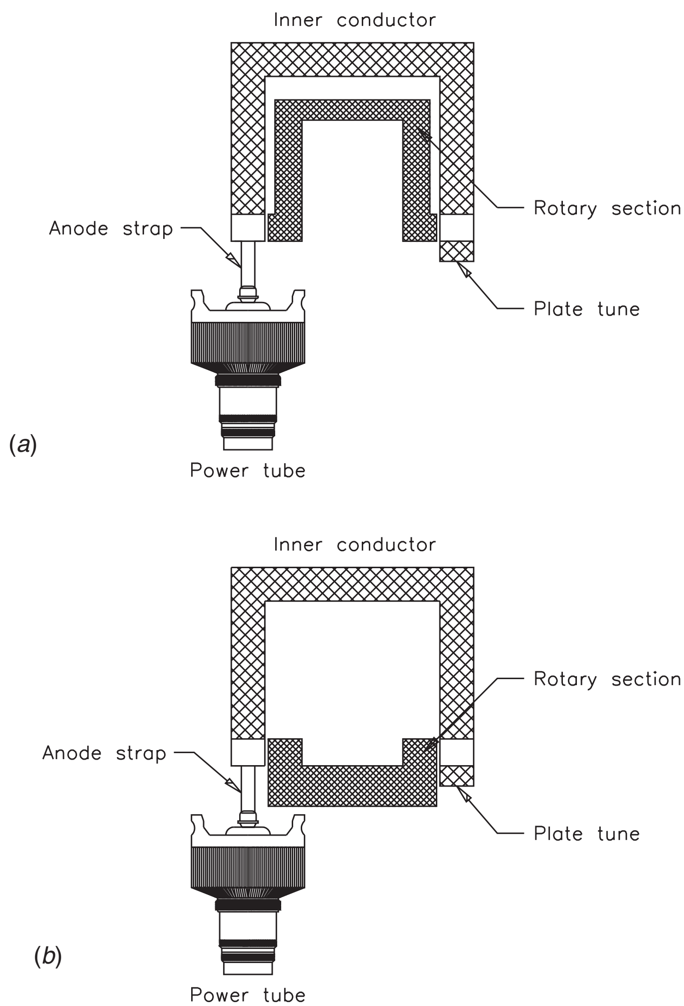


Figure 4.27 Using the cavity's movable section to adjust for resonance: (*a*) the rotary section at maximum height, (*b*) the rotary section at minimum height.

inductance allows operation at a higher resonant frequency. This condition is illustrated in Figure 4.27*b*. The rotary section provides an infinite number of coarse settings for various operating frequencies.

The movable plate-tune assembly is located at the end of the inner-plate transmission line. It can be moved up and down to change the physical length of the inner conductor by about 4-3/4 in. This assembly is linked to the front-panel plate-tuning knob, providing a fine adjustment for cavity resonance.

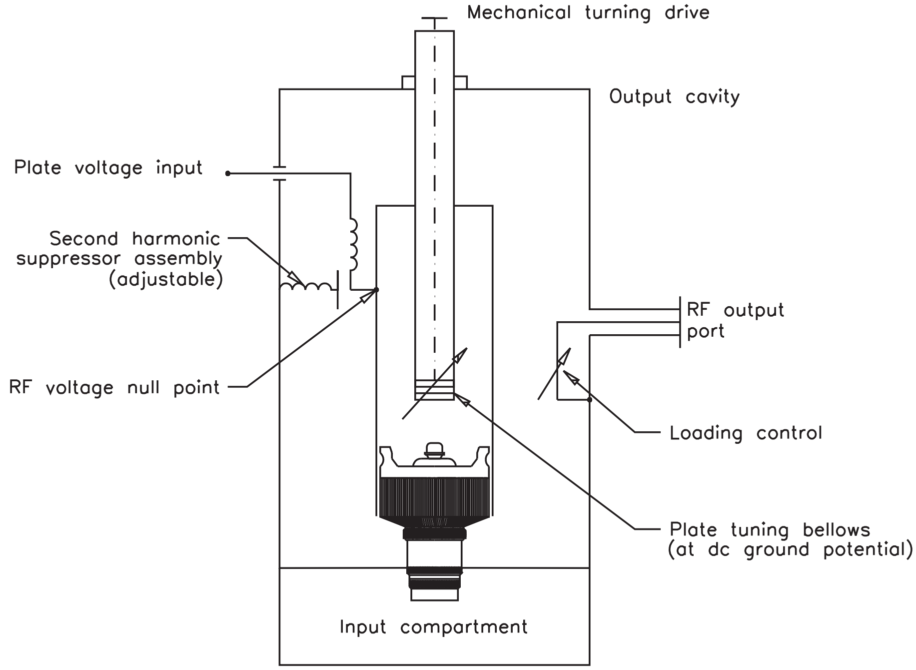


Figure 4.28 The basic design of a folded 1/2-wavelength cavity.

4.3.5 Folded 1/2-Wavelength Cavity

A special case of the 1/2-wavelength PA cavity is shown in Figure 4.28. The design employs a folded 1/2-wave resonator constructed with coaxial aluminum and copper tubing. This cavity arrangement eliminates the high-voltage blocking capacitor and high-current shorting contacts of conventional designs by connecting the main transmission-line resonant circuit conductor directly to the anode of the power tube. A grounded, concentric transmission-line center conductor tunes the cavity with a variable reentrant length inserted into the end of the main conductor opposite the tube.

The main conductor (the fixed portion of the plate line) is insulated from ground and carries the anode dc potential. High-voltage power is fed at the fundamental-frequency RF voltage null point, approximately 1/4 wavelength from the anode, for easy RF decoupling. A large surface area without sliding contacts results in minimal loss.

Incorporated into the tank design is a second harmonic suppressor. Rather than attenuating the second harmonic after the signal has been generated and amplified, this design essentially prevents formation of second-harmonic energy by series-*LC* trapping the second-harmonic waveform at the point where the wave exhibits a high impedance (approximately 1/4 wavelength from the anode). The second harmonic will peak in voltage at the same point that the dc-plate potential is applied.

Plate tuning is accomplished by an adjustable bellows on the center portion of the plate line, which is maintained at chassis ground potential. Output coupling is accomplished with an untuned loop intercepting the magnetic field concentration at the voltage null (maximum RF current) point of the main line. The PA loading control varies the angular position of the plane of the loop with respect to the plate line, changing the amount of magnetic field that it intercepts. Multiple phosphor-bronze leaves connect one side of the output loop to ground and the other side to the center conductor of the output transmission line. This allows for mechanical movement of the loop by the PA loading control without using sliding contacts. The grounded loop improves immunity to lightning and to static buildup on the antenna.

4.3.6 Wideband Cavity

The cavity systems discussed so far in this section provide adequate bandwidth for many applications. Some uses, however, dictate an operating bandwidth beyond what a conventional cavity amplifier can provide. One method of achieving wider operating bandwidth involves the use of a double-tuned overcoupled amplifier, shown in Figure 4.29. The system includes four controls to accomplish tuning:

- *Primary tune*, which resonates the plate circuit and tends to tilt the response and slide it up and down the bandpass.
- *Coupling*, which sets the bandwidth of the PA. Increased coupling increases the operating bandwidth and lowers the PA efficiency. When the coupling is adjusted, it can tilt the response and change the center of the bandpass, necessitating readjustment of the plate-tune control.
- *Secondary tune*, which resonates the secondary cavity and will tilt the response if so adjusted. The secondary tune control typically can be used to slide the response up and down the bandpass, similar to the primary tune control.
- *Loading*, which determines the value of ripple in the bandpass response. Adjustment of the loading control usually tilts the response and changes the bandwidth, necessitating readjustment of the secondary tune and coupling controls.

4.3.7 Output Coupling

Coupling is the process by which RF energy is transferred from the amplifier cavity to the output transmission line. Wideband cavity systems use coupling to transfer energy from the primary cavity to the secondary cavity. Coupling in tube-type power amplifiers usually transforms a high (plate or cavity) impedance to a lower output (transmission line) impedance. Both capacitive (electrostatic) and inductive (magnetic) coupling methods are used in cavity RF amplifiers. In some designs, combinations of the two methods are used.

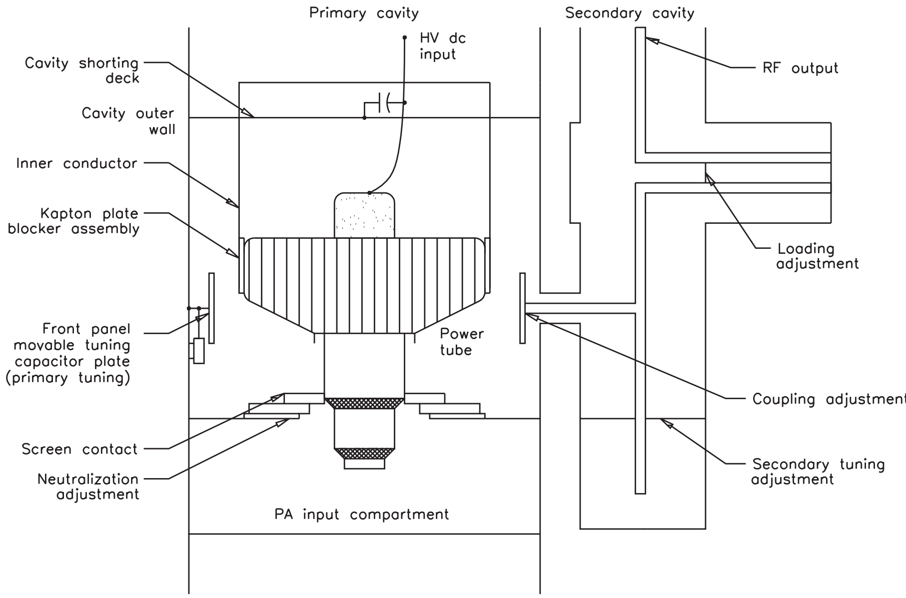


Figure 4.29 Double-tuned overcoupled broadband cavity amplifier.

Inductive Coupling

Inductive (magnetic) coupling employs the principles of transformer action. The efficiency of the coupling depends upon three conditions:

- The cross-sectional area under the coupling loop, compared to the cross-sectional area of the cavity (see Figure 4.30). This effect can be compared to the turns ratio principle of a transformer.
- The orientation of the coupling loop to the axis of the magnetic field. Coupling from the cavity is proportional to the cosine of the angle at which the coupling loop is rotated away from the axis of the magnetic field, as illustrated in Figure 4.31.
- The amount of magnetic field that the coupling loop intercepts (see Figure 4.32). The strongest magnetic field will be found at the point of maximum RF current in the cavity. This is the area where maximum inductive coupling is obtained. Greater magnetic field strength also is found closer to the center conductor of the cavity. Coupling, therefore, is inversely proportional to the distance of the coupling loop from the center conductor.

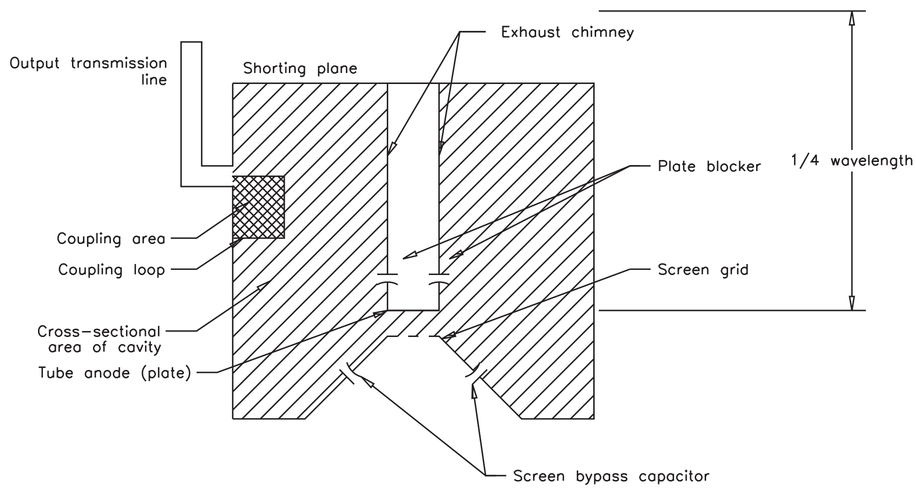


Figure 4.30 The use of inductive coupling in a 1/4-wavelength PA stage.

