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Chapter

5

Applying Vacuum Tube Devices

5.1 Introduction

Numerous circuits have been developed to meet the varied needs of high-power RF amplification. Designers have adapted standard vacuum tube devices to specific applications that require unique performance criteria. Most of these circuits are built upon a few basic concepts.

5.2 AM Power Amplification Systems

Class C plate modulation is the classic method of producing an AM waveform [1]. Figure 5.1 shows the basic circuit. Triodes and tetrodes may be used as the modulator or carrier tube. Triodes offer the simplest and most common configuration.

Numerous variations on the basic design exist, including a combination of plate and screen grid modulation. The carrier signal is applied to the control grid, and the modulating signal is applied to the screen and plate. The plate is fully modulated, and the screen is modulated 70 to 100 percent to achieve 100 percent carrier modulation. Modulation of the screen can be accomplished using one of the following methods:

- Screen voltage supplied through a dropping resistor connected to the unmodulated dc plate supply, shown in Figure 5.2*a*
- An additional (third) winding on the modulation transformer, illustrated in Figure 5.2*b*
- A modulation choke placed in series with a low-voltage fixed screen supply, shown in Figure 5.2*c*

Depending on the design, screen/plate modulation also may require partial modulation of the control grid to achieve the desired performance characteristics.

During the portion of the modulation cycle when the plate voltage is increased, the screen current decreases. If the screen is supplied through an impedance, such as the screen dropping resistance of a modulation choke, the voltage drop in this series impedance becomes less and the screen voltage rises in the desired manner. During the part of the modulation cycle when the plate voltage is decreased, the screen current increases,

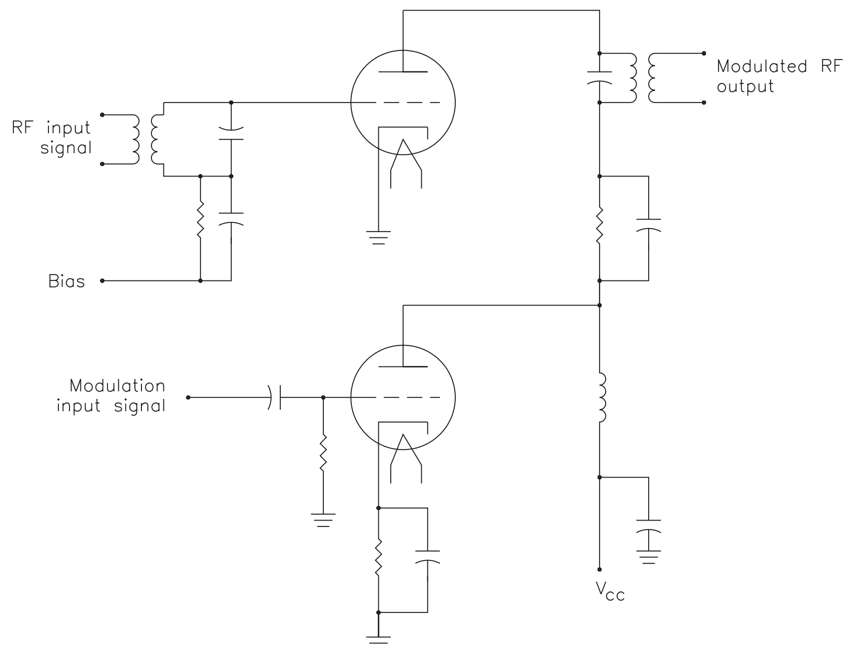

Figure 5.1 The classic plate-modulated PA circuit for AM applications.

causing a greater voltage drop in the screen series impedance, and lowering the voltage on the screen of the tube. The value of the screen bypass capacitor in the class C stage is a compromise between good RF bypassing and the shunting effect of this capacitance on the screen modulation circuit.

5.2.1 Control Grid Modulation

Although not as common as class C plate modulation, a class C RF amplifier may be modulated by varying the voltage on the control grid of a triode [1]. This approach, shown in Figure 5.3, produces a change in the magnitude of the plate current pulses and, therefore, variations in the output waveform taken from the plate tank. Both the carrier and modulation signals are applied to the grid.

The primary benefit of grid modulation is that a modulator is not required. Grid modulation, however, requires a fixed plate supply voltage that is twice the peak RF voltage without modulation. The result is higher plate dissipation at lower modulation levels, including carrier level. The typical plate efficiency of a grid-modulated stage may range from only 35 to 45 percent at carrier. Grid modulation typically is used only in systems where plate modulation transformers cannot provide adequate bandwidth for the intended application.

When grid modulation is used, the screen voltage and grid bias must be taken from sources with good regulation. This usually means separate low-voltage power supplies.

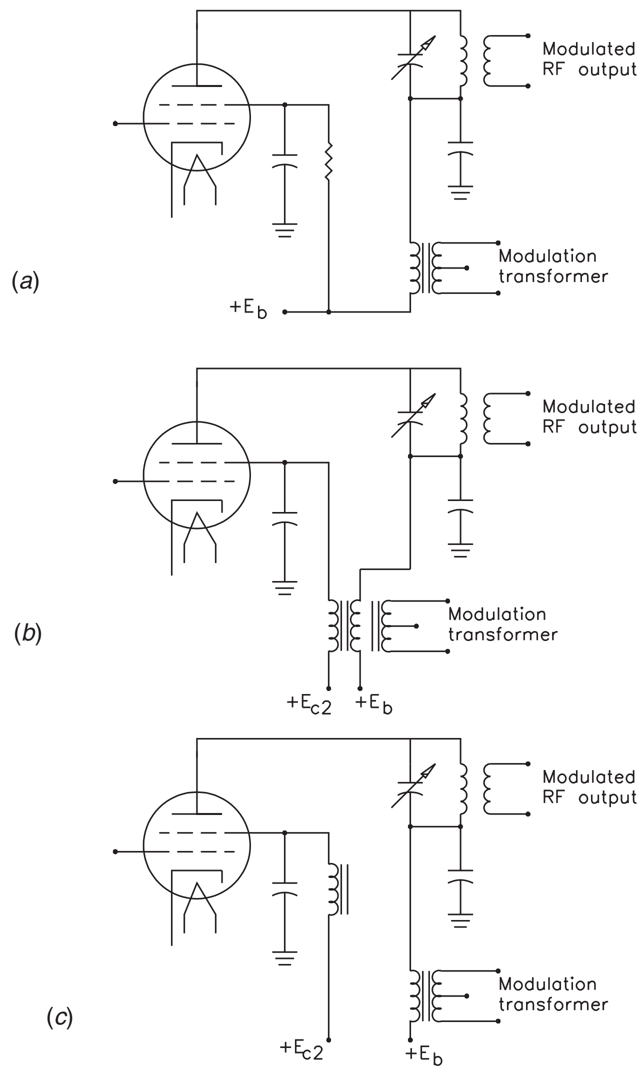

Figure 5.2 Basic methods of screen and plate modulation: (*a*) plate modulation, (*b*) plate and screen modulation, (*c*) plate modulation using separate plate/screen supplies.

5.2.2 Suppressor Grid Modulation

The output of a class C pentode amplifier may be controlled by applying a modulating voltage, superimposed on a suitable bias, to the suppressor grid to produce an AM waveform [1]. As the suppressor grid becomes more negative, the minimum instanta-

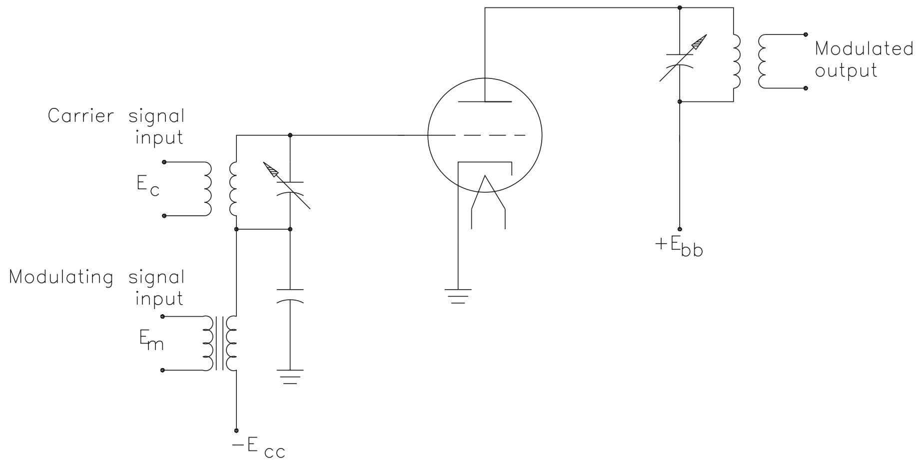

Figure 5.3 Grid modulation of a class C RF amplifier.

neous potential at which current can be drawn to the plate is increased. Thus, as modulation varies the suppressor grid potential, the output voltage changes.

This method of modulating an RF amplifier provides about the same plate efficiency as a grid-modulated stage. The overall operating efficiency, however, is slightly lower because of increased screen grid losses associated with the design. Linearity of the circuit is not usually high.

5.2.3 Cathode Modulation

Cathode modulation incorporates the principles of both control grid and plate modulation [1]. As shown in Figure 5.4, a modulation transformer in the cathode circuit varies the grid-cathode potential, as well as the plate-cathode potential. The ratio of grid modulation to plate modulation is set by adjustment of the tap shown in the figure. Grid leak bias typically is employed to improve linearity of the stage.

5.2.4 High-Level AM Amplification

The basic tuned-anode vacuum tube power amplifier is described in graphical form in Figure 5.5 [2]. The tube may be a triode, tetrode, or pentode. Tetrode final amplifiers are the most common. The RF excitation voltage is applied to the grid of the PA tube, and the ratio of dc grid bias voltage to peak RF excitation voltage (shown sinusoidal in Figure 5.5) determines the conduction angle θ_c of anode current:

$$\theta_c = 2 \arccos \left\{ \frac{E_{cc}}{E_g - E_{cc}} \right\} \tag{5.1}$$

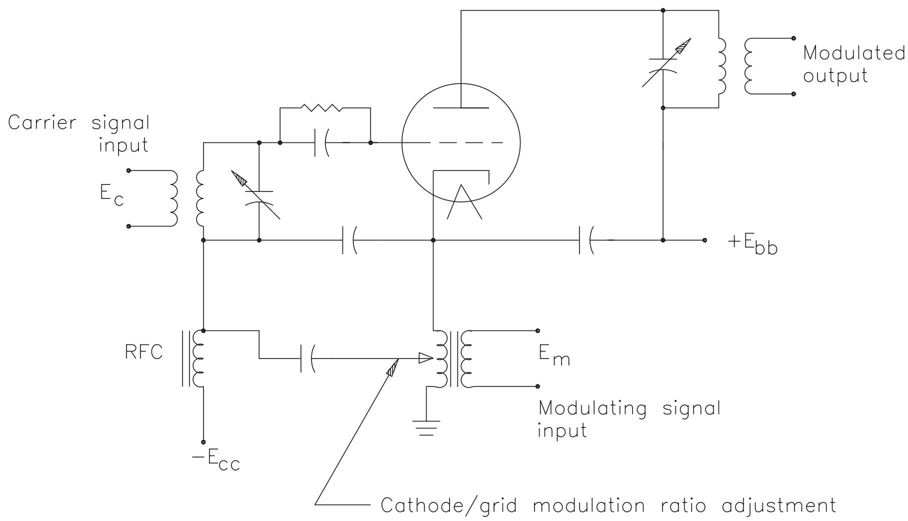

Figure 5.4 Cathode modulation of a class C RF amplifier.

Where:
E_g = exciting grid signal
E_{cc} = dc grid bias voltage

The shape of the anode current pulse is determined by the vacuum tube transfer characteristics and the input waveshape. The pulse of current thus generated is supplied by the dc power supply E_{bb} and passes through the resonant anode tank circuit. The tank is assumed to have sufficient operating Q to force anode voltage e_p to be essentially sinusoidal and of the same periodic frequency as the RF excitation voltage and resultant anode current pulse. The instantaneous anode dissipation is the product of instantaneous tube anode voltage drop and anode current. The tube transfer characteristic is a variable, dependent upon device geometry and other factors, including the maximum drive signal E_g. The exact shape and magnitude of the current waveform is normally obtained from a load-line plot on the constant-current characteristic tube curves supplied by the device manufacturer. The resonant anode load impedance is chosen and adjusted to allow $E_{p(min)}$ to be as low as possible without causing excessive screen grid (tetrode case) or control grid dissipation.

