10. Cathode Pulsers: Hard-Tube Modulators

The hydrogen thyratron, the half-control device most commonly associated with the line-type pulser, is often simply called a "soft-tube" modulator because it is filled with low-pressure hydrogen. If the soft tube is filled with low-pressure hydrogen, what is a "hard" tube filled with? The answer is: as close to nothing as possible. In short, the tube is "filled" with as "hard" a vacuum as can be achieved and maintained, typically 10^{-8} to 10^{-9} torr. (Be assured that the external envelopes of both tube types are equally hard.)

The fewer the number of residual gas molecules in a hard tube the better. A plasma arc in a soft tube is its conduction mechanism. Once the arc is initiated, however, it can be terminated only by means external to the tube, hence its "half-control" designator. A plasma arc in a hard tube, on the other hand, is a fault condition. The conduction mechanism is the flight of free electrons from the cathode, or virtual cathode, to the anode. This mechanism is sometimes called space-charge-limited operation. (The charge carrier in this case is the cloud of electrons that hovers near the thermionic emitter.) Each electron that leaves the cathode must make it to the anode in order to contribute to useful load current. Therefore, any electron collision with residual gas molecules must be minimized.

The number of electrons that are permitted to make the trip per unit of time and per unit of area is called current density. Current density is controlled by means of the voltage applied between the cathode and a helical or meshlike grid (or grids) positioned somewhere between the cathode and the anode. This grid or control electrode of whatever type is placed physically closer to the cathode than to the anode—usually much closer, so that the electric field produced at the surface of the cathode by voltage applied between grid and cathode will be proportionately greater than the field intensity produced by voltage applied between anode and cathode.

10.1 The full-control switch tube

A vacuum tube—or "valve" as the English insist on calling it—that has a cathode, an anode, and a control grid is called a triode. A tube with an additional grid located between the control grid and the anode is called a tetrode. (Other electrodes such as the suppressor grid in the pentode, the shield grid, and even an arc-shield electrode are sometimes employed in hard tubes. But the triode and the tetrode are the basic gridded-tube types.) Their geometries can be planar or coaxial. In the planar configuration, the cathode, grids, and anode take the form of circular discs. In the coaxial configuration, which is also referred to as a radial-beam tube, the electrodes take the form of coaxial cylinders, with the anode being either external or internal (either the outermost or innermost cylinder).

In any case, the total current that leaves the cathode is shared between the anode and the grid(s). The portion of the total current shared with the anode is considered the useful current. Even though grids are perforated so that electrons can pass through them, they are not electronically invisible. If the voltage applied to a grid is positive with respect to the cathode, some electrons will be intercepted. (This is almost always true for the screen grid and, in a high-power gridded tube when full current is being demanded, for the control grid as well.) One of the goals of tube design is to minimize the control-electrode current because it serves no useful purpose and subtracts from the current that can be useful. The total cathode current in a tetrode can be expressed as:

$$I_{cathode} = I_{anode} + I_{control\ grid} + I_{screen\ grid} = KV^{\frac{3}{2}},$$

where K is the effective perveance. K can also be evaluated as

$$K = \frac{2.33 \times 10^{-6}(A)c}{\left(S_{diode}\right)^2},$$

where A is the cathode emitting area in cm^2, c is a constant dependent on tube geometry, S_{diode} is the equivalent diode spacing and given in cm, and V is the effective voltage. The fractional exponent 3/2 applies to most tube geometries. Effective voltage can be further defined as

$$V = V_{control\ grid} + \frac{V_{screen\ grid}}{\mu_{sg}} + \frac{V_{anode}}{\mu_a},$$

where μ_{sg} and μ_a are the amplification factors of the screen grid and anode, respectively.

The perveance of electron tubes can vary from less than 10^{-6} perv, or 1 μperv (the equivalent to 1 $\mu A/V^{3/2}$), for a high-performance millimeter-wave TWT, to 0.01 perv, or 10,000 μpervs (the equivalent to 0.01 $A/V^{3/2}$), for a large, gridded power tube. The screen-grid and anode amplification factors, or μ_{sg} and μ_a, are factors that quantify the strengths of the electric-field contributions at the cathode produced by voltages applied to screen grid and anode as they relate to the electric field produced by the control grid. If, for instance, a tetrode had a screen amplification factor of 10, it would require a change of screen-grid voltage 10 times as great as that of the control grid to have the same effect on cathode current change. Similarly, if the anode amplification factor were 100, the change in anode voltage would have to be 100 times as great as the change in control-grid voltage for the same cathode-current change. The screen-grid and anode-μ factors reflect the facts that not only are the screen grid and anode farther away from the cathode than the control grid, but the field produced by voltage on the screen grid must penetrate the shielding effect of the control grid, and the field produced by voltage on the anode must penetrate the shielding effects of both grids. We can see, then, that for positive voltage applied to both screen grid and anode, there will be a negative voltage that, when applied to the control grid, will cancel the field-strength contributions of the others and result in no current leaving the cathode. This balance point is called the cut-off grid bias for a given

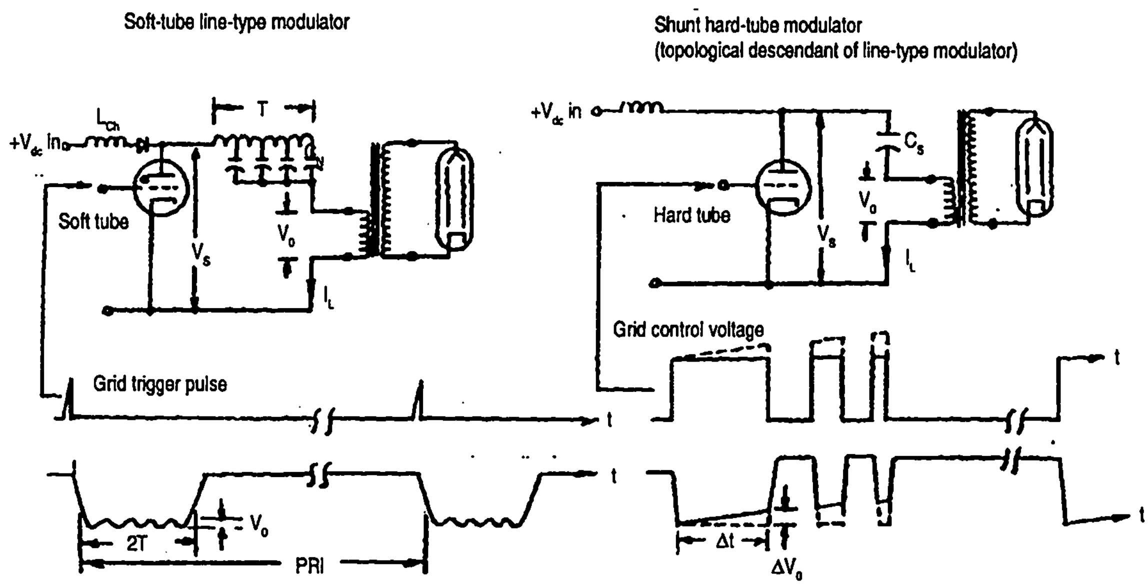


Figure 10-1. Progression of soft-tube, line-type modulator into shunt-type, hard-tube modulator.

combination of screen and anode voltages.

Because of cut-off grid bias, the hard tube can be turned off as well as on. Tube current, therefore, can be fully controlled by the voltage applied to the control grid from zero to the maximum. (The maximum current is determined by cathode emissive capability, a factor that varies with cathode area, cathode material, and cathode temperature.) This feature, then, makes the tube a "proportional and reciprocal" full-control modulator switch. It is called proportional and reciprocal because the factors that will reduce current are simply the reverse of the factors that can increase it. More will be said about hard-tube technology later.

10.2 Hard-tube modulator topology

Figure 10-1 shows how the replacement of the half-control switch in the line-type modulator with a full-control switch allows the pulse-forming network to be replaced with a quasi-infinite energy store. This is now possible because the time delay of an artificial transmission line is no longer required to externally turn the switch tube off through commutation. Initiated by a trigger, the pulse duration in the line-type modulator was set by network constants: the PRF was limited by the rate of rise of charging voltage across the soft switch tube, while the pulse-repetition interval (PRI) had to exceed

$$\pi\sqrt{L_{Ch}C_N}\,.$$

The pulses of current removed from the hard-tube modulator are now replicas of the voltage applied to the grid of the hard switch tube. The output voltage V_0 of the hard-tube modulator is the inverted replica of the circuit's switch tube grid drive voltage. Unlike the line-type modulator, the output-pulse duration and

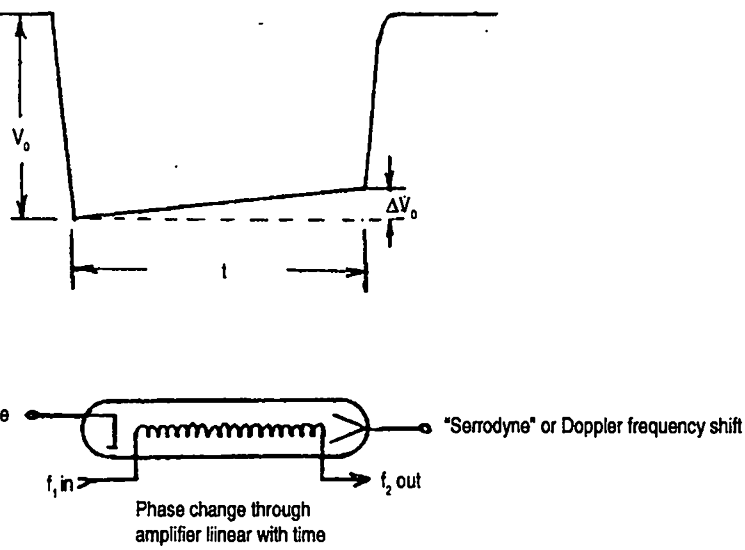
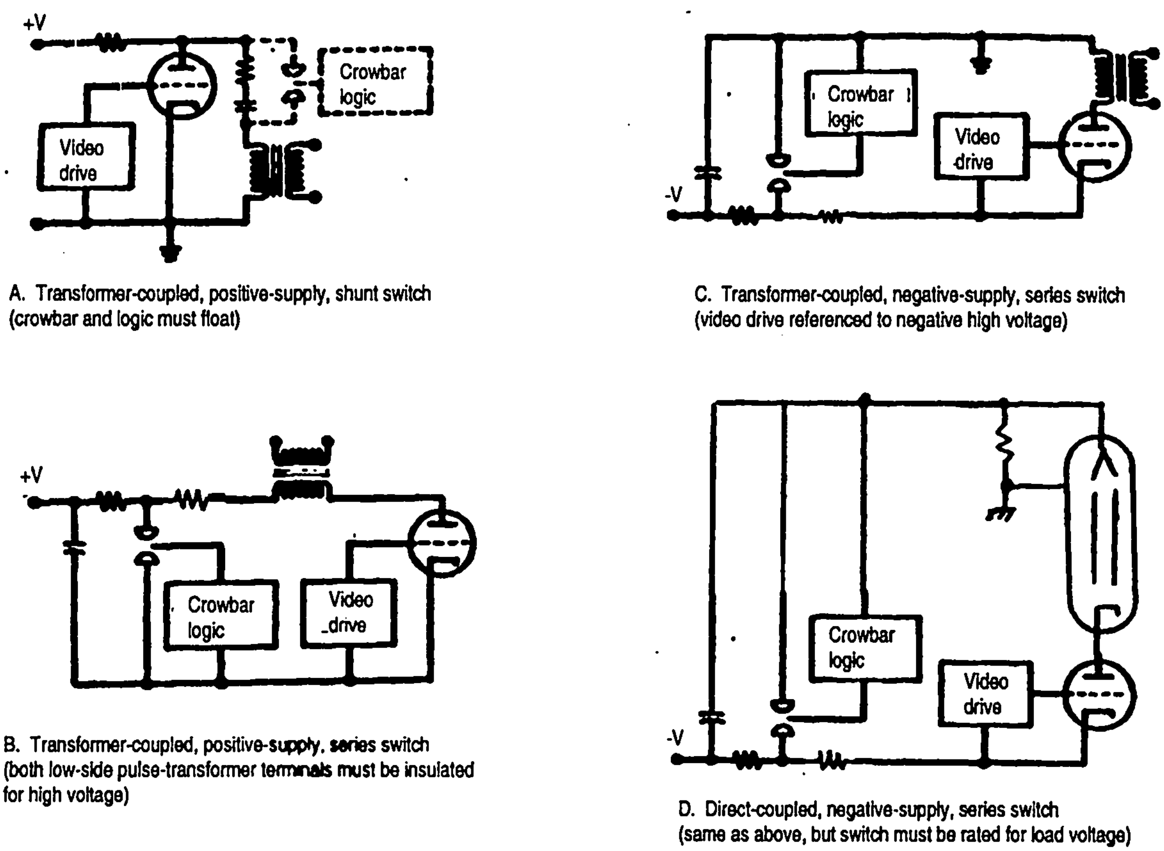


Figure 10-2. Consequences of pulse voltage decrement, or "droop," on power and phase shift.

PRF of the hard-tube modulator is versatile. On the other hand, the pulse-top ripple of the line-type modulator is replaced by the voltage decrement, DV_0, of the hard-tube modulator, but this condition can be compensated for by ramping the input. Changes in the terminal voltage of the energy-storage capacitor—especially the decrement, or "droop," in voltage that any finite-sized capacitor will experience when a charge is removed from it—can be compensated for by the proportional property of the grid-voltage/plate-current transfer characteristic and the ability of the anode to dissipate significant pulse energy and continuous power.

The capacitance C_S (for storage capacitance) for the hard-tube circuit is many times larger than C_N (for network capacitance) for the soft-tube circuit for a given amount of pulse energy because C_S is only partially discharged each pulse, whereas C_N, as we recall, is completely discharged with each pulse. Even though the loss in capacitor voltage experienced during a pulse can be compensated for by a complementary change in switch-tube anode voltage, as mentioned above, this is an energy- and power-wasteful process. To maximize the efficiency of a hard-tube modulator—which, as we will see, will be difficult enough—the switch tube is typically driven to the minimum anode-voltage drop that is still consistent with the peak current required so that capacitor-voltage droop will be passed on the load voltage. How much droop is tolerable? The answer, of course, depends upon the nature of the microwave tube load. (Refer back to Fig. 9-28 for some idea.)

Figure 10-2 shows the effect of incremental voltage change on pulse energy and phase angle. For a microwave tube having a linear-beam electron gun with space-charge-limited electron flow, beam power will vary as the 5/2 power of beam voltage. For a small value of $\Delta V_0/V_0$, the loss in electron beam power, $\Delta P_B/P_B$, from the beginning to the end of the pulse will be $2.5\Delta V_0/V_0$. The loss in per-unit electron-beam pulse energy will be the time average of the power loss over the pulse, which, for small values, can be treated as linearly varying. This figure amounts to approximately half of the end-of-pulse power loss, or $1.25\Delta V_0/V_0$.



A. Transformer-coupled, positive-supply, shunt switch (crowbar and logic must float)

C. Transformer-coupled, negative-supply, series switch (video drive referenced to negative high voltage)

B. Transformer-coupled, positive-supply, series switch (both low-side pulse-transformer terminals must be insulated for high voltage)

D. Direct-coupled, negative-supply, series switch (same as above, but switch must be rated for load voltage)

Figure 10-3. Alternative circuit arrangements for hard-tube modulators.

For a crossed-field-type device, the effect of voltage change will be far more pronounced. In this case, $\Delta P_B/P_B$ will be as much as $11\Delta V_0/V_0$, and the approximate per-unit loss in pulse energy will be $5.5\Delta V_0/V_0$. (These values refer only to loss of dc beam.) Other factors listed in Fig. 9-28 relate to how RF power output will be affected.

Because beam-voltage change for small values of $\Delta V/V$ is very nearly linear with time, the change in RF phase angle between input and output of the RF amplifier tube will be nearly linear. (This was noted in the section that discussed phase-pushing effects.) However, linear phase change is indistinguishable from pure frequency translation: the frequency coming out is offset from the one going in. The amount of frequency difference is equal to the percentage of voltage change times the phase-pushing factor expressed in degrees per percentage of voltage change. To find the differential phase change over the duration of the pulse, this product is divided by the pulse duration, or Δt times 360°/cycle. In the TWT example shown, the phase-pushing factor is $40°/1\%\Delta V/V$, the voltage droop ΔV is 1%, the pulse duration is 1 ms (0.001 second). The frequency offset, Δf (absolute, not per-unit), will therefore be

$$\frac{1\% \times 40°/1\%}{360 \times 0.001s} = 111\,{}^{cyc}\!/_{s} = 111\ \mathrm{Hz}\ .$$

The phase slope with time is 40,000°/s. Because there are 360° per cycle, the frequency shift, which is exactly the same as a Doppler shift, is 40,000/360, or 111 Hz. This effect has been referred to by some as the "serrodyne" effect because of its serrated, or sawtooth, waveform, which is caused in this case by voltage removal from a capacitor. Before there were modern frequency synthesizers with phase-locked loops and other means of assuring phase coherency between transmitter and receiver local-oscillator frequencies, the serrodyne effect was actually used to achieve the frequency offset between them. This was done by applying a linear ramp of voltage with its short-duration reset timed so that the period of the resulting sawtooth waveform was the same as the desired receiver intermediate frequency. This sawtooth waveform was applied between helix and cathode of a low-power, continuous-wave TWT that had a relatively high phase-pushing factor. With proper adjustment of the amplitude of the sawtooth waveform, the RF output would be at the desired local-oscillator frequency.