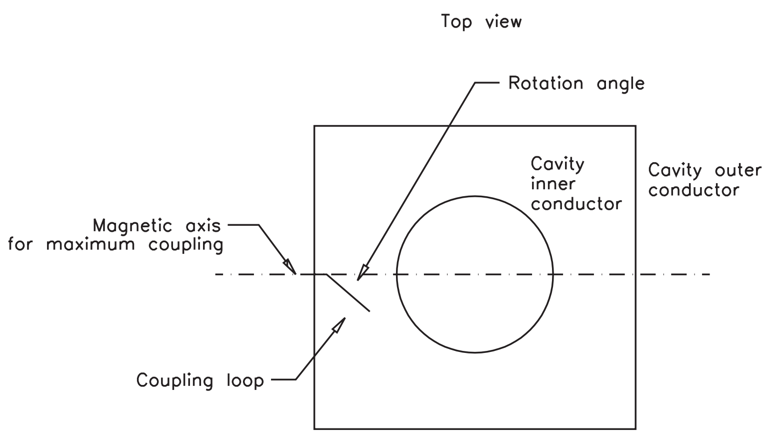


Figure 4.31 Top view of a cavity showing the inductive coupling loop rotated away from the axis of the magnetic field of the cavity.

In both 1/4- and 1/2-wavelength cavities, the coupling loop generally feeds a 50 Ω transmission line (the load). The loop is in series with the load and has considerable inductance at very high frequencies. This inductance will reduce the RF current that flows into the load, thus reducing power output. This effect can be overcome by placing a variable capacitor in series with the output coupling loop. The load is connected to one end of the coupling loop, and the variable capacitor ties the other end of the loop to

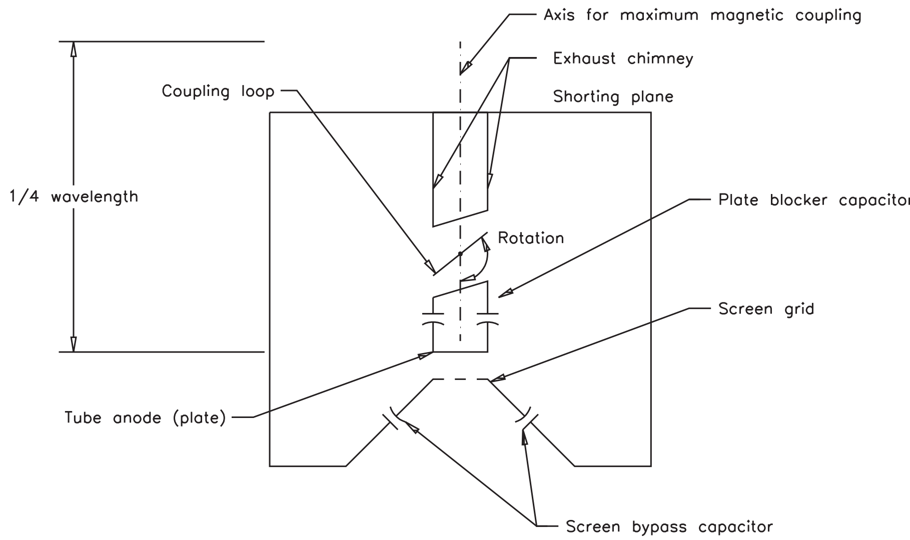


Figure 4.32 Cutaway view of the cavity showing the coupling loop rotated away from the axis of the magnetic field.

ground. The variable capacitor cancels some or all of the loop inductance. It functions as the PA-stage loading control.

Maximum loop current and output power occurs when the loading capacitor cancels all of the inductance of the loading loop. This lowers the plate impedance and results in heavier loading.

Light loading results if the loading capacitance does not cancel all of the loop inductance. The loop inductance that is not canceled causes a decrease in load current and power output, and an increase in plate impedance.

Capacitive Coupling

Capacitive (electrostatic) coupling, which physically appears to be straightforward, often baffles the applications engineer because of its unique characteristics. Figure 4.33 shows a cavity amplifier with a capacitive coupling plate positioned near its center conductor. This coupling plate is connected to the output load, which may be a transmission line or a secondary cavity (for wideband operation). The parameters that control the amount of capacitive coupling include:

- The area of the coupling capacitor plate (the larger the area, the greater the coupling)
- The distance from the coupling plate to the center conductor (the greater the distance, the lighter the coupling)

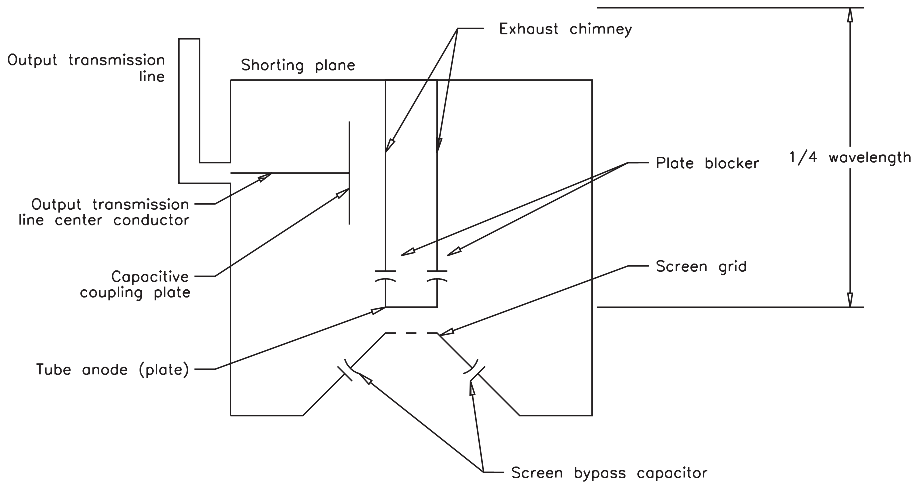


Figure 4.33 A 1/4-wavelength cavity with capacitive coupling to the output load.

Maximum capacitive coupling occurs when the coupling plate is at the maximum voltage point on the cavity center conductor.

To understand the effects of capacitive coupling, observe the equivalent circuit of the cavity. Figure 4.34 shows the PA tube, cavity (functioning as a parallel resonant circuit), and output section. The plate-blocking capacitor isolates the tube's dc voltage from the cavity. The coupling capacitor and output load are physically in series, but electrically they appear to be in parallel, as shown in Figure 4.35. The resistive component of the equivalent parallel circuit is increased by the coupling reactance. The equivalent parallel coupling reactance is absorbed into the parallel resonant circuit. This explains the need to retune the plate after changing the PA stage coupling (loading). The coupling reactance may be a series capacitor or inductor.

The series-to-parallel transformations are explained by the following formula:

$$R_p = \frac{R_s^{\,2} + X_s^{\,2}}{R_s} \tag{4.12}$$

$$X_p = \frac{R_s^{\,2} + X_s^{\,2}}{X_s} \tag{4.13}$$

Where:
R_p = effective parallel resistance
R_s = actual series resistance

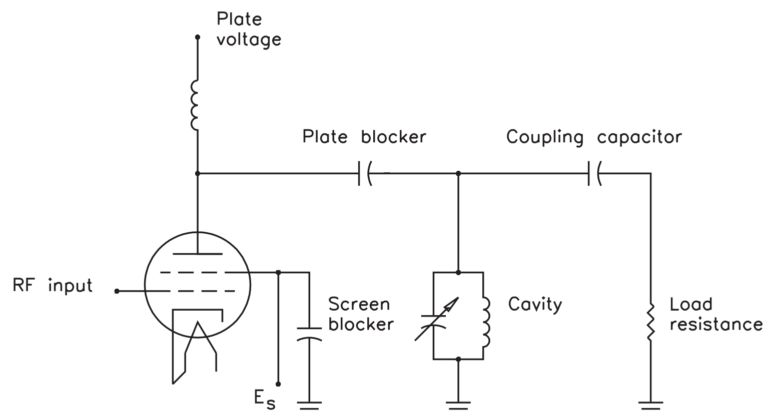


Figure 4.34 The equivalent circuit of a 1/4-wavelength cavity amplifier with capacitive coupling.

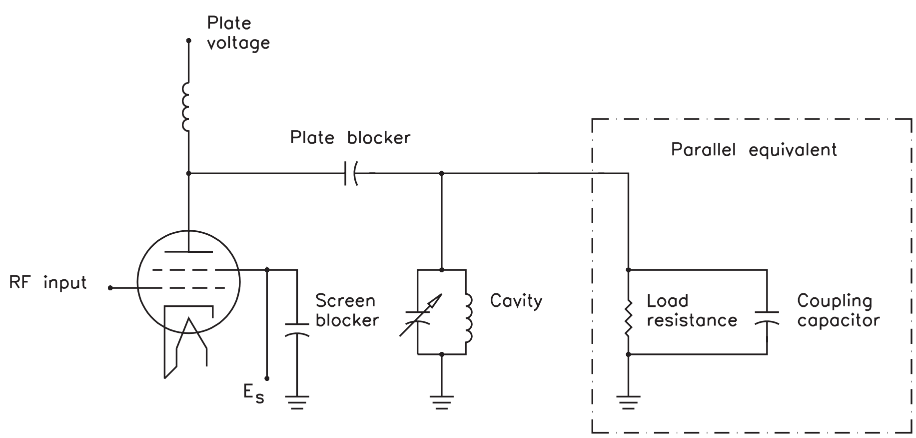


Figure 4.35 The equivalent circuit of a 1/4-wavelength cavity amplifier showing how series capacitive coupling appears electrically as a parallel circuit to the PA tube.

X_s = actual series reactance
X_p = effective parallel reactance

PA Loading

Proper loading of a cavity PA stage to the output transmission line is critical to dependable operation and optimum efficiency of the overall system. Light coupling produces light loading and results in a high plate impedance; conversely, heavy coupling results in heavier loading and a lower plate impedance. Maximum output power, coinciding with maximum efficiency and acceptable dissipation, dictates a specific plate impedance for a cavity of given design. This plate impedance is also dependent upon the operating values of dc plate voltage (E_p) and plate current (I_p).

Plate impedance dictates the cavity parameters of loaded Q, RF circulating current, and bandwidth. The relationships can be expressed as follows:

- Loaded Q is directly proportional to the plate impedance and controls the other two cavity parameters. Loaded $Q = Z_p/X_l$, where Z_p = cavity plate impedance and X_l = cavity inductive reactance.
- Circulating current in the cavity is much greater (by a factor of the loaded Q) than the RF current supplied by the tube. Circulating current $= Q \times I_p$, where I_p = the RF current supplied to the cavity by the tube.
- The cavity bandwidth is dependent on the loaded Q and operating frequency. Bandwidth $= F_r/Q$, where F_r = the cavity resonant frequency.

Heavy loading lowers the PA plate impedance and cavity Q. A lower Q reduces the cavity RF circulating currents. In some cavities, high circulating currents can cause cavity heating and premature failure of the plate or screen blocking capacitors. The effects of lower plate impedance—a by-product of heavy loading—are higher RF and dc plate currents and reduced RF plate voltage. The instantaneous plate voltage is the result of the RF plate voltage added to the dc plate voltage. The reduced swing of plate voltage causes less positive dc screen current to flow. Positive screen current flows only when the plate voltage swings close to or below the value of the positive screen grid.

Light loading raises the plate impedance and cavity Q. A higher Q will increase the cavity circulating currents, raising the possibility of component overheating and failure. The effects of higher plate impedance are reduced RF and dc plate current and increased RF and dc plate voltage excursions. The higher cavity RF or peak dc voltages may cause arcing in the cavity.

There is one value of plate impedance that will yield optimum output power, efficiency, dissipation, and dependable operation. It is dictated by cavity design and the values of the various dc and RF voltages and currents supplied to the stage.

Depending on the cavity design, light loading may seem deceptively attractive. The dc plate voltage is constant (set by the power supply), and the lower dc plate current resulting from light loading reduces the tube's overall dc input power. The RF output power may change with light loading, depending on the plate impedance and cavity design, while efficiency will probably increase or, at worst, remain constant. Caution must be exercised with light loading, however, because of the increased RF voltages and circulating currents that such operation creates. Possible adverse effects include

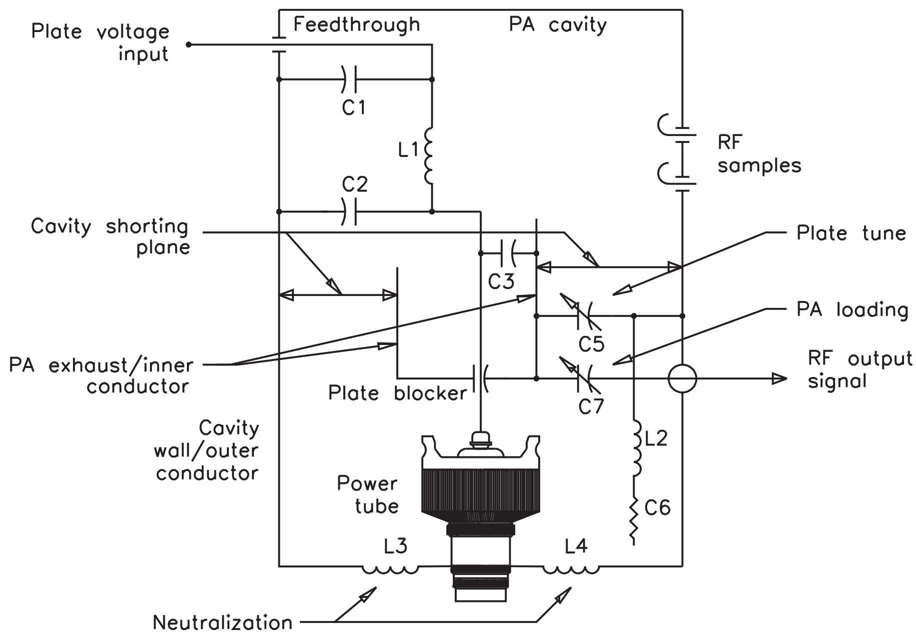


Figure 4.36 Typical VHF cavity amplifier.

cavity arcing and overheating of cavity components, such as capacitors. The manufacturer's recommendations on PA tube loading should, therefore, be carefully observed.