The overall anode efficiency of this circuit can be extended beyond the normal limits for typical class C amplification by employing a third-harmonic resonator between the output anode connection and the fundamental resonant circuit. In some cases, fifth-harmonic resonators are also employed. This technique has the effect of squaring up the anode voltage waveform e_p, thus causing the integral of the $e_p \times i_p$ product (anode dissipation) to be smaller, resulting in lower dissipation for a given RF power output. An amplifier employing the third-harmonic anode trap commonly is referred to as a class C-D stage, suggesting efficiency ranging between conventional class C operation

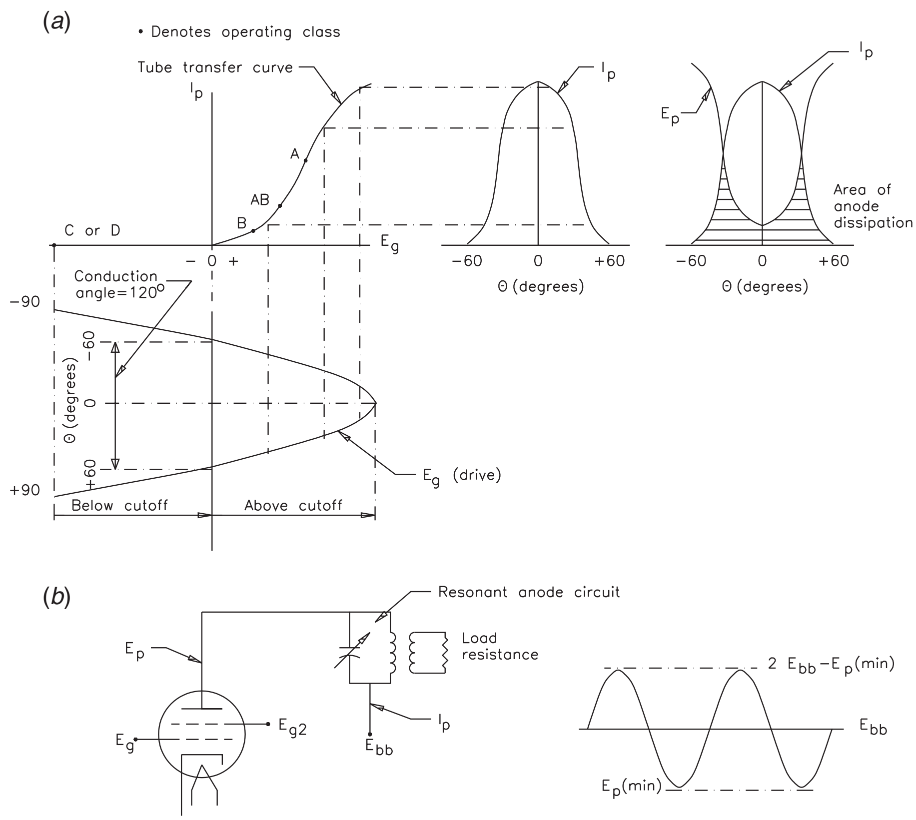

Figure 5.5 Classic vacuum tube class C amplifier: (*a*) representative waveforms, (*b*) schematic diagram. Conditions are as follows: sinusoidal grid drive, 120° anode current, resonant anode load. (*After* [2].)

and class D operation. The use of the third-harmonic technique permits efficiencies of 85 percent or more for transmitters of 10 kW power output and above.

Class B Modulator

The most common method of applying low-frequency (typically audio) intelligence to a high-level amplitude-modulated amplifier is the class B push-pull system illustrated in Figure 5.6 [1]. The vacuum tubes used in such a circuit may be triodes, tetrodes, or pentodes. The output circuit includes a modulation transformer, audio coupling capacitor, and dc shunt feed inductor. The capacitor and shunt inductor network is used to prevent unbalanced dc from magnetizing the modulation transformer core, which would result in poor low-frequency performance. Advanced core materials and improved transformer design have permitted elimination of the coupling ca-

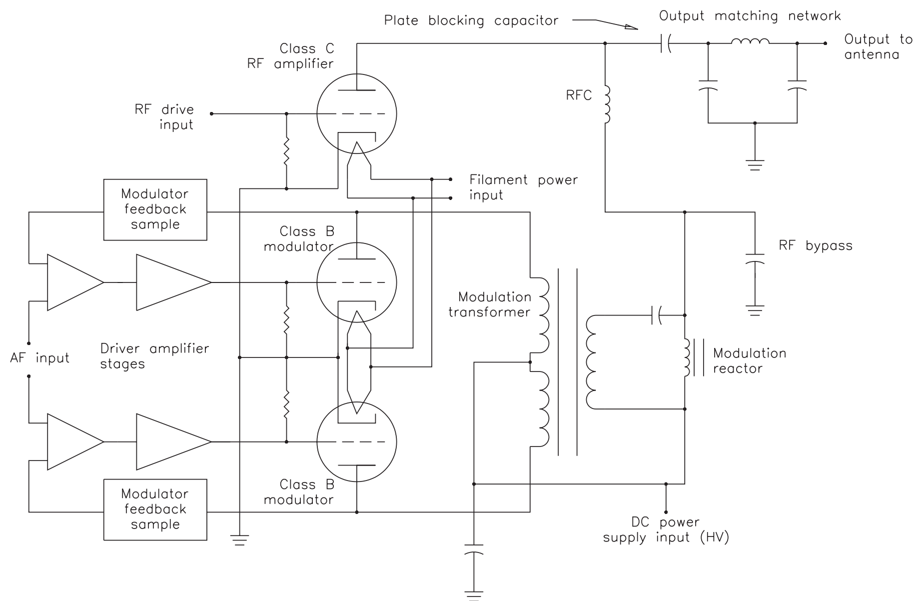

Figure 5.6 Class B push-pull modulator stage for a high-level AM amplifier.

pacitor and the shunt reactor in modern systems. The direct current to the modulated RF amplifier anode, therefore, flows directly through the secondary of the modulation transformer.

Elimination of the coupling capacitor and shunt reactor results in improved performance from the modulator circuit. The *L* and *C* components effectively form a three-pole high-pass filter that causes low-frequency transient distortion to be generated when driven with complex audio waveforms. The simplified circuit constitutes a single-pole high-pass filter, greatly reducing low-frequency transients.

Even in the improved design, the transient distortion performance of the class B push-pull modulator can be insufficient for some applications. Complex modulating waveforms with sharp transitions present special problems. Stray internal winding capacitances and leakage inductances effectively form a multipole low-pass filter at the high-frequency end of the audio spectrum. This filter can produce overshoots when driven with complex waveforms. Such distortion can be reduced by filtering the input signal, at the cost of somewhat lower high-frequency response.

Negative feedback is used to reduce nonlinear distortion and noise in the push-pull circuit. Negative feedback, however, usually increases high frequency transient distortion.

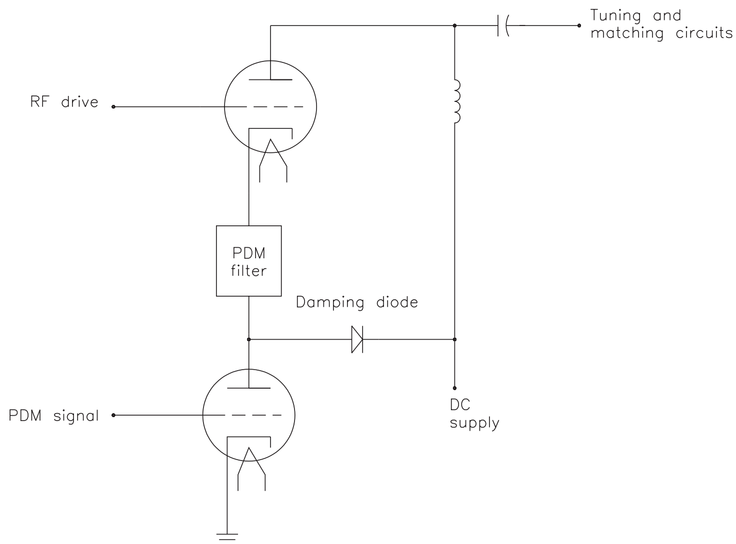

Figure 5.7 The pulse duration modulation (PDM) method of pulse width modulation.

5.2.5 Pulse Width Modulation

Pulse width modulation (PWM), also known as *pulse duration modulation* (PDM), is a common transmission system for modern vacuum tube transmitters. Figure 5.7 shows the PDM scheme (Harris) applied to an amplitude-modulated PA. The PDM system works by utilizing a square wave switching system, illustrated in Figure 5.8.

The PDM process begins with a signal generator (Figure 5.9). A 75 kHz sine wave is produced by an oscillator and used to drive a square wave generator, resulting in a simple 75 kHz square wave. The square wave is then integrated, resulting in a triangular waveform that is mixed with the input audio in a summing circuit. The resulting signal is a triangular waveform that rides on the incoming signal, typically audio. This composite signal is then applied to a threshold amplifier, which functions as a switch that is turned on whenever the value of the input signal exceeds a certain limit. The result is a string of pulses in which the width of the pulse is proportional to the period of time the triangular waveform exceeds the threshold. The pulse output is applied to an amplifier to obtain the necessary power to drive subsequent stages. A filter eliminates whatever transients may exist after the switching process is complete.

The PDM scheme is, in effect, a digital modulation system with the input information being sampled at a 75 kHz rate. The width of the pulses contains all the intelligence. The pulse-width-modulated signal is applied to a *switch* or *modulator tube*. The tube is simply turned *on*, to a fully saturated state, or *off* in accordance with the instanta-

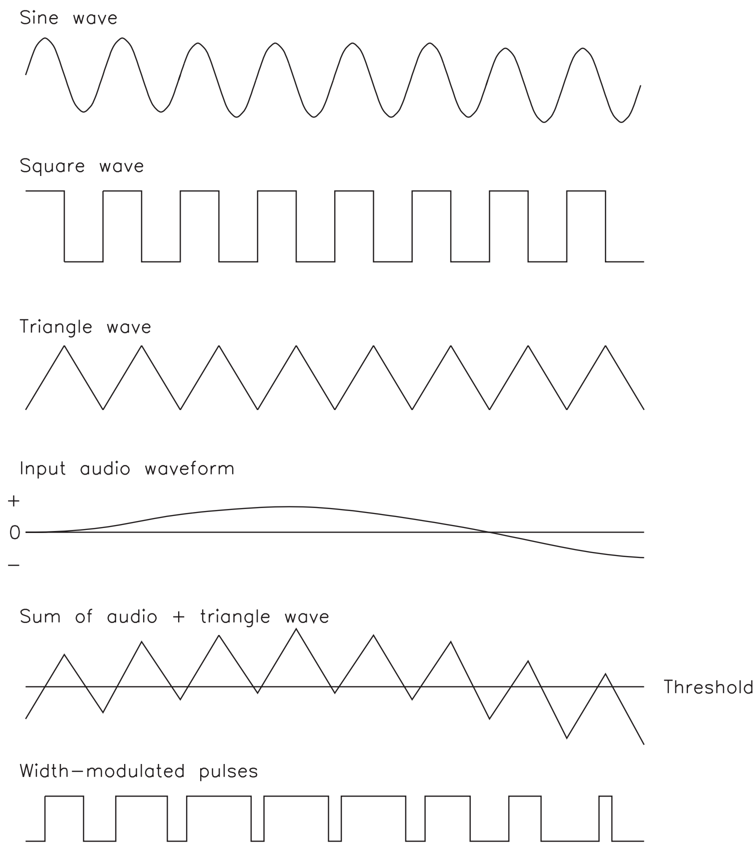

Figure 5.8 The principal waveforms of the PDM system.

neous value of the pulse. When the pulse goes positive, the modulator tube is turned on and the voltage across the tube drops to a minimum. When the pulse returns to its minimum value, the modulator tube turns off.

This PDM signal becomes the power supply to the final RF amplifier tube. When the modulator is switched on, the final amplifier will experience current flow and RF will be generated. When the switch or modulator tube goes off, the final amplifier current will cease. This system causes the final amplifier to operate in a highly efficient class D switching mode. A dc offset voltage to the summing amplifier is used to set the carrier (no modulation) level of the transmitter.

A high degree of third-harmonic energy will exist at the output of the final amplifier because of the switching-mode operation. This energy is eliminated by a third-harmonic trap. The result is a stable amplifier that normally operates in excess of 90 percent efficiency. The power consumed by the modulator and its driver is usually a fraction of a full class B amplifier stage.

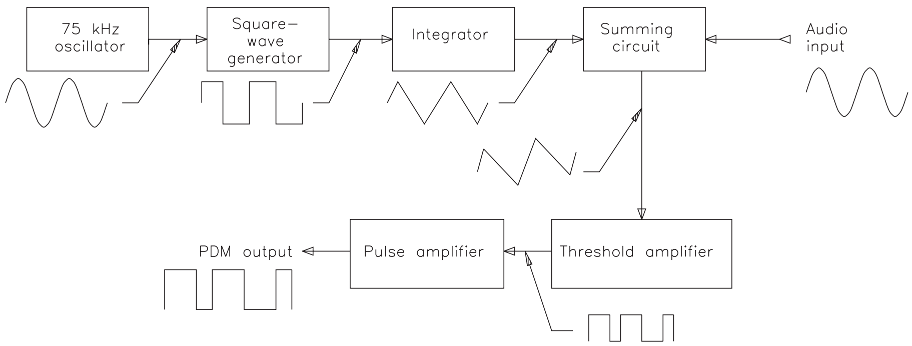

Figure 5.9 Block diagram of a PDM waveform generator.

The *damping diode* shown in Figure 5.9 is included to prevent potentially damaging transient overvoltages during the switching process. When the switching tube turns off the supply current during a period when the final amplifier is conducting, the high current through the inductors contained in the PDM filters could cause a large transient voltage to be generated. The energy in the PDM filter is returned to the power supply by the damping diode. If no alternative route were established, the energy would return by arcing through the modulator tube itself.

The PWM system makes it possible to eliminate audio frequency transformers in the transmitter. The result is wide frequency response and low distortion. Note that variations on this amplifier and modulation scheme have been used by other manufacturers for standard broadcast, shortwave service, and other applications.

5.3 Linear Amplification

The following features are desirable for vacuum tubes used in RF linear amplifier service [1]:

- High power gain
- Low plate-to-grid capacitance
- Good efficiency
- Linear characteristics that are maintained without degradation across the desired operating band

For linear service, RF amplifiers may be operated in class A, AB_1, AB_2, or B modes. Triode, tetrode, or pentode devices may be used, either grid- or cathode-driven. The choice of mode, tube, and driving method depends upon the operational specifications of each application.

5.3.1 Device Selection

The triode tube, having a large plate-to-grid interelectrode capacitance, requires neutralization in grid-driven linear service to prevent oscillation [1]. A triode having a low amplification factor is suitable for class AB and AB_2 grid-driven operation. The RF grid excitation voltage for this type of service will be high; grid excursions into the positive region are normal for class AB_2. A swamping resistor should be used across the input tuned circuit to maintain a constant input impedance to the stage and to provide for stability. With a low value of swamping resistance, the grid current drawn is only a small part of the total grid load, and the driver load impedance is relatively constant. The swamping resistor improves RF stability by providing a low impedance to ground for regenerative feedback through the plate-to-grid capacitance.

