3
MODULATORS

The purpose of the modulator, or pulser, is to provide a voltage or current waveform that will permit the selected microwave source to operate in a proper manner [8,21]. There are a number of important considerations in the selection and design of a modulator for a particular transmitter tube, including:

- Pulse length
- PRF
- Operating voltage and current
- Tube protection (from arcs)
- Spurious modes
- Pulse flatness (amplitude and phase)
- Cost
- Size and weight
- Efficiency
- Reliability and maintainability

There are several basic types of modulators, and numerous offshoots of each type. All of the modulators have one characteristic in common: they contain some means for storing energy and a switch to control the discharge of energy into the load, as is shown in Figure 3–1(*a*). The energy-storage element may store energy in a magnetic field or in an electric field. The energy in the energy-storage element must be replenished from the power supply, and an isolating element to prevent interaction is often included. A block diagram of such a complete system is shown in Figure 3–1(*b*).

There are two principal types of modulators: *line-type* and *hard-tube.* In the line-type modulator all the energy stored in the energy-storage device is dissipated in the load during each pulse, while the hard-tube

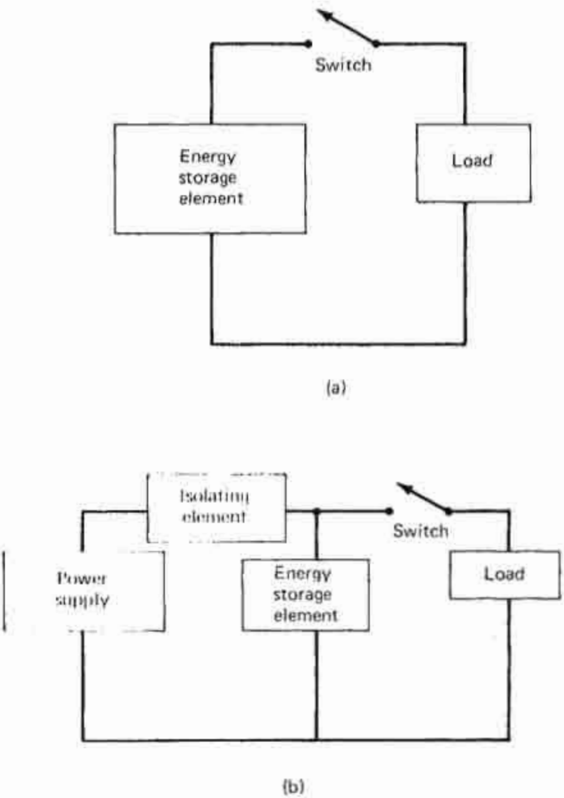


FIGURE 3–1 Basic modulator configurations, illustrating discharge of stored energy into the load.

modulator dissipates only some fraction of the stored energy during each pulse. These modulator types will now be discussed in greater detail.

3–1 HARD-TUBE MODULATORS

The hard-tube modulator is essentially a class C pulsed amplifier. A simplified diagram of a hard-tube modulator is shown in Figure

3–2. A positive pulse on the grid turns on the tube, and the capacitor couples the resulting step in plate voltage to the load. The hard-tube modulator is usually larger and more complex than other types of modulators, but it is also more versatile, not being as strongly influenced by load characteristics, and its output pulse length may be easily changed.

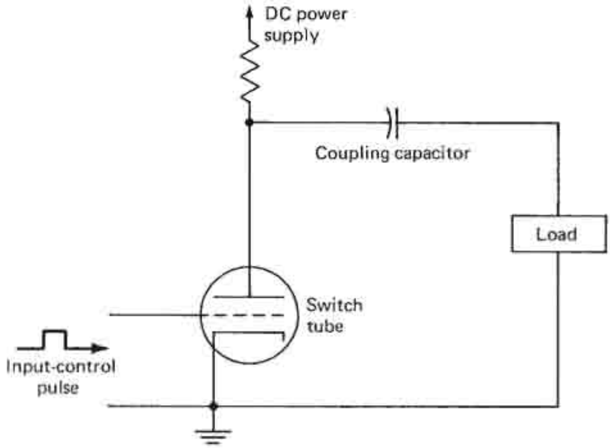


FIGURE 3–2 Simplified hard-tube modulator schematic diagram.

Hard-Tube Modulator Configurations

There are a number of variations on the basic hard-tube modulator, several of which are shown in Figure 3–3. Figure 3–3(*a*) is the conventional capacitor-coupled hard-tube modulator already outlined earlier; Figure 3–3(*b*) is the transformer-coupled hard-tube modulator; while Figure 3–3(*c*) is the capacitor- and transformer-coupled hard-tube modulator; and Figure 3–3(*d*) is the series-discharge, parallel-charge hard-tube modulator with capacitive coupling. The conventional capacitively coupled hard-tube modulator is by far the most common configuration encountered, and one which we will consider in some detail [5]. A number of different loads may be attached to such a modulator, including a resistive load (which is usually a good approximation of a klystron or TWT); a biased-diode load, such as a magnetron, with resistive charging; or a biased-diode load with an inductive recharging path.

A special form of hard-tube modulator is sometimes used with tubes containing a modulating anode, such as TWTs and klystrons. This is the *floating-deck* modulator; it is illustrated schematically in Figure 3–4. When the microwave tube is off, tube 1 is turned on and tube 2 is

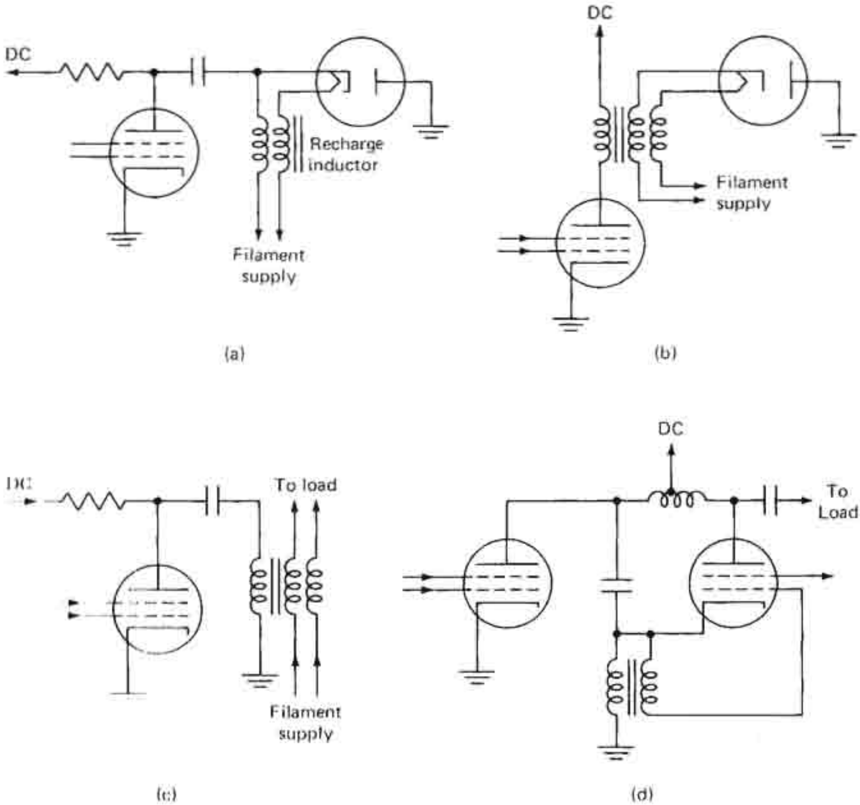


FIGURE 3–3 Some commonly encountered types of hard-tube modulator configurations; (*a*) capacitor-coupled; (*b*) transformer-coupled; (*c*) capacitor- and transformer-coupled; (*d*) parallel charge, series discharge.

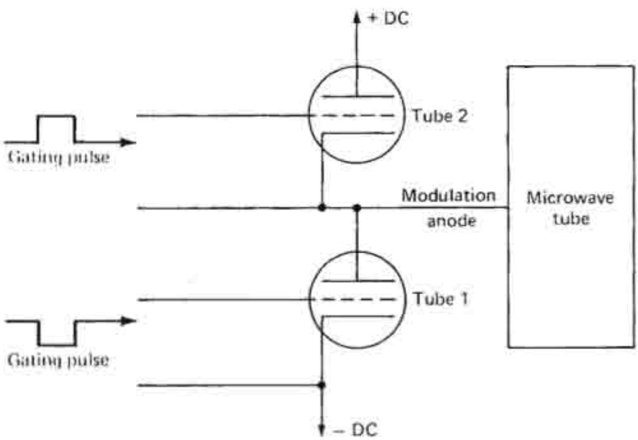


FIGURE 3–4 Simplified representation of a floating-deck modulator.

nonconducting. In order to turn on the microwave tube, tube 1 is turned off and tube 2 turned on. At the end of the pulse, tube 2 turns off and tube 1 turns on. The gating pulses to each tube may be capacitively coupled, transformer-coupled, optically coupled, or coupled by means of RF energy. Floating-deck modulators are typically utilized with large klystrons, and swings of 200 kV are not uncommon for this application. There are a number of different means of implementing such floating-deck modulators; several of the more common implementations are summarized in Figure 3–5. Also, floating-deck modulators are sometimes

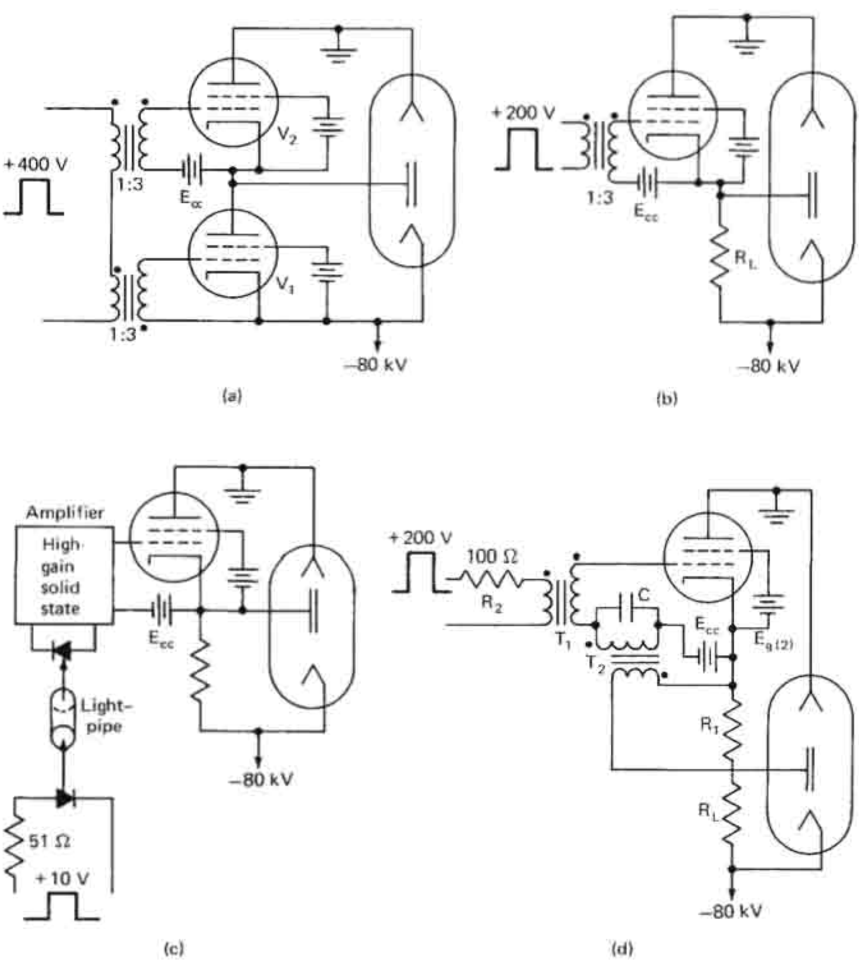


FIGURE 3–5 Various floating-deck modulator configurations [6]: (*a*) dual-tube, high-*L* coupling transformers; (*b*) single-tube, high-*L* coupling transformers; (*c*) light-pipe interface; (*d*) "blocking-oscillator" regenerative modulator.

used to cathode-pulse tubes directly when a very wide range of pulse widths with rapid rise and fall times must be accommodated.

Types of Hard-Tube Modulator Switches [5,19]

The switches that are typically utilized for hard-tube modulators may be triodes, tetrodes, or pentodes; however, mainly triode and tetrode tube configurations have been utilized for such applications.

Triodes

A triode vacuum tube consists of a cathode, which emits electrons, a grid, which controls the flow of electrons, and a plate, or anode, which receives the electrons. The total current flow in a triode tube is determined by the electrostatic field in the region of the cathode, which is in turn controlled by both the grid potential and the anode potential. The total cathode current of an ideal triode may be approximated as

$$I_k = k\left(E_c + \frac{E_b}{\mu}\right)^{3/2}$$

where I_k = cathode current
k = perveance
E_c = grid-cathode voltage
E_b = plate-to-cathode voltage
μ = amplification factor

One of the more important parameters associated with a triode is its voltage amplification factor μ. The μ of a triode can be approximated as follows:

$$\mu = \frac{\Delta E_b}{\Delta E_c} \qquad \text{with plate current held constant}$$

where ΔE_b = change in plate voltage
ΔE_c = change in grid voltage

Tetrodes

A tetrode is a tube containing four elements: the cathode, the control grid, the screen grid, and the plate or anode. Tetrodes are often utilized in high-power modulators as switch tubes because relatively small control voltage swings are required in order to control large amounts of power in the tube. This is due to the presence of the additional or screen grid between the control grid and the anode; the screen grid serves as an accelerating element and also screens the control grid from the effects of the anode. As in any vacuum tube, the field near the cathode controls the current through the tube; the current is given by

$$I_k = k\left(E_{c1} + \frac{E_{c2}}{\mu_s} + \frac{E_b}{\mu_p}\right)^{3/2}$$

where E_{c1} = control-grid voltage
E_{c2} = screen-grid voltage
μ_s = screen amplification factor
μ_p = plate amplification factor

The effect of the additional grid is to largely decouple the effect of changes in plate voltage from the cathode current; thus, a tetrode tends to operate as a constant-current device for a given value of control-grid voltage, regardless of the particular plate voltage applied. It should also be noticed that variations in screen-grid voltage can affect the current through the tube; therefore, the screen grid must either be driven from a well-regulated source, or bypassed sufficiently with a low-inductance capacitor so as to minimize screen-grid voltage variations during the time the tube is turned on.

Under certain conditions not uncommon in high-power tubes, the screen grid can actually become an emitter, and current will appear to flow out of the screen. It is essential that the tube designer provide a sufficiently low-impedance path for any such reverse electron flow in order to prevent a resultant increase in screen-grid voltage and possible damage to the tube. Thus, the screen-grid voltage supply must be well regulated, must be thoroughly bypassed, and must represent a low impedance to any reverse current that might flow out of the screen grid.

It is also possible for electrons to be emitted from the grid, particularly for tubes operated under conditions of high average power. For that reason, one should be prepared to accommodate reverse currents from the control grid when designing the control-grid driving circuit. Often tube elements are gold-plated in order to reduce secondary emission, and it should be kept in mind that the amount of secondary emission from both the control grid and the screen grid may increase during the life of the tube, particularly when the tube operates with oxide cathodes.