Figure 10-1 showed the most gradual of topological changes that could ease a designer from the line-type, half-control-switch modulator technology into the hard-tube, full-control-switch modulator technology. Figure 10-3 shows four further refinements. The first refinement, illustrated in Fig. 10-3a, simultaneously replaces the soft-tube switch with the hard-tube switch *and* the artificial transmission line PFN with a capacitor energy store, which is coupled with a series fault-current-limiting resistance. In the case of a switch-tube arc, however, all of the charge and energy stored in the capacitor will be dumped. Charge and energy will be transferred to the load connected to the pulse-transformer secondary until the transformer core saturates, which, in a well-engineered application, will not be much longer than the longest normal pulse. Up until this time, the current in the discharge loop will only be about 15% greater than the normal load current, assuming that the normal switch-tube voltage drop was approximately 10% of the initial capacitor voltage and that the load is a 3/2-power, space-charge-limited, diode-type electron gun.

When the pulse transformer core saturates, the transfer of energy and charge to the load terminates, but things get much worse for the faulted switch tube. Remember from the discussion on half-control switches that an arc in vacuum is a very good conductor with almost negligible voltage drop. Current is now limited only by the series resistance of the capacitor. What is more, the action integral of the discharge, or

$$\int i^2 dt,$$

is equal to the stored energy divided by the value of series resistance. (joules/ohm is the equivalent of ampere^2-seconds.) If the arc in the switch tube was between the anode and one of the small-diameter wires that make up the control grid, the action integral determines whether the grid wire will fuse. If it does and if enough grid wire is involved, the voltage on the grid will no longer effectively cut off anode current in the interpulse interval.

For low intrapulse voltage droop, the capacitor's stored energy should be many times the amount delivered to the load over the course of the longest normal pulse. ($Q = CV$, so $\Delta V/V = \Delta Q/Q$. For a 10% voltage droop, the stored charge must be 10 times charge delivered per pulse. For a 5% droop, the stored charge must be 20 times charge per pulse. Because the load voltage, V, is almost constant, both stored and load energies are proportional to charge, $W = QV$.) With 10 to 20 times more energy stored in this type of modulator, a means of charge diversion from the faulted switch tube is often necessary. This is usually handled by an electronic crowbar. For physically and electrically large capacitive storage, the topology illustrated in Fig. 10-3a already suffers from the fact that both terminals of the storage system must not only float at high voltage but they must support rapid pulse-voltage change as well. When the added complexity of an electronic crowbar system is factored into this configuration, an alternative architecture becomes attractive.

One advantage of the Fig. 10-3a topology is that the switch-tube grid drive and filament circuits are returned to ground. If we keep this advantage and reconnect the capacitor bank and crowbar so that they are also referenced to ground, the topology of Fig. 10-3b emerges. Although the storage terminals no longer float, both terminals of the pulse-transformer primary float. This means that during the interpulse interval both terminals must be insulated from the core for the power-supply voltage. This introduces the possibility of a failure mode in a component that should normally be one of the most reliable in the entire transmitter. Therefore, this is not a popular solution among those who make their livings designing and building transformers. For this reason, the arrangement shown in Fig. 10-3c is the most common.

In Fig. 10-3c the pulse-transformer primary and capacitor-bank/crowbar are referenced to ground, and the power-supply polarity is reversed. The switch-tube drive and filament circuits are now referenced to the negative high voltage. The filament power supply for the switch tube, which is often no more than a step-down power transformer (albeit one that may be rated for kilowatts of heater power), must now also be insulated for the operating anode-cathode voltage of the switch tube. This is not, however, a high-performance requirement, in that leakage inductance and interwinding capacitance of the transformer are of little or no concern. What might, under other circumstances, be considered excessive leakage inductance can even be advantageous in this case, in that it will automatically limit cold-filament in-rush current. (The resistance of a switch-tube filament, like that of most wires, follows a positive temperature-coefficient characteristic, where resistance increases with temperature—except that the temperature of this wire will eventually reach about 1000°C.)

The grid-drive circuits also will require primary power that is isolated from ground for the full anode-cathode operating voltage. This power is often derived from the isolated filament power (or vice versa). More important, the low-level gate, or on/off trigger signals, must also be coupled from ground-level reference to high-voltage deck reference. Modern designs rely almost exclusively on optically coupled links for this function, even though the common-mode rejection requirements are far less for this application than for the "floating-deck" type modulator, which will be discussed later. In this case, even though the reference deck is at high voltage with respect to ground, this voltage is relatively constant with time. Any changes consist mostly of the capacitor-bank droop and the voltage drop across the surge-current-limiting resistors caused by the normal peak-load current. Before there were dependable high-performance optically coupled signal links, special pulse-transformer designs and even pulsed RF links were used to transmit low-level timing signals across the high-voltage gap, examples of which we will see later. As complicated as these circuits are, they are in most cases preferable to solving the problem of insulating the pulse-transformer primary winding for the full switch-tube operating voltage—especially because the performance of the pulse transformer is either the most important issue or second only to that of the switch tube itself.

Circuit complications are also unavoidable in the final configuration, which is shown in Fig. 10-3d. This is the direct-drive hard-tube modulator—the top of the line, as it were. In this design, there is no pulse transformer at all. This feature eliminates the limitations on pulse duration, rise-and-fall times, intrapulse droop, and leading-edge overshoot and ringing. The price paid for this improvement is increased circuit complexity and the fact that the switch tube itself must handle the full operating voltage and current of the microwave tube load as separate and independent ratings—not as a peak-power product, where lack of voltage hold-off can be compensated for by increased current-handling (or vice-versa) with any mismatch handled by an appropriate pulse-transformer turns ratio. This direct-drive connection is capable of the highest performance of all modulator types. Given enough excess capability, or head-room, it can be programmed to compensate for almost every shortcoming of the rest of a transmitter, including power-supply ripple, noise, and capacitor-bank droop. In fact, there is a class of switch tube, which will be described later, that will do much of this with no programming at all. Not surprisingly, therefore, this class of modu-

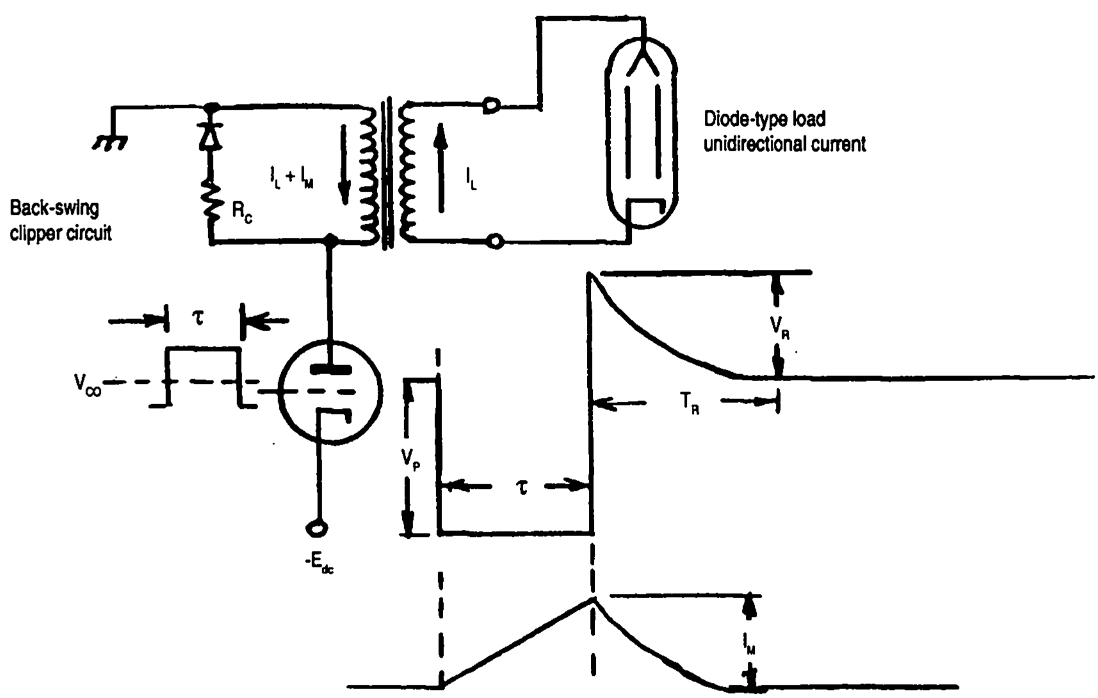


Figure 10-4. The back-swing clipper circuit as used with transformer-coupled unidirectional loads.

lator is almost always the most expensive, if not always the largest.

Before leaving the high-power output-pulse transformer behind, it would be negligent not to discuss its use in a hard-tube modulator that has a diode-type, or unidirectional, conduction (dc) load. We have already mentioned what can be done with energy stored in the transformer magnetizing inductance when used in a line-type modulator. So long as the switch tube is still in the forward-conduction state, the energy will manifest itself as reverse voltage on the pulse-forming network. This is not possible in a hard-tube modulator, because the means of terminating a pulse is to stop switch-tube conduction. This situation leaves no place for the flow of current, which has built up in the magnetizing inductance over the course of the pulse. For this reason, a back-swing clipper circuit, as shown in Fig. 10-4, is required. Without it, the voltage across the transformer primary will jump up at the end of the pulse in a direction that increases the forward voltage across the switch tube (just as it is trying to recover its voltage-blocking state and after its electrodes have been heated to their maximum transient temperatures). The result will be a switch-tube arc. The clipper circuit provides a path for the magnetizing current after the switch tube is cut off. This clipper current can be defined as

$$i_{clipper} = \int I_M e^{-tR_C/L_M} = \frac{I_M L_M}{R_C},$$

where R_C is the clipper resistance, L_M is the transformer magnetizing inductance, and I_M is the peak value of magnetizing current that occurs at the end of the pulse of voltage applied to the transformer. The energy stored in the transformer magnetizing inductance, W_M, can be defined as

$$W_M = \frac{1}{2} L_M I_M^2 = \frac{1}{2} L_M \left(\frac{V_p \tau}{L_M} \right)^2 = \frac{V_p^2 \tau^2}{2L_M},$$

where V_p is the voltage pulse output of the switch tube and τ is the pulse length.

There will still be a transient overvoltage due to the clipper resistance in series with the clipper diode. Without series resistance, the clipper time-constant could theoretically be infinite because it is L_M/R_C. The greater the value of R_C, the shorter the recovery time will be but the greater will be the peak reverse voltage. The volt-time products across a transformer winding must be equal in the positive-going and negative-going polarities. If they are not, the result is time-averaged voltage across the winding—which ultimately is just a length of wire—and it is difficult to maintain average voltage across it. The larger the value of clipper resistance, the shorter the decay time of magnetizing current but the greater the transient overvoltage, which is equal to $I_M \times R_C$. There can be an advantage to using a non-linear resistance such as Thyrite, or the more modern metal-oxide varistor (MOV), as the clipper load . A properly selected non-linear resistance will have nearly constant voltage across it throughout the interpulse interval, thus minimizing the stress on the switch tube immediately following the pulse, which is the time when the tube is least able to handle it. The MOV clamp voltage should be chosen so that the product of it and the minimum interpulse interval is as close to the product of pulse voltage and pulse duration as possible, making positive- and negative-going volt-time products equal across the transformer winding.

10.3 Storage capacitors

We all know that a capacitor is a device that consists of no more than two plates separated by dielectric material and is capable of storing energy in the electrostatic field between the plates. Nevertheless, entire books have been written on capacitor technology. What a transmitter designer is most concerned with is what the ratings of a capacitor mean and how the stress levels imposed affect capacitor performance.

Generally speaking, the factor most important in determining a capacitor's volume and cost is the energy it is capable of storing. This assumes, however, that once a capacitor is charged, its repetitive discharges will only be partial. (This was clearly not the case with the capacitors used to store energy in the PFNs of the line-type modulators. These capacitors were completely discharged each pulse during normal operation.) In addition, under some fault conditions the voltage can be reversed across the capacitor, which is one of the most stressful events a capacitor can experience—and this applies to capacitors that are *not* inherently polarized, such as electrolytic types. These caveats both hint at the fact that even though the energy stored in a hard-tube modulator capacitor bank might be 10 to 20 times that stored in the PFN capacitance for comparable pulse energy, we are not talking about the same kind of capacitors. Pulse-forming-network capacitors are high-stress components whose failure rates cannot be overlooked. (Water-cooling such capacitors is not unheard of, and most of them are more correctly rated in shot life or number of pulses in a lifetime rather than number of operating hours in a lifetime.)

On the other hand, capacitors for partial-discharge, or dc-filter, service are usually rated for a given operational life, which can be tens of thousands of hours at a given terminal voltage. They come in standard case sizes that differ in energy-storage capability. The height of their bushings also varies, depending on operating voltage independent of stored energy. These capacitors are expected to be operated at their rated voltages. The deratings for life expectancy, which primarily affect voltage stress in the dielectric material, have already been designed in. Operating them at reduced voltage will, of course, result in great extension of life but at the expense of their stored energy, which varies as the square of the voltage.

However, these capacitors are no more tolerant of voltage reversal than any other kind. They will be fully discharged by a short-circuit load fault or by the operation of an electronic crowbar switch. It is extremely important that there be sufficient resistance in series with the discharge loop so that the discharge is critically, or overly, damped in order to preclude voltage reversal. This means that the series inductance of the loop cannot be ignored. In most cases, the value of series resistance will depend upon the permissible peak fault current and will be safely above the value required for critical damping,

$$R = 2\sqrt{\frac{L}{C}}.$$

But it is always prudent to calculate, or measure, the actual value of discharge-loop series inductance just to be sure. (This includes the internal series inductance of the capacitors themselves, which can vary depending upon how they are internally constructed.)

In at least once case, an energy-storage system with a greatly undermatched artificial transmission line was used in lieu of a low-inductance capacitor bank having the same value of capacitance. In this design, characteristic impedance was 10% of the load impedance. The delay-time of the network was made equal to 1/2 of the longest pulse duration. The load—a super-power, long-pulse klystron—was connected in series with a hard-tube modulator switch, much like the one shown in Fig. 10-3d. When the modulator switch was gated on to initiate a pulse of load current, there was an instantaneous drop in network voltage of approximately $0.1V$, where V is the initial voltage across the network. The voltage across the klystron and modulator-switch combination was $0.9V$. This voltage would persist for a time of $2T$, where T is the delay-time of the network. At $2T$, the voltage would drop again by an additional $0.1V$, to $0.8V$. If, however, the modulator switch tube was turned off after time T but before time $2T$ had elapsed, most of the network would have been discharged by an amount $0.2V$, but not all of it.

The load current was 0.9 V/R, where R is the effective resistance of the klystron in series with the conducting-state switch tube. If the storage system used the same value of capacitance but only the minimum-achievable value of inductance, the terminal voltage of the bank would start to droop in essentially linear fashion from the beginning of the current pulse to the end. If the starting voltage is V and the initial load current was V/R, the total droop would equal to the charge removed divided by the total capacitance, or $\Delta V = I\Delta t/C$. (This discussion is simplified by assuming that current is constant throughout the pulse, which it isn't.) The time increment, dt, is the pulse length, $2T$, which was the same as $\sqrt{(LC)}$. The total load resistance was $10Z_0$, where Z_0 was the characteristic impedance of the network described above, or $\sqrt{(L/C)}$. So $R = 10\sqrt{(L/C)}$ and $\sqrt{(L/C)} = 0.1R$. Combining things,

$$\Delta V = \frac{I\Delta t}{C} = \frac{V}{R}\times 2\frac{\sqrt{LC}}{C} = \frac{V}{R}\times 2\sqrt{\frac{L}{C}} = \frac{V}{R}\times 2\times 0.1R = 0.2V,$$

which was the same as the end-of-pulse step from the distributed network described above. The charge and energy removed from the capacitance was the same in both cases, except the finite-Z_0 network voltage changes in stair-step fashion while the minimum-Z_0 capacitor bank changes in continuous ramp fashion.

The purpose for creating this artificial network was to minimize the anode dissipation of the modulator switch because the source voltage was constant throughout the duration of the pulse, at $0.9V$. In order to produce a flat-top load voltage pulse in the latter case, the ramplike, or droop, component would have to be subtracted from the output. To keep klystron voltage constant at $0.8V$, the switch-tube anode voltage would vary from $0.2V$ at the beginning of the pulse to a theoretical minimum of zero (assuming no head-room requirement). The additional anode dissipation per pulse would be $0.1V \times I$, where $0.1V$ is the time-averaged anode drop and I was the load current, as before.