High-amplification-factor triodes perform exceptionally well in circuits where the grid is grounded and the cathode is driven. Under these conditions, the control grid acts as a shield between the input and output circuits. Neutralization, therefore, is not normally required. Zero-bias triodes operate in the class AB_2 mode and require only filament, plate, and drive power. For optimum linear operation, a tuned circuit is placed in the cathode RF return path to maintain a sinusoidal waveshape over the drive cycle. The tuned circuit will reduce the intermodulation distortion produced by the amplifier and also will reduce drive power requirements.

If the driver and PA stages are located close to each other, the tuned cathode circuit can be a part of the output circuit of the driver. If, however, the amplifiers are far removed and coupled by a length of coaxial cable, it is recommended that a tuned cathode circuit with a Q of between 2 and 4 be used.

Most tetrode and pentode amplifiers are designed to be grid-driven to take advantage of the high power gain of these devices.

In all linear amplifier systems, the driver output impedance should be kept low because of the nonlinear input loading characteristics of the amplifier tube as it approaches maximum power. The lower the driver amplifier impedance, the smaller the effect of the nonlinear input loading.

5.3.2 Grid-Driven Linear Amplifier

A linear amplifier utilizing a tetrode or pentode is usually grid-driven to take advantage of the inherent high gain of the tube [1]. Such an amplifier can be driven into the grid-current region under the proper circumstances. In any case, the input circuit will be loaded by the tube grid. For the no-grid-current case, the driver will work into the input conductance loading; for the grid-current case, the driver will work into the input conductance loading plus grid-current loading. It is therefore desirable (and necessary in the grid-current case) to swamp the input circuit with an appropriate noninductive resistor. The resistor will maintain an almost constant load on the driver and minimize the effects of any nonlinearity in grid loading.

5.3.3 Cathode-Driven Linear Amplifier

The cathode-driven amplifier may use triode, tetrode, or pentode tubes [1]. The drive signal is applied to the cathode in this class of operation. The cathode-driven amplifier is particularly suitable for high-power stages using high-μ triodes in the HF and VHF region. This class of operation normally eliminates the need for neutralization because the control grid screens the plate from the input circuit. The power gain for suitable triode class AB cathode-driven amplifiers is on the order of 7 to 20. The actual tube power gain is approximately equal to the ratio of RF plate voltage to RF cathode voltage. This relationship is true because the fundamental component of the plate current is common to the input and output circuits.

Tetrode tubes also can be used in cathode-driven operation. Power gain is considerably higher than that of the triode, on the order of 20 to 50. It is important to note that screen grid current loads the input circuit, just as control-grid current does.

For an amplifier located some distance from the driver, an improvement in intermodulation distortion can be realized by tuning the cathode circuit. When the driver is located close to the amplifier (0.1 wavelength or so) other means may be used to minimize the nonlinear loading of the cathode-driven stage.

5.3.4 Intermodulation Distortion

When an RF signal with varying amplitude is passed through a nonlinear device, many new products are produced [1]. (Intermodulation distortion is discussed in detail in Section 10.3.6.) The frequency and amplitude of each component can be determined mathematically because the nonlinear device can be represented by a power series expanded about the zero-signal operating point. The example of a typical two-tone signal serves to summarize this mathematical relationship. Assume that two equal-amplitude test signals (f_1 = 2.001 MHz, and f_2 = 2.003 MHz) are applied to a linear amplifier. The output spectrum of the amplifier is shown in Figure 5.10. Many of the distortion products fall outside the passband of the amplifier tuned circuits. If no impedance exists at the frequencies of the distortion components, then no voltage can be developed. Further study of this spectrum discloses that no even-order products fall near the two desired signals. Some odd-order products, however, fall near the desired frequencies and possibly within the passband of the tuned circuits.

The distortion products usually given in tube data sheets are the third- and fifth-order intermodulation distortion products that can fall within the amplifier passband. Using the same f_1 and f_2 frequencies of the previous example, the frequencies of the third-order products are:

$$2f_1 - f_2 = 1.999 \text{ MHz}$$

$$2f_2 - f_1 = 2.005 \text{ Mhz}$$

The frequencies of the fifth-order products are:

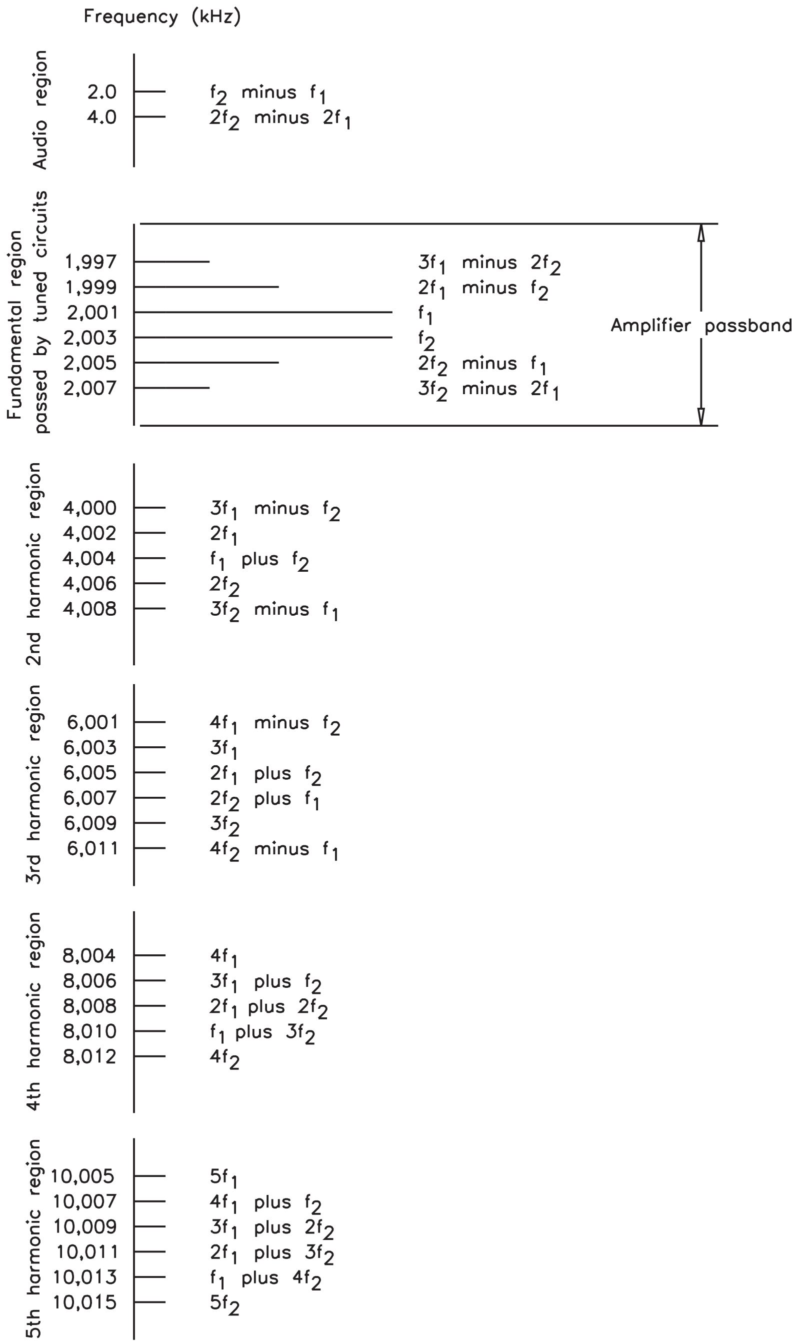

Figure 5.10 Spectrum at the output of a nonlinear device with an input of two equal-amplitude sine waves of f_1 = 2.001 MHz and f_2 = 2.003 MHz.

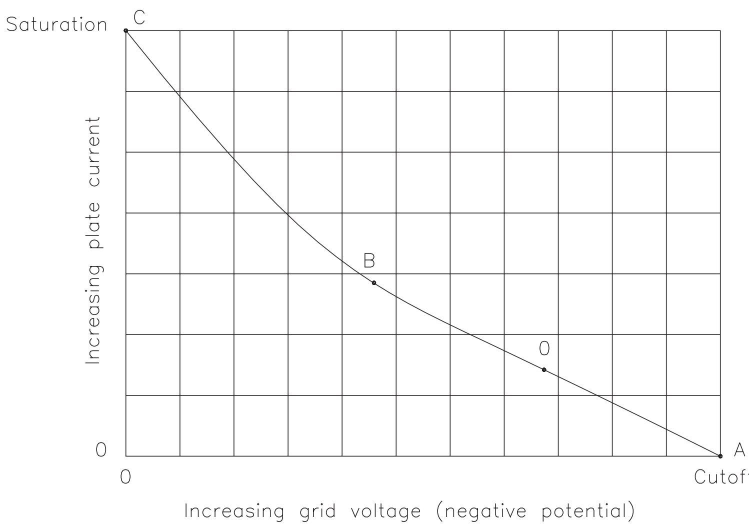

Figure 5.11 Ideal grid-plate transfer curve for class AB operation.

$$3f_1 - 2f_2 = 1.997 \text{ MHz}$$

$$3f_2 - 2f_1 = 2.007 \text{ MHz}$$

These frequencies are well within the passband of a tuned circuit intended to pass voice frequencies and, therefore, power will be delivered to the antenna at these frequencies. All intermodulation distortion power is wasted and serves no purpose other than to cause interference to adjacent channels.

Intermodulation distortion in a power amplifier tube is the result of variations of the transfer characteristics of the device from the ideal case. An ideal transfer characteristic curve is shown in Figure 5.11. Even-order products do not contribute to the intermodulation distortion problem because they fall outside the amplifier passband. Therefore, if the transfer characteristic produces an even-order curvature at the small-signal end of the curve (from point *A* to point *B*) and the remaining portion of the curve (point *B* to point *C*) is linear, the tube is considered to have an ideal transfer characteristic. If the operating point of the amplifier is set at point *0* (located midway horizontally between point *A* and point *B*), there will be no distortion in a class AB amplifier. However, no tube has this idealized transfer characteristic. It is possible, by manipulation of the electron ballistics within a given tube structure, to alter the transfer characteristic and minimize the distortion products.

5.4 High-Efficiency Linear Amplification

The class B RF amplifier is the classic means of achieving linear amplification of an AM signal. A significant efficiency penalty, however, is paid for linear operation. The plate efficiency at carrier level in a practical linear amplifier is about 33 percent, while a high-level (plate)-modulated stage can deliver a basic efficiency of 65 to 80 percent. In a real-world transmitter, power consumed by the modulator stage must also be taken into account. However, because the output power of the modulator is, at most, one-third of the total power output, the net efficiency of the plate-modulated amplifier is still higher than a conventional linear amplifier of comparable power.

Still, linear amplification is attractive to the users of high-power transmitters because low-level modulation can be employed. This eliminates the need for modulation transformers and reactors, which—at certain power levels and for particular applications—may be difficult to design and expensive to construct. To overcome the efficiency penalty of linear amplification, special systems have been designed that take advantage of the benefits of a linear mode, while not imposing an excessive efficiency burden.

5.4.1 Chireix Outphasing Modulated Amplifier

The *Chireix* amplifier employs two output stage tubes. The grids of the tubes are driven with signals whose relative phase varies with the applied modulating signal. The output waveforms of the tubes are then applied to a summing network, and finally to the load. Figure 5.12 shows a simplified schematic diagram.

Operation of the circuit is straightforward. When the grids of the tubes are driven with signals that are 180° out of phase, power output will be zero. This corresponds to a negative AM modulation peak of 100 percent. Full (100 percent) positive modulation occurs when both grids are driven with signals that are in phase. The AM modulating signal is produced by varying the relative phase of the drive signals.

The phase-modulated carrier-frequency grid waveforms are generated in a low-level balanced modulator and amplified to the level required by the PA tubes. Stability of the exciter in the Chireix system is critical. Any shift in the relative phase of the exciter outputs will translate to amplitude modulation of the transmitted waveform.

Because of the outphasing method employed to produce the AM signal, the power factor associated with the output stage is of special significance. The basic modulating scheme would produce a unity power factor at zero output. The power factor would then decrease with increasing RF output. To avoid this undesirable characteristic, the output plate circuits are detuned, one above resonance and one below, to produce a specified offset. This *dephasing*, coupled with adjustment of the relative phase of the driving signals, maintains the power factor within a reasonable range.

There are, however, two major disadvantages to this system. First, the efficiency of the amplifier is compromised because at all instantaneous levels of modulation except one, the anode circuits must work into a reactive load. Typical overall efficiency is on the order of 60 percent. Second, output carrier power adjustment is sensitive to tuning of any stage in the chain. This problem can be reduced by employing broadband amplifying circuits at lower-level stages.