Operation of Vacuum Tubes in Hard-Tube Modulators [5]

Important in determining the suitability of a given tube for a particular application are the maximum values of plate voltage, grid voltage, grid dissipation, and plate dissipation. However, since it is difficult to give an analytical representation that connects grid voltage and current and plate voltage and current over a wide range of operating parameters, graphical representations are often utilized. In order to determine the interaction of the tube with the load, *static-characteristic* curves are normally utilized representing (1) plate current as a function of plate voltage with constant grid-drive curves as parameters, or (2) grid voltage as a function of plate voltage with constant plate and grid current curves as parameters.

In addition to designing for the interaction of the tube with the load, it is necessary to control the plate current during the interpulse period in order to control the overall anode dissipation of the tube. For this reason, a curve of plate voltage as a function of negative grid bias for cutoff conditions is normally given, as shown in Figure 3–6. This curve is particularly important, since even a relatively small current flow during the interpulse period may contribute substantially to overall anode dissipation, and because the negative bias voltage must be added to the positive grid-drive voltage in order to determine the total required grid

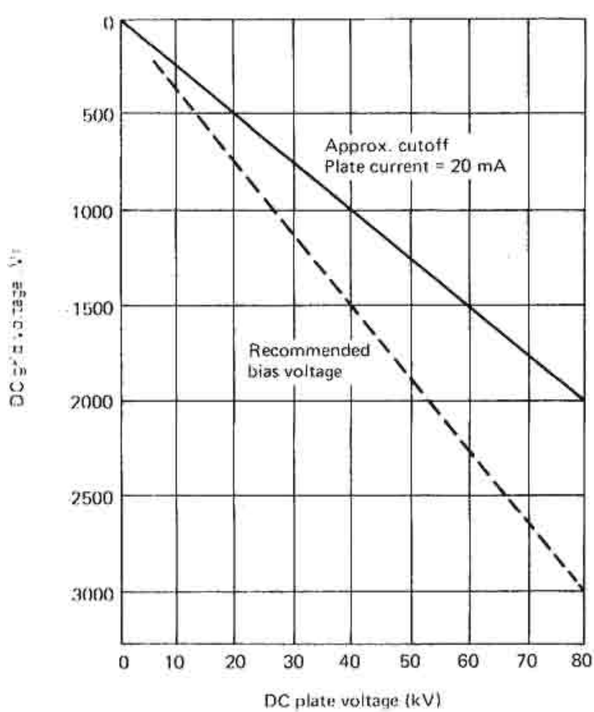


FIGURE 3–6 Grid bias versus plate voltage for cutoff condition for the ML-7560 triode tube. [5]

voltage swing. The required cutoff-bias voltage will vary with the plate voltage of a triode and with both the plate and the screen voltages of a tetrode. All high-voltage tubes may exhibit some field emission from the grid, and it should be noted that this field emission is independent of grid bias. Plate current flowing during the interpulse period may give rise to appreciable X-radiation, necessitating adequate shielding for personnel protection.

The constant–grid-drive voltage curves for a typical triode, the ML-7560 tube, are shown in Figure 3–7, which gives plate current and grid current as functions of plate voltage. Operation in the region at the left of the constant–grid-drive characteristic curves, on the so-called *diode line,* represents *saturated* operation of the tube.

The use of so-called *shielded grid tubes* can result in an appreciable reduction in grid current, as shown in the characteristic curves of Figure

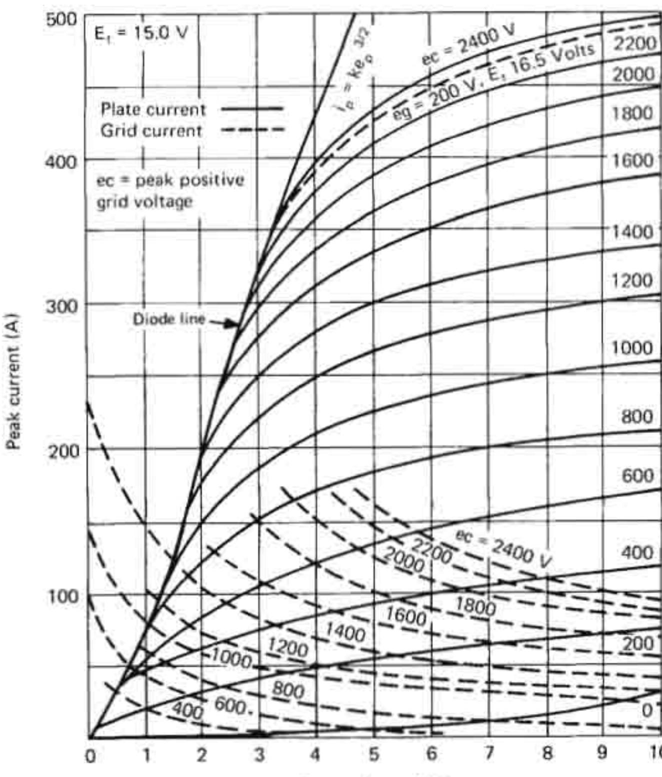


FIGURE 3–7 Constant–grid-voltage characteristic curves of the ML-7560 triode. [5]

3–8 for a shielded-grid triode. There are a number of advantages that favor the use of shielded-grid tubes at higher power levels, including the following: (1) grid-control current is reduced; (2) the shield-grid current is zero; (3) an arc to the shield grid will not transfer to the cathode; (4) grid dissipation is relatively low; (5) the structure is rugged;

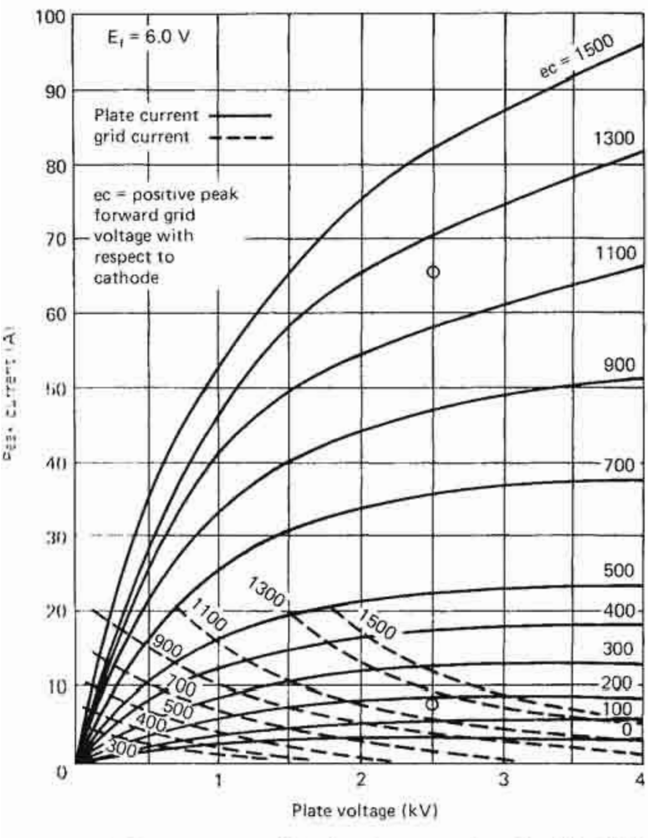


FIGURE 3–8 Constant-grid-voltage curves for the ML-6544 shielded-grid triode. [5]

(6) no screen-grid power supply is required; (7) the amplification factor is high; and (8) negative–grid-bias requirements are relatively low.

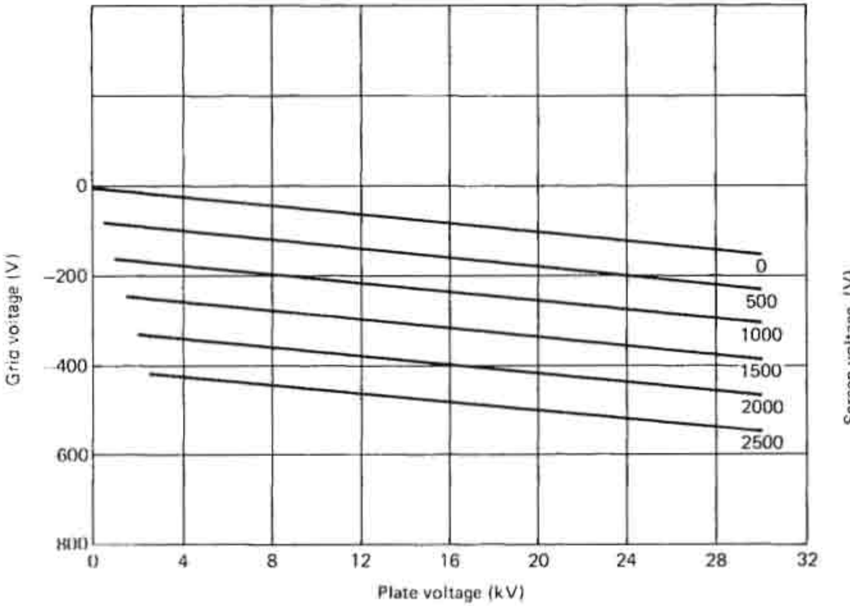
At low and moderate power levels, tetrode-type vacuum tubes are more often utilized as switches. Figure 3–9 shows a set of tube characteristics for the 4PR1000A tetrode; a screen voltage of 1500 V was used for the constant–grid-voltage and constant-current characteristics. Note that the required grid bias for cutoff is a function of both the screen and plate voltages. The screen grid shields the cathode from the plate potential, with the result that cutoff bias is determined primarily by the screen-grid voltage. The total grid voltage required to obtain rated cathode emission is less in a tetrode than in a triode, and relatively little control-grid current is drawn. The grid current in a tetrode is not appreciably affected by plate voltage variations, but if the plate voltage is reduced sufficiently, this may no longer be the case. Of course, in a tetrode, appreciable screen-grid current is usually drawn when the control grid is driven positive, and both a low-impedance power supply and sufficient capacitive decoupling are required in order to achieve stable operation over a wide range of conditions.

Occasionally, pentodes or specially designed tetrodes are utilized in high-power modulators, usually for those applications where their approximately constant-current capabilities are a particular advantage [1]. In such applications, the relatively high tube drop required in order to achieve constant-current operation often necessitates an undesirable amount of switch-tube power dissipation and considerably increased dc supply voltages.

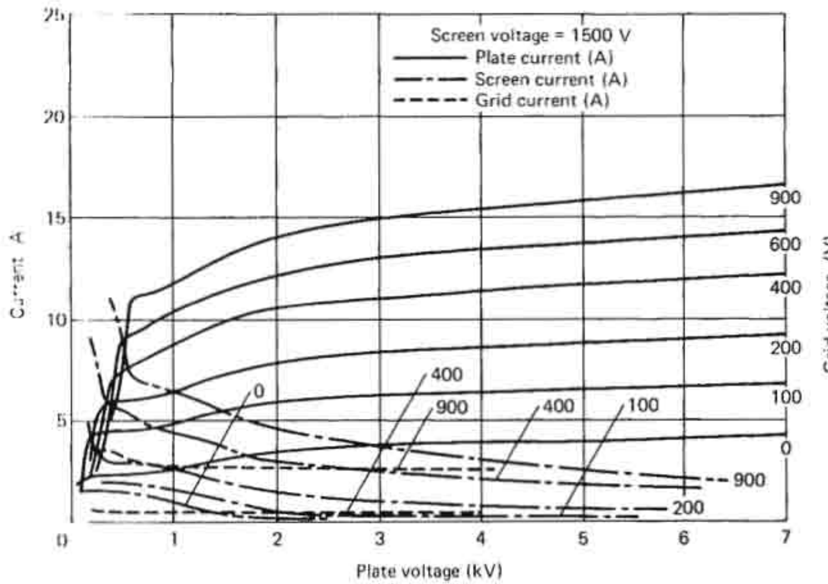
Whether a triode, tetrode, or pentode is selected for the switch tube during the design process, it is first necessary to specify the required pulse voltage and pulse current for the load. If a pulse transformer is utilized, then the peak pulse power can be used to select the tube and the turns ratio adjusted to match the tube ratings. If it is desired to capacitively couple the pulse-modulator tube directly to the load, then a modulator tube must be selected such that the maximum dc plate voltage is approximately 10% greater than the required pulse voltage delivered to the load, and the tube must be capable of switching the full output pulse current.

Once a tube has been selected, the next step is to determine the cutoff bias required during the interpulse period in order to limit plate dissipation to a reasonable value. This value then determines the cutoff condition for the tube. One next picks a positive grid drive from the grid voltage versus plate voltage characteristic curves so that the tube can deliver the required plate current to the load; this is the switch ON condition, and it involves a compromise between low plate voltage and grid-drive power requirements. One should check carefully at this point to ensure that neither the control- nor screen-grid dissipation has been exceeded. For triodes, the operating point will normally be where the grid current is from 10 to 30% of the peak plate current. In tetrodes, the operating points should be picked so that the screen-grid current is not excessive and so that screen-grid dissipation remains within safe limits.

The choice of operating point is particularly critical in considering

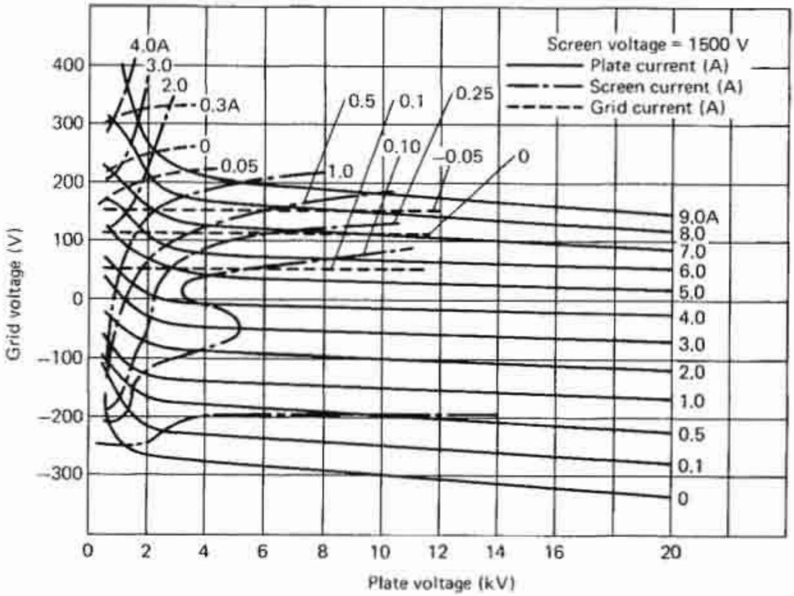


(a)



(b)

FIGURE 3–9 Tube characteristics for the 4PR1000A tetrode switch tube: (*a, top*) typical plate cutoff characteristics for less than 50 μa plate current; (*b, bottom*) typical constant–grid-voltage characteristics; (*c, top right*) typical constant-current characteristics. (*Eimac*)



(c)

the effects of changes in high voltage and grid drive on the pulse delivered to the load. These dependences may be treated by using conventional load-line analysis, using a simplified set of static tube characteristics. Such a set of simplified static characteristics and load lines is shown in Figure 3–10 for saturated operation below the knee of the plate-current curve, and for an unsaturated operation above the knee of the plate-current curve. The two sets of load lines shown in each of the diagrams of Figure 3–10 are the straight load line, which connects E_{bb} and OP 1, valid for resistive loads, and the bent load line, connecting E_{bb}, E_d, and OP 1, which is valid for a biased-diode load such as a magnetron.