Despite the many similarities among various cavity designs, each imposes its own set of operational requirements and limitations. No two cavity systems will tune up in exactly the same fashion. Given proper maintenance, a cavity amplifier will provide years of reliable service.

4.3.8 Mechanical Design

The operation of a cavity amplifier can be puzzling because of the nature of the major component elements. It is often difficult to relate the electrical schematic diagram to the mechanical assembly that exists within the transmitter. Consider the PA cavity schematic diagram shown in Figure 4.36. The grounded screen stage is of conventional design. Decoupling of the high-voltage power supply is accomplished by C1, C2, C3, and L1. Capacitor C3 is located inside the PA chimney (cavity inner conductor). The RF sample lines provide two low-power RF outputs for a modulation monitor or other test instrument. Neutralization inductors L3 and L4 consist of adjustable grounding bars on the screen grid ring assembly.

Figure 4.37 shows the electrical equivalent of the PA cavity schematic diagram. The 1/4-wavelength cavity acts as the resonant tank for the PA. Coarse-tuning of the cavity

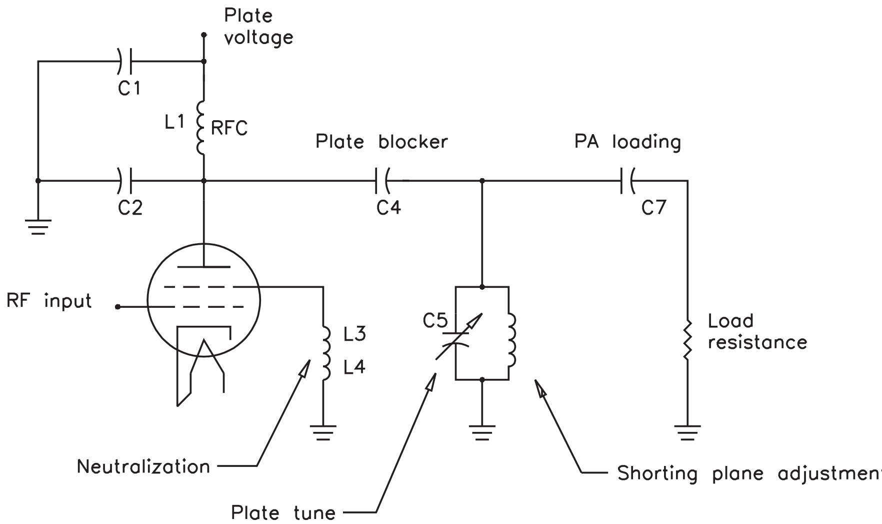
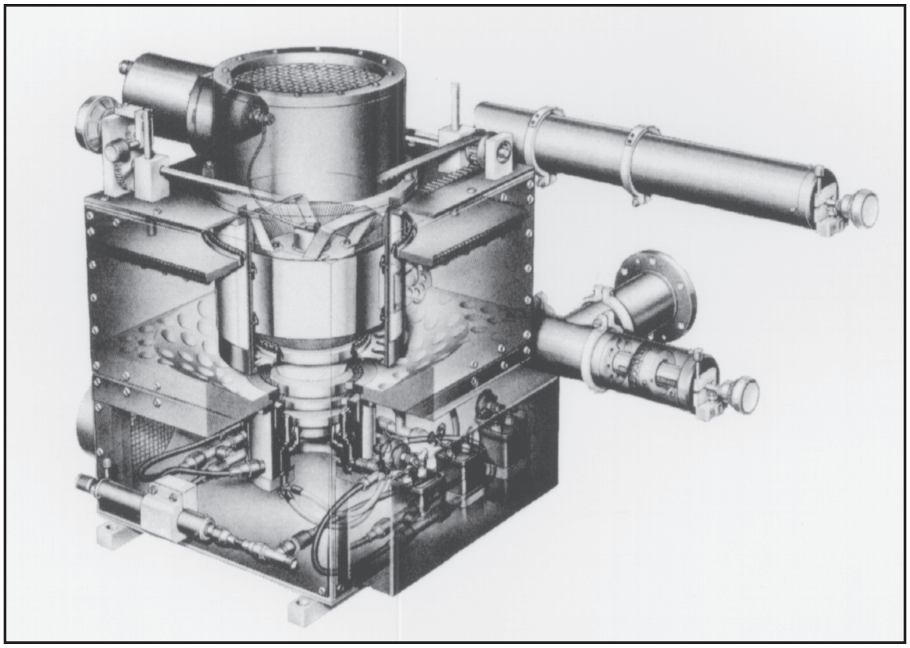


Figure 4.37 Electrical equivalent of the cavity amplifier shown in Figure 4.36.

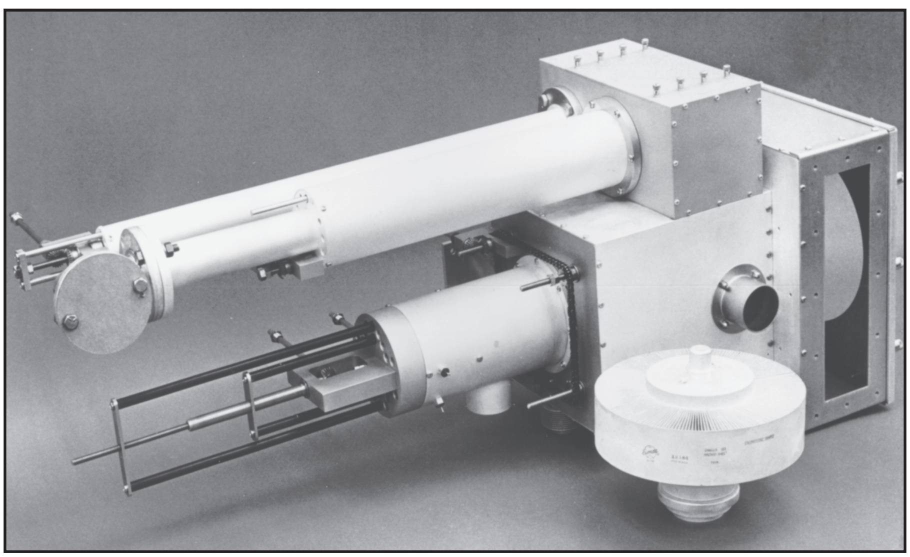
is accomplished by adjustment of the shorting plane. Fine-tuning is performed by the PA tuning control, which acts as a variable capacitor to bring the cavity into resonance. The PA loading control consists of a variable capacitor that matches the cavity to the load. The assembly made up of L2 and C6 prevents spurious oscillations within the cavity.

Blocking capacitor C4 is constructed of a roll of *Kapton* insulating material sandwiched between two circular sections of aluminum. (Kapton is a registered trademark of Du Pont.) PA plate-tuning control C5 consists of an aluminum plate of large surface area that can be moved in or out of the cavity to reach resonance. PA loading control C7 is constructed much the same as the PA tuning assembly, with a large-area paddle feeding the harmonic filter, located external to the cavity. The loading paddle may be moved toward the PA tube or away from it to achieve the required loading. The L2-C6 damper assembly actually consists of a 50 Ω noninductive resistor mounted on the side of the cavity wall. Component L2 is formed by the inductance of the connecting strap between the plate-tuning paddle and the resistor. Component C6 is the equivalent stray capacitance between the resistor and the surrounding cavity box.

It can be seen that cavity amplifiers involve as much mechanical engineering as they do electrical engineering. The photographs of Figure 4.38 show graphically the level of complexity that a cavity amplifier may involve. Figure 4.38*a* depicts a VHF power amplifier (Philips) with broadband input circuitry. Figure 4.38*b* shows a wideband VHF amplifier intended for television service (Varian). Figure 4.38*c* illustrates a VHF cavity amplifier designed for FM broadcast service (Varian).



(*a*)



(*b*)

Figure 4.38 VHF cavity amplifiers: (*a) cross-sectional view of a broadband design, (b*) *cavity designed for television service, (c*, next page) FM broadcast cavity amplifier.

(*c*)

Figure 4.38*c*

4.4 High-Voltage Power Supplies

Virtually all power grid and microwave vacuum tubes require one or more sources of high voltage to operate. The usual source of operating power is a single-phase or multiphase supply operating from ac line current. Most supplies use silicon rectifiers as the primary ac-to-dc converting device.

4.4.1 Silicon Rectifiers

Rectifier parameters generally are expressed in terms of reverse-voltage ratings and mean forward-current ratings in a -wave rectifier circuit operating from a 60 Hz supply and feeding a purely resistive load. The three principle reverse-voltage ratings are:

- *Peak transient reverse voltage* (V_{rm}). The maximum value of any nonrecurrent surge voltage. This value must never be exceeded.
- *Maximum repetitive reverse voltage* ($V_{RM(rep)}$). The maximum value of reverse voltage that may be applied recurrently (in every cycle of 60 Hz power). This includes oscillatory voltages that may appear on the sinusoidal supply.
- *Working peak reverse voltage* ($V_{RM(wkg)}$). The crest value of the sinusoidal voltage of the ac supply at its maximum limit. Rectifier manufacturers generally recommend a value that has a significant safety margin, relative to the peak transient reverse voltage (V_{rm}), to allow for transient overvoltages on the supply lines.

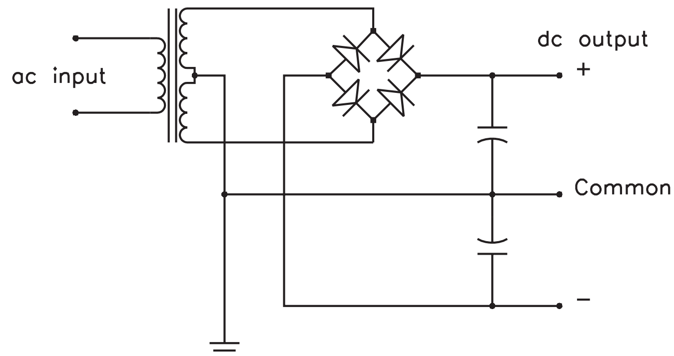


Figure 4.39 Conventional capacitor input filter full-wave bridge.

There are three forward-current ratings of similar importance in the application of silicon rectifiers:

- *Nonrecurrent surge current* ($I_{FM,surge}$). The maximum device transient current that must not be exceeded at any time. $I_{FM,surge}$ is sometimes given as a single value, but more often is presented in the form of a graph of permissible surge-current values vs. time. Because silicon diodes have a relatively small thermal mass, the potential for short-term current overloads must be given careful consideration.
- *Repetitive peak forward current* ($I_{FM,rep}$). The maximum value of forward-current reached in each cycle of the 60 Hz waveform. This value does not include random peaks caused by transient disturbances.
- *Average forward current* ($I_{FM,av}$). The upper limit for average load current through the device. This limit is always well below the repetitive peak forward-current rating to ensure an adequate margin of safety.

Rectifier manufacturers generally supply curves of the instantaneous forward voltage vs. instantaneous forward current at one or more specific operating temperatures. These curves establish the forward-mode upper operating parameters of the device.

Figure 4.39 shows a typical rectifier application in a bridge circuit.

4.4.2 Operating Rectifiers in Series

High-voltage power supplies (5 kV and greater) often require rectifier voltage ratings well beyond those typically available from the semiconductor industry. To meet the requirements of the application, manufacturers commonly use silicon diodes in a series configuration to give the required working peak reverse voltage. For such a configuration to work properly, the voltage across any one diode must not exceed the rated peak transient reverse voltage (V_{rm}) at any time. The dissimilarity commonly found between the reverse leakage current characteristics of different diodes of the same type number makes this objective difficult to achieve. The problem normally is

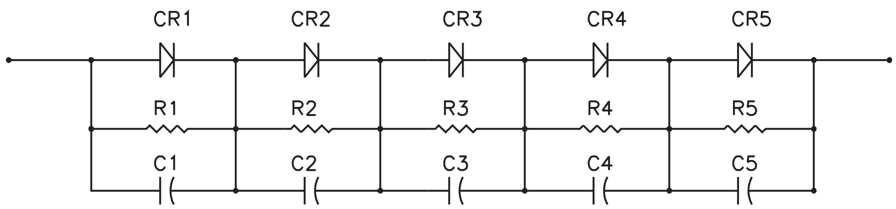


Figure 4.40 A portion of a high-voltage series-connected rectifier stack.