This may have sounded like a good idea, but it wasn't. Its fatal flaw was that it didn't account for what happens under the major fault condition: simultaneous klystron and switch-tube arc-breakdown, or load shoot-through. The only good news was that the fault current was limited by the network characteristic impedance to 10 times I, where I is the normal peak load current. Without series resistance, however, the network stored energy would tend to oscillate back and forth, completely reversing the voltage on the network capacitance periodically until the load arc was finally extinguished. To completely absorb network energy without capacitor-voltage reversal, a series resistance equal to Z_0, or $0.1R$, was required between network and load. But this completely subverted the reason for the network in the first place. (This is but another example of having to design for the worst-case scenario rather than the hoped-for "normal" operating conditions.)

Rarely, if ever, are the capacitive energy-storage requirements of even a modest-sized transmitter met by a single capacitor (or "can," as it is sometimes called). Multiple cans connected in parallel or series-parallel are the rule rather than the exception. The danger in multiple-capacitor arrays is what can happen if a single capacitor internally short circuits. If the total surge-current-limiting resistance is distributed throughout the array by having at least one resistor capable of dissipating the energy stored in a capacitor in series with each capacitor, a short-circuit failure will cause no damage because virtually no energy will be dissipated in the faulted capacitor. Such an arrangement is so fault tolerant that the failed capacitor may not immediately reveal itself. (If the capacitors are all in parallel, the resistor in series with the failed capacitor will be connected across the dc power supply when an attempt is made to turn it on again. What is left of that resistor will usually be an adequate tell-tale sign for where the trouble lies.)

Series-parallel connections are often employed when the system's operating voltage is much above 50 kV. This is because energy storage density tends to degrade at much higher voltage, and more volumetrically efficient banks can be assembled from capacitors rated at less than 50 kV—by no means a hard and fast number—than those rated at, say, 100 kV. (In fact, probably the most volumetrically efficient energy storage bank assembled for 100kV-plus operation used a multitude of electrolytic capacitors that were individually rated at 450 Vdc. They were assembled in over 200 series-connected tiers and protected from individual overvoltage by a string of Zener diodes shunting them.) When there is a large number of series-connected tiers—say, at least five—the presence of a shorted capacitor in one tier will be far from obvious, even though it will short-circuit all of the parallel-connected capacitors in that tier through its series resistor. This short will cause the overall bank capacitance to increase, but all of the remaining capacitors will be subjected to 25% overvoltage, or less, at the operating voltage. For this reason, some capacitor banks are interconnected with fusible links in-

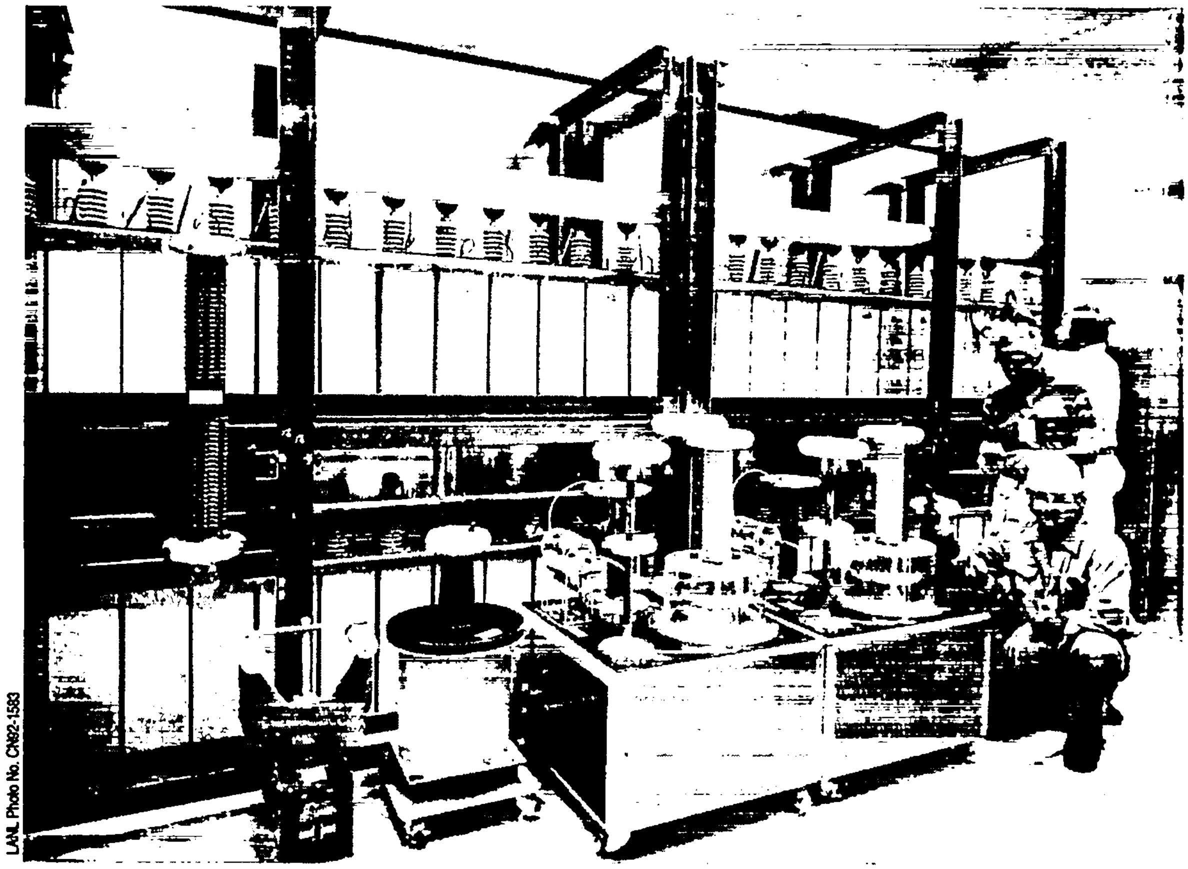


Figure 10-5. Typical series/parallel capacitor bank (two groups of 25 parallel-connected capacitors) totalling 70 μF, with triggered air-gap crowbar switches in foreground.

stead of individual series resistors. Fusible links have the advantage of automatically disconnecting a failed capacitor. The fuse, however, must be of the type that permits very little let-through energy, meaning that it must be arc-quenching. A capacitor subjected to too much let-through energy is likely to see its can rupture or its bushing explosively expelled—a spectacular but very hazardous event. When a capacitor bank has only two tiers, the effect of an individual shorted capacitor is the most easily observable, but the penalty for not observing it is the greatest: the possibility of a 100% overvoltage of the remaining tier. The voltage rating of a capacitor is associated with life expectancy more than external arc-over, and many high-voltage capacitors will endure 100% overvoltage for a short time. With only two or three tiers to the bank, it is advisable to measure voltage balance between tiers continuously and provide this information to the high-voltage dc interlock system of the transmitter.

Stacked, multi-tiered capacitor banks also require corona shielding. A capacitor designed to operate at 50 kV may (or may not) have can and bushing geometries shaped so as to be corona-free (see Chapter 7). The capacitor cans of the first tier will typically be connected to the high-voltage return bus, the high-voltage-system "neutral," which will be at or very near ground potential. The cans of the capacitors in the next tier will be at 50 kV with respect to ground, and their high-voltage-bushing terminals will be at 100 kV with respect to ground, and so on up the stack. The radii of curvature that are required for corona-free operation continually increase, but the capacitor can and bushing geometries do not, because most practical banks are built up of identical capacitors. Capacitors, therefore, are usually mounted on "rafts" that are surrounded by oblong, donut-shaped corona shields whose cross-sectional diameters increase as the rafts ascend in height and voltage. An example of such construction is shown in Fig. 10-5. Please note that if spacings between conductors are adequate, very little shielding is required to prevent voltage breakdown. This is because corona under these conditions is self-healing; the exterior of the ionization sheath tends to expand so as to have the same effect as a large radius-of-curvature surface. However, in today's sophisticated systems, which involve ever-more-sensitive and susceptible low-level circuitry, the electrical noise resulting from the corona dis-

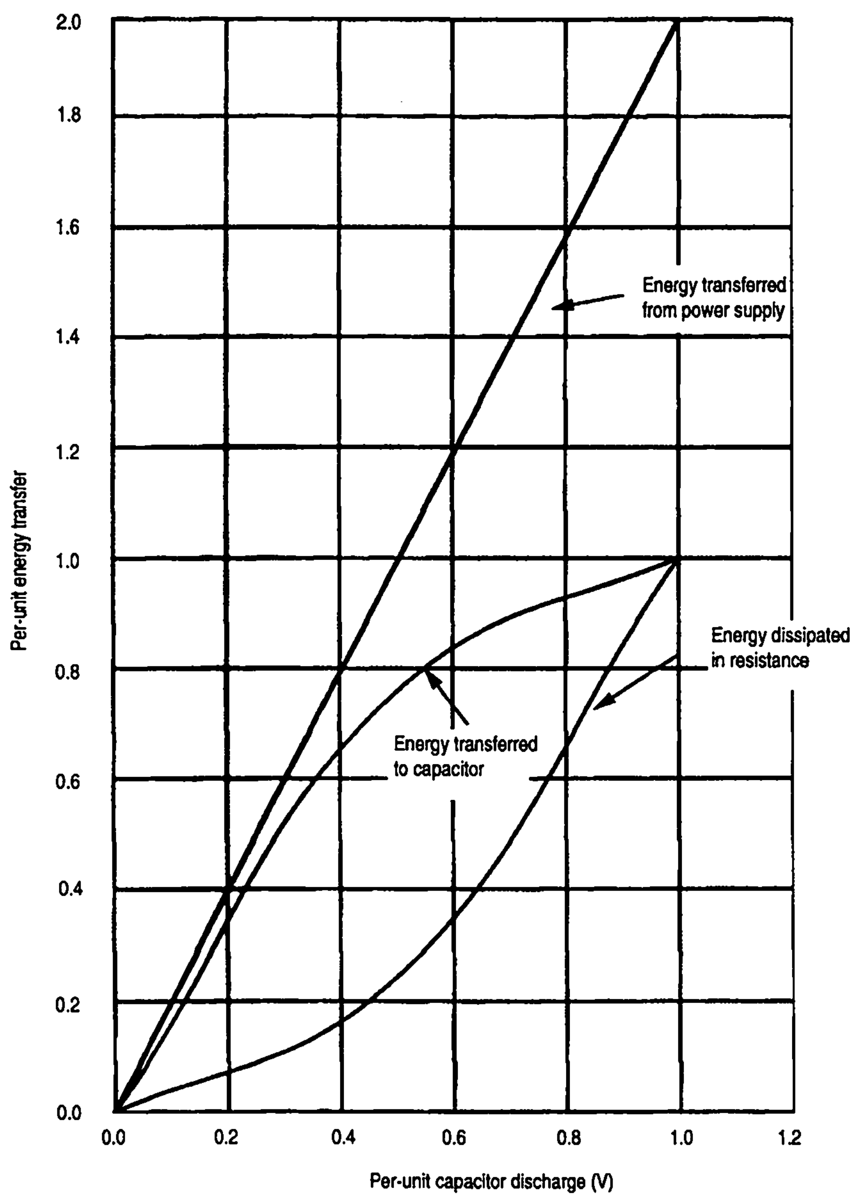


Figure 10-6. Capacitor charging energy relationships as a function of amount of capacitor discharge.

charge is increasingly intolerable. (Remember what we used to pick up on the AM car radio when we drove near high-tension wires, especially on a foggy day?)

We have discussed the fact that the removal of charge from a capacitor bank, no matter how large, will result in its terminal voltage being smaller than it was before the charge-removal interval, or pulse. In the interpulse interval, therefore, it will be necessary to replace the charge removed during the pulse. In the discussion of line-type modulators, much was made of the issue of recharging without incurring energy loss if the current was limited by resistance. Is there a similar concern when it comes to replacing partial discharge? The answer is not nearly as much. The reason for this is shown graphically in Fig. 10-6, which plots energy removed from the dc power supply, energy transferred to the capacitor bank, and energy dissipated in series resistance versus the per-unit amount that the capacitor is discharged each pulse. (This graph assumes that the resistance is small enough so that the *RC* time-constant is less than approximately 1/3 of the minimum interpulse interval, assuring that the capacitor will recharge to approximately the power-supply terminal voltage between pulses.) If we look at full discharge, or per-unit discharge of unity, the results are familiar. The per-unit capacitor energy transfer and resistor energy dissipation are both unity, and the per-unit energy removed from the power supply is 2. Notice, however, that only the energy removed from the power supply is a linear function of per-unit discharge voltage. For less-than-unity per-unit discharge, energy transferred to the capacitor falls off less rapidly than dissipation in the resistor. Therefore, the efficiency of energy transfer increases as $\Delta V/V$ decreases.

The reason for this can be seen if we note the change in charge during a pulse of current, $\Delta Q = C\Delta V$, and the change in capacitor energy, $\Delta W = \Delta Q$, times average capacitor voltage during discharge, which can be expressed as

$$C\Delta V \times \frac{V + V - \Delta V}{2}$$

or

$$CV\Delta V - \frac{C(\Delta V)^2}{2} \ .$$

The energy removed from the power supply during recharge is ΔQ times the power supply voltage, which can be expressed as $C\Delta V$ x V. Recharging efficiency is the ratio of energy transferred to the capacitor to the energy removed from the power supply, or

$$\frac{C\left(V\Delta V - \dfrac{\Delta V^2}{2}\right)}{CV\Delta V} = 1 - \frac{1}{2}\left(\frac{\Delta V}{V}\right).$$

When $\Delta V/V = 1$, full discharge efficiency is 1/2, which we knew. For small per-unit discharge—say, 0.1, or 10% droop—efficiency is 1 - 1/2 x 0.1 = 1 - 0.05 = 0.95, or 95%.

In most high-power applications, however, resistance is not used between power supply and capacitor bank. Recharging current from the power supply is always limited by the equivalent ac series impedance of the power supply, most of which is the leakage reactance of the inductively coupled components, especially the high-voltage rectifier transformer.

10.4 Vacuum tubes as switch tubes

Power grid vacuum tubes have numerous characteristics that define their total performance capabilities as pulse-modulator switching devices. The first characteristic of importance is peak-pulse power output, which is the product of the simultaneous maxima of load voltage and current that can be switched. Of course, the maximum load voltage is related to the continuous voltage that can be applied between anode and cathode of the tube while maintaining an accept-

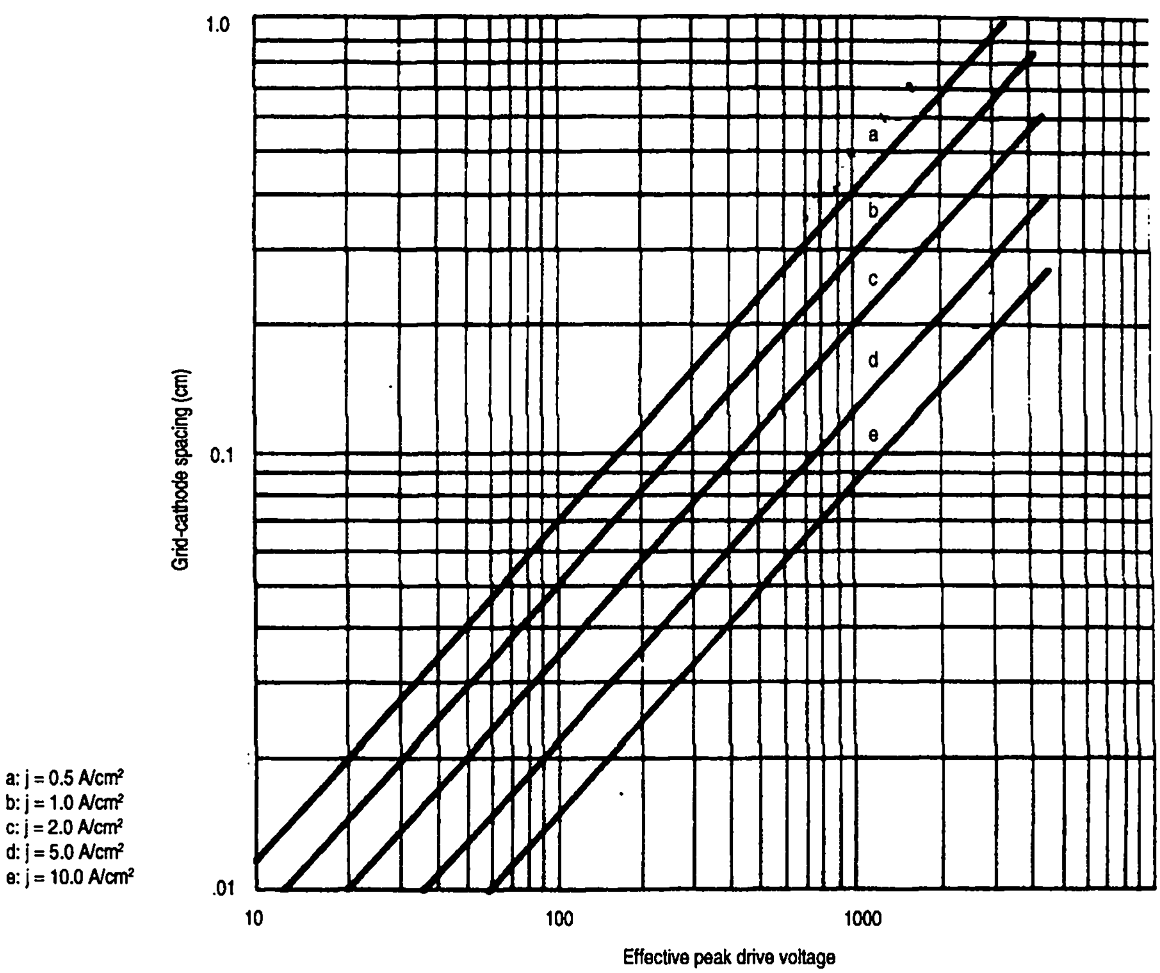


Figure 10-7. How cathode emission density is affected by spacing between grid and cathode (effect on perveance).