If adequate positive grid-drive voltage is available to keep the operating point for the tube OP 1 saturated [below the knee of the characteristic curve, as is shown in Figure 3–10(*a*)], irregularities in the top of the grid-driving pulse will not be observed in the output pulse. However, since such operation results in high grid currents in a triode and high screen-grid currents in a tetrode, care must be taken to keep the grid dissipation within appropriate limits. The disadvantage of operating the tube saturated is that any changes in E_{bb} are transmitted directly to the load, thus necessitating a well-regulated high-voltage supply.

If the tube is operated in an unsaturated region, as shown in Figure 3–10(*b*), and the grid-drive voltage changes over the range from C to

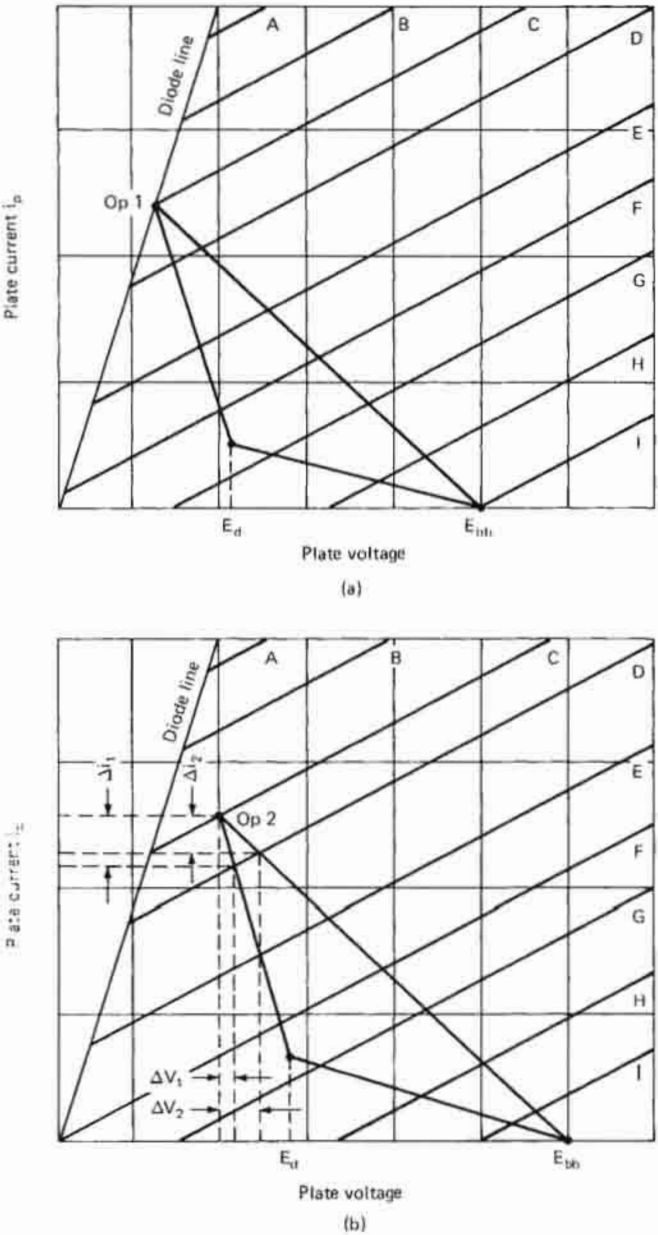


FIGURE 3–10 Idealized tube characteristics and load lines for hard-tube modulator operation with resistive and biased-diode loads showing (*a*) saturated (below knee of curve) and (*b*) unsaturated (above knee of curve) operation. [5]

D, then the operating point changes by an amount that depends upon the nature of the load (whether biased-diode or resistive). Changes in load current are greater for the low-dynamic-resistance biased-diode load ($\Delta i_1 > \Delta i_2$), while the change in the output voltage is less than would be the case for a purely resistive load ($\Delta V_1 < \Delta V_2$). Thus, when the switch tube is operated in an unsaturated condition, irregularities in the grid-voltage drive pulse are transferred to the load.

Expressions for plate current as a function of various tube parameters may be differentiated in order to obtain the sensitivities of plate current to changes in other circuit parameters. An important utilization of these relationships is to determine the changes in load current associated with changes in plate voltage. For the triode operating above the knee of the circuit and a magnetron load, the appropriate relationship is

$$\frac{di_p}{i_p}=\frac{1}{1+(\mu_p e_g-E_d)/E_{bb}}\,\frac{dE_{bb}}{E_{bb}}$$

and for a tetrode,

$$\frac{di_p}{i_p}=\frac{1}{\dfrac{\mu_p(e_g+E_{sg}/\mu_{sg}-E_d/\mu_p)}{E_{bb}}+1}\,\frac{dE_{bb}}{E_{bb}}$$

where i_p = instantaneous value of plate current
e_g = instantaneous value of grid voltage
μ_p = plate voltage-amplification factor
μ_{sg} = screen-grid voltage-amplification factor
E_d = magnetron diode voltage
E_{sg} = screen-grid voltage
E_{bb} = plate supply voltage

For the triode or tetrode operating below the knee of the curve, the appropriate equation becomes

$$\frac{di_p}{i_p}=\frac{1}{1-E_d/E_{bb}}\,\frac{dE_{bb}}{E_{bb}}$$

To compare hard-tube modulator regulation for saturated and unsaturated switch tube operation, consider a magnetron load that requires at 11 kV and 10 A. Assume E_d will be 9.7 kV. For a 4PR60 switch tube operating below the knee of the curve, the plate voltage E_{bb} will be approximately 11.7 kV, the screen-grid voltage 1250 V, and the grid voltage 0; then

$$\frac{di_p}{i_p}\approx 5.85\,\frac{dE_{bb}}{E_{bb}}$$

For operation above the knee of the curve, E_{bb} will be approximately 13 kV, μ_{sg} approximately equal to 4, and μ_p approximately equal to 70. E_g will equal −35 V, and

$$\frac{di_p}{i_p} \approx 0.57 \frac{dE_{bb}}{E_{bb}}$$

Thus, regulation in plate current due to variations in supply voltage is improved by a factor considerably greater than 1 when operating in an unsaturated condition. Of course, for operation in an unsaturated mode it is necessary to carefully regulate both the control-grid and screen-grid voltages. A similar comparison with a pure-resistance load ($E_d = 0$) would result in a ratio of approximately 2.5, rather than the ratio of slightly greater than 10 achieved above with a biased-diode load.

The use of grid-drive saturation to give flat-top pulses is normally confined to low- to medium-power applications, since it often produces excessive screen-grid dissipation in tetrodes and excessive control-grid dissipation in triodes. However, if a shielded-grid triode is utilized, grid-drive saturation is usually satisfactory even in higher-power tubes, because the grid is capable of handling the required dissipation. In a tetrode, the flow of pulse current to the screen grid and the plate–screen-grid capacitances necessitate the use of a large bypass capacitor between the screen grid and the cathode. If fast-rising pulses are utilized, the self-inductance of the capacitor must be small enough to present a low impedance at the maximum useful frequency component of the pulse. The sensitivity of output voltage to screen-grid voltage in a tetrode usually necessitates the regulation of this power-supply voltage; it is sometimes possible to compensate for changes in plate voltage by adjusting the screen-grid voltage, which may be simpler than regulation of the plate power-supply voltage.

In the event of arcing, it is necessary to limit arc dissipation within the tube to a few joules or less, and at low to medium power levels sufficient series resistance may suffice for this purpose; however, at power levels above 10 kW, it is essential to use fast-acting crowbar circuits to divert stored energy from the tube to a safe discharge path. This crowbar circuit must act in less than approximately 10 μs, to divert energy from an arcing tube to a shunt circuit and to prevent damage to the switch tube. This energy diverter must, in general, be some form of gaseous-discharge device such as a thyratron or a spark gap, so that it will have a low internal impedance. It is also necessary to utilize fast-acting circuit breakers, since once the crowbar fires, energy will be fed in from the lines until the primary power is removed. It should be emphasized that the design of the crowbar circuitry must be such that the discharge through the crowbar is critically damped. If a resonant, underdamped condition exists, damage to the tube may be encountered even if a crowbar circuit is utilized, because the oscillatory discharge may cause the crowbar to deionize before the stored energy is dissipated.

In order to protect the switch tube from transients produced when the load arcs, the control grid may be clamped back to the bias whenever an arcing condition is sensed. Of course, the plate current should not be cut off too abruptly; otherwise large transient plate voltages associated with the interruption of current through an inductor may be generated. A thyratron in the switch-tube grid circuit, as shown in Figure 3–11,

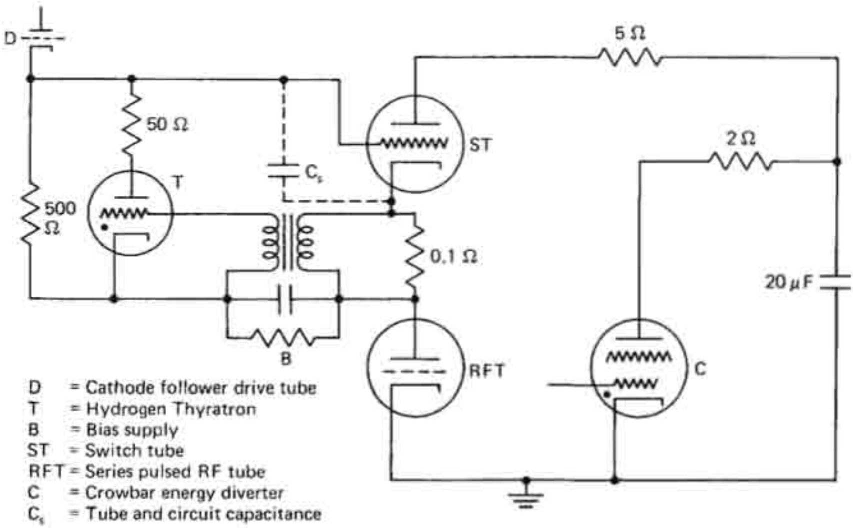


FIGURE 3–11 Circuit with thyratron in switch-tube grid circuit to ensure tube cutoff in the event of an arc in the load. [5]

with a proper time constant can be of considerable value in protecting high-power switch tubes. In fact, with this circuit it is sometimes possible to shut off the switch tube without using the crowbar to short the plate power supply when the load device arcs.

Output Waveforms

The loads analyzed up to this point were either purely resistive or of a biased-diode nature; of course, any realistic application involves the consideration of the effects of various reactive elements associated with or connected to the load. Various stray capacitances and inductances and additional reactive elements affect peak currents and voltages in the switch tube and the load and also strongly influence the achievable rise times, fall times, and flatness of pulse. The analysis of the output-pulse shape is simplified by subdividing the pulse into regions, as shown in Figure 3–12.

A particularly simple case is that of a resistive load with some associated stray capacitance, giving rise to the equivalent circuit shown in

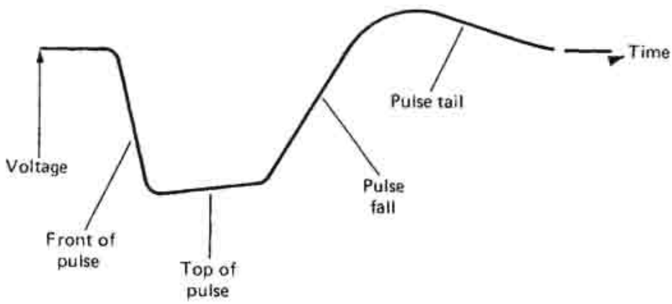


FIGURE 3–12 Hard-tube modulator output pulse, showing principal parts of the pulse.

Figure 3–13. In Figure 3–13, the vacuum switch tube is represented as an equivalent plate resistance R_p and an ideal switch, while the load is represented by a load resistance R_l in parallel with some stray capacitance C_s. For calculation of the rise time, assume that the value of C_c, the coupling capacitor, is much larger than C_s, and that R_p is much

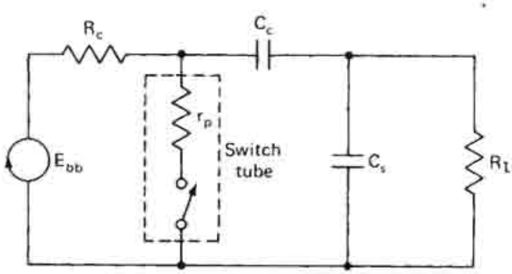


FIGURE 3–13 Simplified equivalent circuit for a hard-tube modulator.

less than the value of the recharging resistor R_c. For this configuration, the equivalent charging time constant is

$$C_s \frac{r_p R_l}{r_p + R_l}$$

which gives a rise time (from 10% to 90% of full amplitude) of

$$t_r = 2.2 C_s \frac{r_p R_l}{r_p + R_l}$$

The top of the pulse can be analyzed by assuming that the voltage droop is relatively small, enough that C_s can be essentially disregarded, and the appropriate time constant is then given by

$$C_e(R_l + r_p)$$

and the corresponding current droop is given by

$$\text{Droop} \approx \frac{\tau}{C_e(R_l + r_p)}$$

where τ is the pulse duration.

Once the switch opens, the fall time may be calculated by noting again that the value of C_e is much greater than C_s, so that the time constant for discharge is dominated by the time constant determined by R_l and R_e in parallel and C_s. Thus the fall time (from 90% to 10% of full charge) is given by

$$t_f = 2.2\,C_s \frac{R_l R_e}{R_l + R_e}$$

Note that in general the fall time is substantially longer than the rise time for such a resistive load.

Next, consider a biased diode with a resistor R_s in parallel with the load in order to permit the recharge of the coupling capacitor. An equivalent circuit for the modulator is given in Figure 3–14(a). If one assumes that during the rise time R_e has a value much greater than R_p and that the biased-diode switch does not close until almost the full output voltage is reached, then an equivalent circuit for the rise time is as given by Figure 3–14(b). The time constant for the rise time is dominated by the stray capacitance and an equivalent resistance given by the parallel combination of R_s, R', and r_p. In general, r_p is by far the smallest of these resistances, so that the 10% to 90% rise time is given by

$$t_r = 2.2\,C_s r_p$$

The analysis of the top of the pulse is critically dependent upon particular circuit parameters. However, if the change in output load current is small, then this change in current is

$$\text{Current droop} = \frac{\Delta I}{I_o} \approx \frac{\tau}{C_e r_d}$$

For the fall of the pulse, the stray capacity and the parallel combination of R_s, R_e, and r' dominate, since the switch is open. Thus, the fall time is

$$t_f = 2.2\,C_s(R_s \,\|\, r' \,\|\, R_e)$$

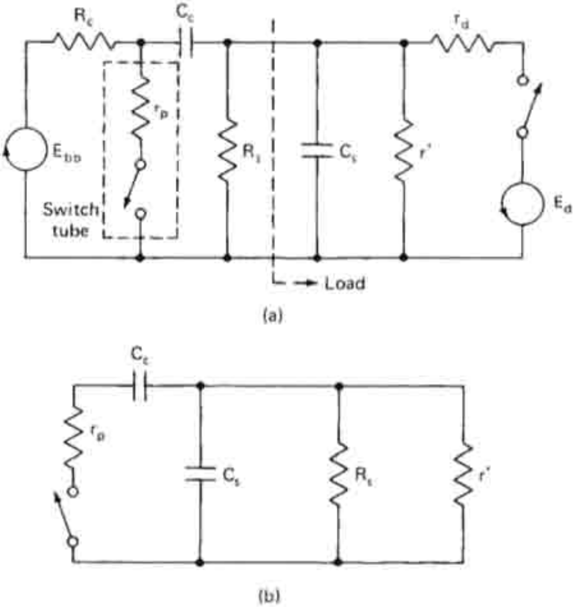


FIGURE 3–14 Equivalent circuit for (*a*) hard-tube modulator with biased-diode load and resistive recharging and (*b*) rise time calculation.