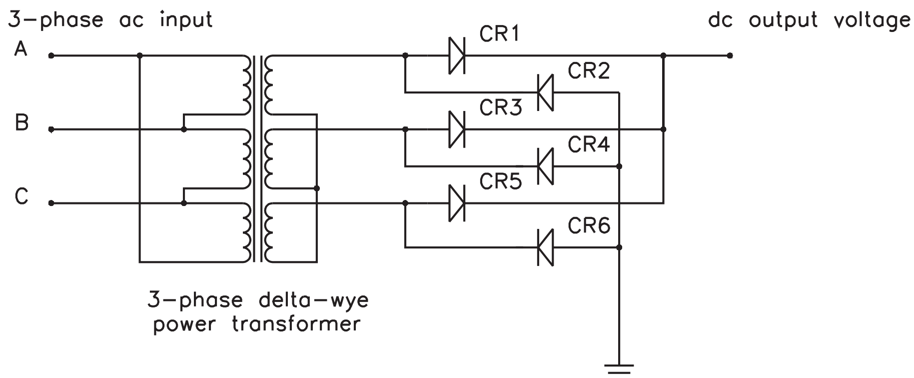


Figure 4.41 A 3-phase delta-connected high-voltage rectifier.

overcome by connecting shunt resistors across each rectifier in the chain, as shown in Figure 4.40. The resistors are chosen so that the current through the shunt elements (when the diodes are reverse-biased) will be several times greater than the leakage current of the diodes themselves.

The *carrier storage* effect also must be considered in the use of a series-connected rectifier stack. If precautions are not taken, different diode recovery times (caused by the carrier storage phenomenon) will effectively force the full applied reverse voltage across a small number of diodes, or even a single diode. This problem can be prevented by connecting small-value capacitors across each diode in the rectifier stack. The capacitors equalize the transient reverse voltages during the carrier storage recovery periods of the individual diodes.

Figure 4.41 illustrates a common circuit configuration for a high-voltage 3-phase rectifier bank. A photograph of a high-voltage series-connected 3-phase rectifier assembly is shown in Figure 4.42.

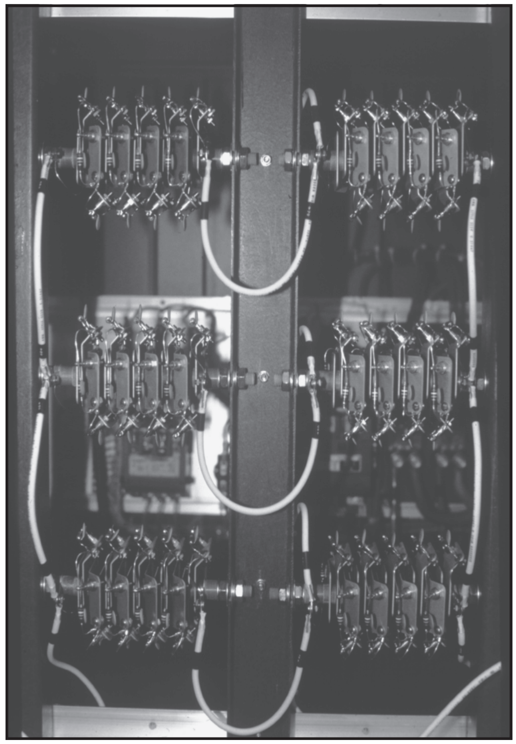


Figure 4.42 High-voltage rectifier assembly for a 3-phase delta-connected circuit.

4.4.3 Operating Rectifiers in Parallel

Silicon rectifiers are used in a parallel configuration when a large amount of current is required from the power supply. *Current sharing* is the major design problem with a parallel rectifier assembly because diodes of the same type number do not necessarily exhibit the same forward characteristics.

Semiconductor manufacturers often divide production runs of rectifiers into tolerance groups, matching forward characteristics of the various devices. When parallel diodes are used, devices from the same tolerance group must be used to avoid unequal sharing of the load current. As a margin of safety, designers allow a substantial derating factor for devices in a parallel assembly to ensure that the maximum operating limits of any one component are not exceeded.

The problems inherent in a parallel rectifier assembly can be reduced through the use of a resistance or reactance in series with each component, as shown in Figure 4.43. The buildout resistances (R1 through R4) force the diodes to share the load current equally. Such assemblies can, however, be difficult to construct and may be more expensive than simply adding diodes or going to higher-rated components.

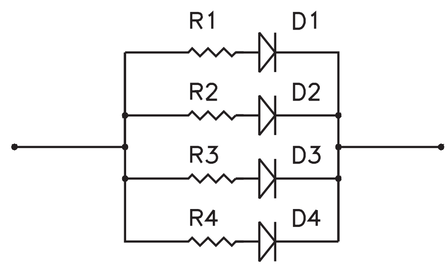


Figure 4.43 Using buildout resistances to force current sharing in a parallel rectifier assembly.

4.4.4 Silicon Avalanche Rectifiers

The *silicon avalanche diode* is a special type of rectifier that can withstand high reverse power dissipation. For example, an avalanche diode with a normal forward rating of 10 A can dissipate a reverse transient of 8 kW for 10 ms without damage. This characteristic of the device allows elimination of the surge-absorption capacitor and voltage-dividing resistor networks needed when conventional silicon diodes are used in a series rectifier assembly. Because fewer diodes are needed for a given applied reverse voltage, significant underrating of the device (to allow for reverse-voltage transient peaks) is not required.

When an extra-high-voltage rectifier stack is used, it is still advisable to install shunt capacitors—but not resistors—in an avalanche diode assembly. The capacitors are designed to compensate for the effects of carrier storage and stray capacitance in a long series assembly.

4.4.5 Thyristor Servo Systems

Thyristor control of ac power has become a common method of regulating high-voltage power supplies. The type of servo system employed depends on the application. Figure 4.44 shows a basic single-phase ac control circuit using discrete thyristors. The rms load current (I_{rms}) at any specific phase delay angle (α) is given in terms of the normal full-load rms current at a phase delay of zero ($I_{rms\text{-}0}$):

$$I_{rms} = I_{rms-\theta} \left\{ 1 - \frac{\alpha}{\pi} + (2\pi) - 1 \sin 2\alpha \right\}^{1/2} \tag{4.14}$$

The load rms voltage at any particular phase-delay angle bears the same relationship to the full-load rms voltage at zero phase delay as the previous equation illustrates for load current. An analysis of the mathematics shows that although the theoretical delay range for complete control of a resistive load is 0 to 180°, a practical span of 20 to 160° gives a power-control range of approximately 99 to 1 percent of maximum output to the load. Figure 4.45 illustrates typical phase-control waveforms.

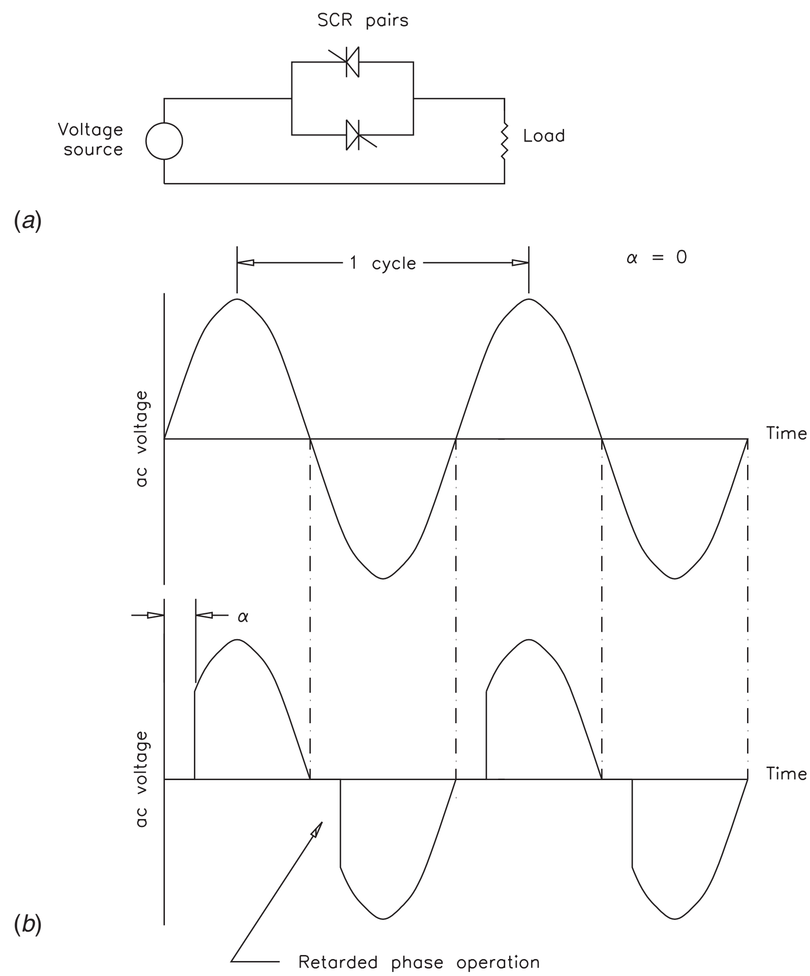


Figure 4.44 Inverse-parallel thyristor ac power control: (*a*) circuit diagram, (*b*) voltage and current waveforms for full and reduced thyristor control angles. The waveforms apply for a purely resistive load.

The circuit shown in Figure 4.44 requires a source of gate trigger pulses that must be isolated from each other by at least the peak value of the applied ac voltage. The two gate pulse trains must also be phased 180° with respect to each other. Also, the gate pulse trains must shift together with respect to the ac supply voltage phase when power throughput is adjusted.

Some power-control systems use two identical, but isolated, gate pulse trains operating at a frequency of twice the applied supply voltage (120 Hz for a 60 Hz system). Under such an arrangement, the forward-biased thyristor will fire when the gate pulses are applied to the silicon controlled rectifier (SCR) pair. The reverse-biased thyristor will

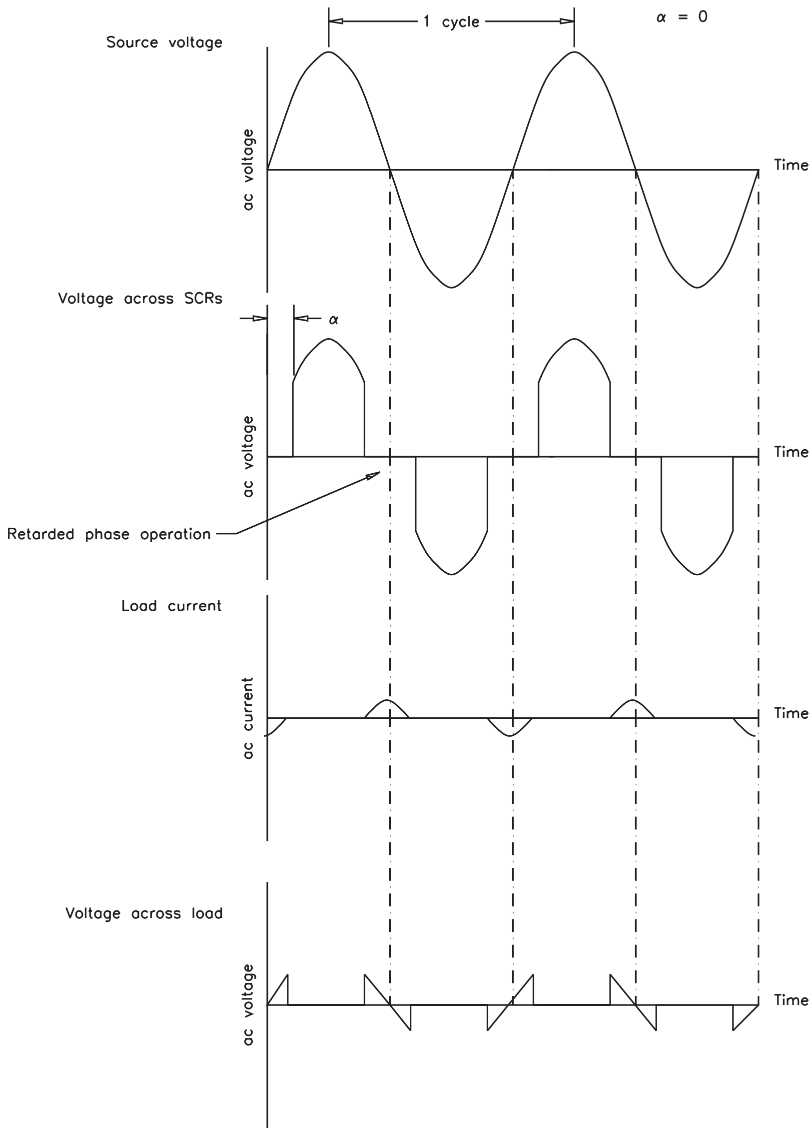


Figure 4.45 Waveforms in an ac circuit using thyristor power control.

not fire. Normally, it is considered unsafe to drive a thyristor gate positive while its anode-cathode is reverse-biased. In this case, however, it may be permissible because the thyristor that is fired immediately conducts and removes the reverse voltage from the other device. The gate of the reverse-biased device is then being triggered on a thyristor that essentially has no applied voltage.