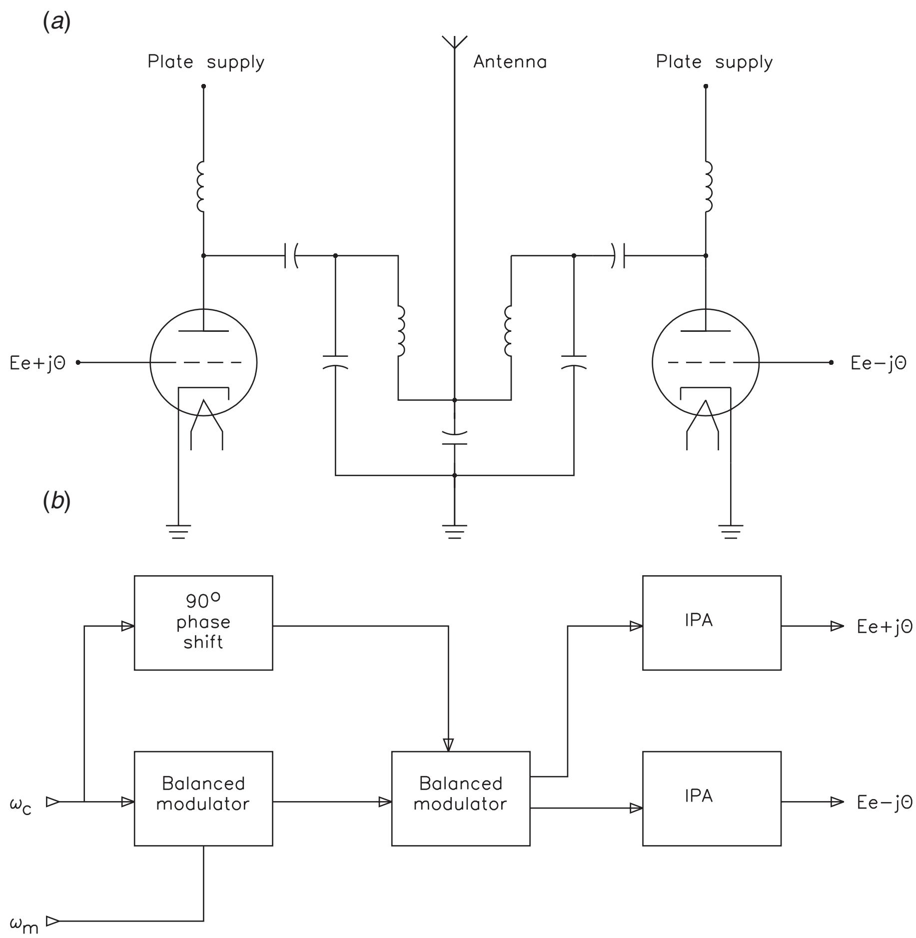

Figure 5.12 Simplified schematic diagram of the Chireix outphasing modulated amplifier: (*a*) output circuit, (*b*) drive signal generation circuit.

5.4.2 Doherty Amplifier

The *Doherty* modulated amplifier has seen considerable use in high-efficiency linear AM transmitters. The system is a true high-efficiency linear amplifier, rather than a hybrid amplifier/modulator. The Doherty amplifier employs two tubes and a 90° network as an *impedance inverter* to achieve load-line modification as a function of power level. The basic circuit is shown in Figure 5.13.

Two tubes are used in the Doherty amplifier, a *carrier tube* (V1) and a *peak tube* (V2). The carrier tube is biased class B. Loading and drive for the tube are set to provide

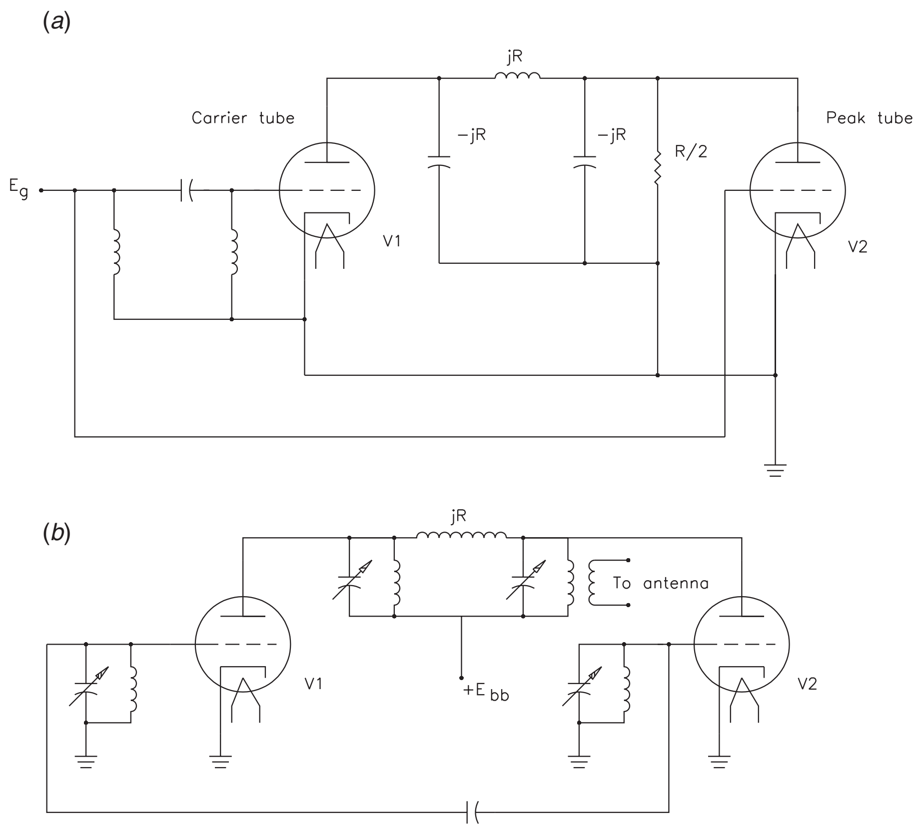

Figure 5.13 Doherty high-efficiency AM amplifier: (*a*) operating theory, (*b*) schematic diagram of a typical system.

maximum linear voltage swing at carrier level. The peak tube is biased class C; at carrier level it just begins to conduct plate current. Each tube delivers an output power equal to twice the carrier power when working into a load impedance of R Ω. At carrier, the reflected load impedance at the plate of V1 is equal to 2R Ω. This is the correct value of load impedance for carrier-level output at full plate voltage swing. (The impedance-inverting property of the 90° network is similar to a 1/4-wave transmission line.)

A 90° phase lead network is included in the grid circuit of V1 to compensate for the 90° phase lag produced by the impedance-inverting network in the plate circuit. For negative modulation swings, the carrier tube performs as a linear amplifier with a load impedance of 2R Ω; the peak tube contributes no power to the output.

On positive modulation swings, the peak tube conducts and contributes power to the R/2 load resistance. This is equivalent to connecting a negative resistance in shunt with the R/2 load resistance, so that the value seen at the load end of the 90° network increases. This increase is reflected through the network as a decrease in load resistance at the plate of V1, causing an increase in output current from the tube, and an increase in output power. The drive levels on the tubes are adjusted so that each contributes the same power (equal to twice the carrier power) at a positive modulation peak. For this condition, a load of R Ω is presented to each tube.

The key aspect of the Doherty circuit is the change in load impedance on tube V1 with modulation. This property enables the device to deliver increased output power at a constant plate voltage swing. The result is high efficiency and good linearity. Overall efficiency of the Doherty amplifier ranges from 60 to 65 percent.

A practical application of the Doherty circuit is shown in Figure 5.13*b*. The shunt reactances of the phase-shift networks are achieved by detuning the related tuned circuits. The tuned circuits in the grid are tuned above the operating frequency; those in the plate are tuned below the operating frequency.

The Doherty amplifier has two important advantages in high-power applications. First, the peak anode voltage at either tube is (approximately) only one-half that required for an equivalent carrier power high-level pulse width modulation (PWM) or class B anode-modulated transmitter, thus significantly increasing reliability and usable tube life. Second, no large modulation transformer or special filtering components are used in the final amplifier stages that might cause transient overshoot distortion.

Disadvantages of the Doherty amplifier include nonlinear distortion performance and increased complexity of tuning. The major sources of nonlinear distortion include the nonlinearity of the carrier tube at or near the 100 percent negative modulation crest, and the nonlinearity of the peak tube at or near carrier level when it is just beginning to conduct. Both sources of distortion may be effectively reduced by using moderate amounts of overall envelope feedback.

5.4.3 Screen-Modulated Doherty-Type Amplifier

A variation on the basic Doherty scheme can be found in the screen-modulated power amplifier. The design is unique in that the system does not have to function as a linear amplifier. Instead, modulation is applied to the screen grids of both the carrier and peak tubes. The peak tube is modulated upward during the positive half of the modulation cycle, and the carrier tube is modulated downward during the negative half of the cycle. This technique results in performance improvements over the classic Doherty design in that RF excitation voltages and modulating voltages are isolated from each other, thereby eliminating a troublesome source of tuning-vs.-modulation interaction.

The basic screen modulation system is shown in Figure 5.14. The peak and carrier tubes are biased and driven in quadrature as class C amplifiers from the continuous wave RF drive source. At carrier level, the screen voltage of the carrier tube is adjusted so that the carrier device is near anode saturation and delivering approximately 96 percent of the carrier power. The screen voltage of the peak tube is adjusted so that it is just into conduction, and is supplying the remaining 4 percent (approximately) of carrier power. The combined anode efficiency at carrier level is greater than 77 percent.

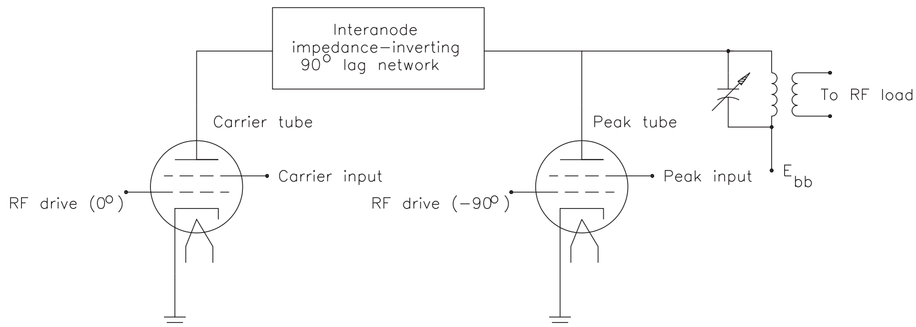

Figure 5.14 Screen-modulated Doherty-type circuit.

Modulation of the carrier occurs when the screen voltage of the peak tube begins to rise during the positive modulation half-cycle, thus causing the peak tube to supply more RF current to the output load. This increase of current into the output network causes the resistance seen by the interanode network to increase and, because of the impedance-inverting characteristics of the 90° interanode network, causes a proportional decrease in the load impedance presented to the carrier tube.

The carrier-tube resonant anode voltage drop is fully saturated over the entire positive modulation half-cycle, and therefore is effectively a constant voltage source. The power output of the carrier tube thus increases during the positive modulation half-cycle, caused by the modulated decreasing impedance at its anode, until both peak and carrier tubes deliver twice carrier power at the 100 percent positive modulation crest. During the negative modulation half-cycle, the peak tube is held out of conduction, while the carrier-tube output voltage decreases linearly to zero output at the 100 percent modulation trough.

A screen-modulated amplifier designed for MF operation at 1 MW is shown in Figure 5.15. The carrier tube delivers all of the 1 MW carrier power to the load through the 90° impedance-inverting network. When modulation is applied to the screen of the carrier tube (during the negative half-cycle), the carrier tube output follows the modulation linearly. Because the carrier tube is driven to full swing, its voltage excursion does not increase during the positive half-cycle. Positive modulation is needed on the screen only to maintain the full plate swing as the impedance load changes on the carrier tube.

The peak tube normally is operated with a negative screen voltage that maintains plate current at near cutoff. During the positive peak of modulation, the screen of the peak tube is modulated upward. As the peak tube is modulated positive, it delivers power into the output circuit. As the delivered power increases, the load seen by the interplate network changes as described for the general case.

The advantages of screen modulation are the same as those identified for the Doherty linear amplifier, plus higher efficiency at all depths of modulation. Further-

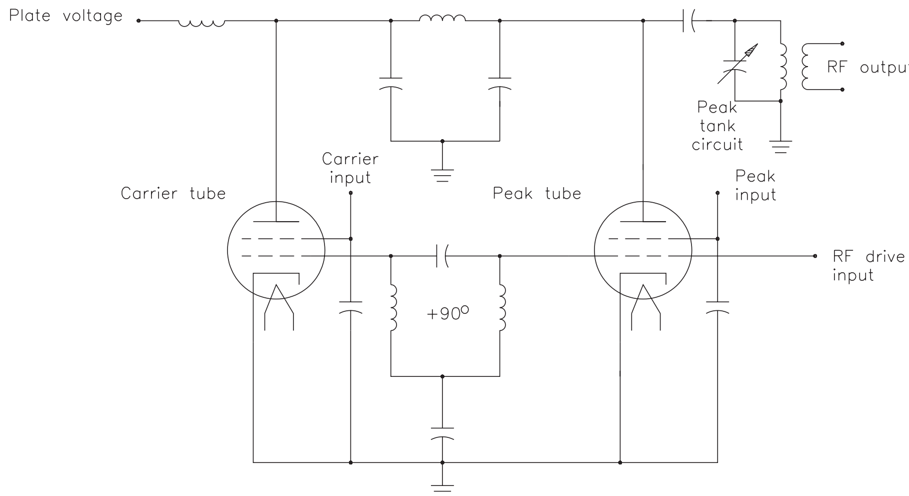

Figure 5.15 Practical implementation of a screen-modulated Doherty-type 1 MW power amplifier (Continental Electronics).

more, the screen modulation system is less sensitive to misadjustment of RF amplifier tuning.

5.4.4 Terman-Woodyard Modulated Amplifier

The *Terman-Woodyard* configuration uses the basic scheme of Doherty for achieving high efficiency (the impedance-inverting property of a 90° phase-shift network). However, the Terman-Woodyard design also employs grid modulation of both the carrier tube and the peak tube, allowing both tubes to be operated class C. The result is an increase in efficiency over the Doherty configuration.

The basic circuit is shown in Figure 5.16. With no modulation, V1 operates as a class C amplifier, supplying the carrier power, and V2 is biased so that it is just beginning to conduct. The efficiency at carrier is, therefore, essentially that of a class C amplifier.

During positive modulation swings, V2 conducts. At 100 percent modulation, both tubes supply equal amounts of power to the load, as in the Doherty design. During negative modulation swings, V2 is cut off, and V1 performs as a standard grid-modulated amplifier.

The Terman-Woodyard system rates efficiency as a function of modulation percentage for sinusoidal waveforms. Typical efficiency at carrier is 80 percent, falling to a minimum of 68 percent at 50 percent modulation and 73 percent at 100 percent modulation.