The third commonly encountered configuration utilizes a biased-diode load with an inductive recharge path. An equivalent circuit for such a configuration is given by Figure 3–15(*a*). The equivalent circuit for the rise time is given by Figure 3–15(*b*); since it is usually good design practice to choose L_s so that its initial current is small, its contribution during the rise time may be ignored. Thus the rise time is

$$t_r = 2.2 C_s \frac{r_p r'}{r_p + r'}$$

During the top of the pulse, again assume that the droop is small enough that the equivalent circuit is that given by Figure 3–15(*c*). Then the current drawn from the capacitor may be approximated by a constant I_o and the change in output current may be approximated by

$$I_o = \tau \frac{I_o + I_m/2}{r_d C_c}$$

The performance of the equivalent circuit valid for the tail of the pulse given by Figure 3–16 is a relatively complex function of a number of parameters. If an air-core inductor is used so that linear circuit analysis

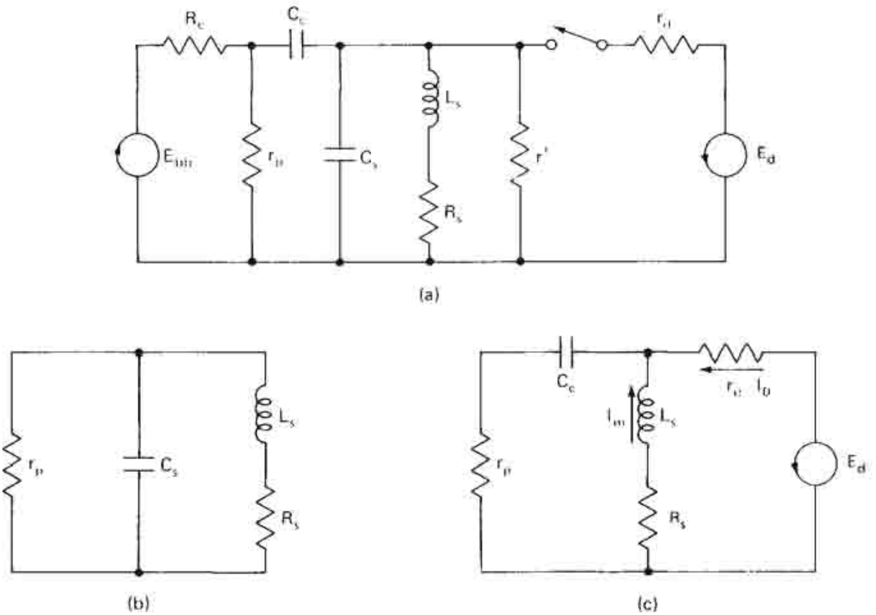


FIGURE 3–15 (*a*) **Equivalent circuit for hard-tube modulator with biased-diode load and inductive recharging path. Simplified circuits are given in** (*b*) **valid for the rise time and in** (*c*) **valid for the top of pulse.**

may be employed, then the tail-of-pulse performance is primarily a function of circuit damping and the ratio of current in the inductor to the load current at the end of the pulse. Figure 3–17 shows underdamped (oscillating), critically damped, and overdamped responses for increasing values of Δ, which is the ratio I_m/I_o, illustrating the decreased fall time but increased value of backswing associated with increased values of Δ. In general, the underdamped condition should be avoided to prevent any spurious outputs if the cathode voltage should go negative, and large values of backswing voltage should be avoided to reduce any chance of tube arcing. It is desirable to have the voltage fall to zero quickly

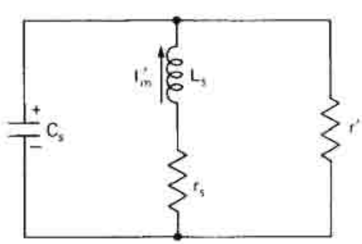


FIGURE 3–16 Equivalent circuit for the hard-tube modulator of Figure 3–15 which is useful for calculation of the fall of the pulse.

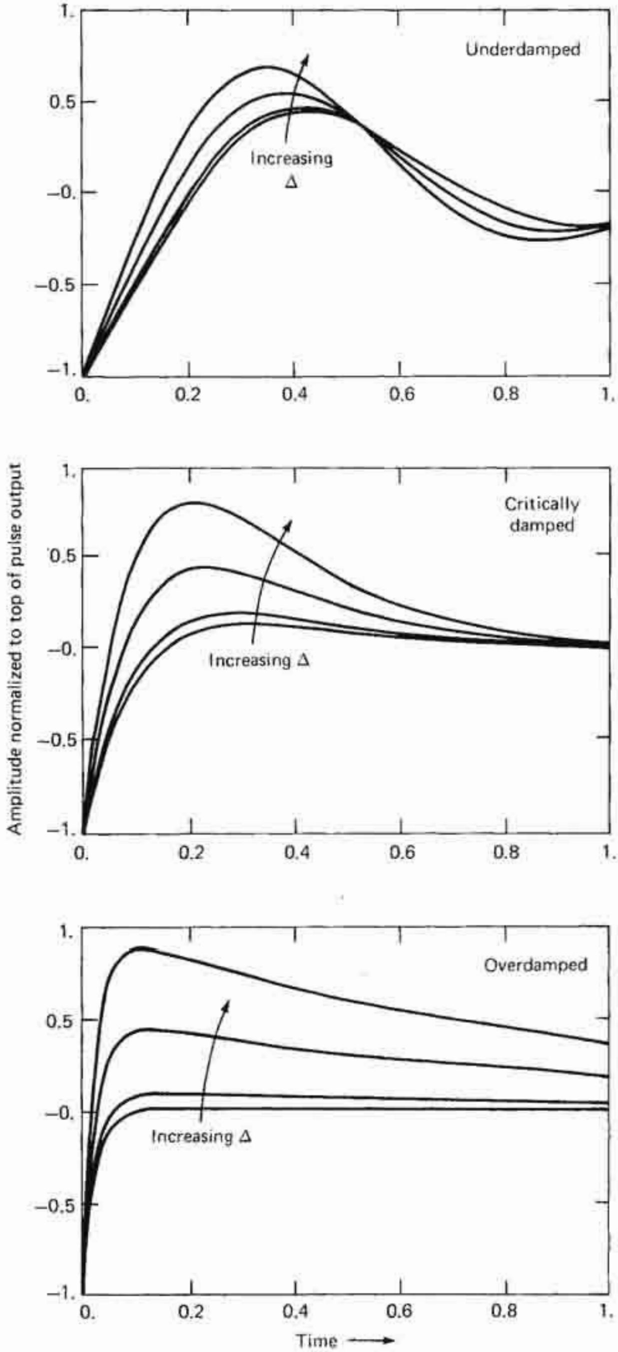


FIGURE 3–17 Tail-of-pulse responses for the circuit of Figure 3–16 for underdamped, critically damped, and overdamped conditions, illustrating effects of increasing Δ, the ratio of inductor current to load current at the end of the pulse top. Plots are for a negative pulse, and are normalized to the pulse voltage at the end of the pulse.

but to have little backswing and not to become negative again (to prevent spurious oscillations). One approach to achieve this is to provide enough damping that the voltage never recrosses the baseline; this corresponds to the critically damped condition

$$\frac{R_s}{L_s} = \frac{2}{\sqrt{C_s L_s}}$$

In practice, achieving this desirable condition often requires an excessively long fall time. Thus, some small amount of negative overshoot is typically allowed. The amount of negative overshoot is dependent on the specific circuit parameters and the initial current flowing in the inductor. Smaller values of L_s, and higher values of inductor current at the end of the pulse, result in faster fall times but larger values of overshoot, as shown in Figure 3–17.

It may be that requirements for fast fall time and minimum negative overshoot with no recrossing of the axis cannot be met simultaneously. In this case, it is often necessary to incorporate a clipping diode in order to limit the backswing on the output voltage. In some cases, it may be desirable to incorporate additional coupling elements to improve output pulse shape [16,17]. In cases where extremely rapid fall time is required, a "tail biter," which is an active switch connected directly across the load, may be used to rapidly discharge the stray capacity.

3–2 LINE-TYPE MODULATORS

The Pulse-Forming Network (PFN)

The line-type modulator derives its name from the similarity of the behavior of its energy storage element to that of an open-circuited transmission line [8,11]. If a length of transmission line having one-way propagation $\tau/2$ is connected as in Figure 3–18, charged to voltage V, and then discharged through its characteristic impendance Z_0, a pulse length of τ and amplitude $V/2$ is generated across the load. For practical

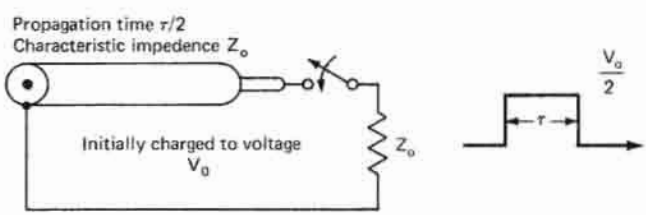
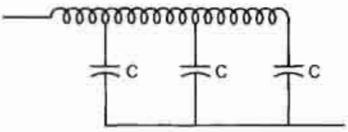


FIGURE 3–18 Generation of a rectangular pulse by discharge of a charged open-circuited transmission line into its characteristic impedance.

pulse lengths and voltages, the required bulk of cable often becomes excessive, and in practice a network of lumped inductors and capacitors is often used. Such a network is shown in Figure 3–19 and is called a *pulse-forming network* (PFN). The PFN resembles the lumped-circuit approximation of the actual transmission line, but there are some significant



Important relationships are:

$$Z_o = \sqrt{\frac{L_n}{C_n}} \qquad L_n = \frac{\tau Z_o}{2}$$

$$\tau = \sqrt{L_n C_n} \qquad \frac{l}{d} = \frac{4}{3}$$

$$C_n = \frac{\tau}{2Z_o} \qquad \frac{L_e}{L} = 1.1 \text{ to } 1.2$$

where C_n = total network capacitance
L_n = total network inductance
τ = pulse width at 50% points
Z_o = characteristic impedance
n = number of sections
L = inductance per section L_n/n
L_e = inductance of section on closed end
C = capacitance per section C_n/n
l = length of coil in one section
d = coil diameter

FIGURE 3–19 Guillemin E-type pulse-forming network circuit arrangement and key design parameters.

differences. The PFN shown in Figure 3–19 is sometimes called the Guillemin E-type network. It consists of equal-valued capacitors and a continuously wound, tapped coil whose physical dimensions are chosen so as to provide the proper mutual coupling at each mesh. The total capacitance and inductance are given by

$$C_n = \frac{\tau}{2Z_o}$$

$$L_n = \frac{\tau Z_o}{2}$$

where C_n = total network capacitance
L_n = total network inductance
τ = pulse width at 50% points
Z_o = characteristic impedance

The number of sections is chosen to provide the desired rise time.

Operation of a Line-Type Modulator

The load for a line-type modulator often requires extremely high voltages. In order to reduce the voltages on the PFN, a step-up transformer is often inserted between the PFN and the load. Bifilar secondary windings on the step-up transformer often provide a convenient means to provide heater current for the transmitting device.

A device must be used to isolate the switch from the power source. A resistor could be utilized, but it would limit the maximum efficiency to 50%. An inductor is often utilized because of the increased efficiency obtained, and because it is then possible to charge the PFN to approximately twice the dc supply voltage. A charging diode is often utilized to prevent discharge of the PFN once it is charged. A more typical schematic diagram of a line-type pulser is shown in Figure 3–20; it includes a shunt diode and resistor to damp out any reflected voltage due to load mismatch.

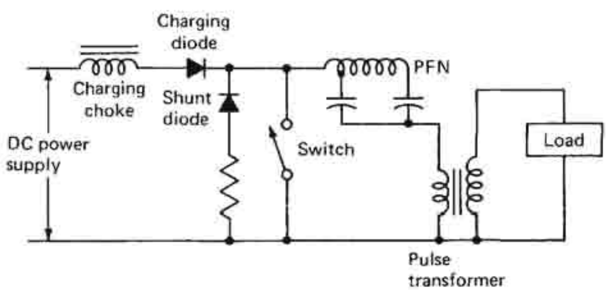


FIGURE 3–20 Simplified schematic representation of a typical line-type modulator. PFN = pulse-forming network.

Analysis of the operation of a line-type modulator can be perhaps best understood by breaking the operation into subintervals: a recharge interval, during which time the energy stored in the PFN is replenished, and the discharge interval, during which time the energy stored in the PFN is discharged into the load, normally through the pulse transformer.

Recharge Interval

Since the energy-storage element is completely discharged on each pulse, some means of efficiently replenishing this energy must be provided. It is conceptually possible to incorporate charging through a resistor from a dc power supply, but efficiency with such an arrangement is limited to no more than 50%. A more customary means of replenishing the PFN charge is by so-called *inductive charging.*

Analysis of the charging circuit of a line-type modulator may be accomplished by assuming that the equivalent series inductance of the energy-storage element is relatively small so that the equivalent circuit shown in Figure 3–21 suffices for analysis. If one assumes that C_n is initially

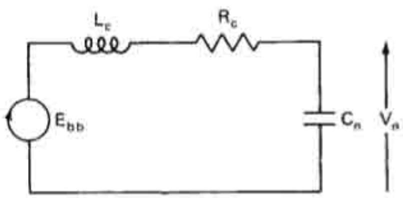


FIGURE 3–21 Equivalent circuit for inductive charging of PFN.

discharged and the current in L_c has not yet begun to appreciably increase at the beginning of the recharge interval, then an expression for the voltage is given by

$$V_n(t) = E_{bb} + e^{-at}[-E_{bb}(\cos \omega t + \frac{a}{\omega} \sin \omega t)$$

where $a = \dfrac{R_c}{2L_c}$

$$\omega_o = \frac{1}{\sqrt{L_n C_n}}$$

$$\omega = \sqrt{\omega_o^2 - a^2} = \sqrt{\frac{1}{L_n C_n} - \frac{R_c^2}{4L_c^2}}$$

The natural oscillations are thus damped sinusoids; however, note that the first peak is only slightly less than twice the supply voltage and that the resonant frequency is given by

$$\frac{1}{2\pi \sqrt{L_c C_n}}$$

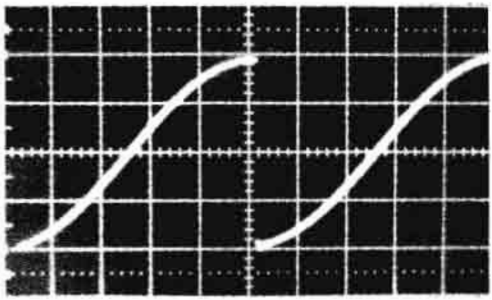


FIGURE 3–22 PFN voltage charging waveform for resonant charging condition, when the switch closes at the peak of the charging waveform.

If one picks the pulse-repetition frequency equal to twice the resonant frequency, then

$$\mathrm{PRF} = \frac{1}{\pi\sqrt{L_c C_n}}$$

and the peak voltage on the network is approximately twice the dc supply voltage. In order to achieve maximum voltage output for minimum dc power-supply voltages, it would be desirable to close the switch at the peak of the charging voltage, as shown in Figure 3–22; this is called *resonant charging*. However, this would then require complete control of the interpulse period. More flexibility may be achieved if a series, or *charging diode* is incorporated in series with the charging inductor, giving rise to a *subresonant charging* voltage waveform of the form shown in Figure 3–23. For this condition, the peak voltage is normally between 1.9 and 2.0 E_{bb}, for an inductor Q greater than 10.