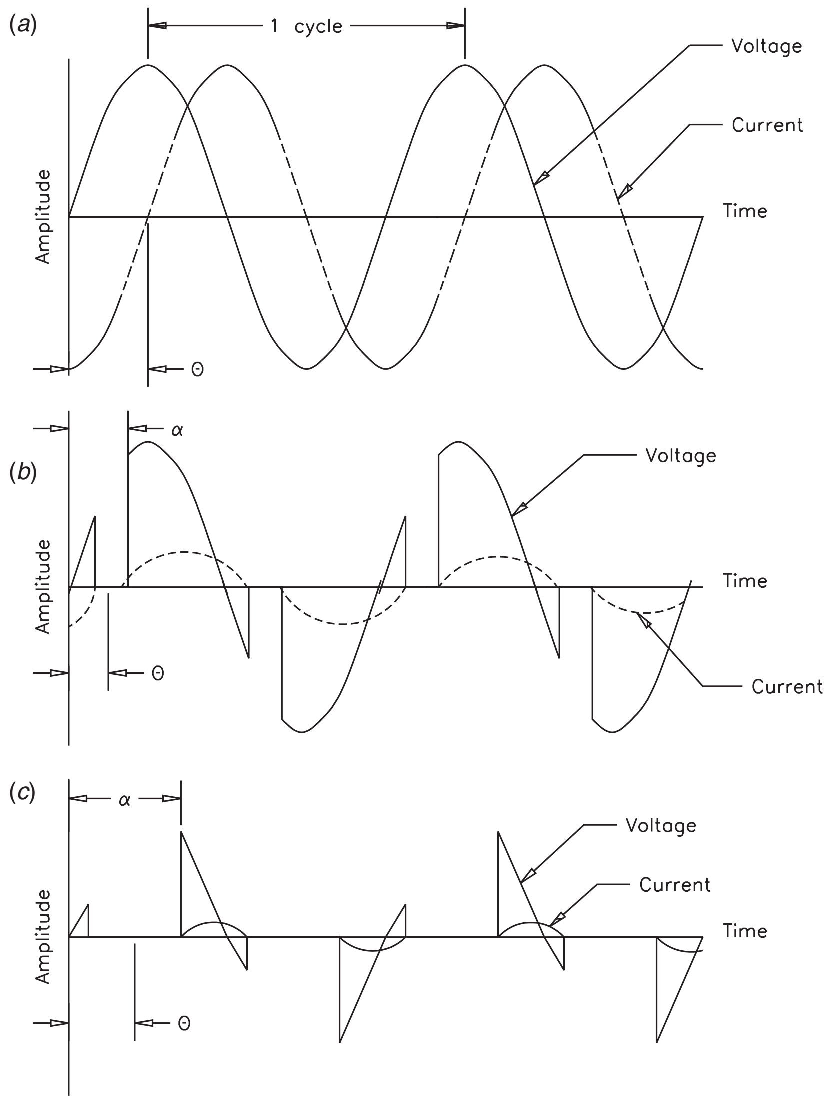


Figure 4.46 Voltage and current waveforms for inverse-parallel thyristor power control with an inductive load: (*a*) full conduction, (*b*) small-angle phase reduction, (*c*) large-angle phase reduction.

Inductive Loads

The waveforms shown in Figure 4.46 illustrate effects of phase control on an inductive load. When inductive loads are driven at a reduced conduction angle, a sharp transient change of load voltage occurs at the end of each current pulse (or loop). The

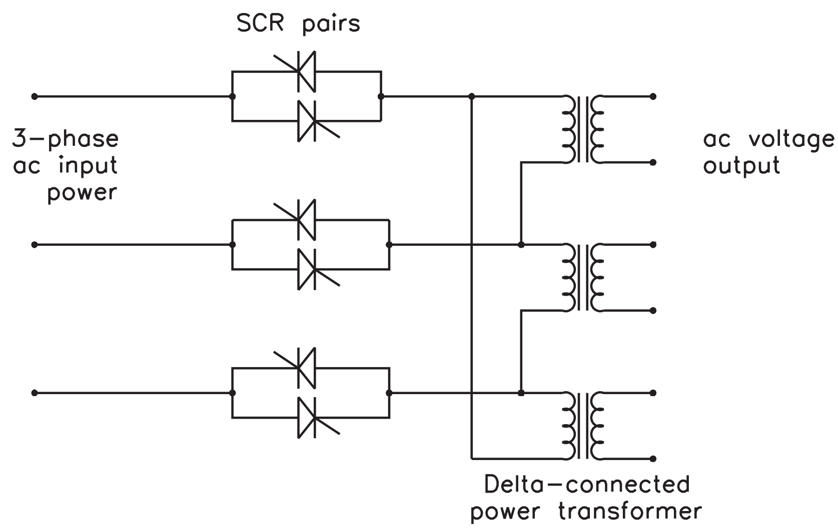


Figure 4.47 Modified full-thyristor 3-phase ac control of an inductive delta load.

transients generally have no effect on the load, but they can be dangerous to proper operation of the thyristors. When the conducting thyristor turns off, thereby disconnecting the load from the ac line supply, the voltage at the load rapidly drops to zero. This rapid voltage change, in effect, applies a sharply rising positive anode voltage to the thyristor opposing the device that has been conducting. If the thyristor *dv/dt* (change in voltage as a function of a change in time) rating is exceeded, the opposing device will turn on and conduction will take place, independent of any gate drive pulse.

A common protective approach involves the addition of a resistor-capacitor (*RC*) snubber circuit to control the rate of voltage change seen across the terminals of the thyristor pair. Whenever a thyristor pair is used to drive an inductive load, such as a power transformer, it is critically important that each device fires at a point in the applied waveform exactly 180° relative to the other. If proper timing is not achieved, the positive and negative current loops will differ in magnitude, causing a direct current to flow through the primary side of the transformer. A common trigger control circuit, therefore, should be used to determine gate timing for thyristor pairs.

Applications

Several approaches are possible for thyristor power control in a 3-phase ac system. The circuit shown in Figure 4.47 consists of essentially three independent, but interlocked, single-phase thyristor controllers. This circuit is probably the most common configuration found in commercial and industrial equipment.

In a typical application, the thyristor pairs feed a power transformer with multitap primary windings, thereby giving the user an adjustment range to compensate for varia-

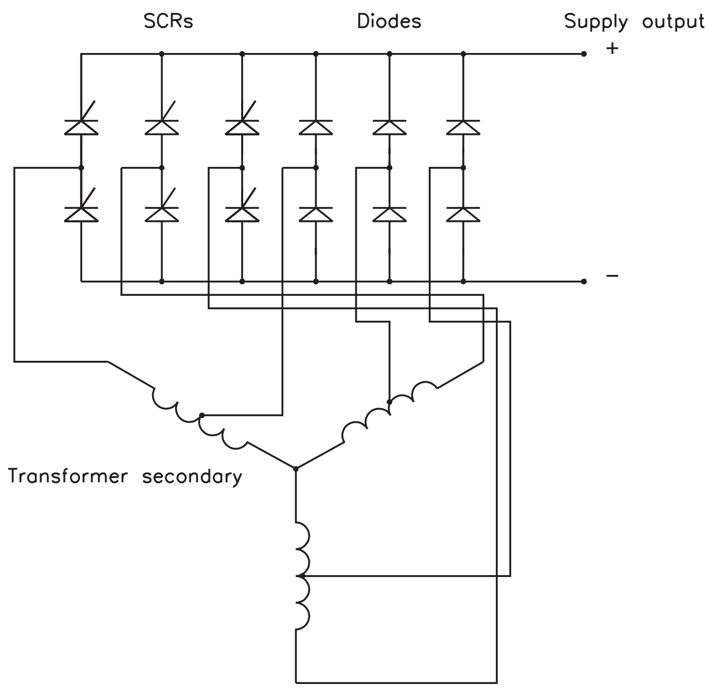


Figure 4.48 A 6-phase boost rectifier circuit.

tions in utility company line voltages from one location to another. A common procedure specifies selection of transformer tap positions that yield a power output of 105 percent when nominal utility company line voltages are present. The thyristor power-control system then is used to reduce the angle of conduction of the SCR pairs as necessary to cause a reduction in line voltage to the power transformer to yield 100 percent rated power output from the power supply. A servo loop from a sample point at the load may be used to compensate automatically for line-voltage variations. With such an arrangement, the thyristors are kept within a reasonable degree of retarded phase operation. Line voltages will be allowed to sag 5 percent or so without affecting the dc supply output. Utility supply voltage excursions above the nominal value simply will result in further delayed triggering of the SCR pairs.

Thyristor control of high-power loads (200 kW and above) typically uses transformers that provide 6- or 12-phase outputs. Although they are more complicated and expensive, such designs allow additional operational control, and filtering requirements are reduced significantly. Figure 4.48 shows a 6-phase *boost rectifier* circuit. The configuration consists basically of a full-wave 3-phase SCR bridge connected to a wye-configured transformer secondary. A second bridge, consisting of six diodes, is connected to low-voltage taps on the same transformer. When the SCRs are fully on, their output is at a higher voltage than the diode bridge. As a result, the diodes are reverse-biased and turned off. When the SCRs are partially on, the diodes are free to con-

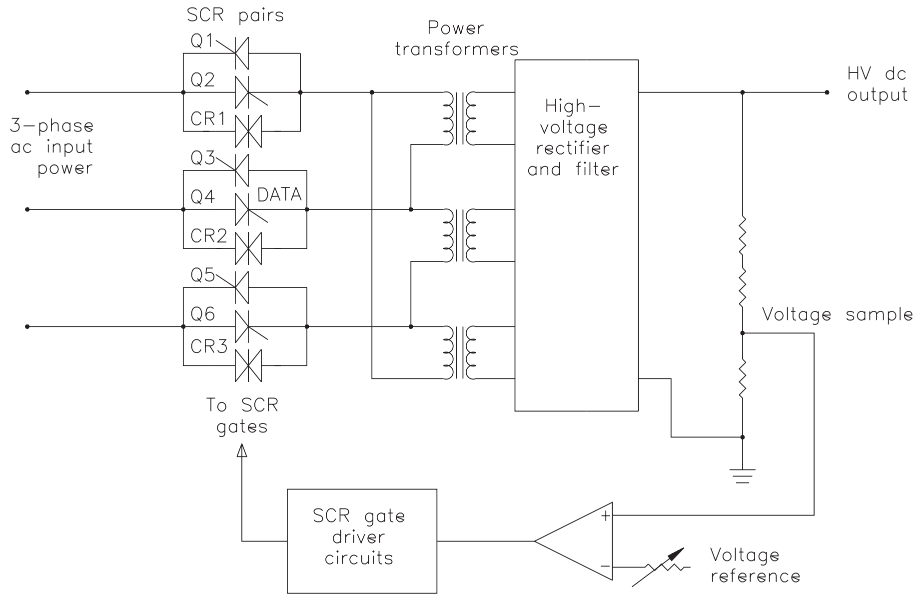


Figure 4.49 Thyristor-controlled high-voltage servo power supply.

duct. The diodes improve the quality of the output waveform during low-voltage (reduced conduction angle) conditions. The minimum output level of the supply is determined by the transformer taps to which the diodes are connected.

A thyristor-driven 3-phase power-control circuit is shown in Figure 4.49. A single-phase power-control circuit is shown in Figure 4.50.

Triggering Circuits

Accurate, synchronized triggering of the gate pulses is a critical element in thyristor control of a 3-phase power supply. The gate signal must be synchronized properly with the phase of the ac line that it is controlling. The pulse also must properly match the phase-delay angle of the gates of other thyristors in the power-control system. Lack of proper synchronization of gate pulse signals between thyristor pairs can result in improper current sharing (*current hogging*) among individual legs of the 3-phase supply.