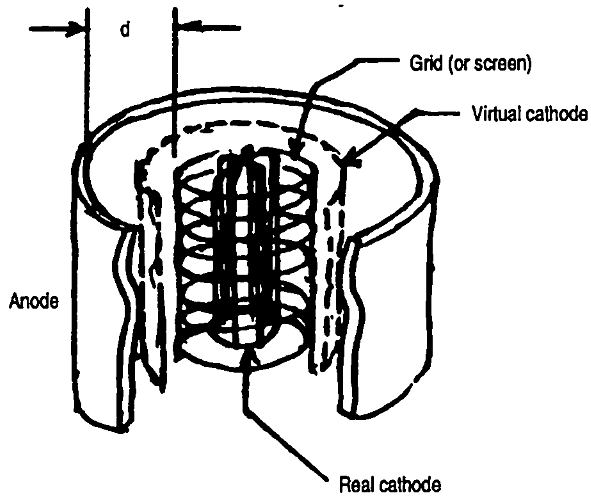


Figure 10-8. How current density in region between anode and outermost grid is affected by the spacing between them and effective drive voltage, due to the creation of a virtual cathode between the electrodes.

ably low spark-down rate. Already it is obvious that there will be a degree of flexibility in what the usable anode hold-off voltage really is. In an application where a switch-tube arc is tantamount to disaster, an extremely conservative value must be used that will probably be even less than the manufacturer's suggested operating voltage. Even this number is already somewhat less than the "absolute-maximum independent rating," which defines voltage-hold-off capability at essentially no plate current.

The useful load voltage is affected by load current as well. Maximum load current depends upon the type of cathode, its surface area, and temperature, which is related to heating power; and the current split between anode and other electrodes, such as the grid, shield grid, and screen grid. Although basic cathode research is ongoing and types capable of sustaining high emission densities over long lifetimes have come forth, power-grid-tube cathodes are still of two basic types: thoriated tungsten (directly heated) or oxide-coated (unipotential, or indirectly heated). (Million-hour emission lifetimes can at least be projected from today's data bases for dispenser-type cathodes.) As current from the cathode increases, so must the minimum voltage between anode and outermost grid in order to avoid formation of a "virtual cathode" in the space between the two. Such an occurrence would prevent the passage of electrons through the region. (The electrons will not even leave the cathode unless there is adequate effective drive voltage.)

Figure 10-7 relates cathode emission density to grid-cathode spacing and effective drive voltage. (The last parameter includes the contributions of anode, screen, and control-grid voltages as they are weighted by their respective amplification factors.) Figure 10-8 shows how the current density in the region between the outermost grid and anode is affected by drive voltage and spacing between the anode and outermost grid. These relationships can be defined by the equation

$$j_0 = \frac{2.33 \times 10^{-6}\left(\sqrt{V_A + V_G}\right)^3}{d^2},$$

where j_0 is the density of current that can traverse the region between the outermost grid and anode, V_A is the voltage between anode and cathode, V_G is the voltage between grid and cathode, and d is the spacing between the outermost grid and anode. Notice that the useful component of current that travels from the grid to the anode is inversely proportional to the square of the distance between them.

Figure 10-9 shows why this is not good news. The figure relates anode-grid hold-off voltage to the spacing between them. The uppermost plotted line, A, is the plane-parallel conductor experimental data, which is of mostly academic interest. Note the intercept at 100 kV for 1-cm spacing. The 100 kV/cm gradient is often cited as a criterion for peak electric field in electron guns that use smooth copper surfaces, such as the ones for large klystrons. (An electric field of 100 kV at 1 cm may sound like a lot, but it is only about three times as great as the corresponding number for the dielectric strength of air at sea level. The actual geometries of power gridded tubes behave in less predictable fashion.) Getting back to Fig. 10-9, element B is a region rather than a line. It defines what one

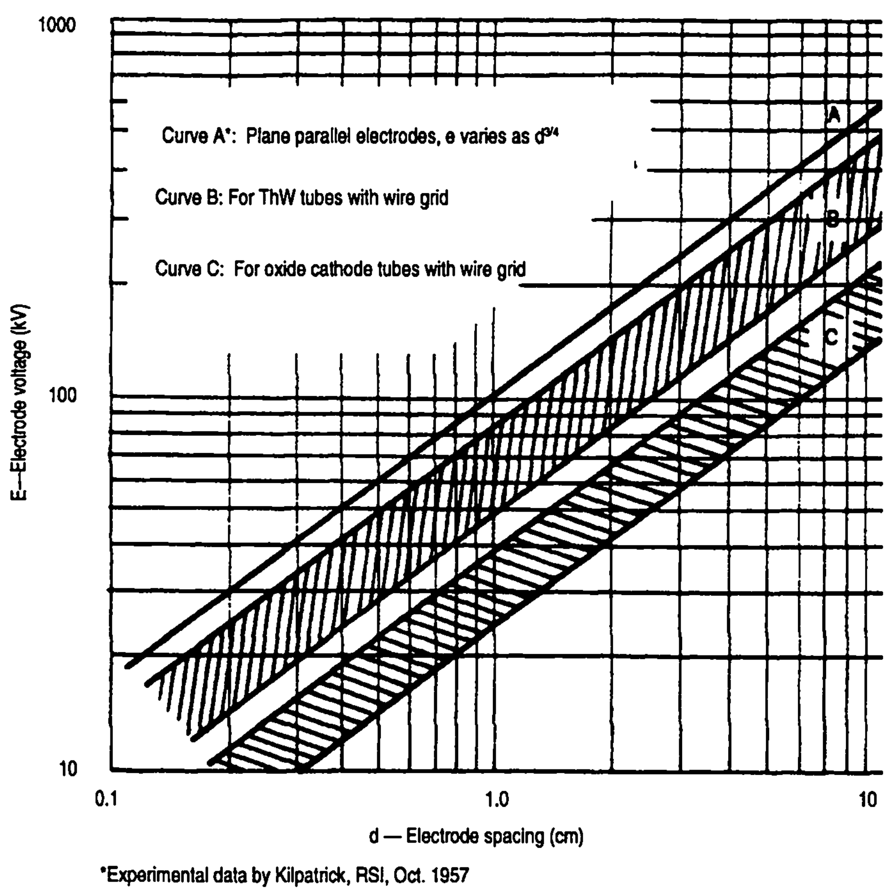


Figure 10-9. Voltage hold-off between electrodes in vacuum as functions of spacing and surface quality.

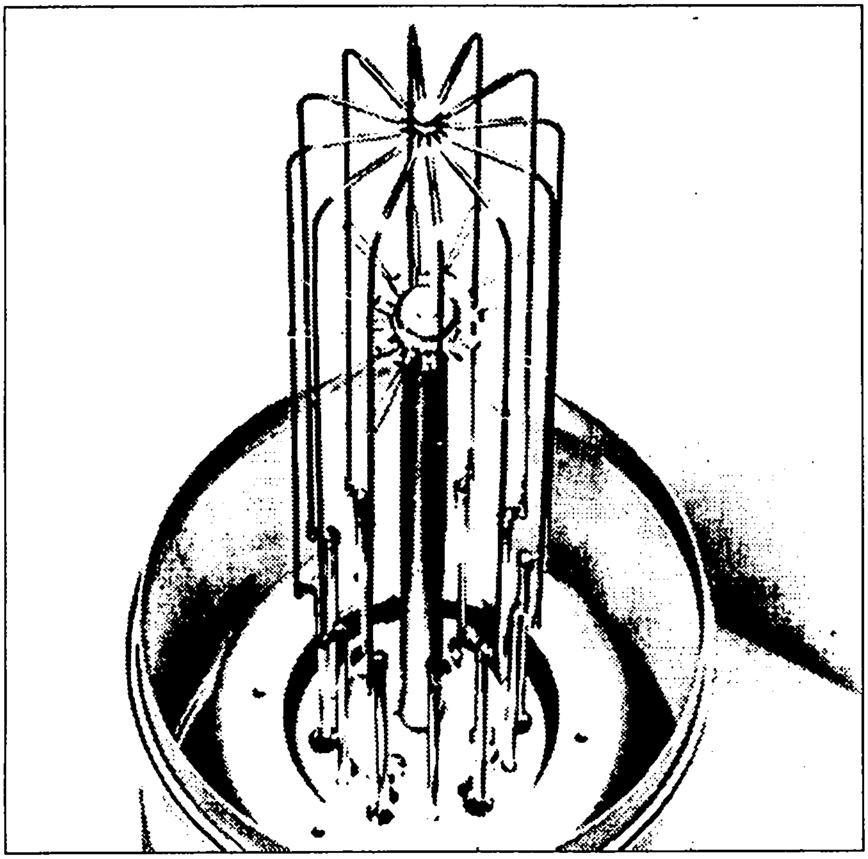


Figure 10-10. Typical directly heated thoriated-tungsten wire-type cathode for power-grid tube.

might expect for a tube with a wire grid and thoriated-tungsten (ThW) cathode. Region C defines the performance to be expected if the cathode is oxide-coated rather than thoriated tungsten. Oxide cathodes are, quite literally, flakier than ones made of thoriated tungsten, and the products of surface deterioration can contaminate the grid wire, producing localized electric field enhancement. Tubes with higher μ factors have more grid wires that are closer to the cathode so that the wire mesh functions more like an equipotential surface rather than individual wires with respect to the electric-field-strength enhancement produced by conductors of small radius-of-curvature. Smoothness of the wire is also important, but once an arc of sufficient action has terminated on a grid (or screen) wire, that wire, even if it is still there, will no longer be smooth.

If oxide-coated cathodes perform less well with respect to high-voltage hold-off, why are they used at all? The reason is that thoriated-tungsten cathodes are not as efficient emitters with respect to heater power. For example, on a thoriated-tungsten cathode, one ampere of peak emission requires more than 15 W of heater power—and this is if a highly regulated filament voltage is used to preclude temperature-limited operation. (Pure tungsten filaments, on the other hand, are often deliberately operated in a temperature-limited fashion in current-control diodes, where current is determined by filament power rather than anode-cathode voltage). There is no pulse-duration limitation for such cathodes.

A comparable oxide-coated cathode may require less than 1/3 as much heater power for the same emission current, but pulse durations are limited to tens of

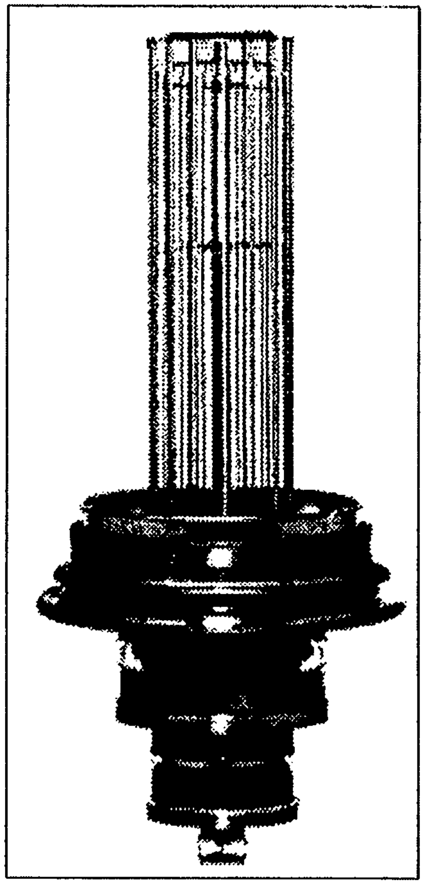


Figure 10-11. Typical helical wire-type control grid for high-power triode vacuum tube.

microseconds and duty factors to 1% or less, which is more than adequate for many applications. Useful emission density for both types is 2 to 3 A/cm^2. Even though emission from an oxide cathode can be 10 to 50 times this great for microsecond-length pulses, the only current that matters is what can get from outermost grid to anode. This current is limited by the virtual-cathode effect mentioned earlier.

The thoriated-tungsten filament can be expected to last more than 10,000 hours (or one year of continuous operation). This longevity is comparable to the life expectancy of the rest of the equipment if utilization is sporadic. The life expectancy of oxide-coated cathodes depends on too many factors to be accurately predicted or even repeated, but it is typically less than half that of thoriated-tungsten.

Figure 10-10 shows how a directly heated thoriated-tungsten wire cathode for a high-power gridded tube is arranged. Such structures can become quite tall, and, when excited with ac heater power, can vibrate like tuning forks due to magnetic force caused by the filament current. Such filaments may require dc or high-frequency ac in order to produce no unbalanced magnetic force. (Or, as we will see later, such filaments may actually be built in three symmetrical sections in order to be excited with balanced 3-phase ac.)

Figure 10-11 shows what a similar tube looks like after a helical grid has been wound around the cathode wires. The closer the grid is to the cathode, the higher the μ factor will be, the higher the cathode-current-density change will be for a given grid-voltage change, and the greater the anode-voltage hold-off is likely to be. The wire-to-wire spacing of the grid, however, will have to be smaller, approximately equal to the spacing between grid and cathode. This is done so the "screening" effect of the grid, or the amount of total grid area that is physically blocked by the presence of grid wire, will be greater than that of a lower-μ grid spaced farther away from the cathode. The greater the screening effect, the more likely the grid is to intercept electrons on their way from cathode to anode when the grid is driven positive with respect to the cathode. This screening effect is reduced if the diameter of the grid wire can be reduced. But the diameter must still be large enough to dissipate heat, including the heating caused by its proximity to the hot cathode and the interception of electrons. Even more importantly, the wire must have cross-sectional area (or AWG gauge) sufficient to absorb the action content of an arc discharge without fusing. (As will be discussed later, the transmitter designer is responsible for limiting the arc action by employing surge-limiting resistance and, when needed, an electronic crowbar charge diverter.) Even in high-μ tubes, the grid screening factor is less than 25%, and it typically varies down to 10% for low-μ tubes. Grid current at anode-saturation voltage is typically 1.5 times the screening factor, so that a 25%-screened, high-μ triode may intercept grid current that is 1.5 times 25%, or 38%, of cathode current, compared with 1.5 times 10%, or 15%, for a low-μ triode.

Grids are almost always wound of tungsten or molybdenum wire that has been coated with gold or platinum. This is done so they can operate at temperatures higher than the cathode—even as high as 1400°C (compared to 1000°C for the cathode)—without becoming primary emitters of electrons themselves. When they are driven positive with respect to the cathode they will attract and intercept electrons. The sum total of these electrons constitutes grid current, and the product of this current and the peak-grid voltage is the peak-grid dissipation. This current and power must be supplied by the external grid-drive circuitry and must be dissipated by the grid as heat.

Because of an effect called secondary emission, not all electrons that collide with the grid are counted as grid current, however. Primary electrons strike the grid surface and knock off others, which continue on to the anode. So long as the secondary emission is less than the primary interception, things look good for both the tube and driver. On the other hand, if more secondary electrons are emitted than primaries are collected, grid current is negative and the grid can run away, unless it is externally swamped by shunt resistance. Pulse stretching is the usual symptom of excessive secondary emission; cathode and plate currents continue even after the grid drive pulse has stopped. Pulse stretching in some thoriated-tungsten tubes can be corrected by a process called grid blackening. In this process, filament voltage is increased to twice its normal value (so filament power approximately trebles). Some of the filament carburization boils off and is deposited on the grid, making its surface stickier and less prone to secondary emission. Secondary emission is less likely from grid wires whose surfaces have been made intentionally rough, but anode-voltage hold-off is degraded as a consequence.

Recently, manufactures have made grids and screens of pyrolytic graphite. This material does not emit secondary electrons, but tubes using it do not always please their users because net grid current is always higher for them than conventional grids. Higher current means that grid drive circuits and screen-grid power supplies have to be rated for more output current. Nevertheless, with pyrolytic-graphite grids and screens there is less likelihood of tube-to-tube variations, or variations within a single tube over its life span.

In oxide-cathode tubes, pulse duration is limited by the cathode. In thoriated-tungsten tubes, pulse duration is usually limited by intrapulse grid heating, which, ignoring heat lost by radiation, is adiabatic, meaning that no heat is dissipated during the pulse itself. Therefore, the temperature rise is proportional to the pulse energy, which is the product of peak power and pulse duration, and inversely proportional to the product of grid-wire mass and specific heat. If power is expressed in watts, pulse length in seconds, and mass in grams, the proportionality constant for temperature rise in degrees Celsius is 0.24. With grid temperature at 1400°C, however, radiation losses are fortunately not insignificant.