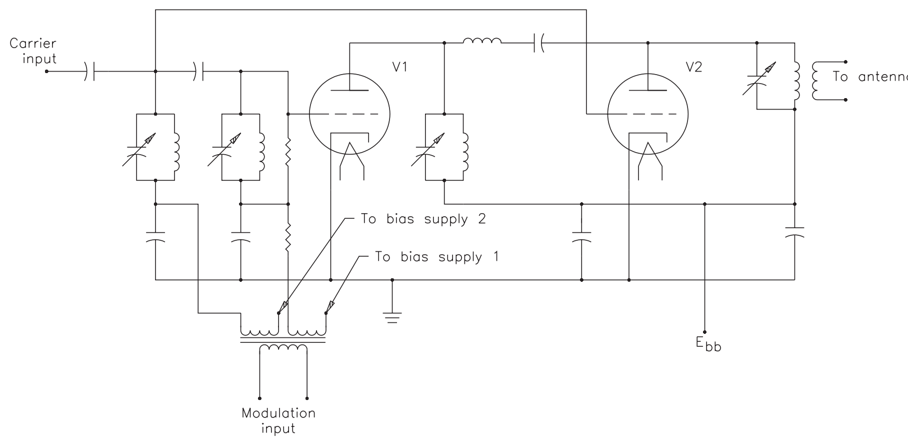

Figure 5.16 Terman-Woodyard high-efficiency modulated amplifier.

5.4.5 Dome Modulated Amplifier

The *Dome* high-efficiency amplifier employs three power output tubes driven by different input signals. Modulation is achieved by load-line modification during positive modulation excursions, and by linear amplification during negative modulation swings. Load-line modification is achieved by absorption of a portion of the generated RF power. However, most of the absorbed power is returned to the plate power supply, rather than being dissipated as heat. High efficiency is the result.

A basic Dome modulated amplifier is shown in Figure 5.17. Tube V1 is used in a plate-modulated configuration, supplying power to the grid of V2 (the power amplifier tube). The load impedance in the plate circuit of V2 is equivalent to the load impedance reflected into the primary of transformer T1 (from the antenna) in series with the impedance appearing across the C8 terminals of the 90° phase-shift network consisting of C8, C9, and L4. The impedance appearing across the C8 terminals of the 90° network is inversely related to the impedance across the other terminals of the network (in other words, the effective ac impedance of V3). Thus, with V3 cut off, a short circuit is reflected across the C8 terminals of the network. Tube V3 performs as a modulated rectifier with intelligence supplied to its grid. This tube is referred to as the *modifier*.

With no modulation (the carrier condition), tube V3 is biased to cutoff, and the drive to V2 is adjusted until power output is equal to four times the carrier power (corresponding to a positive modulation peak). Note that all the plate signal voltage of V2 appears across the primary of transformer T1 for this condition. The bias on V3 is then reduced, lowering the ac impedance of the tube and, therefore, reflecting an increasing

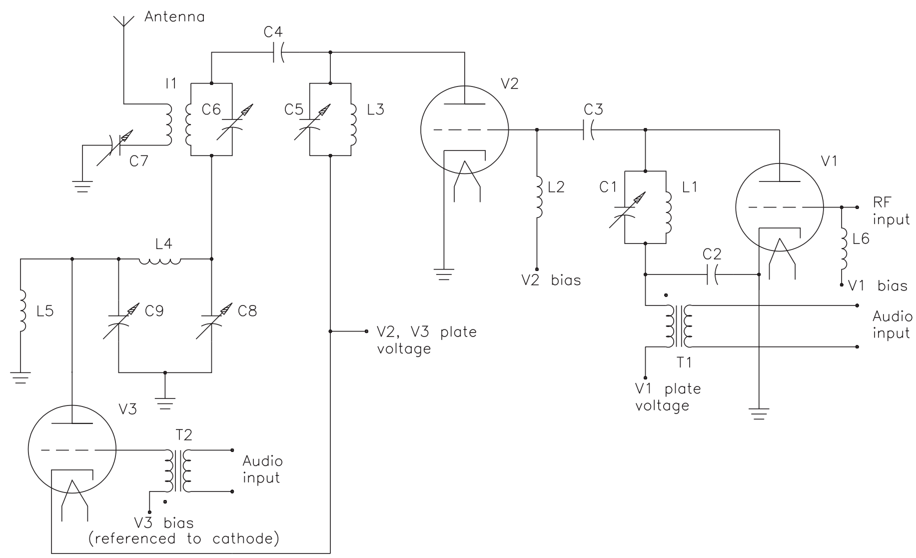

Figure 5.17 Dome high-efficiency modulated amplifier.

impedance at the C8 terminals of the 90° network. The bias on V3 is reduced until the operating carrier power is delivered to the antenna.

An amount of power equal to the carrier is thus rectified by V3 and returned to the plate power supply, except for that portion dissipated on the V3 plate. The drive level to V2 is finally adjusted so that the tube is just out of saturation.

For positive modulation excursions, V3 grid voltage is driven negative, reaching cutoff for a positive modulation peak. For negative modulation swings, V2 acts as a linear amplifier. V3 does not conduct during the negative modulation swing because the peak RF voltage on the plate is less than the dc supply voltage on the cathode.

Because of the serial loss incurred in the power amplifier tube and in the modifier tube for energy returned to the power supply, the Dome circuit is not as efficient as the Doherty configuration. Typical efficiency at carrier ranges from 55 to 60 percent.

5.5 Television Power Amplifier Systems

Although VSBAM is used for a number of varied applications, TV transmission is the most common. A TV transmitter is divided into two basic subsystems:

- The *visual* section, which accepts the video input, amplitude-modulates an RF carrier, and amplifies the signal to feed the antenna system

- The *aural* section, which accepts the audio input, frequency-modulates a separate RF carrier, and amplifies the signal to feed the antenna system

The visual and aural signals usually are combined to feed a single radiating antenna. Different transmitter manufacturers have different philosophies with regard to the design and construction of a transmitter. Some generalizations are possible, however, with respect to basic system configurations. Transmitters can be divided into categories based on the following criteria:

- Output power
- Final-stage design
- Modulation system

5.5.1 System Considerations

When the power output of a TV transmitter is discussed, the visual section is the primary consideration. Output power refers to the peak power of the visual stage of the transmitter (*peak of sync*). The FCC-licensed ERP is equal to the transmitter power output times feedline efficiency times the power gain of the antenna.

A VHF station can achieve its maximum power output through a wide range of transmitter and antenna combinations. Reasonable pairings for a high-band VHF station range from a transmitter with a power output of 50 kW feeding an antenna with a gain of 8, to a 30 kW transmitter connected to a gain-of-12 antenna. These combinations assume reasonably low feedline losses. To reach the exact power level, minor adjustments are made to the power output of the transmitter, usually by a front-panel power control.

UHF stations that want to achieve their maximum licensed power output are faced with installing a very high power transmitter. Typical pairings include a transmitter rated for 220 kW and an antenna with a gain of 25, or a 110 kW transmitter and a gain-of-50 antenna. In the latter case, the antenna could pose a significant problem. UHF antennas with gains in the region of 50 are possible, but not advisable for most installations because of coverage problems that can result.

The amount of output power required of the transmitter will have a fundamental effect on system design. Power levels dictate the following parameters:

- Whether the unit will be of solid-state or vacuum tube design
- Whether air, water, or vapor cooling must be used
- The type of power supply required
- The sophistication of the high-voltage control and supervisory circuitry
- Whether *common amplification* of the visual and aural signals (rather than separate visual and aural amplifiers) is practical

Tetrodes generally are used for VHF transmitters above 25 kW, and specialized tetrodes can be found in UHF transmitters at the 15 kW power level and higher. As solid-state technology advances, the power levels possible in a reasonable transmitter design steadily increase, making solid-state systems more attractive options.

In the realm of UHF transmitters, the klystron (and its related devices) reigns supreme. Klystrons use an *electron-bunching* technique to generate high power—55 kW from a single tube is not uncommon—at ultrahigh frequencies. They are currently the first choice for high-power, high-frequency service. Klystrons, however, are not particularly efficient. A stock klystron with no special circuitry might be only 40 percent efficient. Various schemes have been devised to improve klystron efficiency, the best known of which is *beam pulsing*. Two types of pulsing are in common use:

- *Mod-anode pulsing*, a technique designed to reduce power consumption of the device during the color burst and video portion of the signal (and thereby improve overall system efficiency)
- *Annular control electrode* (ACE) pulsing, which accomplishes basically the same goal by incorporating the pulsing signal into a low-voltage stage of the transmitter, rather than a high-voltage stage (as with mod-anode pulsing)

Experience has shown the ACE approach—and other similar designs—to provide greater improvement in operating efficiency than mod-anode pulsing, and better reliability as well.

Several newer technologies offer additional ways to improve UHF transmitter efficiency, including:

- The *inductive output tube* (IOT), also known as the *Klystrode*. This device essentially combines the cathode/grid structure of the tetrode with the drift tube/collector structure of the klystron. (The Klystrode tube is a registered trademark of Varian Associates.)
- The *multistage depressed collector* (MSDC) klystron, a device that achieves greater efficiency through a redesign of the collector assembly. A multistage collector is used to recover energy from the electron stream inside the klystron and return it to the beam power supply.

Improved tetrode devices, featuring higher operating power at UHF and greater efficiency, have also been developed.

A number of approaches may be taken to amplitude modulation of the visual carrier. Current technology systems utilize low-level intermediate-frequency (IF) modulation. This approach allows superior distortion correction, more accurate vestigial sideband shaping, and significant economic advantages to the transmitter manufacturer.

A TV transmitter can be divided into four major subsystems:

- The exciter
- Intermediate power amplifier (IPA)
- Power amplifier
- High-voltage power supply

Figure 5.18 shows the audio, video, and RF paths for a typical design.

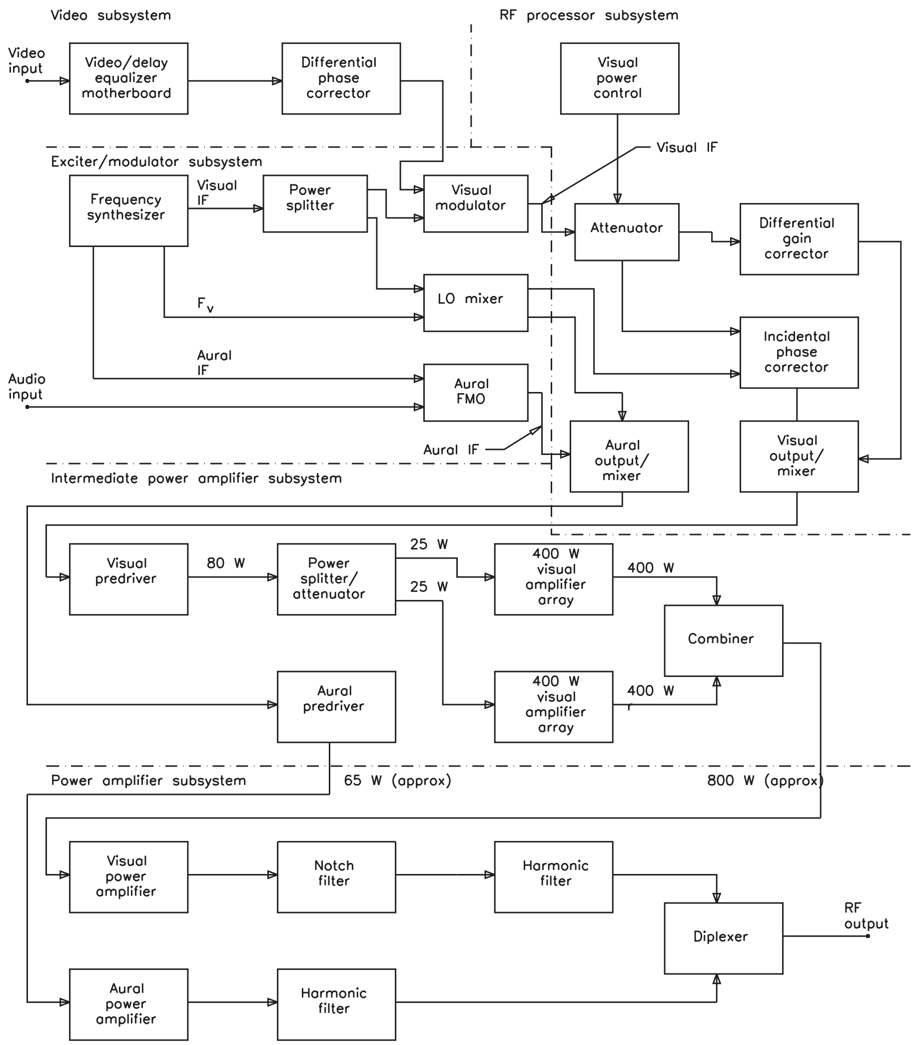

Figure 5.18 Basic block diagram of a TV transmitter. The three major subassemblies are the exciter, IPA, and PA. The power supply provides operating voltages to all sections, and high voltage to the PA stage.

5.5.2 Power Amplifier

The power amplifier raises the output energy of the transmitter to the required RF operating level. Tetrodes in TV service are usually operated in the class B mode to obtain reasonable efficiency while maintaining a linear transfer characteristic. Class B amplifiers, when operated in tuned circuits, provide linear performance because of the flywheel effect of the resonance circuit. This allows a single tube to be used instead of two in push-pull fashion. The bias point of the linear amplifier must be chosen so that the transfer characteristic at low modulation levels matches that at higher modulation levels. Even so, some nonlinearity is generated in the final stage, requiring differential gain correction. The plate (anode) circuit of a tetrode PA usually is built around a coaxial resonant cavity, providing a stable and reliable tank.

UHF transmitters using a klystron in the final output stage must operate class A, the most linear but also most inefficient operating mode for a vacuum tube. The basic efficiency of a nonpulsed klystron is approximately 40 percent. Pulsing, which provides full available beam current only when it is needed (during peak of sync), can improve device efficiency by as much as 25 percent, depending on the type of pulsing used.