The gate circuit must be protected against electrical disturbances that could make proper operation of the power-control system difficult or unreliable. Electrical isolation of the gate is a common approach. Standard practice calls for the use of gate pulse transformers in thyristor servo system gating cards. Pulse transformers are ferrite-cored devices with a single primary winding and (usually) multiple secondary

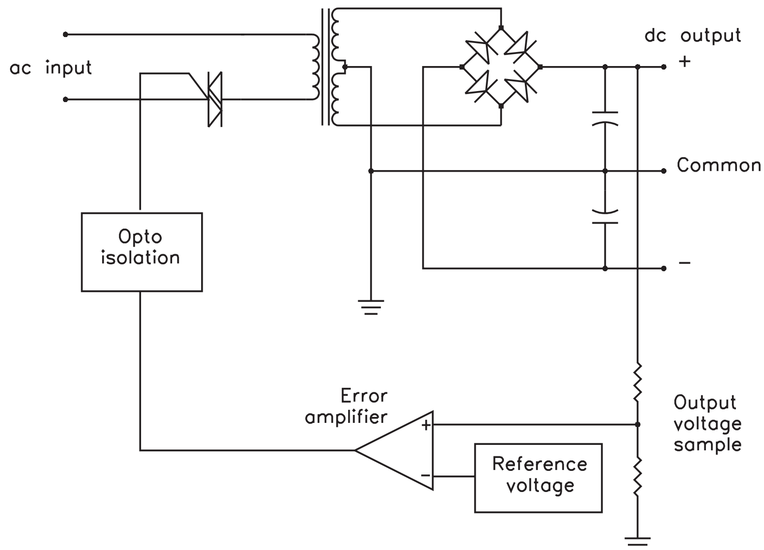


Figure 4.50 Phase-controlled power supply with primary regulation.

windings that feed, or at least control, the individual gates of a back-to-back thyristor pair. This concept is illustrated in Figure 4.51. Newer thyristor designs may use optocouplers to achieve the necessary electrical isolation between the trigger circuit and the gate.

It is common practice to tightly twist together the control leads from the gate and cathode of a thyristor to the gating card assembly. This practice provides a degree of immunity to high-energy pulses that might inadvertently trigger the thyristor gate. The gate circuit must be designed and configured carefully to reduce inductive and capacitive coupling that might occur between power and control circuits. Because of the high *di*/*dt* (change in current as a function of a change in time) conditions commonly found in thyristor-controlled power circuits, power wiring and control (gate) wiring must be separated physically as much as possible. Shielding of gating cards in a metal card cage is advisable.

Equipment manufacturers use various means to decrease gate sensitivity to transient sources, including placement of a series resistor in the gate circuit and/or a shunting capacitor between the gate and cathode. A series resistor has the effect of decreasing gate sensitivity, increasing the allowable *dv*/*dt* of the thyristor and reducing the turn-off time, which simultaneously increases the required holding and latching currents. The use of a shunt capacitor between the gate and cathode leads reduces high-frequency noise components that might be present on the gate lead and increases the *dv*/*dt* withstand capability of the thyristor.

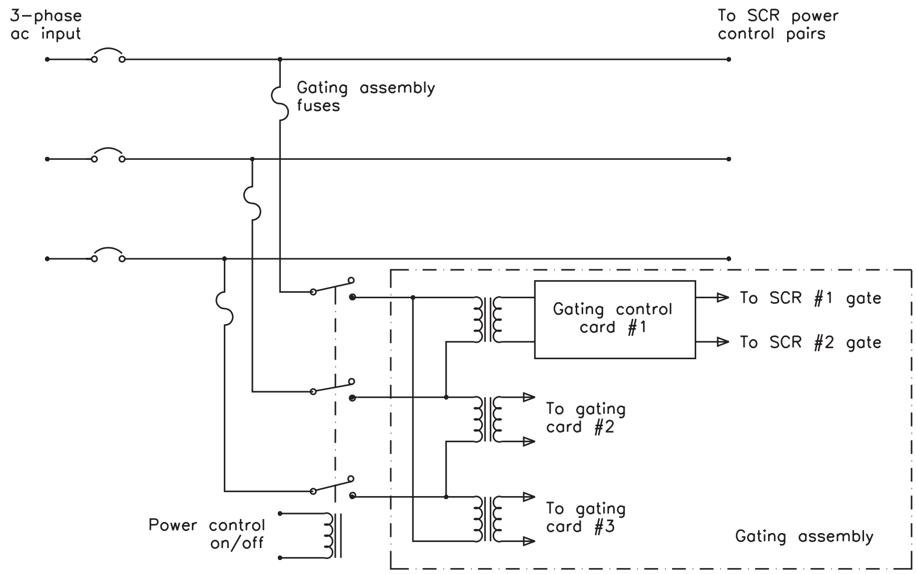


Figure 4.51 Simplified block diagram of the gating circuit for a phase-control system using back-to-back SCRs.

Fusing

Current limiting is a basic method of protection for any piece of equipment operated from the utility ac line. The device typically used for breaking fault currents is either a fuse or a circuit breaker. Some designs incorporate both components. *Semiconductor fuses* often are used in conjunction with a circuit breaker to provide added protection. Semiconductor fuses operate more rapidly (typically within 8.3 ms) and more predictably than common fuses or circuit breakers. Surge currents caused by a fault can destroy a semiconductor device, such as a power thyristor, before the ac line circuit breaker has time to act. Manufacturers of semiconductor fuses and thyristors usually specify in their data sheets the I^2t ratings of each device. Because the thyristor rating normally assumes that the device is operating at maximum rated current and maximum junction temperature (conditions that do not represent normal operation), a safety factor is ensured.

Control Flexibility

Thyristor servo control of a high-voltage power supply is beneficial to the user for a number of reasons, including:

- Wide control over ac input voltages

- Capability to compensate automatically for line-voltage variations
- Capability to soft-start the dc supply

Thyristor control circuits typically include a ramp generator that increases the ac line voltage to the power transformer from zero to full value within 2 to 5 s. This prevents high surge currents through rectifier stacks and filter capacitors during system startup.

Although thyristor servo systems are preferred over other power-control approaches from an operational standpoint, they are not without their drawbacks. The control system is complex and can be damaged by transient activity on the ac power line. Conventional power contactors are simple and straightforward. They either make contact or they do not. For reliable operation of the thyristor servo system, attention must be given to transient suppression at the incoming power lines.

4.4.6 Polyphase Rectifier Circuits

High-voltage power supplies typically used in vacuum tube circuits incorporate multiphase rectification of the ac line voltage. Common configurations include 3-, 6-, and 12-phase, with 3-phase rectification being the most common. Figure 4.52 illustrates four approaches to 3-phase rectification:

- *Three-phase half-wave wye*, Figure 4.52*a*. Three half-wave rectifiers are used in each leg of the secondary *wye* forming one phase. In such an arrangement, each diode carries current one-third of each cycle, and the output wave pulses at three times the frequency of the ac supply. In order to avoid direct-current saturation in the transformer, it is necessary to employ a 3-phase transformer rather than three single-phase transformers.
- *Three-phase full-wave bridge*, Figure 4.52*b*. Six diodes are used in this circuit to produce a low ripple output with a frequency of six times the input ac waveform. It is permissible in this configuration to use three single-phase transformers, if desired.
- *Six-phase star*, Figure 4.52*c*. This circuit, also known as a 3-phase diametric configuration, uses six diodes with a transformer secondary configured as a star, as illustrated in the figure. The output ripple frequency is six times the input ac waveform.
- *Three-phase double-wye*, Figure 4.52*d*. This circuit uses six diodes and a complicated configuration of transformer windings. Note the balance coil (*interphase transformer*) in the circuit.

The relative merits of these rectifier configurations are listed in Table 4.3.

Polyphase rectifiers are used when the dc power required is on the order of 2 kW or more. The main advantages of a polyphase power supply over a single-phase supply include the following:

- Division of the load current between three or more lines to reduce line losses.

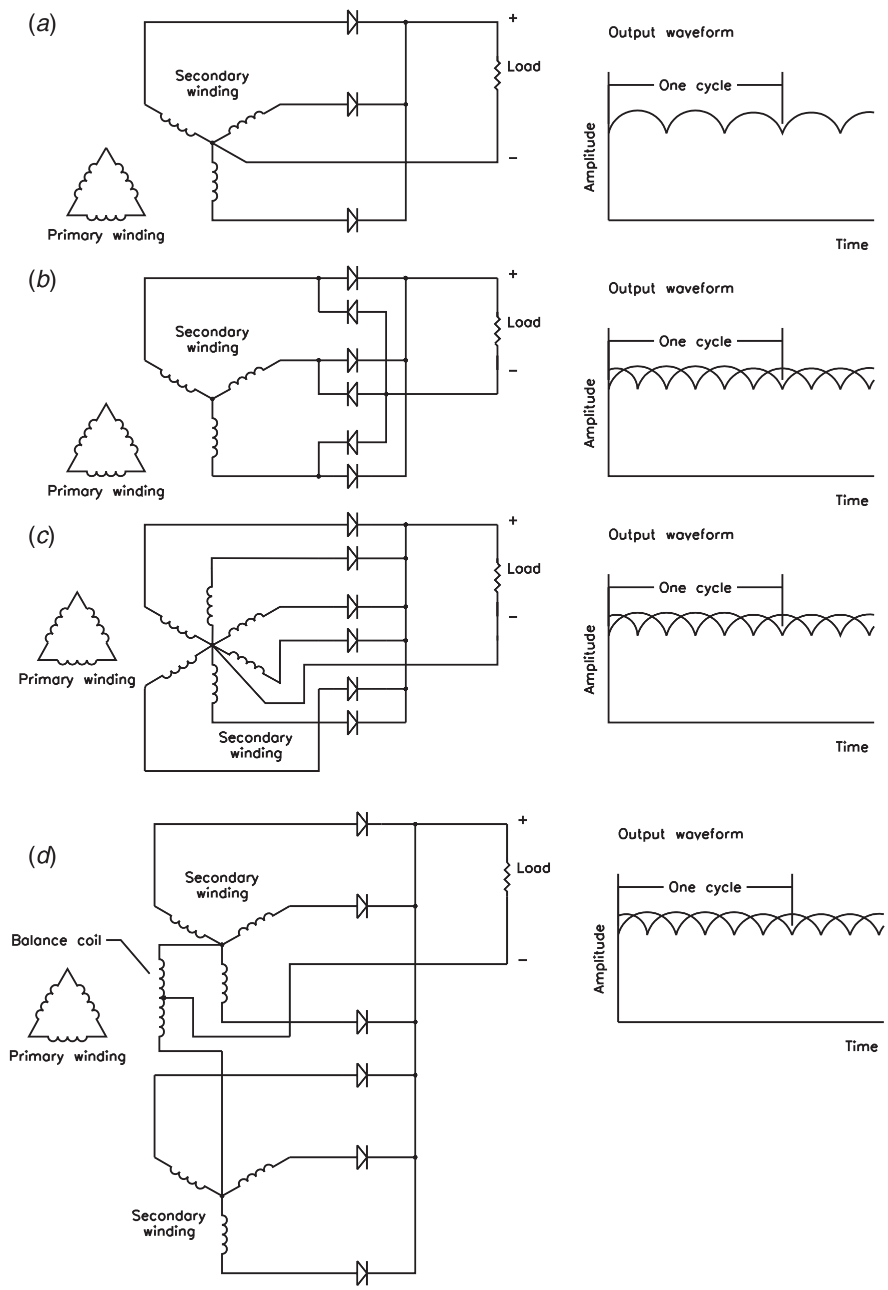


Figure 4.52 Basic 3-phase rectifier circuits: (*a*) half-wave wye, (*b*) full-wave bridge, (*c*) 6-phase star, (*d*) 3-phase double-wye.

Table 4.3 Operating Parameters of 3-Phase Rectifier Configurations

Parameter	3-Phase Star	3-Phase Bridge	6-Phase Star	3-Phase Double-Y	Multiplier[1]
Rectifier elements	3	6	6	6	
Rms dc voltage output	1.02	1.00	1.00	1.00	Average dc voltage output
Peak dc voltage output	1.21	1.05	1.05	1.05	Average dc voltage output
Peak reverse volts per rectifier	2.09	1.05	2.09	2.42	Average dc voltage output
	2.45	2.45	2.83	2.83	Rms secondary volts per transformer leg
	1.41	1.41	1.41	1.41	Rms secondary volts line-to-line
Average dc output current per rectifier	0.333	0.333	0.167	0.167	Average dc output current
Rms current per rectifier, resistive load	0.587	0.579	0.409	0.293	Average dc output current
Rms current per rectifier, inductive load	0.578	0.578	0.408	0.289	Average dc output current
Percent ripple	18.3	4.2	4.2	4.2	
Ripple frequency	3	6	6	6	Line frequency
AC line power factor	0.826	0.955	0.955	0.955	
Transformer secondary rms volts per leg[2]	0.855	0.428	0.740	0.855	Average dc voltage output
Transformer secondary rms volts line-to-line	1.48	0.740	1.48 (max)	1.71 (max, no load)	Average dc voltage output
Secondary line current	0.578	0.816	0.408	0.289	Average dc output current
Transformer secondary volt-amperes	1.48	1.05	1.81	1.48	DC watts output
Primary line current	0.817	1.41	0.817	0.707	(Avg. load I x secondary leg V) / primary line V

[1] To determine the value of a parameter in any column, multiply the factor shown by the value given in this column.
[2] For inductive load or large choke input filter

- Significantly reduced filtering requirements after rectification because of the low ripple output of a polyphase rectifier.
- Improved voltage regulation when using an inductive-input filter. Output voltage *soaring* is typically 6 percent or less from full-load to no-load conditions.
- Greater choice of output voltages from a given transformer by selection of either a *delta* or *wye* configuration.