The last electrode of a power grid tube is, of course, the anode, which is where we would like the majority of our cathode-emitted electrons to end up. A tube designer wants the electrons to arrive at the anode with a kinetic energy that is as small as possible relative to the anode area and the spacing between anode and outermost grid. This kinetic energy and the rate of electron arrival—which is, of course, the anode current—determine the anode dissipation. Only so much can be done to minimize it, and a conduction-anode drop that is 10% of the anode hold-off voltage is considered a good design, resulting in 90% anode efficiency as a modulator switch. The bulk of anode design, other than geometric requirements, falls in the category of heat transfer, which will be touched upon in a later section.

Needless to say, grid and/or screen current does no one any good, and tube designers have exercised a great deal of their creativity in minimizing it. Now we will look at a few actual tube designs to see what has been accomplished in this regard. (Remember, of course, that almost nothing new has been done in this field for almost 30 years.)

10.4.1 The 7560 triode switch tube

Figure 10-12 shows the plate (anode) current characteristics for the 7560 triode modulator switch tube, which once was simultaneously manufactured by three of the largest and most competent power-vacuum-tube sources in the world. Today it may not be made by anyone, because two of the three have ceased operations and the third may no longer be interested. Nevertheless, there is still a host of operating sockets for the tube type, and shops that can rebuild it still flourish. The figure also shows the characteristics of another power triode, the 8547, which is not primarily a modulator switch tube because its nominal anode-voltage rating is 17 kV. Both tubes have water-cooled anodes capable of dissipating 175 kW of power. The 7560 differs from the 8547 only in the increased spacing between its anode and grid, which is required to achieve its 50-kV anode-voltage hold-off rating. Because the anode is farther away from the grid in the 7560, a designer would expect that greater grid-anode voltage drop would be required for a given amount of anode current given its virtual-cathode effect. Indeed, the effect is clearly illustrated in Fig. 10-12a by the slope of the "diode

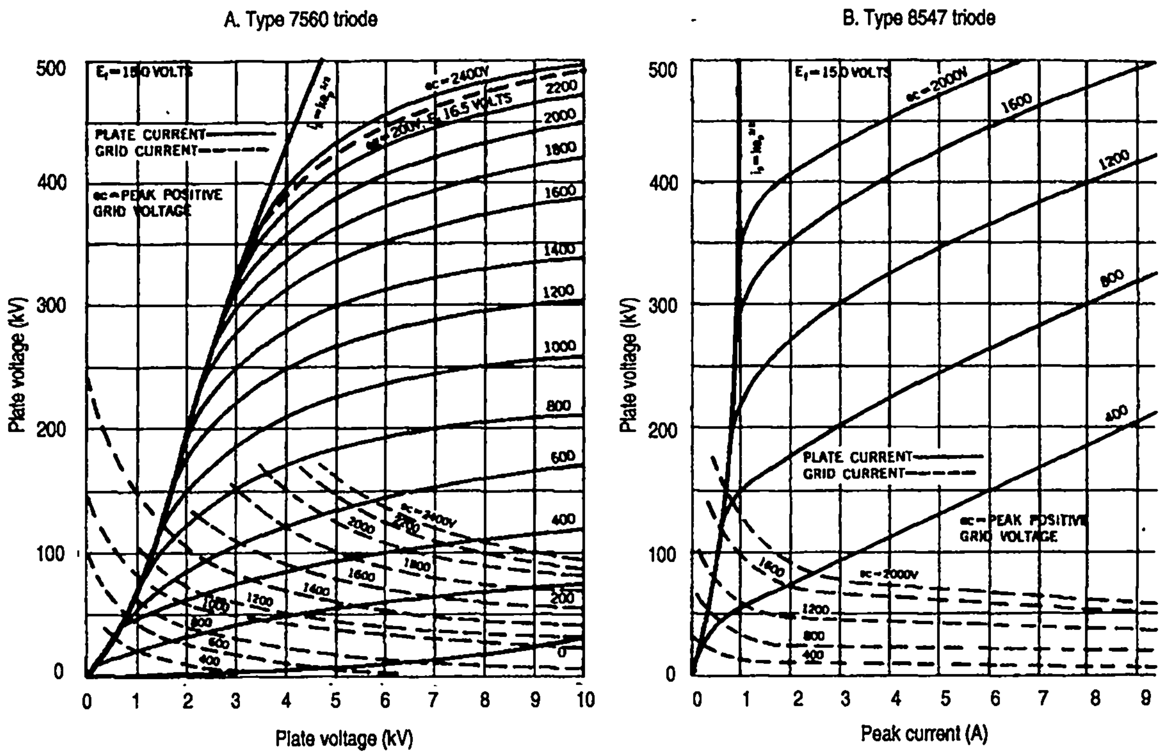
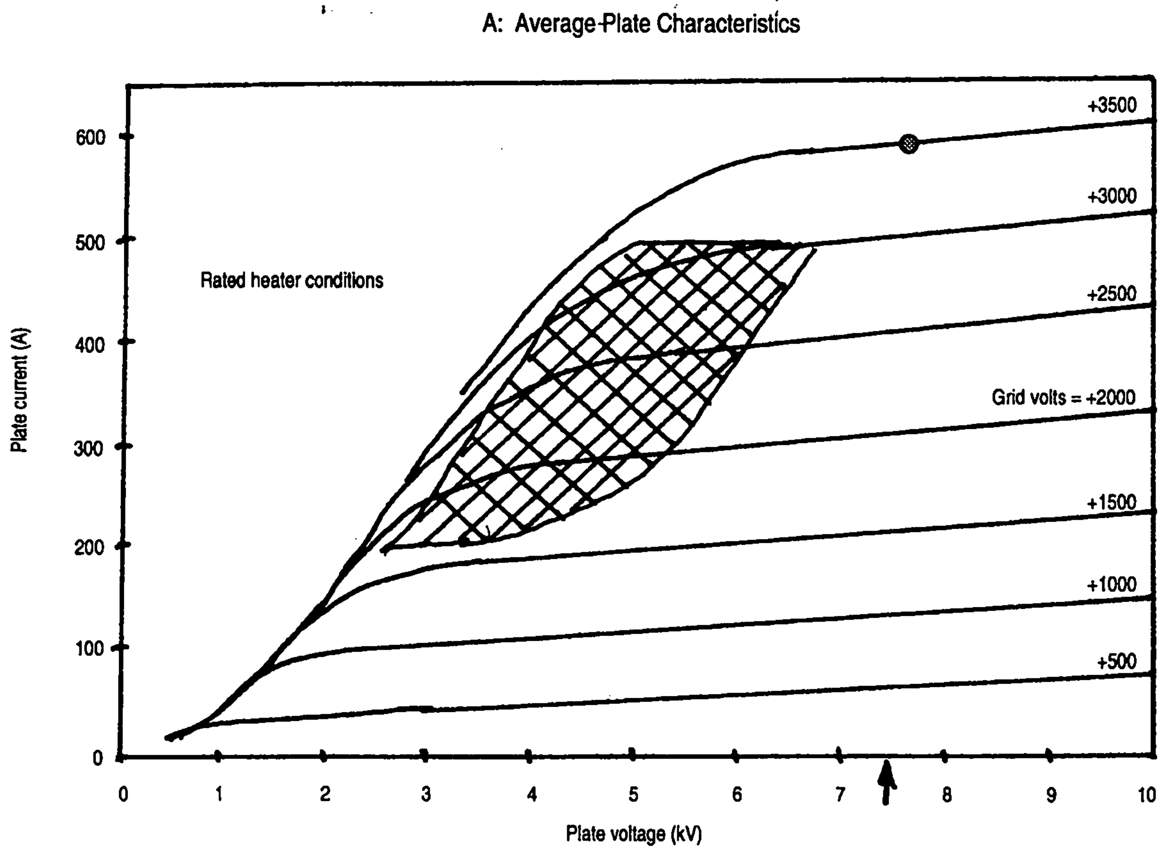


Figure 10-12. Voltage/current characteristics for two power triodes: type 7560 (left) and type 8547 (right), which differ only in that the 7560 anode is farther from its grid.

line," which is the line (not exactly straight) that slopes upward and to the right from the origin. It is also the line where all of the anode current lines merge. This diode line quantifies the virtual cathode and has a much steeper slope in the closely spaced electrodes of the 8547 than in the widely spaced ones of the 7560.

If we select operating points for both tubes that meet the arbitrary but sensible 90%-anode-efficiency criterion—at which the average output power could conceivably be as high as 175 kW/0.1, or 1.75 MW—we can determine how well grid current is managed in the two tubes. For the 7560, the conduction drop selected is 5 kV, which is 10% of 50 kV, and for the 8547 it is 1.7 kV, 10% of 17 kV. With positive grid voltage of 2000 V, the anode current of the 7560 is 380 A, and its grid current is 130 A. Therefore, the ratio of the two values is 2.10. Of the total 513-A cathode current, which is the sum of anode and grid currents, 74% became anode current and 26% became grid current. This breakdown suggests a grid-screening fraction of 17%, which would put the 7560 in the medium-μ category. (Its μ, in fact, is 45.) With 50 kV between anode and cathode, the grid must be negative with respect to cathode by approximately 1.1 kV for "projected" plate-current cutoff. (The actual plate-current cutoff requires an additional safety factor of negative grid bias.) The 8547, with a positive grid voltage of 2000 V, an anode current of 390 A, and a grid current of 100 A, has a ratio of 3.10. Of the 490-A cathode current, 80% made it to the anode, and 20% to the grid, suggesting a 13% grid-screening factor and lower μ. We know, however, that the tubes are identical except for anode-grid spacing, so the grid-screening factors are the same. The μ is lower, however, because the anode is closer to the cathode ($\mu = 14$).



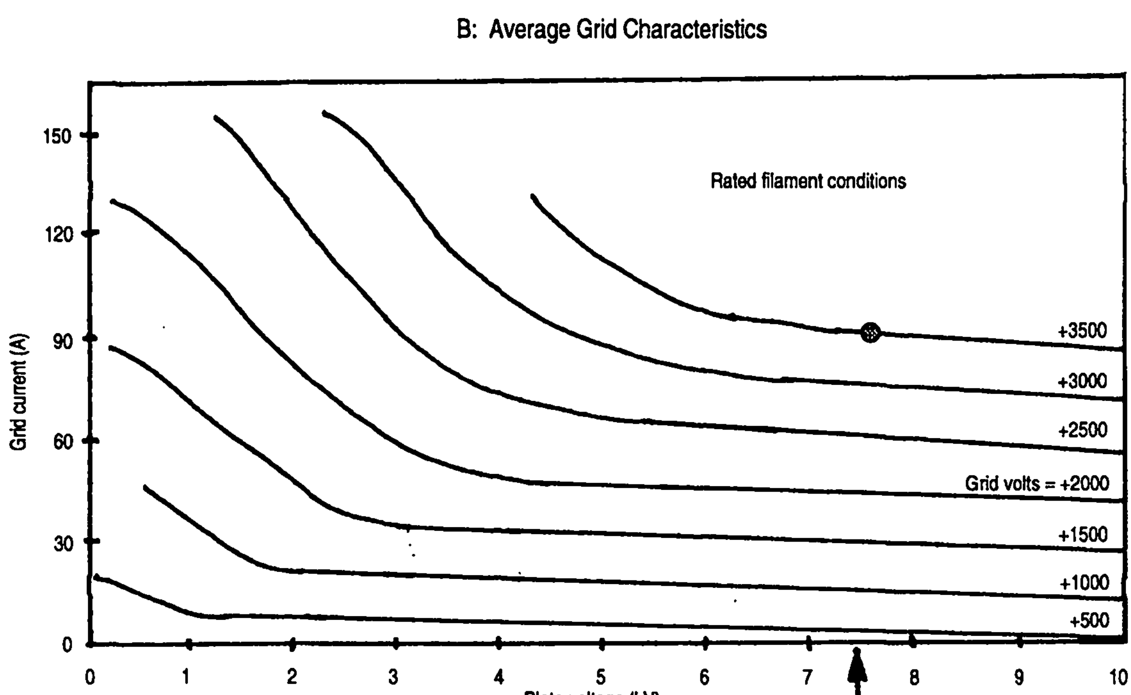


Figure 10-13. Voltage/current characteristics for the type 8461 triode.

The filaments of both tubes operate at 14.5 V and 450 A, or 6250 W. At the rate of 15 W per cathode peak ampere, the available cathode current should be 435 A. The actual peak-cathode current rating is 550 A. Although the 8547 more efficiently converts cathode current to useful anode current, its peak-pulse output power under the conditions evaluated is 6 MW, whereas the peak power from the 7560 is almost 20 MW. Neither of these tubes employs any special strategy, either geometric or electromagnetic, to minimize grid current. The grid is no more sophisticated than a helix of wire wound around the array of cathode wires.

10.4.2. The WL-8461 triode switch tube

This is another example of a thoriated-tungsten cathode. The WL-8461 is a "big but simple" power-grid triode modulator switch tube. Many sockets throughout the world are also filled with this tube, but it is no longer being manufactured. It is, however, being successfully rebuilt.

The WL-8461 is "big" in that its anode-voltage hold-off rating is 75 kV, its peak cathode-current rating is 750 A, and its water-cooled anode-dissipation rating is 200 kW. It is "simple" in that nothing special was designed into this tube to minimize grid current. Figure 10-13 shows its anode- and grid-current characteristic curves. Using the same 90%-anode-efficiency criterion (which yields a conduction-anode drop of 7.5 kV) and a positive grid voltage of 3.5 kV, the anode and grid currents are 600 A and 92 A, respectively, for a total cathode current of 692 A. The current ratio, therefore, is 6.5. Anode current is 88% of cathode current and grid current 12% of it. These values suggest a low μ. (Its μ is actually 25, which is somewhat low.) In Fig. 10-13a, note the cross-hatched area in the anode (plate) current characteristics just to the right of the virtual-cathode diode line. This cross-hatch area is bordered approximately by the +1500-V and +3000-V grid-voltage lines and the 3-kV and 7-kV anode-voltage lines. This is a region where a designer can expect oscillatory instability when the frequency is in the VHF range. The problem derives from the virtual cathode; there is a region of instability between the grid and anode where electrons don't know whether to go forward or backward. (This same effect was deliberately exploited in the Vircator, or virtual-cathode oscillator, an ultra-high-power microwave device.)

Another dimension of 8461 bigness is its directly heated cathode, which is physically so large that it is divided into three sections and excited with 3-phase ac. Each section operates at 15 V and 200 A, for a total of 9 kW input. At 15 W per ampere, the cathode current rating would be 600 A. Figure 10-14 shows the external physical dimensions of the 8461 and the 7560 to give an idea of the large size of tubes in this class.