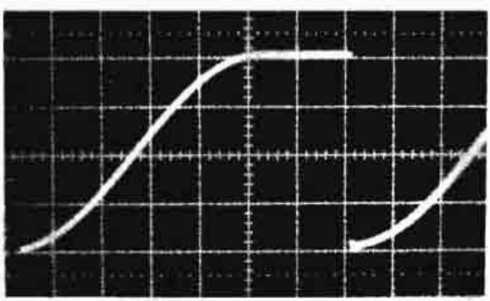


FIGURE 3–23 PFN voltage charging waveform for subresonant charging, with a charging diode used to prevent discharge of the PFN until the switch is closed.

The average current is given by

$$I_{c(\mathrm{av})} = \mathrm{PRF}\, C_n V_{n(\mathrm{peak})}$$

and the ratio of rms to average current for resonant charging by

$$\frac{I_{c\,(\mathrm{rms})}}{I_{c\,(\mathrm{av})}} = 1.11$$

and the ratio of peak to average current by

$$\frac{I_{c\,(\mathrm{max})}}{I_{c\,(\mathrm{av})}} = 1.57$$

If subresonant charging is used, then these values must be adjusted appropriately.

The charging efficiency of such a line-type modulator is given by

$$\eta_c = 1 - \frac{\pi}{4Q}$$

where $Q = \dfrac{\text{energy stored per cycle}}{\text{energy lost per cycle}}$

This charging efficiency is the ratio of the energies extracted from the power supply to the energy transferred to the pulse-forming network.

In the event the load arcs, a negative initial voltage may be reflected to C_n. In this case, the peak charging voltage may exceed twice the dc supply voltage, and may be limited by the Q of the charging choke. Such high voltages are undesirable and may result in damage to the modulator, the load, or both. For this reason, a shunt, or clipping, diode, is often connected across the switch to remove any residual voltage present on C_n.

Discharge Interval

In analysis of the discharge interval, we assume that the capacitors in the pulse-forming network have been charged to their peak voltage, and that the switch is closed. If the PFN then has the basic characteristics of a transmission line, if the pulse transformer is ideal, and if the load is purely resistive and matched to the transmission line, then a rectangular output pulse will be obtained.

In the event that the load does not match the characteristic impedance of the transmission line, the pulse generated will no longer be rectangular, but will have steps, as shown in Figure 3–24. The middle waveform in Figure 3–24 for the condition where the termination is larger than the characteristic impedance of the line, showing a series of same-polarity steps, while the bottom waveform shows the condition where the load impedance is less than the characteristic impedance of the line. It is customary to mismatch the load slightly in order to promote formation of a slight negative voltage on the switching device in order to enhance recovery of the switch.

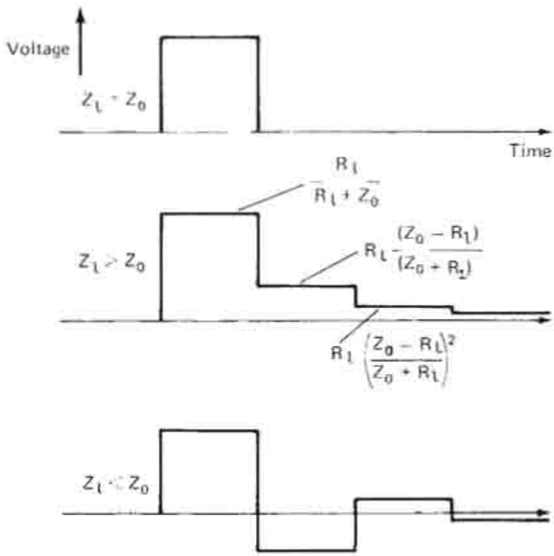


FIGURE 3–24 Discharge waveforms produced by charged transmission lines for various terminating impedances.

Unfortunately, (1) the load is rarely a pure resistance, biased-diode loads such as magnetrons of CFAs being more common; (2) the pulse transformer is not ideal, but has some magnetizing inductance, leakage inductance, and stray capacitance; and (3) the pulse-forming network is not an ideal transmission line, but is a lumped constant network having certain specific properties.

As indicated earlier, the choice of pulse width and impedance level determines the properties of the pulse-forming network. The pulse width is usually set by system requirements, but the impedance level of the PFN can be varied over a fairly wide range by the choice of the output-transformer turns ratio. A choice of too high an impedance level results in high voltages on the pulse-forming network, while utilization of extremely low impedance levels results in high currents, high switch-tube drops, and losses in the pulse transformer primary, which make it difficult to achieve efficient operation. When hydrogen thyratron switches are used, pulse-forming-network impedance levels of 25 and 50 Ω have proved to be a reasonable compromise between high currents and high voltages, and a variety of hydrogen thyratrons have been designed so that their power-handling capability is maximized at the 50- or 25-Ω impedance level. However, if silicon controlled rectifiers (SCRs) are being utilized as the switch for the line-type modulator, a lower-impedance PFN may be indicated, because of the limited peak forward voltage–handling capability of most SCRs. It is normal practice to slightly mismatch the PFN to the load in order to reflect a small negative voltage onto the switch to promote rapid turnoff.

An important characteristic of the modulator is its regulation for changes in line voltage and with the load. Transmission-line theory indicates that a relatively simple equivalent circuit will permit a meaningful calculation of regulation during the principal output pulse; such an equivalent circuit for the line-type modulator is given in Figure 3–25. In

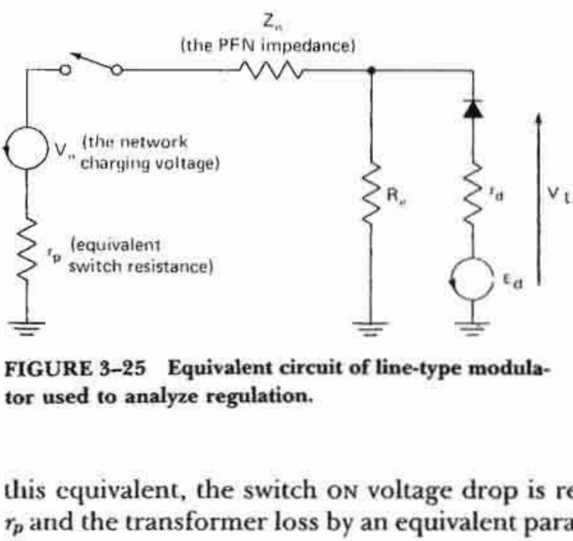


FIGURE 3–25 Equivalent circuit of line-type modulator used to analyze regulation.

this equivalent, the switch ON voltage drop is represented by resistance r_p and the transformer loss by an equivalent parallel resistance R_e. Analysis of this circuit is given in Glascoe and Lebacqz [8], where the expression for the load voltage is given by

$$V = \frac{V_n r_d + E_d Z_n \beta}{Z_n \beta + r_d \alpha}$$

where $\alpha = 1 + \dfrac{Z_n + r_p}{R_e}$

$$\beta = 1 + \frac{r_p}{Z_n}$$

and voltage regulation for changes in network voltage is obtained by taking the derivative

$$\frac{dV_l}{dV_n}=\frac{r_d}{Z_n\beta + r_d\alpha}$$

which can be rearranged to give the change in load voltage for a change in PFN charging voltage as

$$\frac{\Delta V_l}{V_l}=\frac{1}{1+(E_d Z_n/V_n r_d)\beta}\,\frac{\Delta V_n}{V_n}$$

If the modulator is operated under conditions which result in maximum power transfer to the load, i.e., if

$$\frac{V_s}{V_n}=\frac{1-r_d\alpha/Z_n\beta}{2\alpha}$$

the equation for voltage regulation becomes

$$\frac{\Delta V_l}{V_l}=\frac{2}{1+Z_n\beta/r_d\alpha}\,\frac{\Delta V_n}{V_n}$$

In a similar manner, an expression for current regulation for changes in network voltage is given by

$$\frac{\Delta I_l}{I_l}=\frac{1}{1-(E_d/V_n)\alpha}\,\frac{\Delta V_n}{V_n}$$

In the event that no dc regulation or charging regulation is used, this change in network voltage will be equal to the change in line voltage; however, if suitable circuits are incorporated to improve the regulation of the network voltage, this regulation must be included in the calculations to obtain the overall change in voltage and current with changes in line voltage.

It is interesting to evaluate some changes for representative choices of parameters. If

$$Z_n/r_d = 10$$

$$\alpha=\beta=1$$

then

$$\frac{\Delta I_l}{I_l}=1.82\,\frac{\Delta V_n}{V_n}$$

Thus, for a biased-diode load such as a magnetron, the changes in load current may be substantially worse than the changes in input voltage.

The output pulse produced at the load of a line-type modulator deviates from the ideal rectangular pulse because of a number of factors. A photograph of the output pulse associated with a line-type modulator driving a magnetron load is given in Figure 3–26, which illustrates the principal regions to be considered in analyzing the output pulse: the

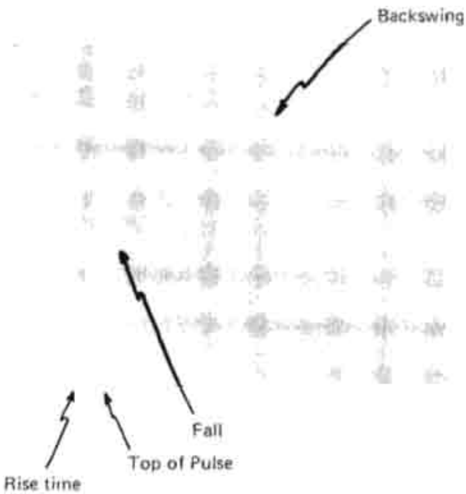


FIGURE 3–26 Line-type modulator output pulse showing principal portions of the pulse.

rise time, the top of the pulse, the fall time, and pulse backswing. It is usually desired that the rise time be sufficiently short for reasonable pulse shape, but not so short as to introduce moding or starting problems in the load. The top of the pulse should be reasonably flat and free from oscillations that would introduce frequency modulation through pushing, and the fall of the pulse should remove the voltage from the magnetron fairly rapidly and should not have excessive backswing, which could break down the tube. The tube voltage should not significantly recross the zero line, since even small negative voltages applied to the cathode may produce low-level RF noise outputs.

Detailed calculations of the pulse shape of a line-type modulator can be quite involved, but for many cases of interest, the particular choice of circuit constraints in a well-designed modulator may permit a considerable simplification. The analysis of the circuit is simplified, as was the case for the hard tube modulator, by breaking the circuit down into three time intervals: the rise of the pulse, the flat top, and the fall of the pulse.

While, in general, the analysis of the rise-time behavior involves the solution of a moderately complex network, for a number of situations one may consider that the pulse-forming network and the pulse transformer are two independent components such that the rise time of the overall combination is given by the square root of the sum of the squares of their rise times.

$$t_r = \sqrt{t^2_{r(\mathrm{PFN})} + t^2_{r(\mathrm{xfmr})}} \tag{3-1}$$

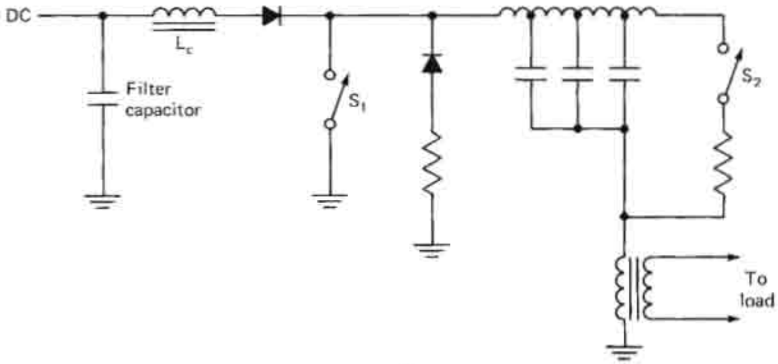
Behavior of the top of the pulse is simplified by the fact that most well-designed modulators produce approximately flat-top pulses, and the dominant circuit element that influences any droop on the top of the pulse is the magnetizing current of the transformer; it can be shown that for a biased-diode load, the change in load current is approximately equal to the magnetizing current. It is generally considered good design practice to design for approximately 10% of the load current to constitute magnetizing current at the end of the pulse; it is usually the case that this amount of current droop can be compensated for by adjustment of the pulse-forming network in order to obtain a reasonably flat top on the pulse.

Analysis of the tailing edge of the pulse is complicated by the nonlinear behavior of the pulse transformer and by the opening of the switch in the primary circuit during the recovery interval; since the details of this analysis are intimately bound with the design of the pulse transformer, detailed consideration of this behavior will be deferred until Chapter 4, on pulse transformers.

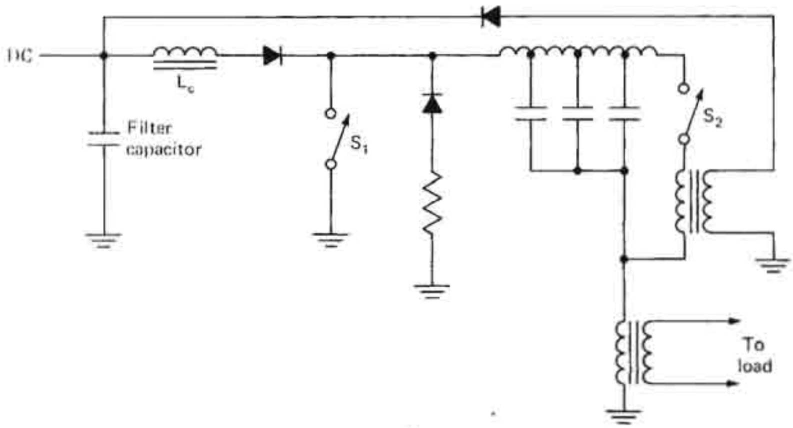
The Variable Pulse-Width Line-Type Modulator

The line-type modulator has been widely used because of its relatively small size and weight and good efficiency. However, some serious disadvantages of a line-type modulator are, first, that the interpulse period may not easily be varied over a wide range and, second, that the pulse width may not easily be varied. While the variation in interpulse period is largely fixed by the values of the recharging components, it is possible to arrive at a number of schemes for varying the pulse width in a line-type modulator.

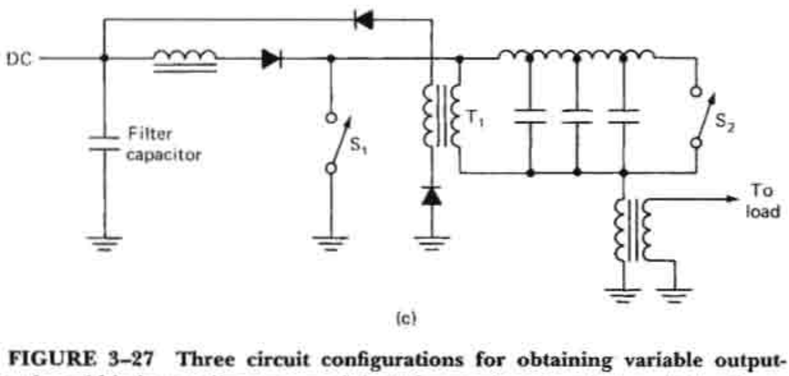
Figure 3–27 illustrates a number of circuit configurations that can result in the ability to vary the output-pulse width of a line-type modulator [14,21]. Each of these schemes essentially involves first a discharge initiation of the line in a manner analogous to normal line-type modulator operation, followed by a second switching action that tends to terminate the output pulse before the line normally completely discharges



(a)



(b)



(c)

FIGURE 3–27 Three circuit configurations for obtaining variable output-pulse width from a line-type modulator by varying the relative closing times of switches S_1 and S_2. Arrangements (*b*) and (*c*) provide for return of a portion of the unused stored energy to the power supply.

itself into the load; the leftover charge may be either dissipated in an auxiliary resistor, as in Figure 3–27(*a*), or returned to the power supply, as shown in Figure 3–27(*b*).