The main disadvantage of a polyphase system is its susceptibility to phase imbalance. Resulting operational problems include increased ripple at the output of the supply and uneven sharing of the load current by the transformer windings.

4.4.7 Power Supply Filter Circuits

A filter network for a high-voltage power supply typically consists of a series inductance and one or more shunt capacitances. Bleeder resistors and various circuit protection devices are usually incorporated as well. Filter systems can be divided into two basic types:

- *Inductive input*: filter circuits that present a series inductance to the rectifier output
- *Capacitive input*: filter circuits that present a shunt capacitance to the rectifier output

Inductive Input Filter

The inductive input filter is the most common configuration found in high-power RF equipment. A common circuit is shown in Figure 4.53, along with typical current waveforms. When the input inductance is infinite, current through the inductance is constant and is carried at any moment by the rectifier anode that has the most positive voltage applied to it at that instant. As the alternating voltage being rectified passes through zero, the current suddenly transfers from one anode to another, producing square current waves through the individual rectifier devices.

When the input inductance is finite (but not too small), the situation changes to that shown by the solid lines of Figure 4.53. The current through the input inductance tends to increase when the output voltage of the rectifier exceeds the average or dc value, and to decrease when the rectifier output voltage is less than the dc value. This causes the current through the individual anodes to be modified as shown. If the input inductance is too small, the current decreases to zero during a portion of the time between the peaks of the rectifier output voltage. The conditions then correspond to a capacitor input filter system.

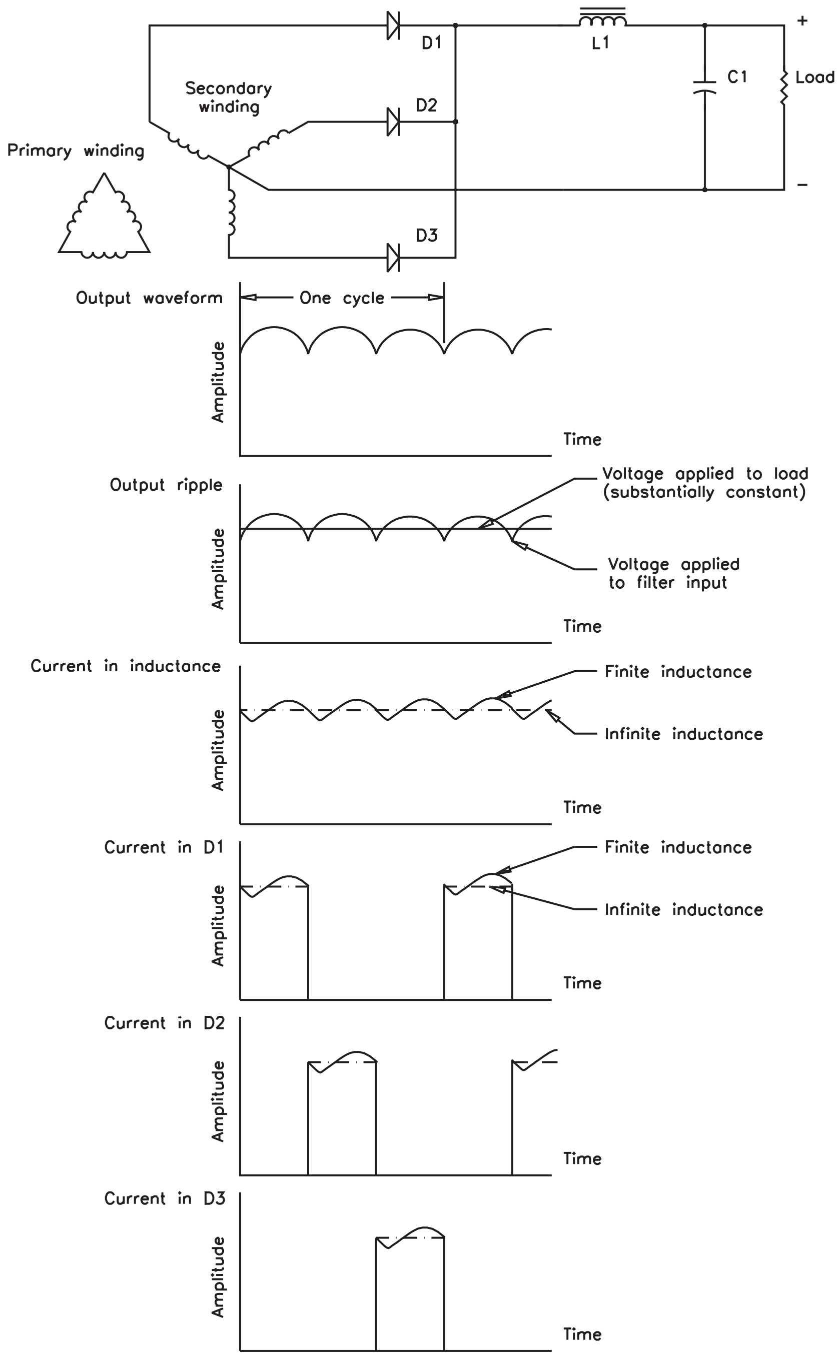


Figure 4.53 Voltage and current waveshapes for an inductive-input filter driven by a 3-phase supply.

The output wave of the rectifier can be considered as consisting of a dc component upon which are superimposed ac voltages (ripple voltages). To a first approximation, the fluctuation in output current resulting from a finite input inductance can be considered as the current resulting from the lowest-frequency component of the ripple voltage acting against the impedance of the input inductance. This assumption is permissible because the higher-frequency components in the ripple voltage are smaller, and at the same time encounter higher impedance. Furthermore, in practical filters the shunting capacitor following the input inductance has a small impedance at the ripple frequency compared with the reactance of the input inductance. The peak current resulting from a finite input inductance, therefore, is given approximately by the relation [1]:

$$\frac{I_f}{I_i} = 1 + \frac{E_1}{E_0}\frac{E_{eff}}{\omega L_1} \tag{4.15}$$

Where:
I_f = peak current with finite input inductance
I_i = peak current with infinite input inductance
E_1/E_0 = ratio of lowest-frequency ripple component to the dc voltage in the rectifier output
R_{eff} = effective load resistance
ωL_1 = reactance of the incremental value of the input inductance at the *lowest* ripple frequency

This equation is derived as follows:

- The peak alternating current through the input inductance is approximately $E_1/\omega L_1$
- The average or dc current is E_0/R_{eff}
- The peak current with finite inductance is, therefore: $(E_1/\omega L_1) + (E_0/R_{eff})$
- The current with infinite inductance is E_0/R_{eff}

The effective load resistance value consists of the actual load resistance plus filter resistance plus equivalent diode and transformer resistances.

The normal operation of an inductive input filter requires that there be a continuous flow of current through the input inductance. The peak alternating current flowing through the input inductance, therefore, must be less than the dc output current of the rectifier. This condition is realized by satisfying the approximate relation:

$$\omega L_1 = R_{eff}\frac{E_1}{E_0} \tag{4.16}$$

In the practical case of a 60 Hz single-phase full-wave rectifier circuit, the foregoing equation becomes:

$$L_1 = \frac{R_{eff}}{1130} \tag{4.17}$$

In a polyphase system, the required value of L_1 is significantly less. The higher the load resistance (the lower the dc load current), the more difficult it is to maintain a continuous flow of current and, with a given L_1, the previous equation will not be satisfied when the load resistance exceeds a critical value.

The minimum allowable input inductance (ωL_1) is termed the *critical inductance*. When the inductance is less than the critical value, the filter acts as a capacitor input circuit. When the dc drawn from the rectifiers varies, it is still necessary to satisfy the ωL_1 equation at all times, particularly if good voltage regulation is to be maintained. To fulfill this requirement at small load currents without excessive inductance, it is necessary to place a bleeder resistance across the output of the filter system in order to limit R_{eff} to a value corresponding to a reasonable value of L_1.

Capacitive Input Filter

When a shunt capacitance rather than a series inductance is presented to the output of a rectifier, the behavior of the circuit is greatly modified. Each time the positive crest alternating voltage of the transformer is applied to one of the rectifier anodes, the input capacitor charges up to just slightly less than this peak voltage. The rectifier then ceases to deliver current to the filter until another anode approaches its peak positive potential, when the capacitor is charged again. During the interval when the voltage across the input capacitor is greater than the potential of any of the anodes, the voltage across the input capacitor drops off nearly linearly with time, because the first filter inductance draws a substantially constant current from the input capacitor. A typical set of voltage and current waves is illustrated in Figure 4.54.

The addition of a shunt capacitor to the input of a filter produces fundamental changes in behavior, including the following:

- The output voltage is appreciably higher than with an inductance input.
- The ripple voltage is less with a capacitance input filter than with an inductance input filter.
- The dc voltage across the filter input drops as the load current increases for the capacitive input case, instead of being substantially constant as for the inductive input case.
- The ratio of peak-to-average anode current at the rectifiers is higher in the capacitive case.
- The utilization factor of the transformer is less with a capacitive input configuration.

Filters incorporating shunt capacitor inputs generally are employed when the amount of dc power required is small. Inductance input filters are used when the amount of power involved is large; the higher utilization factor and lower peak current

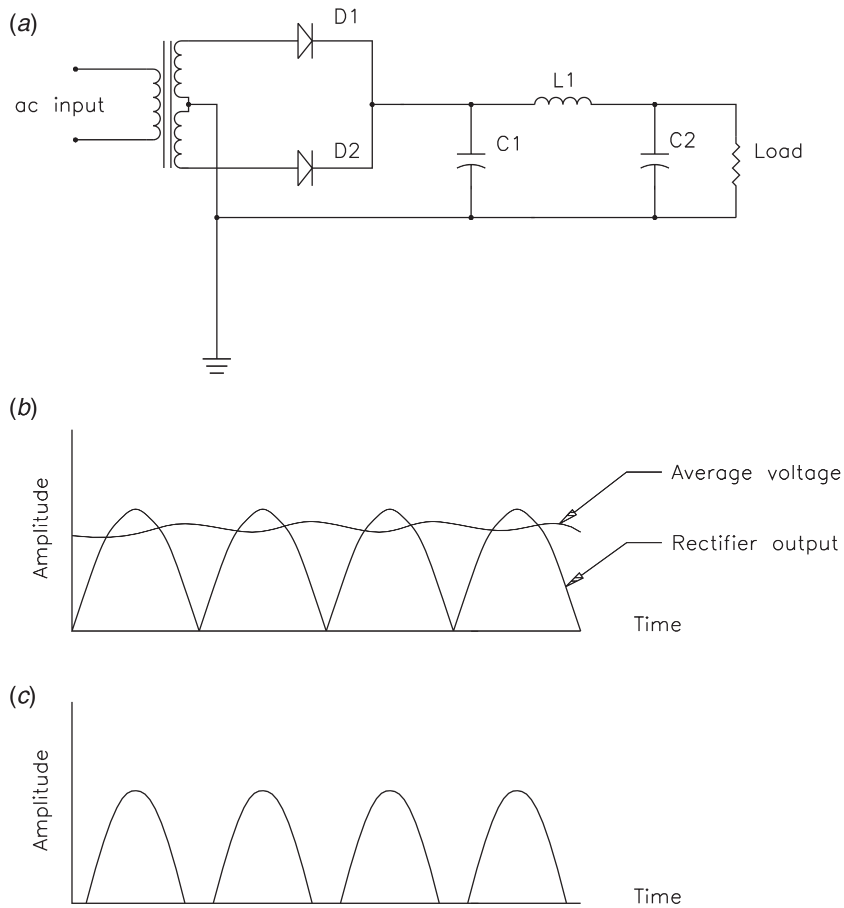


Figure 4.54 Characteristics of a capacitive input filter circuit: (*a*) schematic diagram, (*b*) voltage waveshape across the input capacitor, (*c*) waveshape of current flowing through the diodes.

result in important savings in rectifier and transformer costs under these conditions. Inductance input systems are almost universally employed in polyphase rectifiers.