10.4.3 The ML- 6544 triode switch tube

The objectives in placing the control grid close to the cathode are three-fold:

- to maximize the perveance, or the current density, from the cathode as it relates to the 3/2-power of positive voltage applied between grid and cathode;
- to maximize the μ-factor, which minimizes the negative voltage that must be applied between grid and cathode (or screen-grid and cathode in a

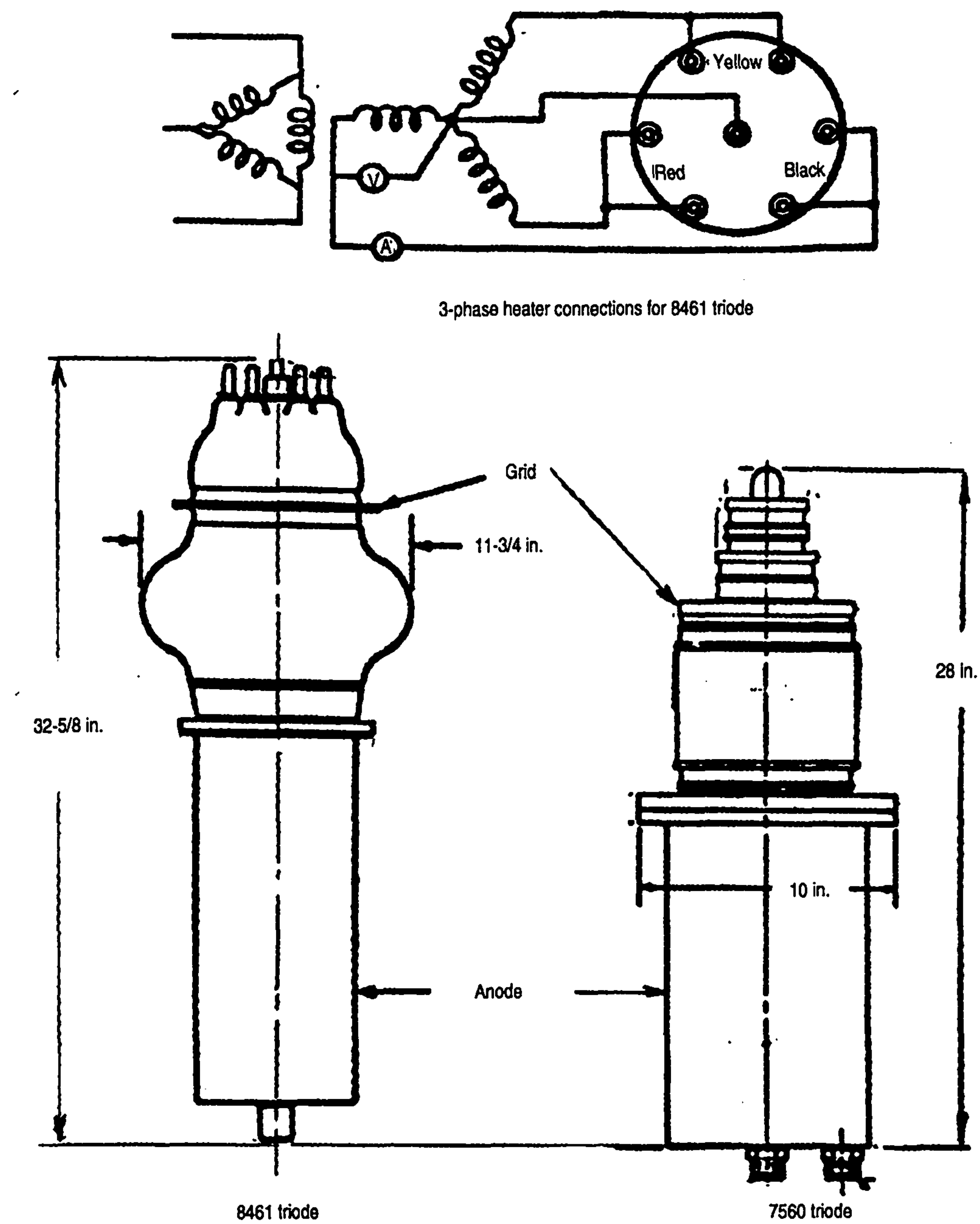


Figure 10-14. Physical attributes of type 8461 and 7560 power-grid triodes.

tetrode) for a given voltage between anode and cathode;

- and to minimize the electric-field enhancement at the anode-facing surface of the grid, which improves the anode-grid voltage-hold-off capability.

The price that must be paid for placing the control grid near the cathode is increased grid heating due to primary electron interception and the proximity of the hot cathode itself.

The type 6544 triode was designed especially for modulator switch-tube service, and it is not a simple triode. Not only does it have a control grid but it has a shield grid as well. In addition, it has an oxide-coated unipotential (indirectly heated) cathode of special shape. All of these features, shown in cross-sectional

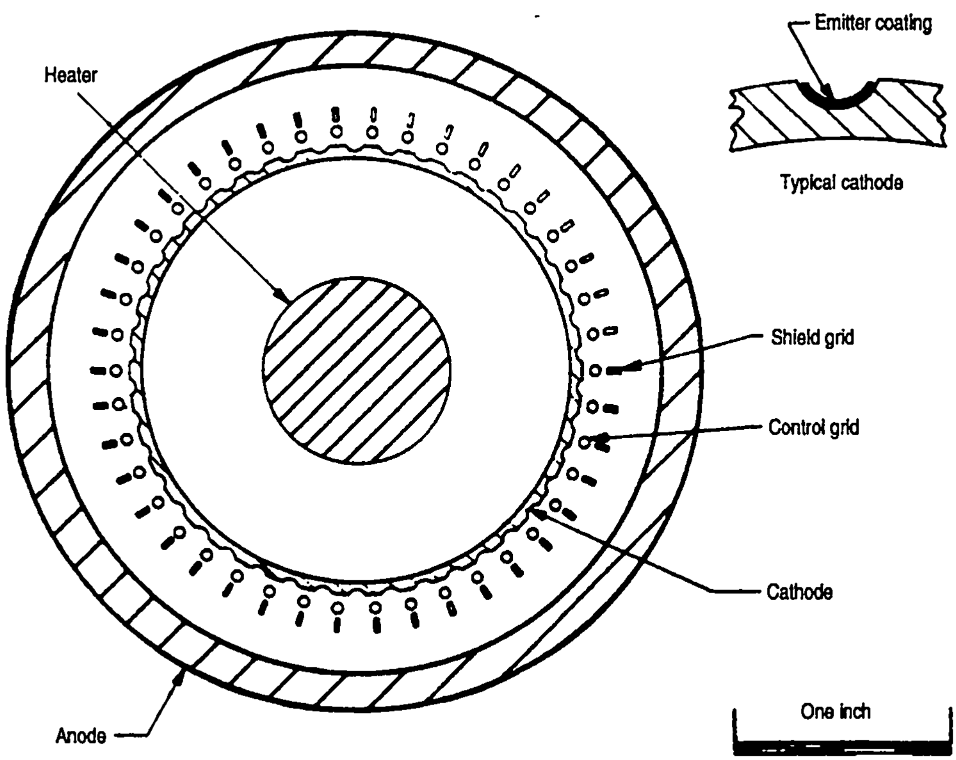


Figure 10-15. Section view through active region of type 6544 shield-grid triode.

view in Fig. 10-15, are intended to contribute to a high-μ triode ($\mu = 90$), which carries the implication that it has high grid-current interception, but without the high grid current. The cathode is of cylindrical shape, but its coated active emitting surfaces by no means cover the entire available surface. In fact, as shown, the individual emitting surfaces are groovelike concavities spaced around the periphery. Control-grid rods, held in place by ringlike headers at each end, run the length of the emitting grooves, but they are staggered in spacing with respect to grooves so that they are in between them and not in the line of fire of electrons leaving the cathode on the way to the anode. The shield grid is another

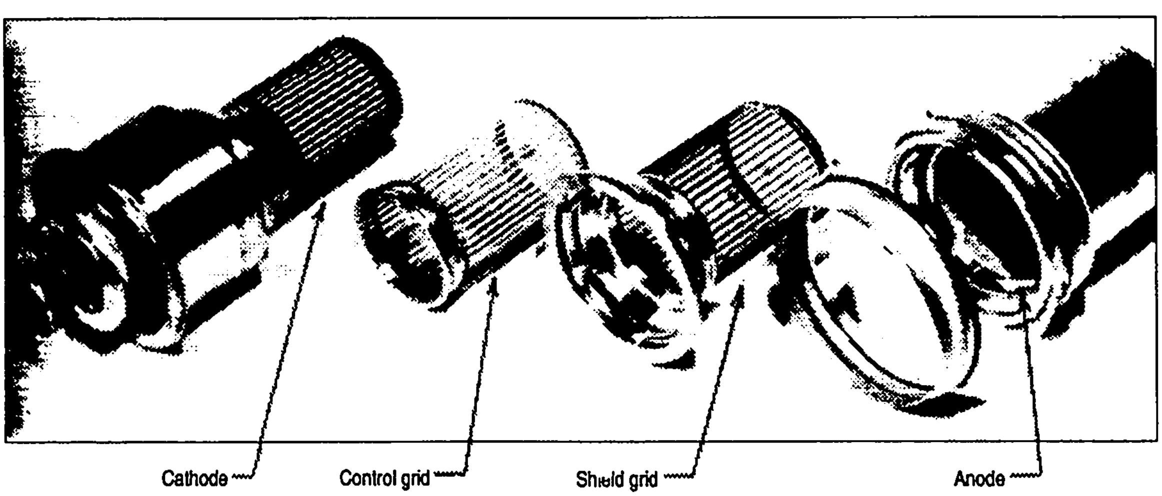


Figure 10-16. View of concentric subassemblies that make up the type 6544 shield-grid triode.

group of rods that has been placed on the same radials as the control-grid rods. The shield rods are located just on the anode side of the control grid on a circle of slightly larger diameter. The shield grid is connected to the cathode. Figure 10-16 shows the normally concentric electrode assemblies lying side by side. Instead of a uniformly emitting cathode surface, the 6544 has a number of individual electrostatically focused electron guns cylindrically arrayed. This geometry "hides" the control grid from the cathode as much as possible in a high-μ tube.

To what extent does this strategy succeed? The anode-voltage rating of the 6544 is 20 kV. Using the same 90%-efficiency criterion, at an anode voltage of 2 kV and a peak-grid voltage of 1300 V, the peak-anode current is 70 A and the grid current is 5.5 A, as shown in the current/voltage characteristics of Fig. 10-17. The ratio is almost 13, which is certainly not bad for a tube with μ of 90. Total cathode current is 75.5 A, which is approximately the peak rating of the tube. Of this, 93% is anode current and 7% is grid current. The filament heating power is 360 W, giving an emission loading of 4.8 W/A.

10.4.4 The S94000E and 4CPW1000KB tetrode switch tubes

The ML-6544 is a triode in that only three electrodes are externally accessible. There is, however, a fourth electrode, the shield grid, but it operates at the same voltage as the cathode and is internally connected to it. A true tetrode has an independently operated fourth electrode, which is physically similar to the 6544 shield-grid and is typically operated at a continuously applied (or sometimes pulsed) voltage positive with respect to the cathode, usually in the range be-

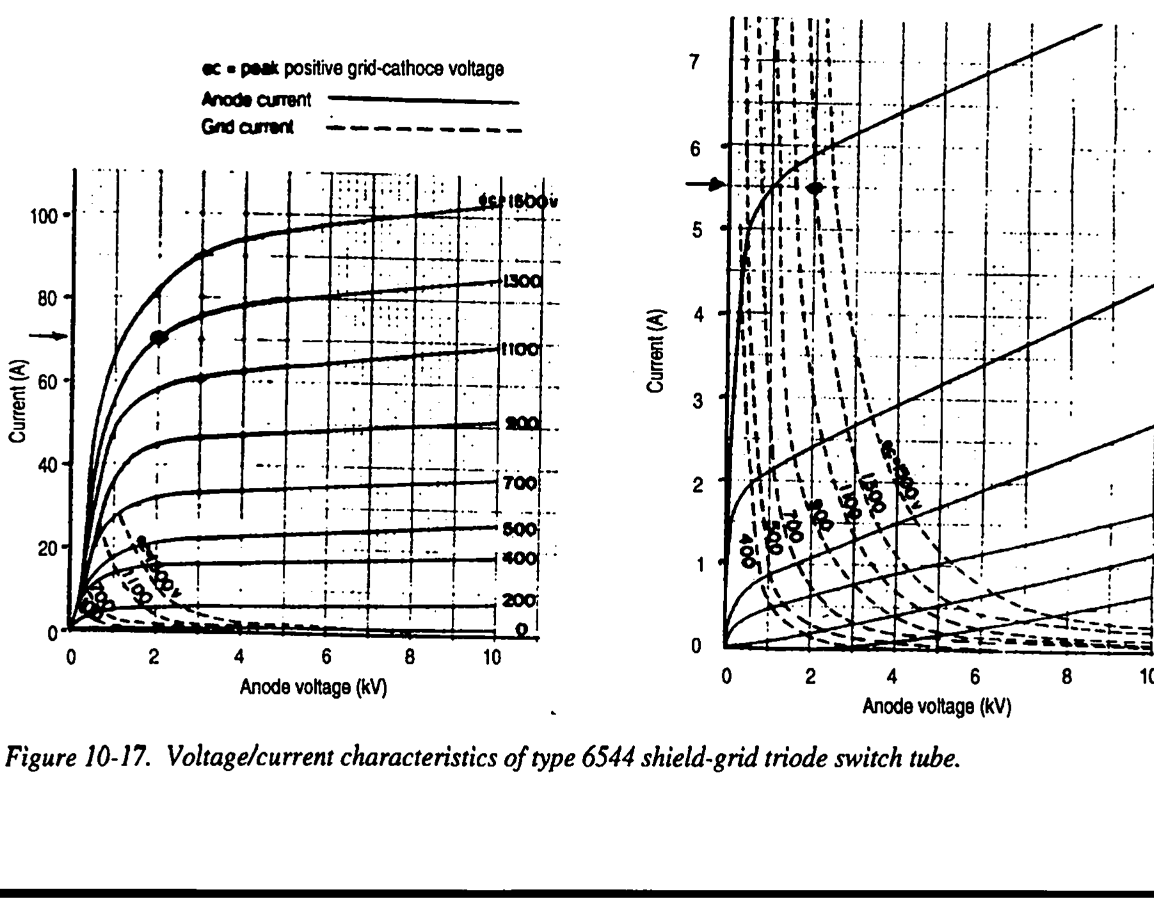


Figure 10-17. Voltage/current characteristics of type 6544 shield-grid triode switch tube.

tween 1000 and 2000 V. This electrode shields, or "screens," the grid from the anode, so it is usually referred to as the screen grid. It is also placed very much closer to the control grid than to the anode. For all intents and purposes, therefore, the screen grid is a virtual anode to the control grid, operating at constant voltage with respect to the cathode. If positive control-grid-to-cathode voltage is required for the desired peak-cathode emission (and it often is not), the amount of current intercepted by the grid for any fixed grid-cathode and screen-cathode voltages will be substantially independent of the anode-cathode voltage. Often, because of the electric-field strength at the cathode that results from screen-cathode voltage, full emission can be obtained with zero or even negative grid-cathode voltage, in which case there will be no grid power dissipation at all due to primary electron interception. However, the grid (and the screen as well) will be heated by radiation from the hot, nearby cathode.

The amplification factor in a tetrode is defined with respect to the screen grid rather than the anode, as is the case with the triode. Tetrode amplification factors range typically from 5 to 10. Projected cathode-current cutoff occurs when a negative control-grid voltage, which is affected only slightly by the anode-cathode voltage, is applied that is between one-tenth to one-fifth of the positive

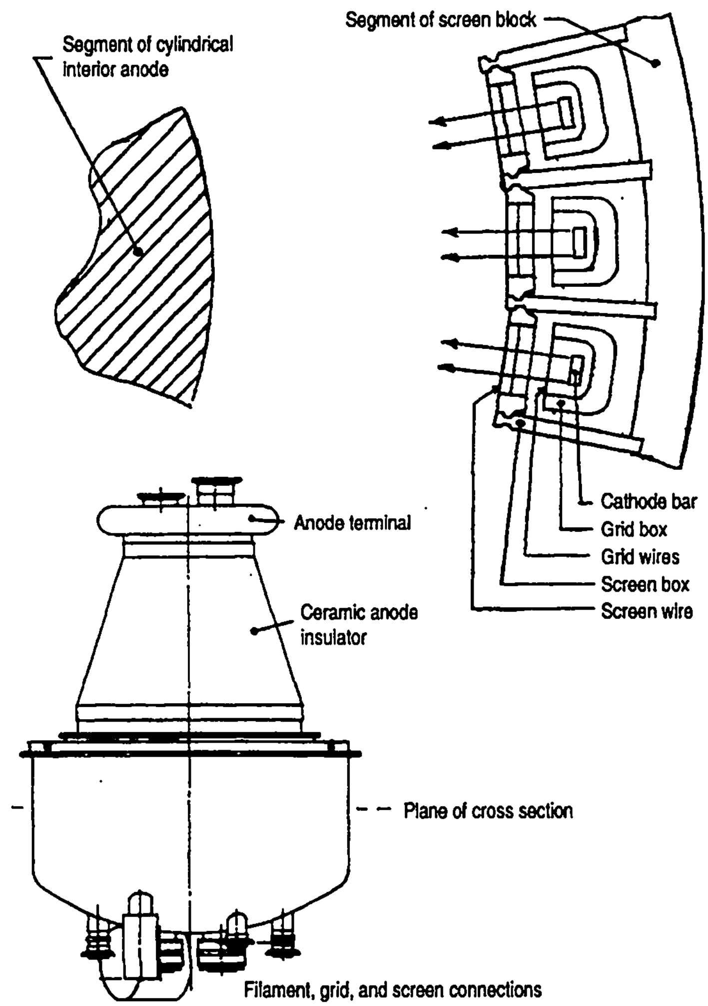


Figure 10-18. Physical aspects of S94000E modulator switch tube.