Figure 3–27(*c*) is a schematic diagram of another successful variable-pulse-width line-modulator concept. Charging of the pulse-forming network and initiation of network discharge by closing S_1 is conventional with this arrangement. However, when it is desired to terminate the pulse, switch S_2 is closed, with the result that the output pulse terminates and the voltage reverses across the PFN; this voltage reversal is coupled to the power supply through transformer T_1 and the diodes, so that a portion of the unused stored energy is returned to the power supply. One of the keys to a successful circuit of this type is the careful design of transformer T_1 in order to minimize losses due to fringing at the air gaps of the core; in some cases, the use of several distributed air gaps has been resorted to in an attempt to increase efficiency.

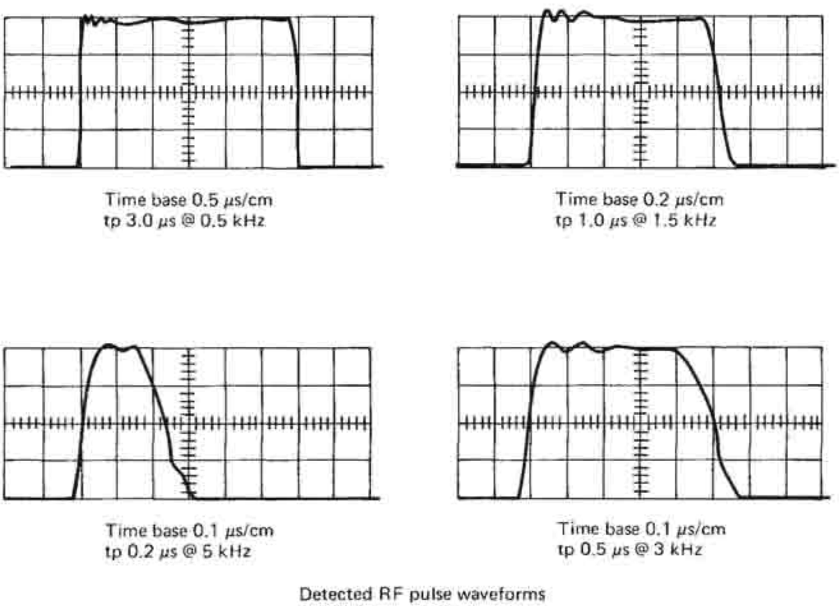


FIGURE 3–28 Detected RF pulse waveforms generated by a magnetron pulsed with a variable-width line-type modulator with energy return. (*Raytheon*)

Figure 3–28 shows the types of waveforms that can be obtained by using this approach, showing detected RF for a range of pulse widths from tenths of a microsecond to several microseconds.

Magnetron-Moding Control in Line-Type Modulators

One of the problems involved with the operation of a microwave magnetron is the tendency for the tube to oscillate in other than the desired mode. This type of operation is highly undesirable for a number of reasons. It has been observed that the rate of rise of voltage applied to the tube at the time when the tube initially begins to conduct is a primary factor in determining the tendency of a particular tube-modulator combination to operate stably and in the desired mode [7]. As will be shown in Chapter 4, the rate of rise of voltage may be controlled to a certain extent by the use of proper design techniques and by achieving specific values of distributed capacitance and leakage inductance in the transformer itself. However, attempts to reduce the rate of rise of voltage by increasing capacity often have the result that undesirably large amounts of energy are stored in the transformer, with a resultant loss in efficiency and an undesirable heating of the modulator components. Another approach that can be used to control the rate of rise of voltage is the connection of an RC network, at either the primary or the secondary side of the pulse transformer. If it is used at the primary side, the RC network is often called a *despiking* network. Such despiking networks are often used to eliminate a large spike of current occurring at the beginning of the pulse, often associated with improper transformer design. An RC network on the output side of the transformer can conveniently be used to control the rate of rise of voltage on the tube, and in fact the resistors are often made selectable to tailor the modulator to a particular tube. Unfortunately, each of these approaches results in a withdrawal of energy from the modulator, which reduces overall modulator efficiency, and such approaches may also make it difficult to achieve a reasonably rectangular pulse shape.

One approach to achieving stable magnetron operation is to use a so-called *pedestal technique* [18]. A pedestal technique involves initially applying a waveform of slowly rising voltage to the tube, bringing the cathode potential to approximately 90% of its normal operating voltage, followed by rapidly increasing the voltage to its full value, as shown in Figure 3–29. Using such techniques, conventional magnetrons have been made to oscillate stably with pulse widths as short as 10 ns.

Another approach, which may be considerably simpler, is the so-called corner-cutter circuit, or pulse-bender arrangement, shown in Figure 3–30 [2,12]. This approach places a large load on the output of the line-type modulator at approximately 75% of the normal operating voltage, slowing the rate of rise of voltage to the tube substantially, and permitting stable operation with reasonably fast rise times, a highly rectangular RF pulse, and rapid fall time.

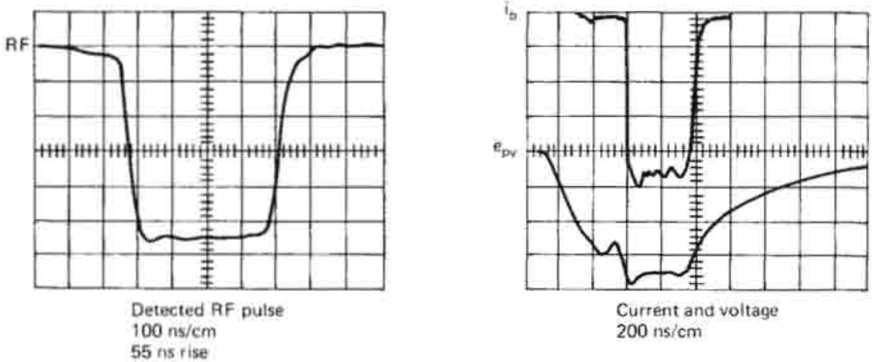


FIGURE 3–29 Detected RF pulse, and voltage and current waveforms of a pedestal-pulsed C-band magnetron. (*Raytheon*)

Operation of the corner-cutter circuit involves the charging of capacitors C_1, C_2, and C_3 to a voltage determined by the zener voltage stack, typically approximately 75% of normal operating voltage. Thus, for voltages less than this voltage, the diodes disconnect the output, and the voltage rises normally. However, once the voltage becomes more negative than the zener voltage, a load consisting of capacitors C_1, C_2, and

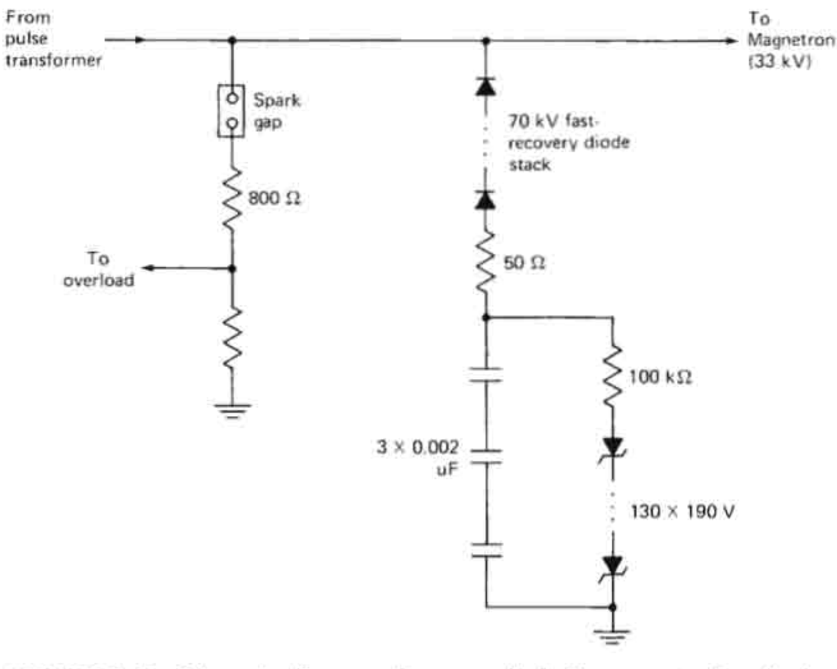
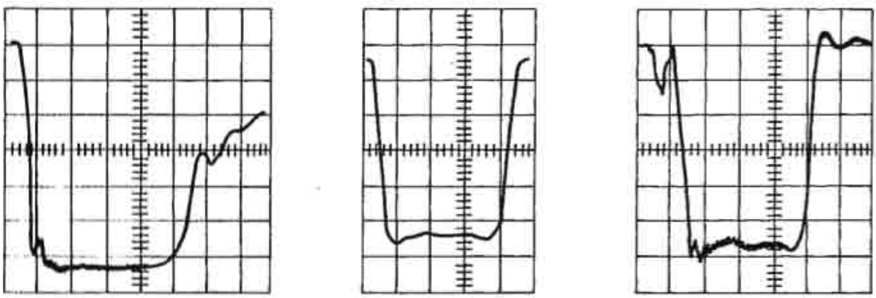


FIGURE 3–30 Schematic diagram of a zener diode "corner-cutter" or "pulse-bender" circuit.

C_3, R_1, and the forward impedance of the diodes is connected across the modulator, substantially reducing the rate of rise of voltage applied to the tube. Once the tube has reached its full operating voltage, the capacitors are charged to this voltage and little current flows through the diodes, so that efficient tube operation results. Between pulses, the capacitor voltage is stabilized at the zener voltage in anticipation of the next pulse. A set of waveforms illustrating operation of the corner cutter circuit is given in Figure 3–31. In the event that variation of the interpulse period is not required, it may be possible to eliminate the zener diode stack and operate with the resistors, the diodes, and the capacitors alone.



Voltage pulse amplifier 5 kV/cm — RF pulse Time base 0.5 μs/div — Current pulse amplifier 10 A/div

FIGURE 3–31 Voltage, current, and RF waveforms associated with a magnetron pulsed by a line-type modulator with the corner-cutter circuit shown in Figure 3–30. (*Raytheon*)

3–3 SCR-MAGNETIC MODULATORS

In recent years, there have been developed a particular class of line-type modulators called *SCR-magnetic* modulators [3,4,13,15]. These systems utilize saturating magnetic cores as switches in order to provide improved reliability over the hydrogen thyratron switch or in order to accommodate higher peak currents or more rapid rate of rise of current than can be accommodated by solid-state controlled rectifiers.

A simplified schematic diagram of one type of magnetic modulator is shown in Figure 3–32. Operation of the system is critically dependent upon the saturable reactors that are involved. These saturable reactors are typically coils wound on gapless toroids, and are designed to have a high unsaturated inductance and a very low saturated inductance, and to make a rapid transition from the unsaturated condition to the satu-

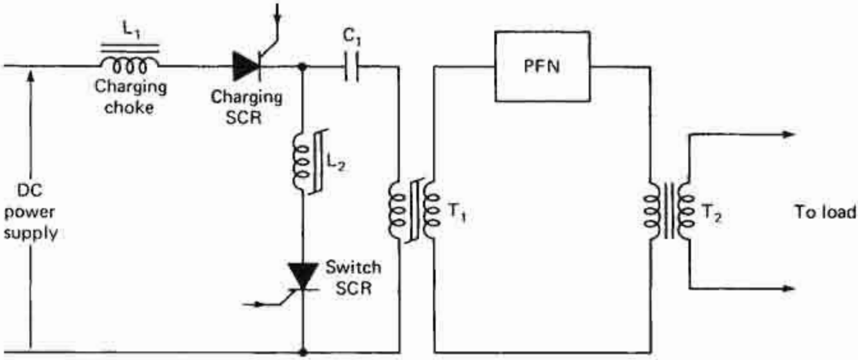


FIGURE 3–32 Magnetic-SCR line-type modulator using SCR switch and saturating magnetic devices. Bias windings have been omitted for clarity, but L_2, T_1, and T_2 are all magnetically biased. T_2 need not be biased, but if it is not, a resistor-diode combination must be connected across the primary to provide a low-impedance path for charging of the PFN.

rated condition. Such a device can act as a switch. There are two purposes for which such devices may be useful: to decrease the rate of rise through the SCR (see Section 7–3), and to reduce the peak-current requirements of the switch.

The operation of a magnetic-SCR modulator may be broken down into a number of separate subintervals: the charging interval for C_1, the transfer of charge from C_1 to the PFN, and the discharge of the PFN into the load.

The operating cycle begins with C_1 and the PFN initially discharged, and both the charging and switch SCRs in the OFF condition. Operation is initiated by triggering the charging SCR, which resonantly charges C_1 through L_1 to twice the dc power-supply voltage, as was the case for the line-type modulator. The charging period is given by $\pi \sqrt{L_1 C_1}$. Once the voltage on C_1 has reached its maximum value, the current through the charging SCR attempts to reverse, turning off the SCR and holding the voltage at C_1 at its maximum value.

The next step in the operation of the modulator is initiated by turning on the switching SCR. Once the switching SCR has been triggered, the full voltage across C_1 is developed across L_2. During this interval, only a small amount of current flows through L_2, permitting the switch SCR to turn on completely; after an interval of time sufficient for the SCR to completely turn on, inductor L_2 saturates, becoming a relatively low impedance, placing much of the voltage across C_1 across the primary T_1 and resonantly transferring charge from C_1 to the PFN through T_1 and through the saturated primary winding of T_2. In order for this transfer to be efficient, T_2 must represent a low impedance for charging of the PFN, and for that reason it is often biased to saturation. If this is not done, a diode may be connected across the T_2 primary.

Figure 3–33 is an equivalent circuit valid for this charge-transfer pe-

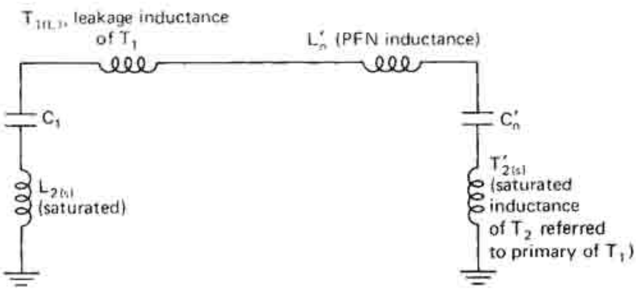


FIGURE 3–33 Equivalent circuit of magnetic modulator during charge transfer from C_1 to C_n.

riod, a resonant transfer of energy from C_1 to the network capacity C_n taking place through an inductance consisting of the saturated inductance of L_2, the leakage inductance of T_1, the saturated inductance of T_2, and the network inductance L_n. In Figure 3–33 L_n', C_n', and $T_{2(s)}$ are all values which are referred to the primary winding of T_1; that is, if L_n is the PFN inductance, C_n the network capacitance, $T_{2(s)}$ the saturated inductance of T_2, and n_1 the turns ratio of T_1,

$$L_n' = \frac{L_n}{n_1^2}$$

$$C_n' = n_1^2 C_n$$

and

$$T_{2(s)}' = \frac{T_{2(s)}}{n_1^2}$$

The transfer of charge from C_1 to the PFN takes place in a time given by

$$\pi \sqrt{(T_{1(l)} + L_{2(s)} + T_{2(s)}' + L_n')\frac{C_1 C_n'}{C_1 + C_n'}} \qquad (3\text{–}2)$$

Any attempt to discharge the pulse-forming network back through T_1 is defeated by back-biasing action of both the switch SCR and the saturated inductor L_2.