Two types of klystrons are presently in service:

- Integral-cavity klystron
- External-cavity klystron

The basic theory of operation is identical for each tube, but the mechanical approach is radically different. In the integral-cavity klystron, the cavities are built into the klystron to form a single unit. In the external-cavity klystron, the cavities are outside the vacuum envelope and bolted around the tube when the klystron is installed in the transmitter.

A number of factors come into play in a discussion of the relative merits of integral- vs. external-cavity designs. Primary considerations include operating efficiency, purchase price, and life expectancy.

The PA stage includes a number of sensors that provide input to supervisory and control circuits. Because of the power levels present in the PA stage, sophisticated fault-detection circuits are required to prevent damage to components in the event of a problem inside or outside the transmitter. An RF sample, obtained from a directional coupler installed at the output of the transmitter, is used to provide automatic power-level control.

The transmitter system discussed in this section assumes separate visual and aural PA stages. This configuration is normally used for high-power transmitters. A combined mode also may be used, however, in which the aural and visual signals are added prior to the PA. This approach offers a simplified system, but at the cost of additional precorrection of the input video signal.

PA stages often are configured so that the circuitry of the visual and aural amplifiers is identical, providing backup protection in the event of a visual PA failure. The aural PA can then be reconfigured to amplify both the aural and the visual signals, at reduced power.

The aural output stage of a TV transmitter is similar in basic design to an FM broadcast transmitter. Tetrode output devices generally operate class C, providing good efficiency. Klystron-based aural PAs are used in UHF transmitters.

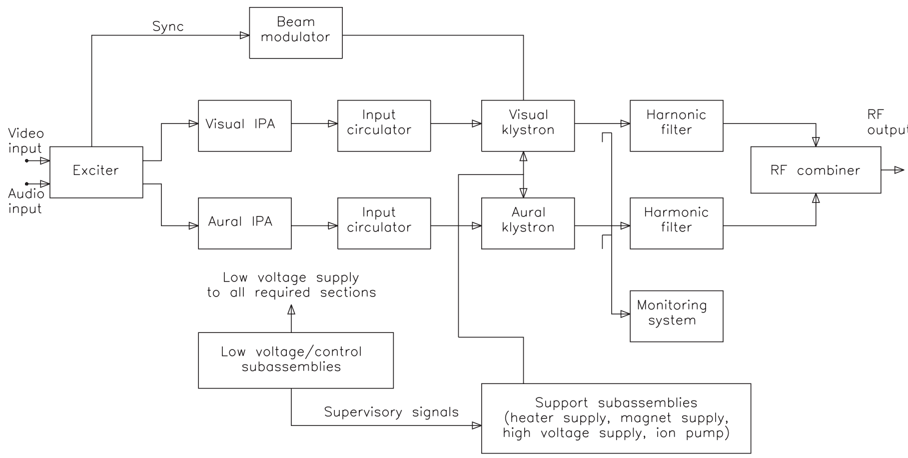

Figure 5.19 Schematic diagram of a 60 kW klystron-based TV transmitter.

Application Example

A 60 kW transmitter is shown in block diagram form in Figure 5.19. A single high-power klystron is used in the visual amplifier, and another is used in the aural amplifier. The tubes are driven from solid-state intermediate power amplifier modules. The transmitter utilizes ACE-type beam control, requiring additional predistortion to compensate for the nonlinearities of the final visual stage. Predistortion is achieved by correction circuitry at an intermediate frequency in the modulator. Both klystrons are driven from the output of a circulator, which ensures a minimum of driver-to-load mismatch problems.

A block diagram of the beam modulator circuit is shown in Figure 5.20. The system receives input signals from the modulator, which synchronizes ACE pulses to the visual tube with the video information. The pulse waveform is developed through a pulse amplifier, rather than a switch. This permits more accurate adjustments of operating conditions of the visual amplifier.

Although the current demand from the beam modulator is low, the bias is near cathode potential, which is at a high voltage relative to ground. The modulator, therefore, must be insulated from the chassis. This is accomplished with optical transmitters and receivers connected via fiber optic cables. The fiber optic lines carry supervisory, gain control, and modulating signals.

The four-cavity external klystrons will tune to any channel in the UHF-TV band. An adjustable *beam perveance* feature enables the effective electrical length of the device to be varied by altering the beam voltage as a function of operating frequency (see Section 6.3). Electromagnetic focusing is used on both tubes. The cavities, body, and gun

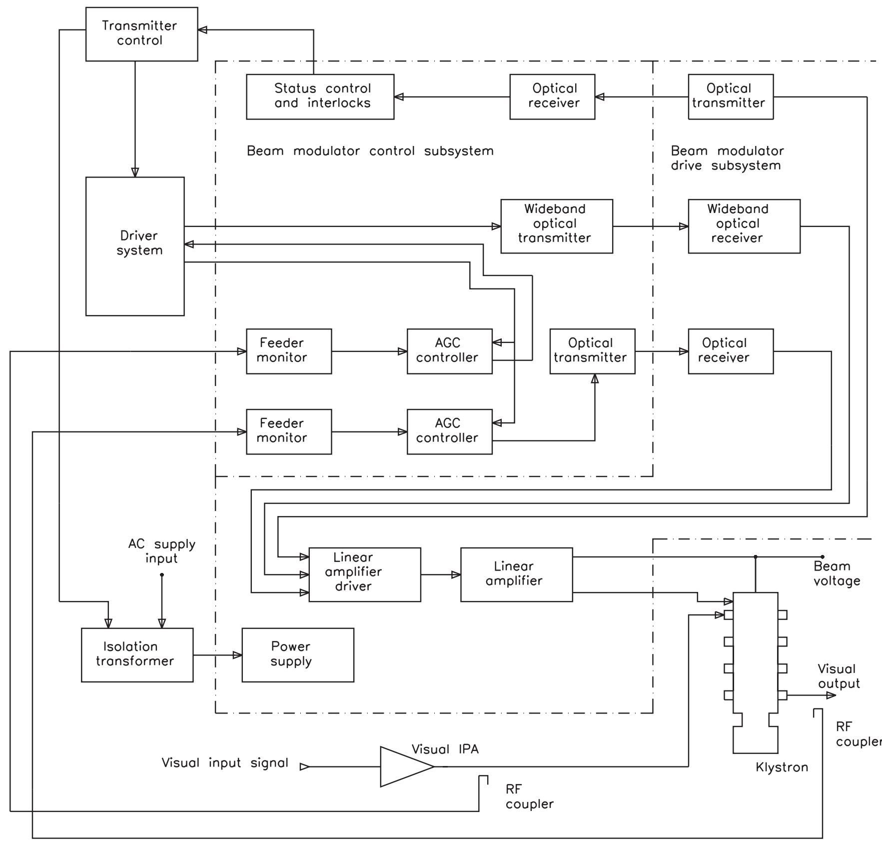

Figure 5.20 Schematic diagram of the modulator section of a 60 kW TV transmitter.

areas of the klystrons are air-cooled. The collectors are vapor-phase-cooled using an external heat exchanger system.

The outputs of the visual and aural klystrons are passed through harmonic filters to an RF combiner before being applied to the antenna system.

5.6 FM Power Amplifier Systems

The amplitude of an FM signal remains constant with modulation so that efficient class B and C amplifiers can be used.[1] Two basic circuit types have evolved:

- Amplifiers based on a tetrode or pentode in a grid-driven configuration

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1 This section was contributed by Geoffrey N. Mendenhall, P. E., Cincinnati, OH, and Warren B. Bruene, P.E., Dallas, TX.

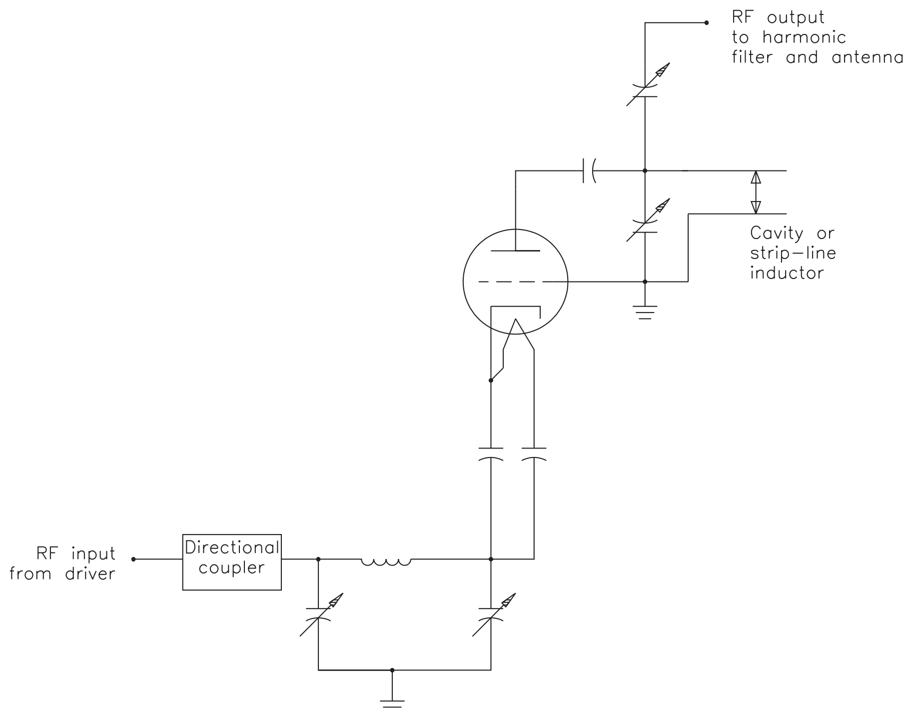

Figure 5.21 Cathode-driven triode power amplifier.

- Amplifiers based on a high-μ triode or tetrode device in a cathode-driven configuration (grounded grid)

High-power FM amplifiers operating in the VHF range typically employ resonant cavity systems. The basic configurations of FM amplifiers are discussed in this section, and amplifiers utilizing cavity resonators are discussed in Section 4.3.

5.6.1 Cathode-Driven Triode Amplifier

The characteristics of high-μ triodes are well adapted to FM power amplifier service. The grounded-grid circuit is simple, and no screen or grid bias power supplies are required. Figure 5.21 shows the basic circuit configuration. In this case, the grid is connected directly to chassis ground. The difference between dc cathode current and dc plate current is dc grid current. The output tank is a shorted coaxial cavity that is capacitively loaded by the tube output and stray circuit capacitance. A small capacitor is used for trimming the tuning, and another small variable capacitor is used for adjusting the loading. A pi network is used to match the 50 Ω input to the tube cathode input impedance.

The triode usually is operated in the class B mode to achieve maximum power gain, which is on the order of 20 (13 dB). The device can be driven into class C operation by providing negative grid bias or positive cathode bias. This increases the plate efficiency, but also requires increased drive power.

Most of the drive into a grounded-grid amplifier is fed through the tube and appears at the output. This increases the apparent efficiency so that the efficiency factor given by the transmitter manufacturer may be higher than the actual plate efficiency of the tube. The true plate efficiency is determined from the following:

$$Eff_p = \frac{P_O}{\left(I_p \times E_p\right) + \left(P_{in} \times \alpha\right)} \tag{5.2}$$

Where:
Eff_p = actual plate efficiency
P_o = measured output power
I_p = dc plate input current to final stage
E_p = dc plate input voltage to final stage
P_{in} = RF input drive power
α = fraction of grid-to-plate transfer (0.9 typical)

Because most of the drive power is fed through the tube, any changes in loading of the output circuit also will affect the input impedance and the driver stage.

Because there is RF drive voltage on the cathode (filament), some means of decoupling must be used to block the signal from the filament transformer. One method employs high-current RF chokes; another feeds the filament power through the input tank circuit inductor.

5.6.2 Grounded-Grid vs. Grid-Driven Tetrode

Tetrodes also may be operated in the grounded-grid configuration by placing both the control grid and the screen grid at RF ground. Higher efficiency and gain can be achieved by placing negative bias on the control grid while placing a positive voltage on the screen grid of a cathode-driven tetrode.

The input capacitance of a tetrode in a grounded-grid configuration is much less than in a grid-driven configuration, and the input impedance is lower, providing wider bandwidth. The approximate input capacitances of some common tube types are listed in Table 5.1.

The typical drive power requirements of a high-gain tetrode, as a function of plate voltage, for a 5 kW power amplifier operating at 100 MHz are as follows:

- Grounded-grid configuration: 340 W input at 4.5 kV plate voltage, 280 W at 5.2 kV
- Grid-driven configuration: 190 W at 4.5 kV plate voltage, 140 W at 5.2 kV

Table 5.1 Approximate Input Capacitance of Common Power Grid Tubes Used in Grounded-Grid and Grounded-Cathode Service

Tube Type	Gounded Grid	Grounded Cathode
4CX3000A	67	140
4CX3500A	59	111
4CX5000A	53	115
4CX15,000A	67	161
4CX20,000A/8990	83	190

There are several tradeoffs in performance between the grounded-grid and the grid-driven modes of a tetrode PA with respect to gain, efficiency, amplitude bandwidth, phase bandwidth, and incidental AM under equivalent operating conditions:

- When the PA is driven into saturation, its bandwidth is limited by the output cavity bandwidth in the grounded-grid amplifier. The PA bandwidth in the grid-driven amplifier is limited by the input circuit Q.
- Ouput bandwidth under saturation can be improved in either configuration by reducing the plate voltage. This involves a tradeoff in efficiency with a smaller voltage swing. The bandwidth improvement can be obtained with a loss of PA gain and efficiency.
- A grounded-grid saturated PA improves bandwidth over a grid-driven saturated PA at the expense of amplifier gain. Best performance for FM operation is obtained when the amplifier is driven into saturation where little change in output power occurs with increasing drive power. Maximum efficiency also occurs at this point.
- The phase linearity in the 0.5 dB bandwidth of the amplifier is better in a grid-driven configuration. The class C grounded-grid PA exhibits a more nonlinear phase slope within the passband, yet has a wider amplitude bandwidth. This phenomenon is the result of interaction of the input and output circuits because they are effectively connected in series in the grounded-grid configuration. The neutralized grid-driven PA provides more isolation between these networks, so they behave more like independent filters.