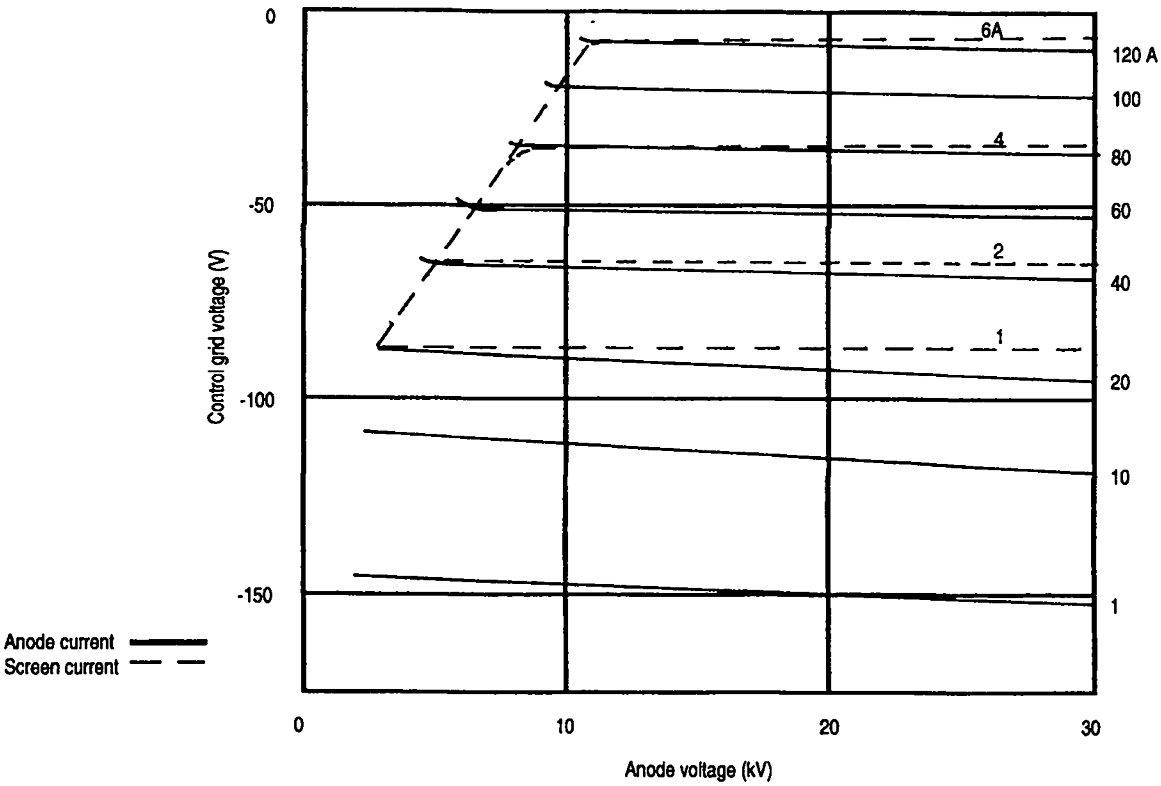


Figure 10-19. Voltage/current relationships in S94000E tetrode switch tube for screen voltage of 1000 V.

screen-cathode voltage.

Many tetrode types (and triode types as well) are used as high-power hard-tube modulator switches, although they were not designed for such service. These tetrodes range in capacity from the 100-kW anode dissipation of the 4CW100,000 right up to megawatt-dissipation tubes. They were originally intended for use in high-power broadcast transmitters or other high-power RF sources at frequencies through the HF band (but usually not above 50 MHz). In RF amplifier service, a dc anode voltage rating in the 20-kV range is common. This may seem inadequate for most hard-tube modulator service, and it probably is. The maximum instantaneous anode voltage that an RF amplifier experiences is nearly twice the dc component, however. Moreover, a tube that is rated for high-level amplitude modulation at 100% positive modulation will see a "dc" or audio-frequency peak that is also twice the average anode voltage. (Remember that the peak instantaneous power from a 100%-amplitude-modulated power amplifier is four times the carrier power, and even a 50-kW-rated transmitter produces 200-kW peak envelope power.) It can be seen that in modulator-switch service, where maximum voltage hold-off occurs during nominally zero-current situations, the true voltage-hold-off capability may be many times the RF-amplifier dc anode-voltage rating. This is especially true if the tube is specified at procurement for such service. In such cases, it can be specially processed during manufacture to enhance high-voltage tolerance.

Two good examples of tetrodes designed expressly for service as high-voltage modulator switch tubes and intrapulse voltage regulators—and not just specially processed RF amplifier tubes—are the S94000E and the 4CPW1000KB. The

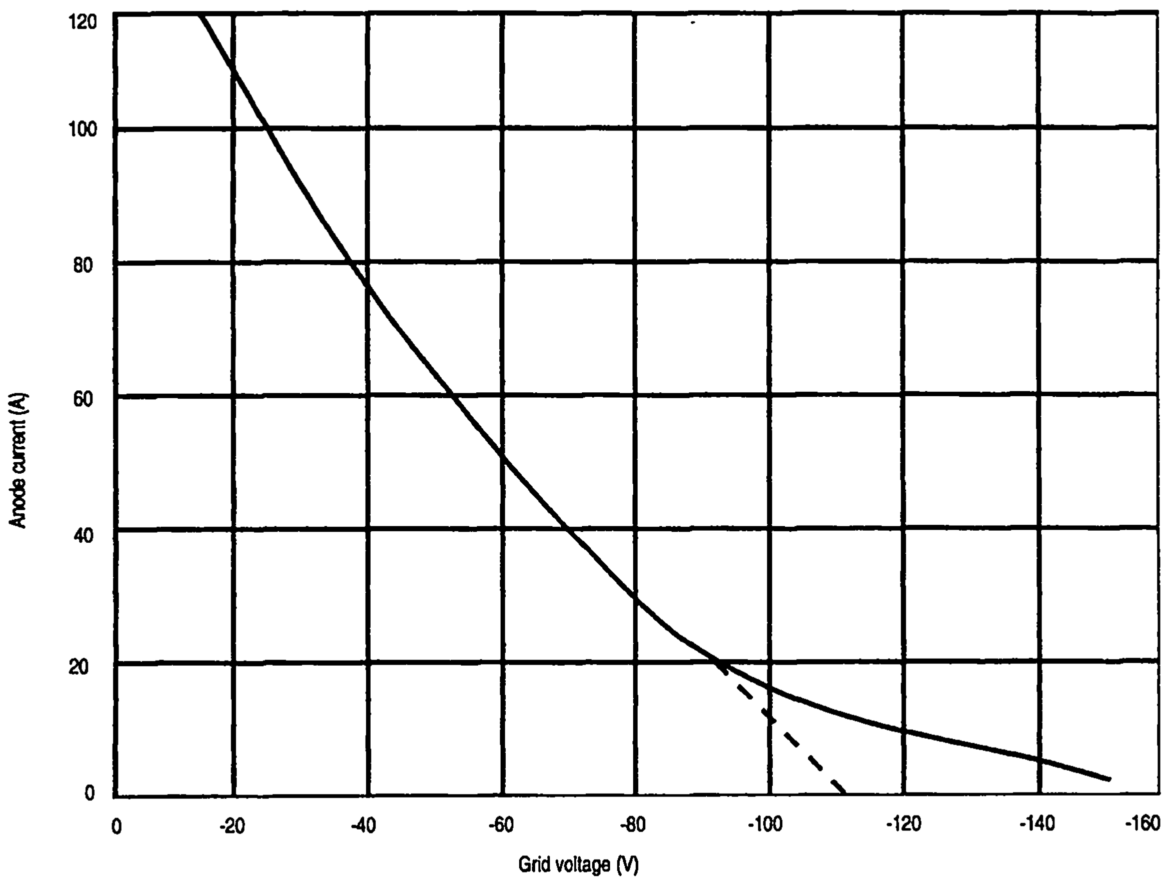


Figure 10-20. Transfer characteristic for S94000E tetrode showing "projected" cutoff.

S94000E has an anode-hold-off voltage rating of 200 kV and a maximum peak-anode current of 125 A. (We will later see that peak-anode current is virtually the same as the peak-cathode current.) The 4CPW1000KB has an anode-voltage rating of 175 kV and a peak-anode-current rating of 75 A. Both tube types evolved from high-power RF amplifier forebears. Like most tubes intended for high-voltage hold-off capability, they differ from their predecessors in the spacing between anode and screen grid. Consequently, the slope of their diode, or virtual-cathode, lines also differ.

The 4CPW1000KB is a radial-beam, external-anode tetrode of more-or-less conventional geometry. It descended from the 4CW250000 and was intended primarily as a megawatt-class RF amplifier in the HF band. Its anode is capable of 1 MW of average dissipation—which is what the "1000 K" in its type designation refers to—giving it extended capability for linear voltage regulation and waveform shaping.

The S94000E, by comparison, is literally inside out, as shown in Fig. 10-18. It is a radial-beam tube, but its cylindrically shaped anode is on the inside of the tube rather than the outside. Around the outer periphery is a circular array of "unitized" electron guns, which are shown conceptually in the figure. Each gun has its own central cathode bar, grid box, and screen box. The grid and screen boxes are solid on three sides. The sides facing the anode are open, except for the very fine wires stretched across them that are the actual control and screen grids.

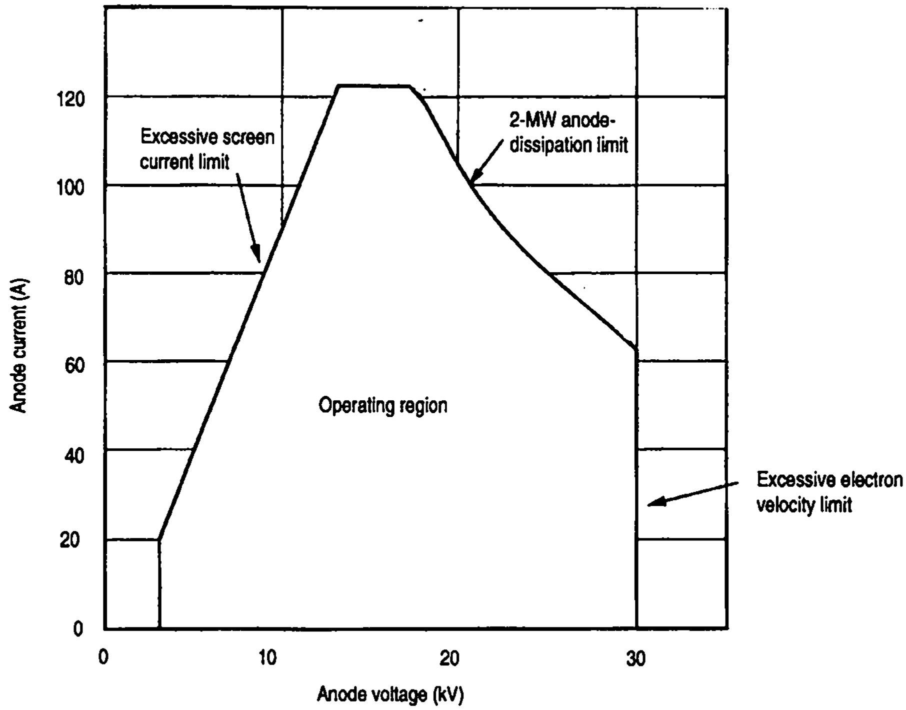


Figure 10-21. Area of permissible operation for S94000E tetrode switch tube.

This electron-optics technique first emerged in a grid-driven power tetrode for UHF television transmitter service and resulted in a many-fold increase in the power-frequency product obtainable from power gridded tubes. The voltage/current characteristics of the tube for a constant screen voltage of 1000 V are shown in Fig. 10-19. Note that these curves have a different format than the ones previously shown. These are constant-current characteristics rather than constant-grid-voltage characteristics. They illustrate the almost-total screening effect of the anode from the cathode. This is due to the fact that anode current is all but independent of anode voltage. The effective perveance, or transconductance, of the electron optics is such that the full current is obtained when the grid voltage is never positive with respect to the cathode, and the total "grid base" (the grid-voltage swing from cutoff to full current) is only slightly more than 150 V. This condition requires virtually zero driver power. With the grid never positive, there is no grid current either. The grid that competes with the anode for electrons is the screen. Note that even with an anode voltage that is only 5% of the hold-off rating (10 kV compared to 200 kV), the current-split ratio between screen and anode is 120 A/6 A, or 20, which is superior to the conventional triodes previously discussed.

Figure 10-20 shows how anode current varies with grid voltage for a constant screen voltage of 1000 V and constant anode voltage of 12 kV. This figure illustrates the concept of "projected" cutoff, which is the dashed line. It "projects" a grid voltage for anode-current cutoff of about -110 V. For a screen voltage of 1000 V, this relates to a screen-grid amplification factor, or μ_s, of 1000 V/110 V, or 9, which is the nominal screen μ. Figure 10-21 shows the dimensions of the tube's permissible operating region, which is the equivalent to the safe operating area (SOA) for a field-effect transistor. The left-hand, sloping limit line is the diode

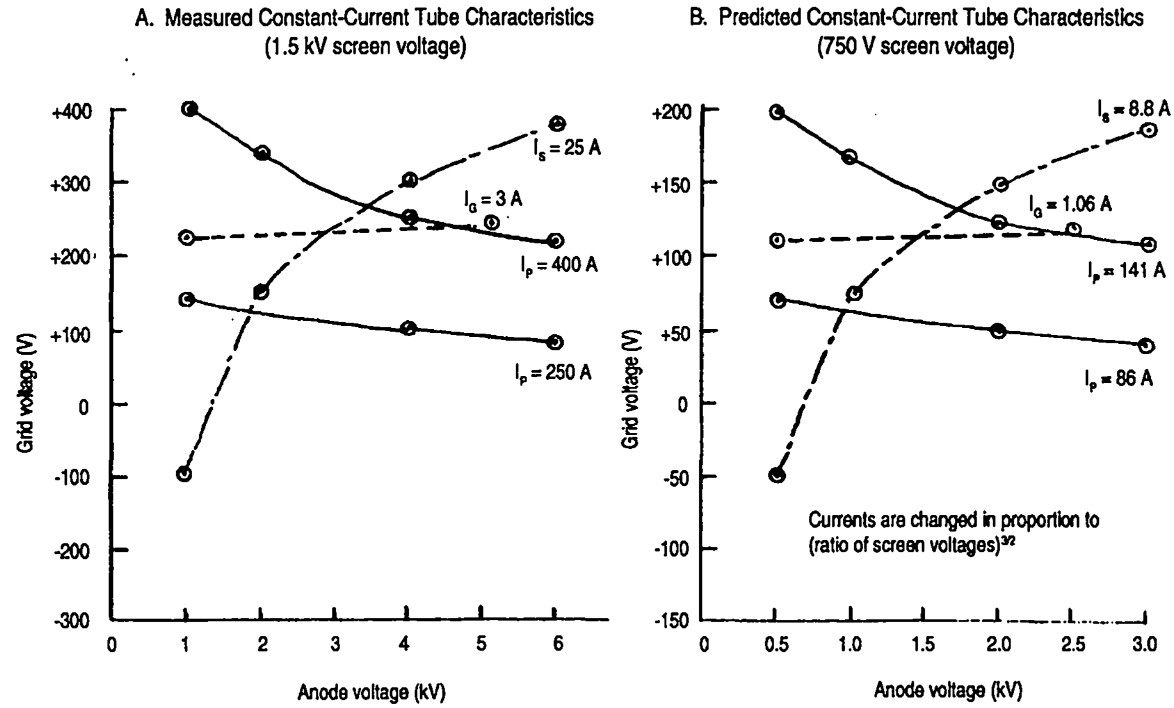


Figure 10-22. Translating tetrode voltage/current characteristics for different screen voltages.

line. To the left of this line screen current dominates anode current, and screen dissipation becomes excessive. The top segment is the maximum anode current limit. The curved-line limit is the actual anode dissipation, beyond which the product of anode voltage and current exceeds 2 MW. The right-hand line, which rises from the 30-kV anode voltage, is a limit on electron energy, or velocity, beyond which electrons will penetrate the anode and cause damage.

The voltage/current characteristics of tetrodes, in general, are more cumbersome to illustrate because there are four degrees of freedom compared to only three in the triode. This means that there is a unique set of data interrelating anode and grid voltages with anode, grid, and screen currents for each and every value of screen voltage. Manufacturer's published data will rarely show more than two such screen voltages. Even if no more than one is given, however, it is possible to synthesize the appropriate data points needed for any other value of screen voltage. How this is done is shown in Fig. 10-22. The manufacturer's information given in Fig. 10-22a and is based on constant-current characteristics for 1500-V screen voltage. But what is wanted is the same data for a screen voltage of 750 V. To derive this information, a new set of axes along with new grid voltage and anode voltage must be generated. In this case, the ratio of 750 V to 1500 V is 1/2. The original current values for grid, screen, and anode are all multiplied by the transfer ratio 1/2 raised to the 3/2 power. In the example shown, the original 250-A anode-current line becomes in Fig. 10-22b 250 A x $(1/2)^{3/2}$, which is 86 A, and the original 400-A line becomes the 141-A line by the same process. Meanwhile the 25-A screen-current line becomes the 8.8-A line and the 3-A grid-current line becomes a 1.06-A line. An entire characteristic can be translated, but plotting only a few points at the full-conduction and cutoff ends usually suffices because we are using these devices primarily as switches.