At this point in time, the pulse-forming network is fully charged, and the full voltage appears across the secondary of transformer T_1. Once the required volt-time integral has been applied to T_1, T_1 saturates, discharging the PFN into the load through transformer T_2. An equivalent circuit for the discharge of the pulse-forming network is given in Figure 3–34. However, if more detailed calculations of output-pulse shape are required, the methods of Kadochnikov [9,10] may be used.

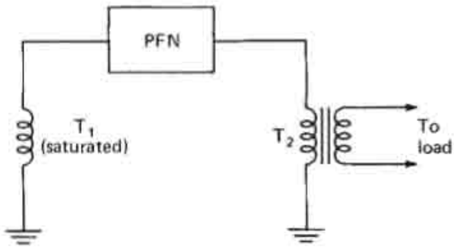


FIGURE 3–34 Equivalent circuit of magnetic modulator during discharge of PFN into load.

There are a number of design considerations that impact the choice of component values in such a modulator. The values of C_1 and the charging inductors are chosen so that resonant charging is accomplished within a relatively short interval, yet the charging current through the charging SCR and transformer T_1 should remain relatively low. L_2 in the unsaturated condition should represent a high impedance, and in the saturated condition an inductance that dominates the resonant transfer of charge from C_1 to the pulse-forming network. This saturated inductance should be small, in order to rapidly charge the pulse-forming network, but too rapid a charge of the pulse-forming network results in excessive peak currents flowing through the switching SCR. Transformer T_1 should be designed to have a low value of saturated reactance, and a volt-time integral product sufficient to permit complete charging of the pulse-forming network with some "guard" interval to accommodate variations in the temperature and voltages. Design equations for L_2 and T_1 are presented in Section 7–5. The design of transformer T_2 is more conventional in nature and follows the guidelines in Chapter 4, but magnetic biasing of the transformer core may permit larger flux excursions. A set of waveforms is given in Figure 3–35, which shows magnetic modulator operation. Figure 3–35(*a*) shows the PFN voltage, illustrating the resonant charge transfer to C_n, the guard interval (the

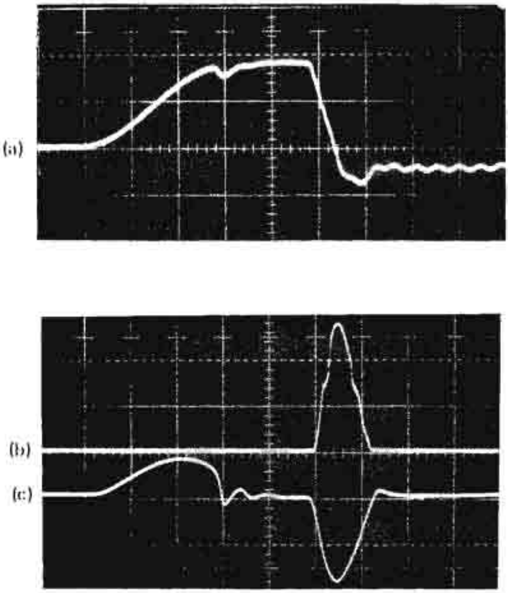


FIGURE 3–35 Magnetic modulator waveforms showing (*a*) PFN voltage, (*b*) current in a TWT load, and (*c*) output voltage.

constant-voltage interval), and the switching period where the voltage rapidly reduces to zero when the saturable inductor saturates. This discharge then produces the output-voltage pulse shown in 3–35(*c*). Note that in this modulator a diode-resistor pair was used across the transformer primary and so a portion of the PFN charging process was reflected in the output voltage.

Figure 3–32 is a much simplified diagram, and additional circuits are necessary in order to provide for proper functioning of the circuit. In order to ensure correct saturation, both L_2 and T_1 require an auxiliary bias supply for resetting; T_2 may require a bias winding also. In order to realize a truly constant peak charging voltage on C_1 (to reduce time jitter), some form of charging regulator, such as those discussed in Chapter 6, must be incorporated in place of the simple charging choke. The biased-diode nature of the load usually results in the reflection of some small voltage at the anode of the switch SCR. This voltage must be monitored, since arcing in the load, transformer, or pulse-forming network can result in excessive voltages reflected back at this point. Of course, dc power-supply overload circuits and appropriate protective circuits for the load (such as magnetron or TWT) must be provided if reliable operation is to be achieved.

The transfer of energy from C_1 to C_n and the discharge of energy stored in C_n into the load is accompanied by an increase in voltage level and a decrease in discharge or transfer time at each stage of the process. Typically, in a single stage, ratios of charging times to discharging times on the order of 10:1 can be achieved. Additional sections may be added in order to provide the additional compression required for shorter pulse lengths, but care must be used to ensure complete transfer of charge between stages in such circuits [3].

3–4 DC POWER SUPPLIES

A common feature of these modulators is a dc power supply. Conventional design practice may be utilized for such systems, but a few special precautions are necessary. For the lower-voltage supplies such as are used with line-type and magnetic modulators, three-phase rectifiers and choke input filters are typically utilized. A simplified schematic diagram of such a system is shown in Figure 3–36, showing provisions for measuring overload current drawn from the supply directly, as well as conventional protection in the primary power lines. It must be emphasized that the power supply should be capable of sustaining

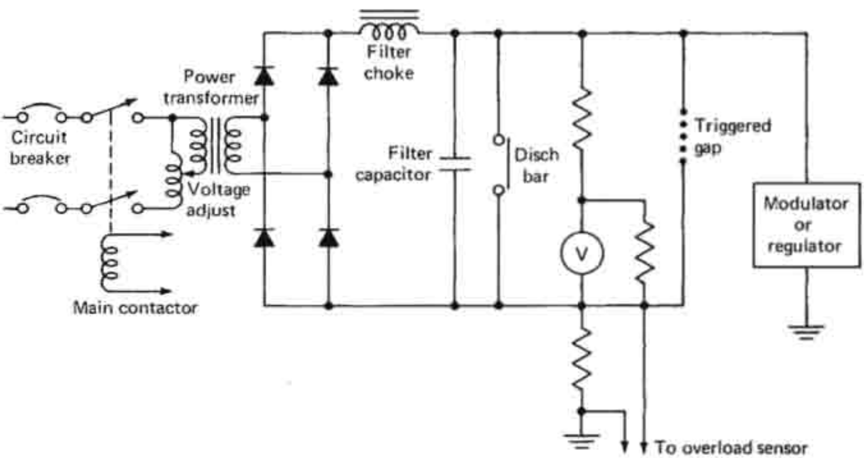


FIGURE 3–36 Representative modulator dc power supply, showing some overload circuits and sensing points.

occasional virtual short-circuit operation and of rapidly restoring normal operation once the short circuit has been removed. In order to provide adequate protection, the overload circuits must be rapid-acting.

Higher-voltage dc supplies such as are encountered with hard-tube systems will often involve the use of conventional voltage doublers in order to reduce output-voltage requirements on the power-supply transformers. With hard-tube modulators, particularly at higher power levels, crowbar circuits are often incorporated across the power-supply output in order to protect the relatively expensive tube, and the power supply must be specially designed in order to withstand the severe stresses imposed by such operation.

Finally, there are cases where the use of a high-frequency inverter is desirable in order to provide a high ripple frequency and to achieve small size and weight, or for operation from available low-voltage dc supplies.

For cases where the filtering provided by choke input filters is not adequate, series vacuum-tube regulators may be incorporated, as shown in Figure 3-37, in order to provide adequate power-supply regulation and ripple reduction. Where filtering is adequate for ripple reduction but additional regulation against line or load variations is desired, it may be adequate to sense a parameter such as tube current or peak

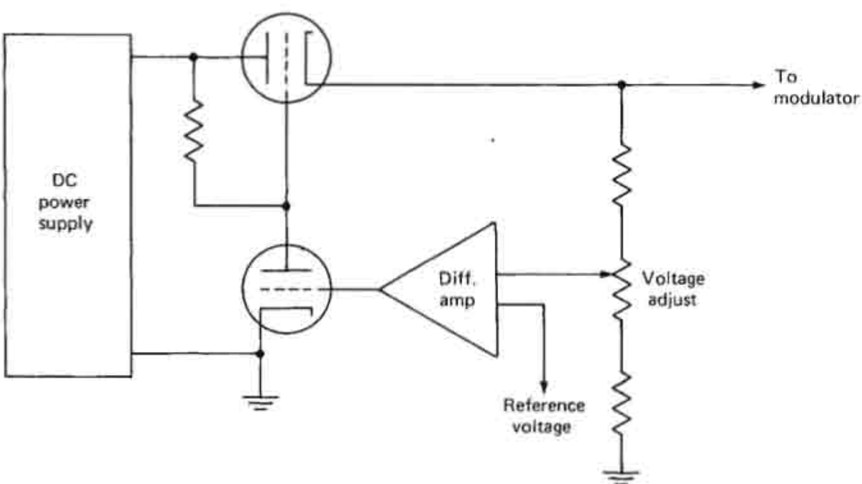


FIGURE 3-37 Series regulator that may be used for a high-stability, low-ripple modulator power supply.

charging voltage and to use a motor-driven autotransformer to vary the input voltage so as to keep this parameter constant.

In any of these high-voltage circuits, adequate provisions for operator safety must be incorporated, including interlocks, shorting switches, and bleeder resistors on all high-voltage capacitors.

3–5 OVERLOADS AND PROTECTIVE CIRCUITS

Throughout the book, the subject of safeties and overloads for various faults is discussed. In this section, a number of these will be brought together, both to have them readily available in one location, and also to indicate the extreme importance of a properly designed overload-protection and performance-monitoring system if a successful and reliable transmitter design is to be achieved. Because of the high voltages and substantial amounts of power associated with a high-power microwave transmitter, a certain number of malfunctions (such as occasional arcs) will normally be encountered, and the modulator and tube must be designed to accommodate these without damage to either the tube or the modulator. In addition, in the event of catastrophic failure either of a modulator component or of the RF tube itself, the overloads and monitoring circuits should be such that damage is confined to that particular unit and other portions of the system are not permanently affected. Finally, transmitter performance should be monitored to give an early indication of problems that might be encountered. In a high-power microwave radar transmitter, the protection requirements can be associated either with the microwave tube itself, or with the modulator utilized to drive the tube; in addition, there are often requirements for overall performance monitoring in order to verify proper operation of the entire transmitter.

Microwave Tube Protection

The requirements for protection of the microwave tube are particularly important, since the microwave tube is often a relatively expensive part of the system—one that requires considerable time to replace—and many of the high-power tubes occasionally evidence occasional arcs or abnormal operation which if not properly monitored and accommodated could permanently damage the device.

The particular type of protection required is dependent upon the type of microwave tube being utilized. For magnetrons, the peak and average current through the tube are particularly sensitive indications of tube operation and should be monitored; operation outside normally acceptable bounds should immediately terminate operation of the modulator. In the event that the magnetron fails to oscillate, if a line-type modulator is utilized, the cathode voltage will rise to unacceptably high levels; in order to avoid this, a spark gap that breaks down at a voltage slightly in excess of the normal operating voltage is often connected to the magnetron cathode. Cooling of high-power magnetrons is quite important; temperature should be monitored at the point specified by the manufacturer, and the presence of appropriate cooling media, either air, liquid, or both, should be monitored and lack of adequate cooling or overtemperature should result in the removal of voltage from the tube. Since the output load has a significant effect on the operation of a magnetron, its status should be monitored by means of a voltage—standing-wave ratio (VSWR) monitor or reflected-power monitor, and excessive reflected power should terminate operation. Waveguide arcs can be extremely damaging if they propagate through the waveguide system to the output window of the tube; optical arc detectors that sense the presence of light within the waveguide system have been developed and are often recommended for monitoring the output waveguide system.

A crossed-field amplifier has protection requirements essentially identical to those of a magnetron, with some additions necessitated by the fact that the CFA is an amplifier rather than an oscillator. In particular, provisions must normally be made to ensure that high voltages are not applied to the CFA if adequate RF drive is not present at the tube input, and the frequency of the RF signal at the input must be such that the CFA is in a region of normal operation. The alignment of RF drive pulses is perhaps one of the more difficult requirements for CFA operation to achieve reliably, particularly for short-pulse operation. While magnetron coolant and temperature monitoring requirements also apply to CFAs, because of their higher average power such monitoring may be much more critical to reliable CFA operation than is the case for many magnetrons.

In the linear-beam tubes, such as the TWT and the klystron, the requirements for tube monitoring are perhaps more severe than is the case for the crossed-field devices. In particular, the monitoring of various features that indicate the defocusing of the electron beam and the corresponding intercept of current by the physical tube structure are particularly important. In many tubes, collector current, beam current, focuscoil or solenoid current, and temperature and coolant flow must be monitored, and modulator operation must be disabled if abnormal values are sensed. The rapid protection of the tube in the event of catastrophic beam defocusing is particularly important, since tube meltdown can occur in a relatively short period of time. Particularly for gridded tubes, the presence of arcs within the tube body must be accommodated. In order to protect the grid-driver circuitry, it is often desirable to include a spark gap from the grid to the cathode in order to prevent excessive grid voltages in the event of a tube arc; in addition, if the tube arcs, the high voltage must be removed from the tube for a sufficient time for the arc to dissipate; this often necessitates the incorporation of a crowbar circuit directly across the tube, and the power supply must be such that it will withstand repeated firings of this crowbar circuit without damage. Also, as was the case in the crossed-field devices, reflected power should often be monitored, to indicate any abnormal VSWR of the load and to prevent the possible self-oscillation of the linear-beam tube.

Modulator Monitoring and Safeties

Monitoring of various test points in the modulator itself is important to ensure proper transmitter operation, since in many cases abnormal operation of the microwave tube is reflected in unusual modulator operation, and also to protect against failures in the modulator components. All modulator designs should have adequate interlocking and safeties to prevent inadvertent operator contact with the potentially lethal high voltages that are often present within the modulator. Complete enclosure of all high-voltage areas, shorting bars with which to discharge all high-voltage capacitors if the enclosure is open, bleeder resistors to discharge capacitors, and interlocks to disable the modulator if access can be achieved by the operator are all important features that are normally an integral part of the modulator design. The specific additional safeties that may be required in a modulator are to a certain extent dependent upon the type that is being protected. Differing requirements exist for line-type, SCR-magnetic, and hard-tube modulators.

In a line-type modulator, if the magnetron fails to oscillate for any reason, the output voltage on the magnetron can double its normal value. Since this might lead to a breakdown of the output transformer or damage to the microwave tube, often a spark gap with a breakdown voltage slightly in excess of the normal operating voltage is connected across the magnetron.

In the event of a mismatch of the load, a portion of the stored energy will be reflected back into the modulator, producing a negative voltage at the switch. In order to remove this voltage, and to prevent the resultant abnormal operation of the modulator, it is customary to connect an inverse diode (also called a *shunt* or a *clipping* diode) and a resistor in series with it across the switch so as to clip off any negative voltage that appears; current through this resistor is often monitored and excessive inverse voltage thus sensed and used to disable the operation of the modulator. This is particularly critical, since a negative voltage at this point at the beginning of the charging cycle results in a corresponding increase in the peak charging voltage, which can in turn further increase the negative voltage at the beginning of the next charging cycle, producing ever-increasing peak charging voltages on the PFN that can permanently damage the modulator.