5.6.3 Grid-Driven Tetrode/Pentode Amplifiers

RF generators using tetrode amplifiers throughout usually have one less stage than those using triodes. Because tetrodes have higher power gain, they are driven into class C operation for high plate efficiency. Countering these advantages is the requirement for neutralization, along with screen and bias power supplies. Figure 5.22

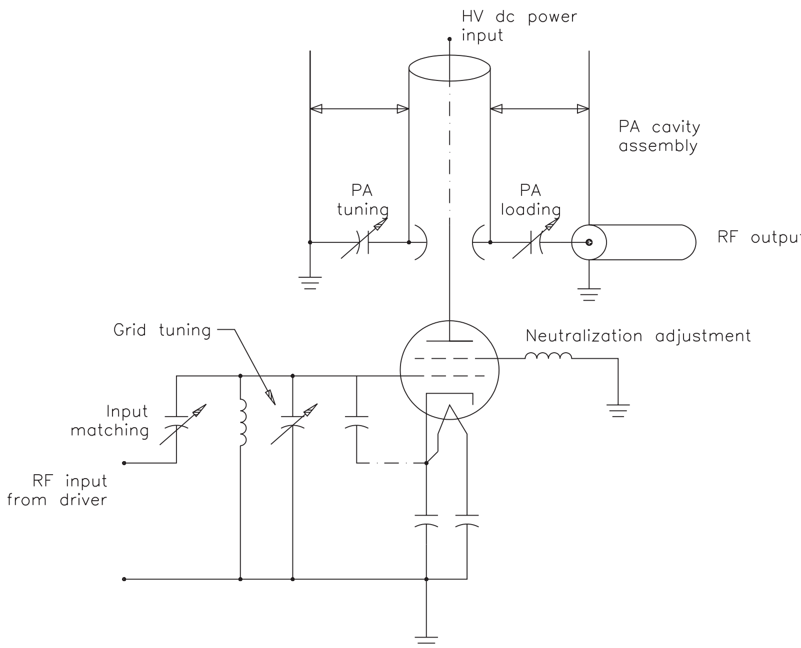

Figure 5.22 Grid-driven grounded-screen tetrode power amplifier.

shows a schematic of a grid-driven tetrode amplifier. In this example, the screen is operated at dc ground potential, and the cathode (filament) is operated below ground by the amount of screen voltage required (grounded-screen operation). The advantage of this configuration is that stability problems resulting from undesired resonances in the screen bypass capacitors are eliminated. With this arrangement, however, the screen supply must be capable of handling the combined plate and screen currents.

With directly heated tubes, it is necessary to use filament bypass capacitors. During grounded-screen operation, these bypass capacitors must have a higher breakdown voltage rating (relative to grounded-grid operation) because they will have the dc screen voltage across them. The filament transformer must have additional insulation to withstand the dc screen voltage. The screen power supply provides a negative voltage in series with the cathode-to-ground, and must have the additional capacity to handle the sum of the plate and screen currents. A coaxial cavity is used in the output circuit so that the circulating current is spread over large surfaces, keeping the losses low. This cavity is a shorted 1/4-wavelength transmission line section that resonates the tube output capacitance. The length is preset to the desired carrier frequency, and a small-value variable capacitor is used to trim the system to resonance. (Amplifiers utilizing cavity resonator systems are discussed in Section 4.3.)

Pentode amplifiers have even higher gain than their tetrode counterparts. The circuit configuration and bias supply requirements for the pentode are similar to the tetrode's because the third (suppressor) grid is tied directly to ground. The additional isolating effect of the suppressor grid eliminates the need for neutralization in the pentode amplifier.

5.6.4 Impedance Matching into the Grid

The grid circuit usually is loaded (swamped) with added resistance. The purpose of this resistance is to broaden the bandwidth of the circuit by lowering the circuit Q and to provide a more constant load to the driver. The resistance also makes neutralization less critical so that the amplifier is less likely to become unstable with varying output circuit loading.

Cathode and filament lead inductance from inside the tube, through the socket and filament bypass capacitors to ground, can increase grid driver requirements. This effect is the result of RF current flowing from grid to filament through the tube capacitance and then through the filament lead inductance to ground. An RF voltage is developed on the filament that, in effect, causes the tube to be partly cathode-driven. This undesirable extra drive power requirement can be minimized by series resonating the cathode return path with the filament bypass capacitors or by minimizing the cathode-to-ground inductance by using a tube socket incorporating thin-film dielectric *sandwich* capacitors for coupling and bypassing.

High-power grid-driven class C amplifiers require a swing of several hundred RF volts on the grid. To develop this high voltage swing, the input impedance of the grid must be increased by the grid input matching circuit. Because the capacitance between the grid and the other tube elements may be 100 pF or more, the capacitive reactance at 100 MHz, for example, will be very low unless the input capacitance is parallel resonated with an inductor. Bandwidth can be maximized by minimizing any additional circuit capacitance and using a portion of the tube input capacitance as part of the impedance-transformation network. Figure 5.23 illustrates two popular methods of resonating and matching into the grid of a high-power tube. Both schemes can be analyzed by recognizing that the desired impedance transformation is produced by an equivalent L network.

In Figure 5.23*a*, a variable inductor (L_{in}) is used to raise the input reactance of the tube by bringing the tube input capacitance (C_{in}) almost to parallel resonance. Parallel resonance is not reached because a small amount of parallel capacitance (C_p) is required by the equivalent L network to transform the high impedance (Z_{in}) of the tube down to a lower value through the series matching inductor (L_s). This configuration has the advantage of providing a low-pass filter by using part of the tube's input capacitance to form C_p.

Figure 5.23*b* uses variable inductor L_{in} to take the input capacitance (C_{in}) past parallel resonance so that the tube's input impedance becomes slightly inductive. The variable series matching capacitor (C_s) forms the rest of the equivalent L network. This configuration constitutes a high-pass filter.

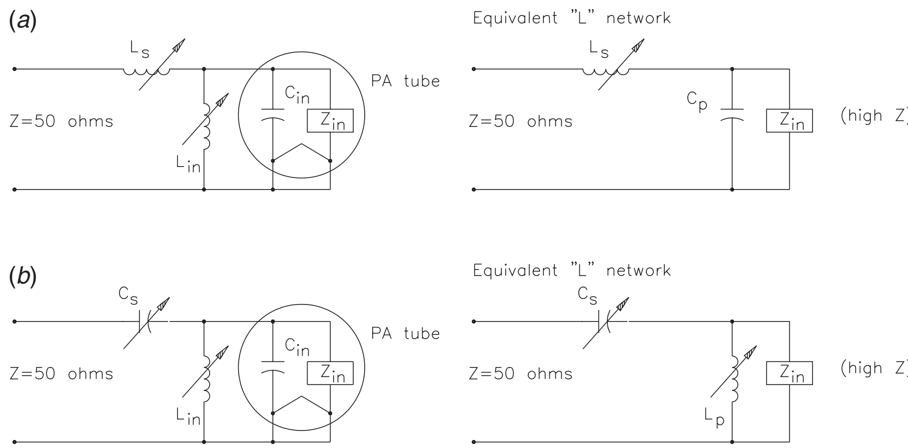

Figure 5.23 Input matching circuits: (*a*) inductive input matching, (*b*) capacitive input matching.

Interstage Coupling

The IPA output circuit and the final amplifier input circuit often are coupled by a coaxial transmission line. Impedance matching usually is accomplished at either end by one of the configurations shown in Figure 5.24. The circuits shown, except for 5.24*d*, require some interactive adjustment of the tuning and loading elements to provide a satisfactory impedance match for each operating frequency and RF drive level. The circuit in Figure 5.24*d* utilizes multiple *LC* sections, with each section providing a small step in the total impedance transformation. This technique provides a broadband impedance match without adjustment, thereby improving transmitter stability, ease of operation, and maintainability. A single grid resonating control is sufficient to tune and match the 50 Ω driver impedance to the high input impedance of the grid over a relatively wide band of frequencies and RF power levels.

The transmission line matching problem is eliminated in some transmitter designs by integrating an IPA stage, using one or more tubes, into the grid circuit of the final amplifier by having the plate of the IPA and the grid of the final tube share a common tuned circuit. This technique has the advantage of simplicity by not transforming the impedance down to 50 Ω and then back up to the grid impedance level.

Solid-state RF power devices possess a low load impedance at the device output terminal, so an impedance transformation that goes through the 50 Ω intermediate impedance of a coaxial cable is necessary to couple these devices into the relatively high impedance of the final amplifier grid.

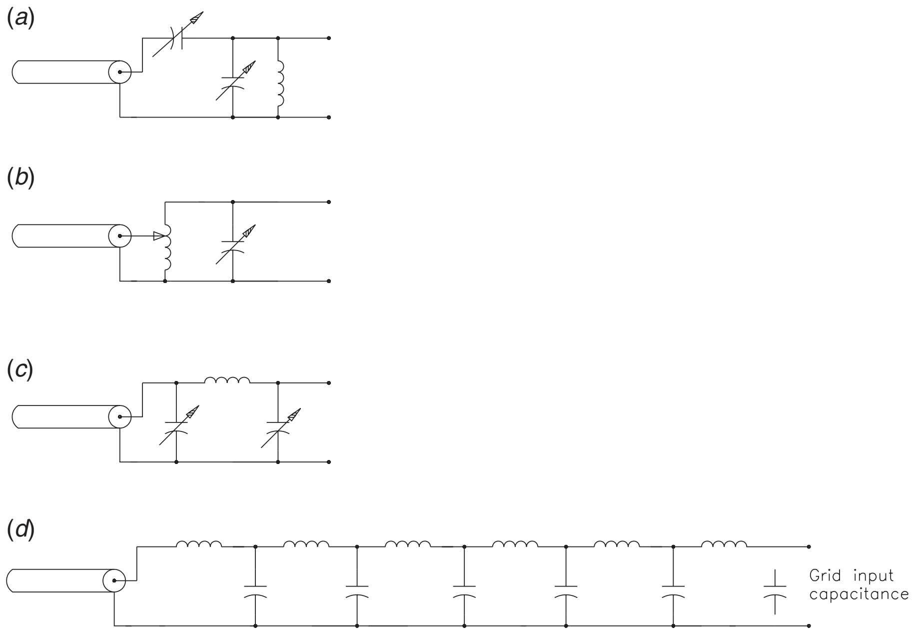

Figure 5.24 Interstage RF coupling circuits: (*a*) variable capacitance matching, (*b*) variable tank matching, (*c*) variable pi matching, (*d*) fixed broadband impedance-transformation network.

5.6.5 Neutralization

Most cathode-driven amplifiers using grounded-grid triodes do not require neutralization. It is necessary that the grid-to-ground inductance, both internal and external to the tube, be kept low to maintain this advantage. Omission of neutralization will allow a small amount of interaction between the output circuit and the input circuit through the plate-to-filament capacitance. This effect is not particularly noticeable because of the large coupling between the input and output circuits through the electron beam of the tube. Cathode-driven tetrodes have higher gain than triodes and, therefore, often require some form of neutralization.

Grid-driven high-gain tetrodes need accurate neutralization for best stability and performance. Self-neutralization can be accomplished simply by placing a small amount of inductance between the tube screen grid and ground. This inductance is usually in the form of several short adjustable-length straps. The RF current flowing from plate to screen in the tube also flows through this screen lead inductance. This develops a small RF voltage on the screen, of the opposite phase, which cancels the voltage fed back through the plate-to-grid capacitance. This method of lowering the self-neutralizing frequency of the tube works only if the self-neutralizing frequency of the tube/socket combination is above the desired operating frequency before the inductance is added. Special attention must be given to minimizing the inductances in the tube socket by integrating distributed bypass capacitors into the socket and cavity deck assembly. Pentodes normally do not require neutralization because the suppressor grid effectively isolates the plate from the grid.

5.7 Special-Application Amplifiers

The operating environment of tubes used in commercial, industrial, and military applications usually is characterized by widely varying supply voltages, heavy and variable vibration, and significant changes in loading. Scientific applications are no less demanding. Research projects span a wide range of powers and frequencies. Typical uses for power vacuum tubes as of this writing include:

- Super proton synchrotron delivering 2.4 MW at 200 MHz
- Fusion reactor using 33 MW of high-frequency heating in the 25 to 60 MHz band
- Plasma research using 12 MW in the 60 to 120 MHz range
- Fusion research using a 500 kW ion cyclotron resonance heating generator operating at 30 to 80 MHz

It can be seen from these examples that heat dissipation is a major consideration in the design of an amplifier or oscillator for scientific uses.

5.7.1 Distributed Amplification

Specialized research and industrial applications often require an amplifier covering several octaves of bandwidth. At microwave frequencies, a traveling wave tube (TWT) may be used. At MF and below, multiple tetrodes usually are combined to achieve the required bandwidth. A *distributed amplifier* consists of multiple tetrodes and lumped-constant transmission lines. The lines are terminated by load resistances with magnitudes equal to their characteristic impedances. A cyclical wave of current is present on the output transmission line circuit, with each tube contributing its part in proper phase. A typical configuration for a distributed amplifier includes 8 to 16 tubes. The efficiency of such a design is low.

5.7.2 Radar

The exciter stage of a radar set comprises oscillators, frequency multipliers, and mixers. The signals produced depend on whether the transmitter output device is operated as a power amplifier or an oscillator.