4.5 Parameter Sampling Circuits

Most RF system parameters can be reduced to a *sample voltage* by using voltage dividers or current-sense resistors in the circuit to be measured [2]. Remote metering is accomplished by sending a signal representing the sample voltage to the remote metering point, and then displaying a value representing the original parameter. In practice, the actual sampling ratio (sample voltage/parameter value) is not particularly important as long as the sampling ratio is stable and the sample voltage is reasonable

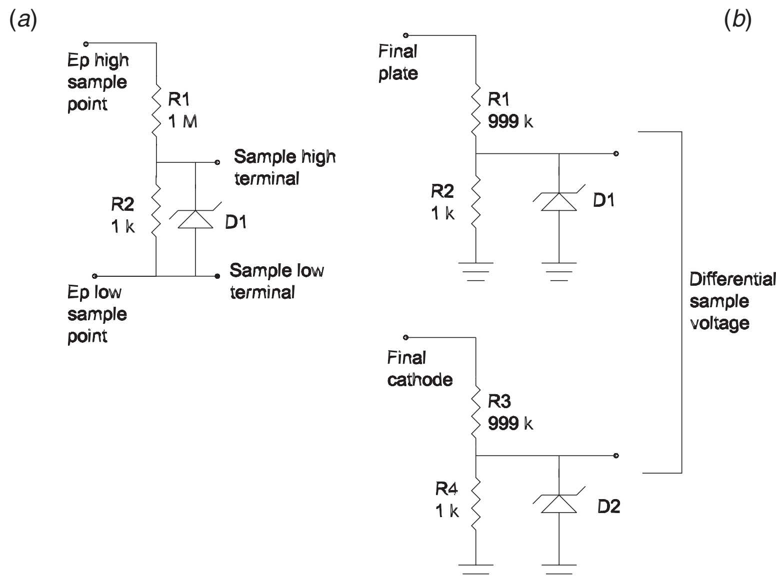


Figure 4.55 Plate voltage sampling circuits: (*a*) single-ended, (*b*) differential.

(i.e., low enough to be easily handled, yet high enough to be immune to noise). Figure 4.55 shows some typical voltage sampling circuits for high-power RF systems.

Figure 4.55*a* shows how a voltage divider can be used to sample plate voltage. The circuit is independent of the transmitter front panel plate voltage meter. A zener clamping diode is placed across the sample line for safety; if the shunt resistor opened, the full plate voltage would be present at the sampling output terminals (assuming minimal loading by the remote indicator). The zener will typically fail in a short-circuit mode, protecting the external sampling equipment.

The sampling circuit of Figure 4.55*a* is adequate for systems that measure plate voltage from a ground reference. In the event that the cathode operates at a potential other than ground, the sampling circuit of Figure 4.55*b* may be used. The sampling line is balanced, with both terminals floating above ground by 1 kΩ The differential output voltage is independent of the common mode input voltage if all the resistors are precisely matched.

Several techniques are available for dealing with differential sampling voltages (where neither side of the sample is grounded). Typical approaches include the following:

- *Isolation amplifier*. Isolation amplifiers provide isolation between the input and output by using magnetic, capacitive, optical, or thermal coupling between the in-

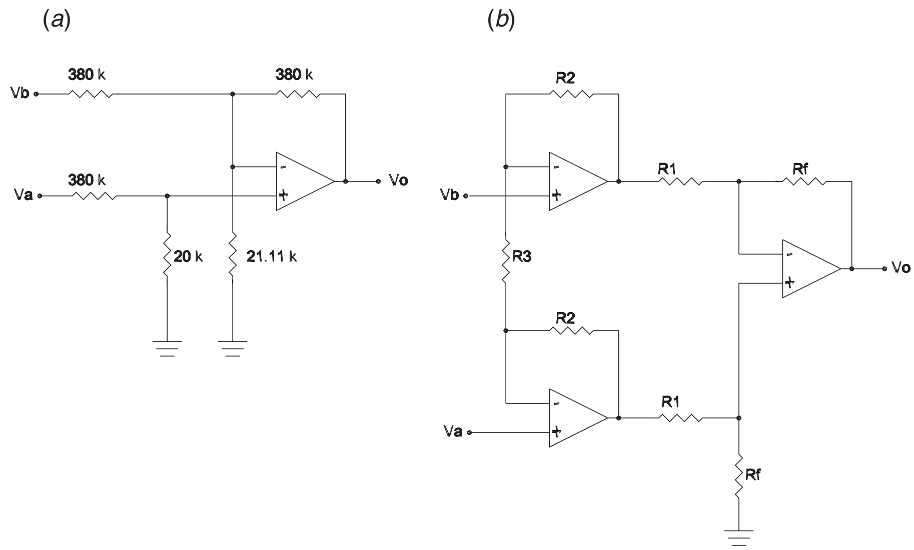


Figure 4.56 Differential amplifier circuits: (*a*) basic circuit, (*b*) instrumentation amplifier.

put and output. Some units include feedback techniques to reduce nonlinearities induced by the coupling method.

- *Differential amplifier*. In most cases, electrical isolation between the sampling circuit and the remote metering equipment is not required. Instead, a differential voltage must be changed to a *single ended* voltage (one side grounded). This can be accomplished using the differential amplifier shown in Figure 4.56*a* or the instrumentation amplifier shown in Figure 4.56*b*.

Measurement of plate and/or screen current is typically done by inserting a sampling resistor into the ground return lead of the respective supplies. This approach is shown in Figure 4.57. Notice that the remote plate current metering is taken across R1, with balancing resistors of 1.2 kΩused to lift the sample above the ground reference. A bidirectional zener diode is placed across the output of the sample terminals for protection. The plate current meter on the transmitter is sampled across R2, and the over-current relay is sampled across R3. Device X1 is a crowbar that will shunt the plate current to ground in the event of a failure in the R1–R3 chain.

The initial tolerance of resistors used as current sense elements or in a voltage divider circuit is usually not particularly critical. An inexact resistance will cause the sample voltage to be something other than that expected. However, whatever sample voltage appears, the remote meter can be calibrated to agree with the local meter. The more important attribute is resistance changes with time and temperature.

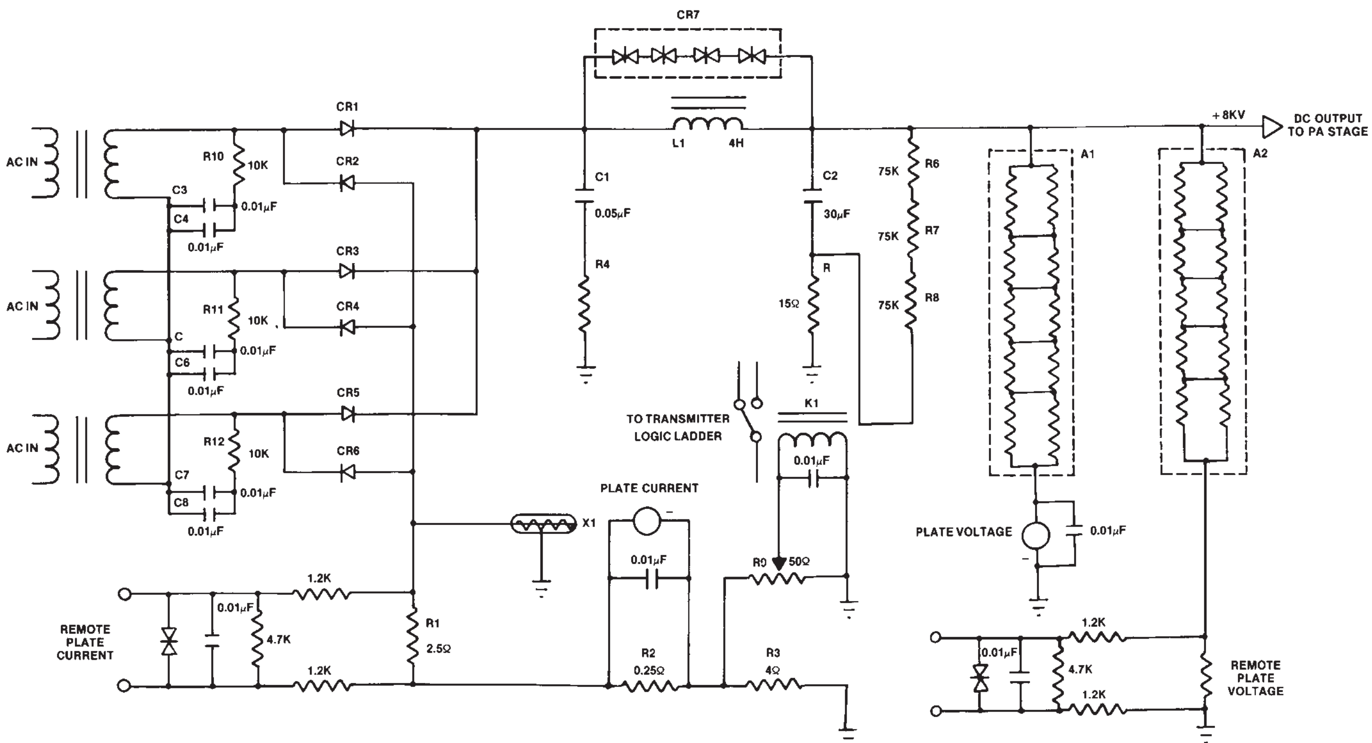


Figure 4.57 Transmitter high-voltage power supply showing remote plate current and voltage sensing circuits.

A carbon composition resistor can commonly have a temperature coefficient of about ±600 ppm/C° for resistors up to 1 KΩ. This increases to ±1875 ppm/C° for resistors up to 1 MΩ. If these resistors were used in a common plate voltage sampling circuit, such as than shown in Figure 4.57, the sample voltage could change as much as 20 percent over a 0 C° to 50 C° temperature range. Metal film resistors are available with temperature coefficients as low as 15 ppm/C°. Bulk metal resistors are available with temperature coefficients of 5 ppm/C°. If the resistors in a voltage divider are at the same temperature, the matching between the *temperature coefficient of resistance* (TCR) is more important than the actual TCR. Because the TCR matching is generally better than the TCR of an individual resistor, sampling systems should be designed to keep the resistors in the network at the same operating temperature.

In differential amplifiers, the matching of resistor values is critical. To insure the resistance values track with temperature, it is common practice to use a resistor network made using hybrid film techniques or integrated circuit techniques. Because the resistors are made at the same time and of the same materials, the TCRs track closely.

4.6 References

1. Terman, F. E., *Radio Engineering*, 3rd. Ed., McGraw-Hill, New York, pg. 560, 1947.
2. Hallikainen, Harold, "Transmitter Control Systems," in *NAB Engineering Handbook*, 9th Ed., Jerry C. Whitaker (ed.), National Association of Broadcasters, Washington, D.C., 1998.

4.7 Bibliography

Crutchfield, E. B. (ed.), *NAB Engineering Handbook*, 8th Ed., National Association of Broadcasters, Washington, DC, 1992.

Gray, T. S., *Applied Electronics*, Massachusetts Institute of Technology, 1954.

Heising, R. A., "Modulation in Radio Telephony," *Proceedings of the IRE*, Vol. 9, no. 3, pg. 305, June 1921.

Heising, R. A., "Transmission System," U.S. Patent no. 1,655,543, January 1928.

Jordan, Edward C. (ed.), *Reference Data for Engineers: Radio, Electronics, Computers, and Communications*, 7th Ed., Howard W. Sams, Indianapolis, IN, 1985.

Laboratory Staff, *The Care and Feeding of Power Grid Tubes*, Varian Eimac, San Carlos, CA, 1984.

Martin, T. L., Jr., *Electronic Circuits*, Prentice Hall, Englewood Cliffs, NJ, 1955.

Mendenhall, G. N., M. B. Shrestha, and E. J. Anthony, "FM Broadcast Transmitters," in *NAB Engineering Handbook*, 8th Ed., E. B. Crutchfield (ed.), National Association of Broadcasters, Washington, DC, pp. 429–474, 1992.

Shrestha, Mukunda B., "Design of Tube Power Amplifiers for Optimum FM Transmitter Performance," *IEEE Transactions on Broadcasting*, Vol. 36, no. 1, IEEE, New York, pp. 46–64, March 1990.

Terman, F. E., and J. R. Woodyard, "A High Efficiency Grid-Modulated Amplifier," *Proceedings of the IRE*, Vol. 26, no. 8, pg. 929, August 1938.

Terman, F. E., *Radio Engineering*, 3rd. Ed., McGraw-Hill, New York, 1947.

Weirather, R. R., G. L. Best, and R. Plonka, "TV Transmitters," in *NAB Engineering Handbook*, 8th Ed., E. B. Crutchfield (ed.), National Association of Broadcasters, Washington, DC, pp. 499–551, 1992.

Whitaker, Jerry C., *AC Power Systems*, 2nd Ed., CRC Press, Boca Raton, FL, 1998.

Whitaker, Jerry C., *Radio Frequency Transmission Systems: Design and Operation*, McGraw-Hill, New York, 1991.

Woodard, George W., "AM Transmitters," in *NAB Engineering Handbook*, 8th Ed., E. B. Crutchfield (ed.), National Association of Broadcasters, Washington, DC, pp. 353–381, 1992.