Power-supply current should be continuously monitored, and excessive power-supply current should disable modulator operation; this is particularly important, since occasionally the switch may go into a condition of continuous conduction, which could rapidly damage either the switch, the power supply, or the charging choke. In addition, breakdown occurring within the modulator will often cause excessive power-supply current to be drawn; normally, the energy stored in the power supply will have to be carefully considered to determine whether disconnection from the power-supply line is sufficient, or whether a more rapidly acting overload device such as a crowbar is necessary. Finally, the temperature of various oil-filled, high-voltage components and the pressure of the oil within these units can be monitored and often provides an early indication of device failure.

For an SCR-magnetic modulator, in addition to providing a spark gap on the magnetron and monitoring temperature at certain critical points, it is normal to monitor inverse voltage on the SCR, peak power-supply current, and magnetron average current. Excessive values of peak power-supply current or magnetron average current will typically be used to interrupt the charging of the modulator by disabling triggers to the charge SCR and turning off the main power-supply contactor. A small amount of inverse voltage on the SCRs at the termination of the pulse is often desirable to aid in SCR recovery; however, excessive voltage at this point indicates abnormal operation of either the modulator or the load and should promptly initiate interruption of the modulator operation.

In a hard-tube modulator, protection is usually centered about the high-voltage switch tube and the load, since arcs can occur in either the switch tube or the output load, particularly for voltages above the 30- to 40-kV region. For these voltages, incorporation of a crowbar circuit across the high-voltage power supply is a virtual necessity for reliable operation. When utilizing such a crowbar, one needs to carefully ensure that an oscillatory discharge will not be encountered if the crowbar is fired, since such a condition could result in crowbar deionization and potential damage to the circuit. Initiation of the crowbar firing can often be accomplished by the sensing of peak currents through either the switch tube or the load, and excessive current should promptly initiate crowbar firing. In some cases, adequate protection can be achieved for the switch tube by connecting the grid voltage to a large negative voltage in the event of a malfunction, as was discussed in Section 3–2. Again, as with any high-power tube, coolant interlocks and temperature monitoring should be an integral part of the protection for the modulator tube.

Transmitter-Performance Monitoring

Perhaps the most important overall performance monitor associated with a radar transmitter is the power output of the device, either peak or average, which can be monitored at the device output. In addition, reflected power or VSWR is normally monitored both as an indication of normal system operation and as protection for the high-power RF tubes. It may be desirable also to monitor a number of additional circuits or functions, but most of these are satisfied by the requirements for tube or modulator protection already outlined. The advent in recent years of microprocessor technology provides a means not only for monitoring a number of these various tube, output, and modulator parameters, but also for providing a permanent record of overall transmitter performance and for monitoring trends or combinations of monitored variables that are indicative of incipient failure, permitting their rectification during normally scheduled maintenance operations, with a resultant increase in system availability.

3–6 LOAD EFFECTS ON MODULATOR DESIGN

The modulator must provide proper operating voltages to the load, but in many instances the characteristics of the particular load will influence the design of the modulator. Tubes requiring extremely high voltages may require special techniques, such as Blumlein or Marx generators, and if short HV pulses are utilized, rise-time enhancement networks may be required. Oscillation and excessive backswing must be avoided after the pulse to prevent the generation of spurious signals. If the tube arcs or shorts regularly, protection must be provided to allow the arcs to extinguish. Also, in line-type modulators, arcing loads may cause excessive voltages to appear on the modulator if the shunt-

TABLE 3-1 STABILITY FACTORS [20]

Type	FM or PM sensitivity	Ratio of dynamic to static impedance	Current or voltage change to 1% change in HVPS voltage	
			Line-type modulator	Hard-tube modulator
Magnetron	$\frac{\Delta F}{F} = 0.001 + 0.003 \frac{\Delta I}{I}$	0.05–0.1	$\Delta I = 2\%$	$\Delta I = 10$–20%
Magnetron (stabilized)	$\frac{\Delta F}{F} = 0.0002 + 0.0005 \frac{\Delta I}{I}$	0.05–0.1	$\Delta I = 2\%$	$\Delta I = 10$–20%
Backward-wave CFA	$\Delta\phi = 0.4$–$1°$ for $1\%\ \Delta I/I$	0.05–0.1	$\Delta I = 2\%$	$\Delta I = 5$–10%
Forward-wave CFA	$\Delta\phi = 1$–$3°$ for $1\%\ \Delta I/I$	0.1 –0.2	$\Delta I = 2\%$	$\Delta I = 5$–10%
Klystron	$\frac{\Delta\phi}{\phi} = 1/2 \frac{\Delta E}{E} \quad \phi \approx 5\lambda$	0.67	$\Delta E = 0.8\%$	$\Delta E = 1\%$
TWT	$\frac{\Delta\phi}{\phi} \approx 1/3 \frac{\Delta E}{E} \quad \phi \approx 15\lambda$ $\Delta\phi = 20°$ for $1\% \frac{\Delta E}{E}$	0.67	$\Delta E = 0.8\%$	$\Delta E = 1\%$
Triode or tetrode	$\Delta\phi = 0$–$0.5°$ for $1\%\ \Delta I/I$	1.0	$\Delta I = 1\%$	$\Delta I = 1\%$

TABLE 3-2 MODULATOR COMPARISON

Type	Switch	Flexibility		Pulse length		Flatness	Working voltages	Application	Size and cost
		Duty	Pulse width	Long	Short				
Line-type	Thyratron	Limited by charging circuit	No	Large PFN	Good	Ripples	High	Most common	Small size
	SCR						Medium		Smallest size higher cost
Hard-tube	Capacitor-coupled	Limited	Yes	Large capacitor	Good	Good	High	Fairly common	Large
	Transformer-coupled	Limited	Yes	Capacitor and xfmr gets large	Good	Fair	High	Not often used	Large
	Modulation-anode (floating-deck)	No limit	Yes	Good	OK	Excellent	High	Usually high-power	Quite large
	Grid	No limit	Yes	Good	Good	Excellent	Low	Widely used at low power	Small and inexpensive
SCR-magnetic	SCR and magnetic cores	Limited by charging circuit and magnetic cores	Normally fixed	Large PFN	Losses may be high	Ripples	Low in initial stages	Becoming more common	Small; initial design costs high

diode circuitry is poorly designed. Crowbar circuits may be necessary in order to fully protect the microwave source and modulator.

Of particular importance are the stability factors achievable by using the various tube-modulator combinations. Some of these data are summarized in Table 3–1 [20].

From this brief discussion, we can appreciate the interaction of tube characteristics with modulator design. An optimum selection of a microwave transmitter should include the modulator requirements as a prime consideration, and optimum modulator design is only achieved when the modulator designer has a thorough knowledge of the characteristics and peculiarities of the modulator load. In summary, the characteristics of a number of important modulator types are compared in Table 3–2.

3–7 COMPARISONS OF COST

Comparisons of cost associated with various modulator types are often of major interest to the radar designer, and are one parameter associated with a radar transmitter that is difficult to accurately define. This difficulty is complicated by the variation in overhead rates, employee efficiency, and materials cost from manufacturer to manufacturer, and from design concept to design concept. Cost also varies with the level of performance required and with the environmental requirements.

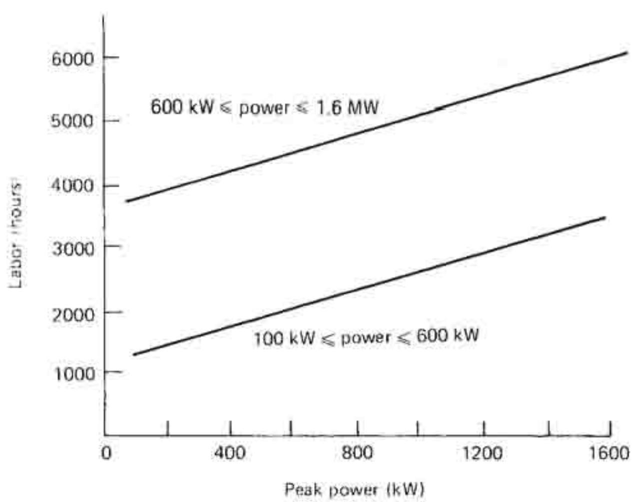


FIGURE 3–38 Engineering labor hours for transmitter design and integration for ground-based radar operating over the 0–55°C temperature range.

However, the estimation of labor hours for various phases of modulator design, fabrication, and checkout may be a useful piece of information for the engineer, and may serve as the basis for more detailed estimates to be generated later in the design process. Estimates of labor requirements for design, fabrication, and checkout, for ground-based radar transmitters designed to operate over the temperature range from 0 to 55°C are presented in Figures 3–38, 3–39, and 3–40. It should be

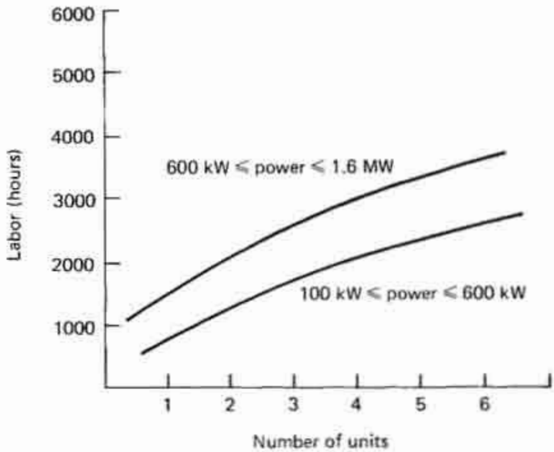


FIGURE 3–39 Labor-hours for transmitter assembly for ground-based radars operating over the 0–55°C temperature range.

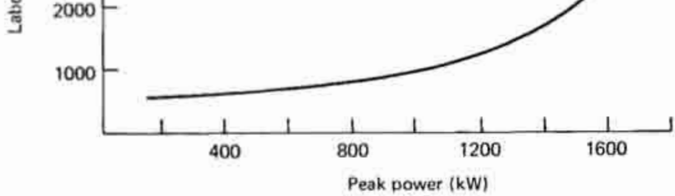


FIGURE 3–40 Labor-hours for transmitter subsystem testing for ground-based radars operating over the 0–55°C temperature range.

emphasized that these are general, representative guidelines only, and that stringent performance or environmental requirements will result in increases, while lenient requirements may result in substantial decreases. Of course, if similar designs can be adapted, there may be a substantial reduction in the design labor requirements.

REFERENCES

1. Alsmeyer, R., "A Constant Current Hard Tube Modulator," in *Conf. Rec. 1973 11th Modulator Symp.*, September 1973, pp. 75–78.

2. Ashworth, J., and R. Browne, Raytheon, personal communication, Nov. 12, 1979.

3. Busch, K. J., et al., "Magnetic Pulse Modulators," *Bell System Technical Journal*, vol. 34, no. 5, September 1955, pp. 943–993.

4. Coate, G. T., and L. T. Swain, Jr., *High-Power Semiconductor-Magnetic Pulse Generators*, MIT Press, Cambridge, Mass., 1966.

5. Doolittle, H. D., "Vacuum Power Tubes for Pulse Modulation," Machlett Laboratories, Stamford, Conn.

6. Doss, J. D., et al., "80 Kilovolt Blocking Oscillator Modulator," in *Proc. 10th Modulator Symp.*, May 1968, pp. 10–21.

7. Gillette, P. R., and K. Oshima, "Pulser Component Design for Proper Magnetron Operation," *IRE Trans. Component Parts*, March 1956, pp. 26–31.

8. Glasoe, G. N., and J. V. Lebacqz, *Pulse Generators*, MIT Rad. Lab. series, vol. 5, McGraw-Hill, New York, 1948. (Also available in Dover and Boston Technical Publishers editions.)

9. Kadochnikov, A. I., I. E., Korobeynikova, and M. A. Kheyfets, "Calculation of the Current Pulse in a Section of a Magnetic Pulşe Generator," *Telecommunications and Radio Engineering*, pt. 2, vol. 29, no. 8, 1974, pp. 90–94.

10. Kheyfets, M. A., and A. I. Kadochnikov, "Switching Properties of Saturable Reactors in Magnetic Pulse Generator Sections," *Telecommunications and Radio Engineering*, pt. 1, vol. 32, no. 2, February 1978, pp. 81–84.

11. Limansky, I., "How to Design a Line-Type Modulator," *Electronic Design*, vol. 8, no. 4, Feb. 17, 1960, pp. 42–45.

12. Main, J. H., "A Self-Biased Solid-State Diode Despiker," in *Proc. 8th Symp. Hydrogen Thyratrons and Modulators*, May 1964, pp. 416–422.

13. Manteuffel, E. W., and R. E. Cooper, "D-C Charged Magnetic Modulator," *Proc. AIEE*, vol. 78, January 1960, pp. 843–850.

14. Martinovitch, V. N., "Generation of Output Pulses Having a Continuously Variable Pulse Width in a Line-Type Modulator," in *Proc. 6th Symp. Hydrogen Thyratrons and Modulators*, May 1960, pp. 106–110.

15. Melville, W. S., "The Use of Saturable Reactors as Discharge Devices for Pulse Generators," *Proc. IEE* (London), vol. 98, pt. III, 1951, pp. 185–207.

16. Nordenberg, H. M., *Electronic Transformers*, Reinhold, New York, 1964.

17. O'Meara, T. R., "Ladder Transformer Coupling Networks for Pulse Amplifiers and Hard-Tube Modulators," *Trans. IRE Gp. on Circuit Theory,* vol. 7, no. 3, September 1960, pp. 239–246.

18. Parker, T. J., "A Modulator Technique for Producing Short Pulses in High Powered Magnetrons," *IRE Nat. Conv. Rec.,* 1954, pp. 142–151.

19. Sutherland, R. I., *Care and Feeding of Power Grid Tubes,* Varian, Eimac Div., San Carlos, Calif., 1967, sec. 2.

20. Weil, T. A., "Transmitters," in M. I. Skolnik, ed., *Radar Handbook,* McGraw-Hill, New York, 1970, chap. 7.

21. Numerous modulator examples and discussions are contained in the proceedings of the modulator symposia sponsored by the U.S. Army Electronics Command and the Pulse Power Conference:

Title	AD number
Proc. 5th Symp. Hydrogen Thyratrons and Modulators, 1958	AD 650 899
Proc. 6th Symp. Hydrogen Thyratrons and Modulators, 1960	AD 254 102
Proc. 7th Symp. Hydrogen Thyratrons and Modulators, 1962	AD 296 002
Proc. 8th Symp. Hydrogen Thyratrons and Modulators, 1964	
Proc. 9th Modulator Symp., 1966	
Proc. 10th Modulator Symp., 1968	AD 676 854
Proc. 11th Modulator Symp., 1973, IEEE 73 CH 0773–2 ED	
Proc. 12th Modulator Symp., 1976, IEEE 76 CH 1045–4 ED	
Proc. 13th Pulse Power Modulator Symp., IEEE 78 CH 1371–4 ED	
Proc. Intern. Pulsed Power Conf., IEEE 76 CH 1197–8 REG 5	
Dig. Tech. Papers, 2d IEEE Intern. Pulsed Power Conf., IEEE 79 CH 1505–7	
Proc. 14th Pulse Power Modulator Symp., IEEE 80 CH 1573–5 ED	