Transmitters using power oscillators such as *magnetrons* determine the radio frequency by tuning of the device itself. In a conventional (*noncoherent*) radar system, the only frequency required is the *local oscillator* (LO). The LO differs from the magnetron frequency by an intermediate frequency, and this difference usually is maintained

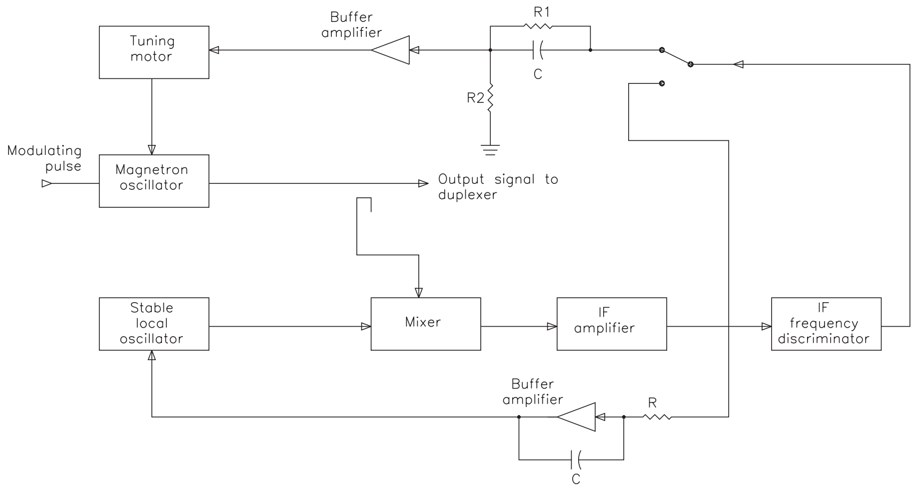

Figure 5.25 Two common approaches to automatic frequency control of a magnetron oscillator.

with an *automatic frequency control* (AFC) loop. Figure 5.25 shows a simple magnetron-based radar system with two methods of tuning:

- Slaving the magnetron to follow the *stable local oscillator* (STALO)
- Slaving the STALO to follow the magnetron

If the radar must use *coherent detection* (such as in Doppler applications), a second oscillator, called a *coherent oscillator* (COHO), is required. The COHO operates at the intermediate frequency and provides a reference output for signal processing circuits.

The synchronizer circuits in the exciter supply timing pulses to various radar subsystems. In a simple marine radar, this may consist of a single multivibrator that triggers the transmitter. In a larger system, 20 to 30 timing pulses may be required. These may turn the beam current on and off in various transmitter stages, start and stop RF pulse time attenuators, start display sweeps, and perform numerous other functions. Newer radar systems generate the required timing signals digitally. A digital synchronizer is illustrated in Figure 5.26.

Modulator

A radar system RF amplifier usually centers on one of two microwave devices: a *crossed-field tube* or *linear-beam tube*. Both are capable of high peak output power at microwave frequencies. To obtain high efficiency from a pulsed radar transmitter, it is necessary to cut off the current in the output tube between pulses. The modulator per-

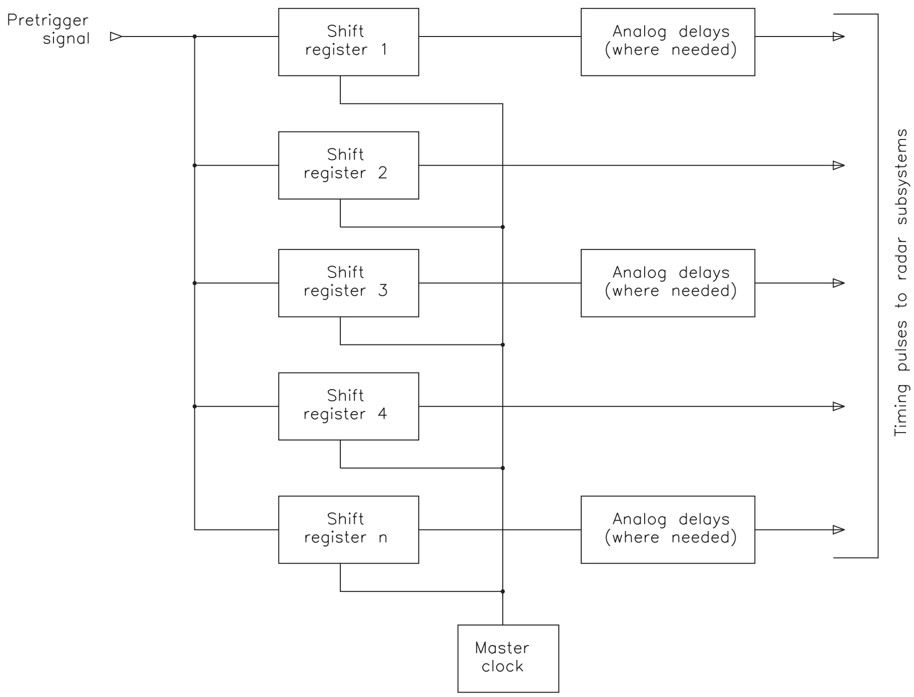

Figure 5.26 Digital synchronizer system for radar applications.

forms this function. Some RF tubes include control electrodes or grids to achieve the same result. Three common types of modulators are used in radar equipment:

- *Line-type modulator* (Figure 5.27). This common radar modulator is used most often to pulse a magnetron. Between pulses, a charge is stored in a *pulse-forming network* (PFN). A trigger signal fires a *thyratron* tube, short-circuiting the input to the PFN, which causes a voltage pulse to appear at the primary of transformer T1. The PFN components are chosen to produce a rectangular pulse at the magnetron cathode, with the proper voltage and current to excite the magnetron to oscillation. An advantage of this design is its simplicity. A drawback is the inability to electronically change the width of the transmitted pulse.
- *Active-switch modulator* (Figure 5.28). This system permits pulse width variation, within the limitations of the energy stored in the high-voltage power supply. A switch tube controls the generation of RF by completing the circuit path from the output tube to the power supply, or by causing stored energy to be dumped to the output device. The figure shows the basic design of an active-switch modulator and three variations on the scheme. The circuits differ in the method of coupling power supply energy to the output tube (capacitor-coupled, transformer-coupled, or a combination of the two methods).

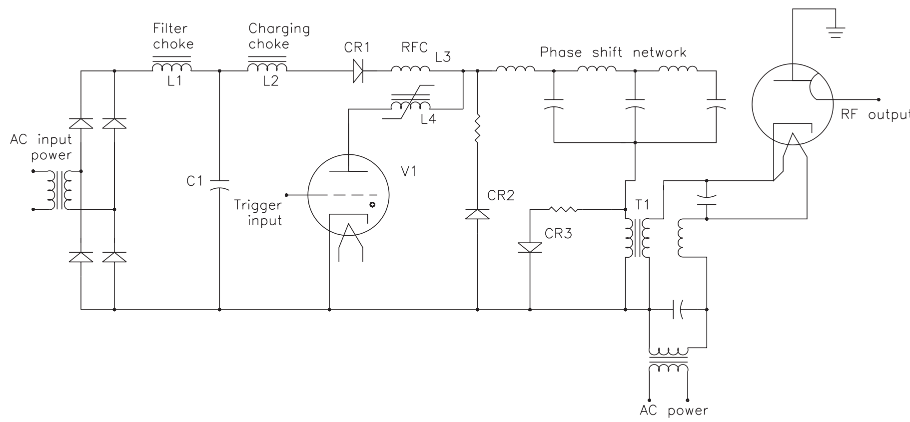

Figure 5.27 A line-type modulator for radar.

- *Magnetic modulator* (Figure 5.29). This design is the simplest of the three modulators discussed. No thyratron or switching device is used. Operation of the modulator is based on the saturation characteristics of inductors L1, L2, and L3. A long-duration low-amplitude pulse is applied to L1, which charges C1. As C1 approaches its fully charged state, L2 saturates and the energy in C1 is transferred in a resonant fashion to C2. This process continues to the next stage (L3 and C3). The transfer time is set by selection of the components to be about one-tenth that of the previous stage. At the end of the chain, a short-duration high-amplitude pulse is generated, exciting the RF output tube.

Because the applications for radar vary widely, so do antenna designs. Sizes range from less than 1 ft to hundreds of feet in diameter. An antenna intended for radar applications must direct radiated power from the transmitter to the azimuth and elevation coordinates of the target. It also must serve as a receive antenna for the echo.

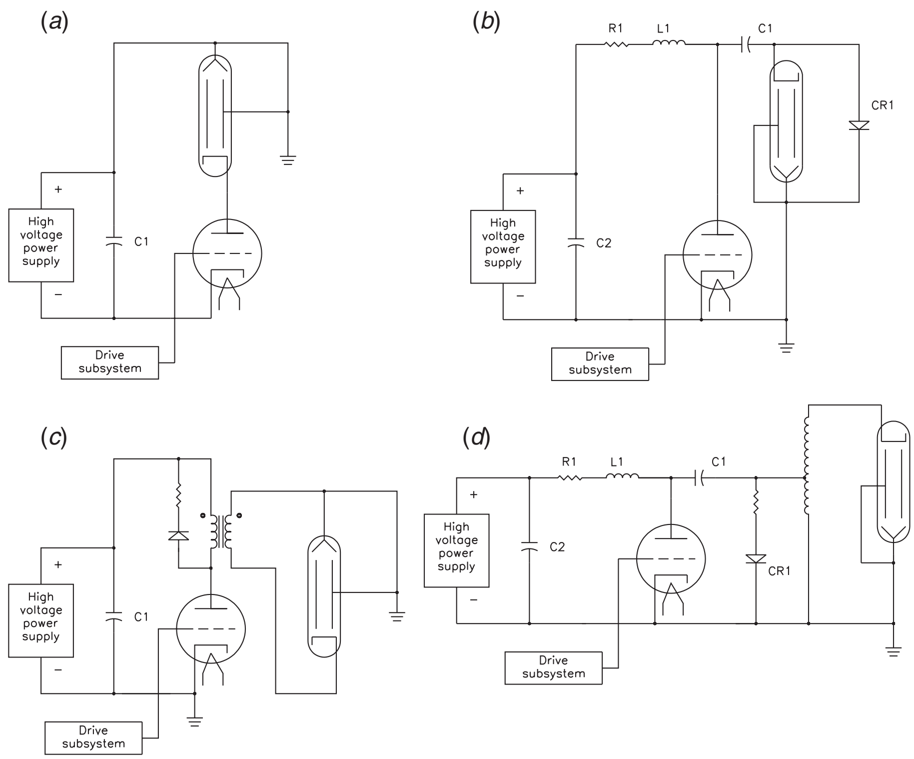

Figure 5.28 Active-switch modulator circuits: (*a*) direct-coupled system, (*b*) capacitor-coupled, (*c*) transformer-coupled, (*d*) capacitor- and transformer-coupled.

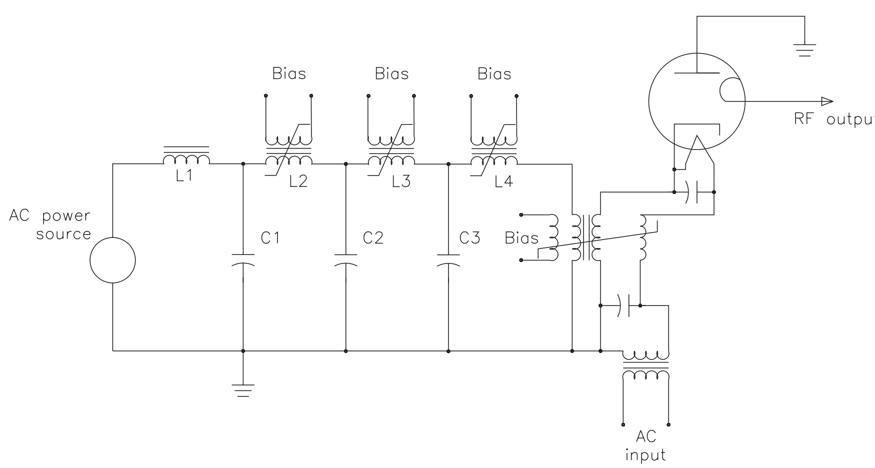

Figure 5.29 A magnetic modulator circuit.

5.8 References

1. Laboratory Staff, *The Care and Feeding of Power Grid Tubes*, Varian Eimac, San Carlos, CA, 1984.
2. Woodard, G. W., “AM Transmitters,” in *NAB Engineering Handbook*, 8th Ed., E. B. Crutchfield (ed.), National Association of Broadcasters, Washington, DC, pg. 356, 1992.)

5.9 Bibliography

Chireix, Henry, “High Power Outphasing Modulation,” *Proceedings of the IRE*, Vol. 23, no. 11, pg. 1370, November 1935.

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

Doherty, W. H., “A New High Efficiency Power Amplifier for Modulated Waves,” *Proceedings of the IRE*, Vol. 24, no. 9, pg. 1163, September 1936.

Fink, D., and D. Christiansen, *Electronic Engineers' Handbook*, 3rd Ed., McGraw-Hill, New York, 1989.

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.

High Power Transmitting Tubes for Broadcasting and Research, Philips Technical Publication, Eindhoven, the Netherlands, 1988.

Honey, J. F., “Performance of AM and SSB Communications,” *Tele-Tech.*, September 1953.

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

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

Mina and Parry, “Broadcasting with Megawatts of Power: The Modern Era of Efficient Powerful Transmitters in the Middle East,” *IEEE Transactions on Broadcasting*, Vol. 35, no. 2, IEEE, Washington, D.C., June 1989.

Pappenfus, E. W., W. B. Bruene, and E. O. Schoenike, *Single Sideband Principles and Circuits*, McGraw-Hill, New York, 1964.

Ridgwell, J. F., “A New Range of Beam-Modulated High-Power UHF Television Transmitters,” *Communications & Broadcasting*, no. 28, Marconi Communications, Clemsford Essex, England, 1988.

Skolnik, M. I. (ed.), *Radar Handbook*, McGraw-Hill, New York, 1980.

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.

Weldon, J. O., “Amplifiers,” U. S. Patent no. 2,836,665, May 1958.

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*, 9th Ed., Jerry C. Whitaker (ed.), National Association of Broadcasters, Washington, D.C., pp. 353-381, 1998.