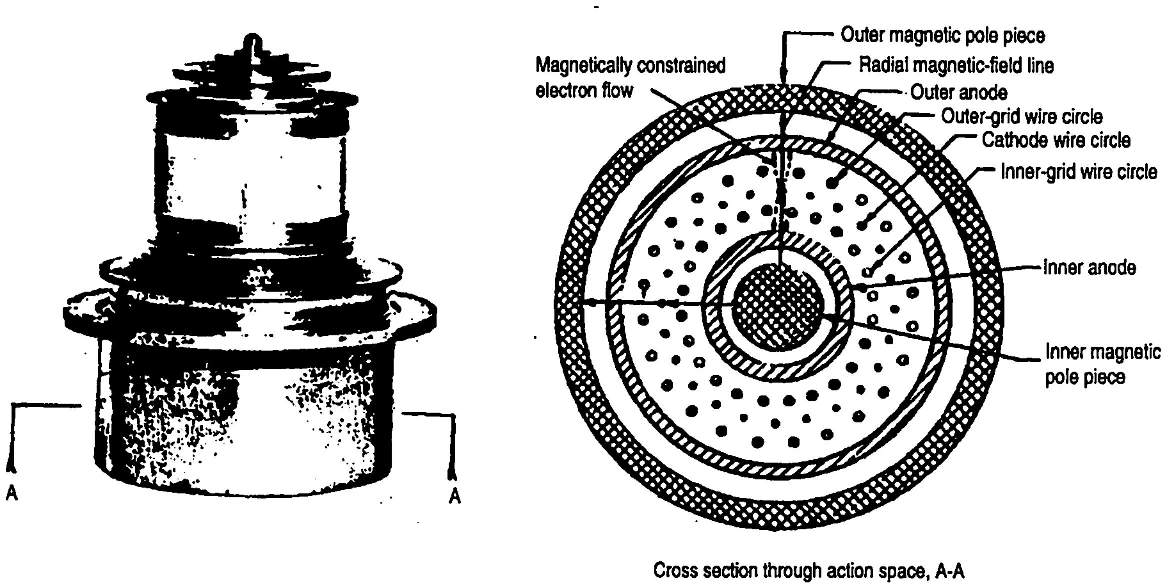


Figure 10-23. Construction of the type ML-8549 magnetically beamed hard-tube switch.

These synthesized points will be as useful to the designer as the original published data, which are offered with no guarantee that they characterize any given tube of the type. What, then, guarantees the performance for any tube type, triode or tetrode? The guarantee lies within the specific quality-assurance test specifications for a given tube type. These specifications are in one of two formats: EIA or MIL-E-1. These specifications typically establish ranges of voltages and/or currents within which a particular tube must fall in order to be sold under its part number. For instance, the negative grid voltage required for the plate current to be less than some specified value at a given anode voltage (or a combination of anode and screen voltages) will not be expressed as a single value but as a range of voltages. The transmitter design had better be able to accommodate the entire range or the worst case. In other words, design to the tube specification not the tube data sheets.

10.4.5 The ML-8549 magnetically beamed triode

The tubes discussed so far have used only geometry and electrostatics to affect electron trajectories. The ML-8549 was one of the first attempts to use a combination of geometric, electrostatic, and electromagnetic influences on electron trajectories in order to maximize the number of them that make it to the anode. The type 8549 triode is illustrated in Fig. 10-23. Note the cross-section view of its inter-electrode action space. It is a radial-beam device. Its elements are cylindrical and they are coaxial. The inner- and outermost elements, however, are not electrodes at all. They are magnetic pole pieces whose orientation is similar to those used in dynamic loudspeakers, in that they support radial magnetic field lines. The space where the voice coil of a loudspeaker would fit is occupied by the electrodes.

But this is no simple triode. There is a wire-type directly heated, thoriated-tungsten cathode. But instead of being the innermost electrode, it is the centermost. Inside and outside of it are two sets of wire control grids that are circumferentially

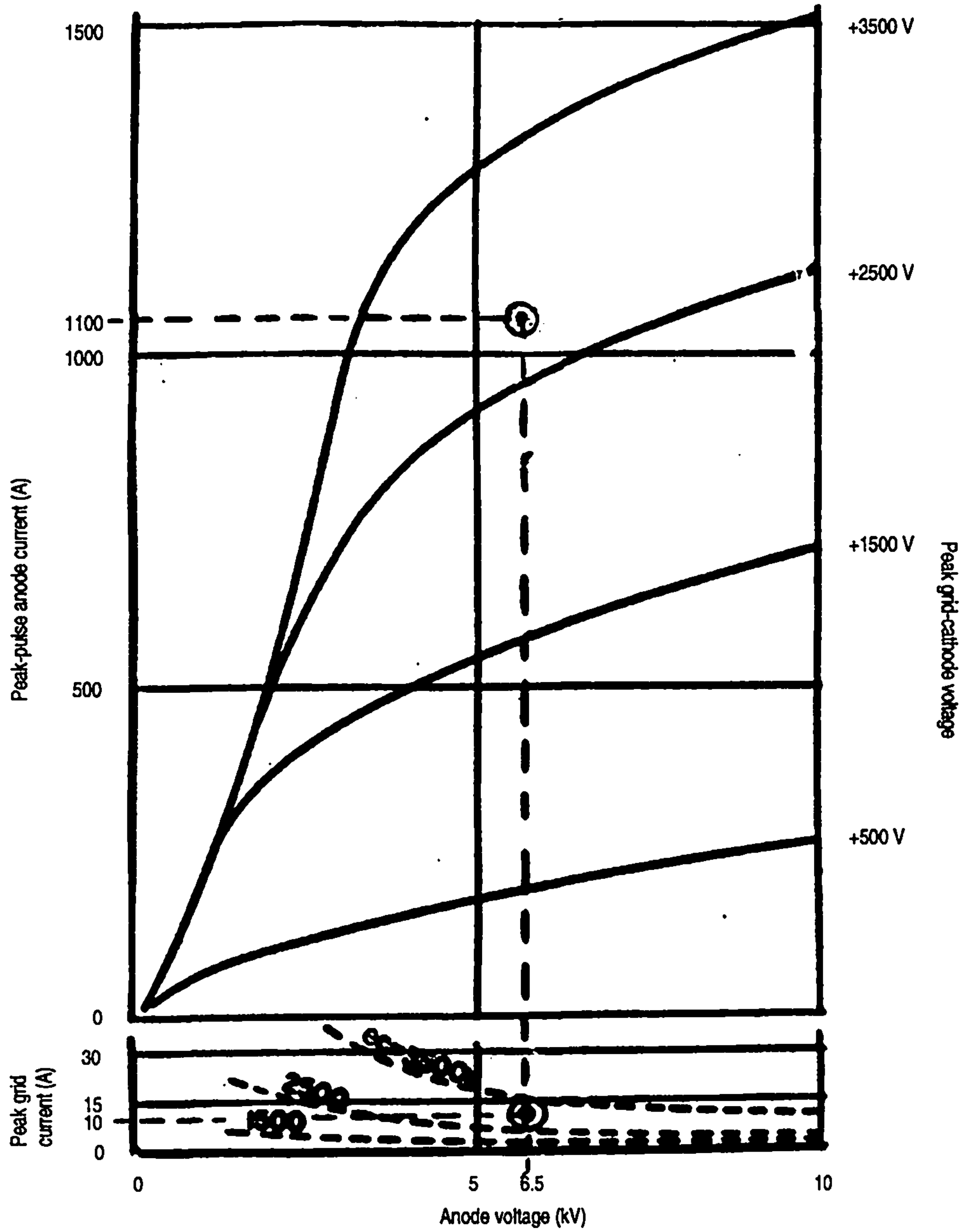


Figure 10-24. Voltage/current relationships in ML-8549 triode.

staggered with respect to the cathode wires so that they are not on the same radial lines as the cathode wires. Inside of the innermost grid wires and outside of the outermost grid wires are an inner and an outer anode. There are, in effect, two triodes sharing the same cathode. Electrons leaving the outer face of the cathode travel outward, and those leaving the inner face travel inward. Moreover, even though they travel past grid wires that are positive with respect to the cathode, and therefore attract them, their paths are constrained by the radial magnetic field to avoid the grid wires and continue on toward the anodes.

How well did all of this work? Figure 10-24 shows the voltage/current characteristics of the 8549. The tube is rated for 65 kV voltage hold-off. Again using the 10%-conduction-voltage drop criterion, or 6.5 kV, we find that the same posi-

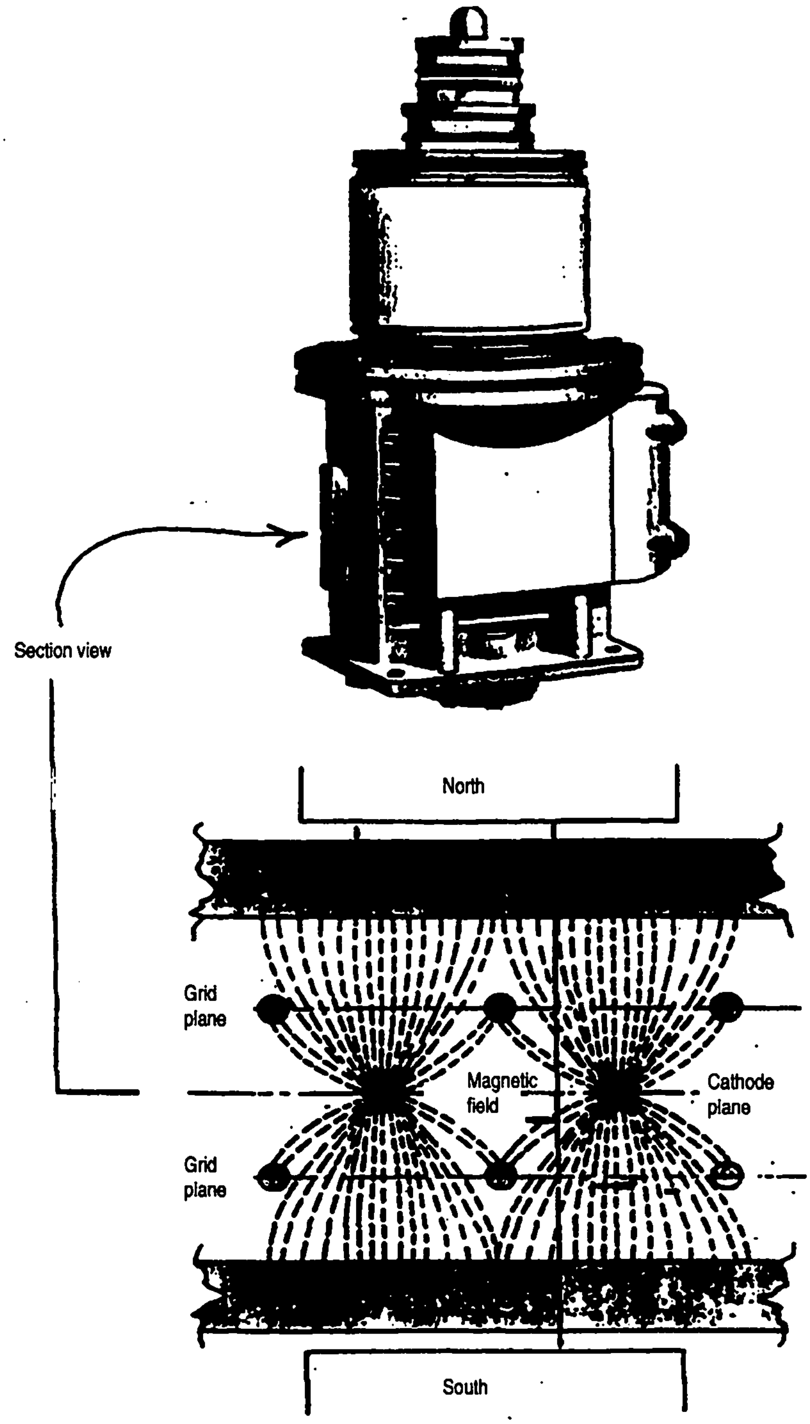


Figure 10-25. Type ML-8618 magnetically beamed triode switch tube.

tive grid voltage that produces 1100-A peak-plate current before (about +2700 V), now produces only 10-A peak-grid current. The ratio is 110:1. More than 99% of the cathode current makes it to the anodes. However, the negative grid voltage required for plate-current cutoff at 65 kV anode voltage is 4000 V, meaning that the amplification factor is only slightly more than 15. The total grid swing, therefore, is almost 7000 V. Even so, a peak-pulse output power of almost 65 MW is obtained from a drive power of approximately 100 kW.

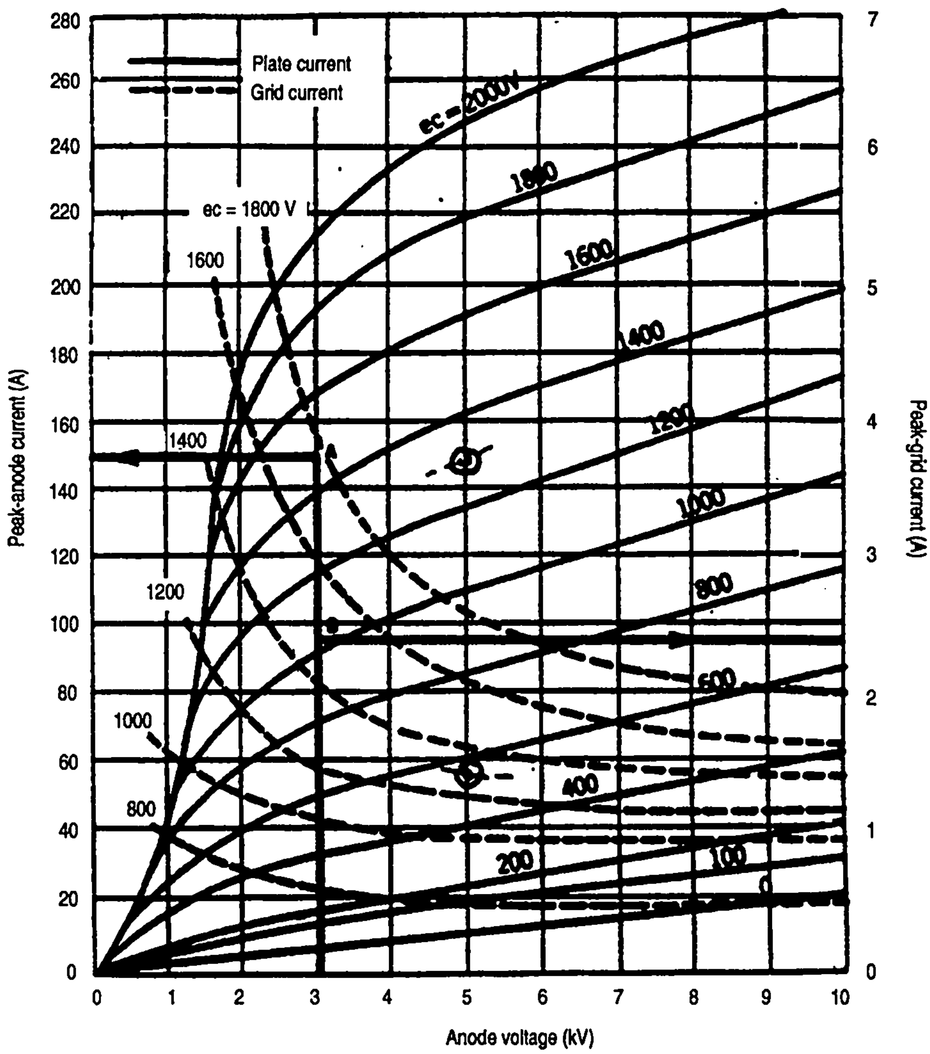


Figure 10-26. Voltage/current relationships of ML-8618 triode.

The tube was not electrically small, either. It had filament power of over 14 kW (7.5 V at 1880 A), which is about 12 W/A of pulse-cathode emission, and an anode-average-power-dissipation rating of 500 kW. Another indication of the reduced grid heating is the maximum pulse duration rating of 0.1 second, or 100,000 μs. (Remember that grid heating, not cathode emission, is the pulse-duration limiting factor in a thoriated-tungsten filament switch tube.)

This tube is spoken of in the past tense because the application for which it was primarily intended—a giant, experimental pulse modulator—was by no means an unqualified success, and the tube type itself suffered by association. The magnetically beamed concept, however, did not die with it.

10.4.6 The ML- 8618 magnetically beamed triode

This tube, as shown in Fig. 10-25, is a very close relative of the type 8549. The electromagnetic strategy is the same, but the geometry is planar instead of coaxially cylindrical. The effect on performance is quite similar, as shown in the voltage/current relationships of Fig. 10-26. The tube has been rated at as much as 50 kV anode-voltage hold-off—although it rarely operated above 40 kV. Using the 50-

kV rating and a corresponding conduction-anode-voltage drop of 5 kV, we see that a typical 150-A value of peak-pulse anode current is obtained with a grid voltage that is 1300-V positive with respect to the cathode along with a peak grid current of about 1.4 A. (The maximum peak-cathode-current rating is 200 A.) The current ratio is slightly more than 100, and over 99% of the electrons emitted by the cathode make it to the anodes.

The grid interception is low—so low that there is no real reason to be limited by the 10%-anode-voltage criterion. Indeed, at a conduction anode voltage of 3 kV, we see that the same 150-A peak-anode current is obtained with a grid voltage 1500 V positive and a grid current of 2.4 A, which is still a ratio greater than 60.

Typical of a thoriated-tungsten-cathode tube with low grid-power dissipation, the 8618's pulse duration limit is 10,000 μs at a duty factor of 6%. The anode, when properly water-cooled, will dissipate up to 80 kW. Unlike the larger 8549, hundreds of 8618s were built and used and they are capable of being

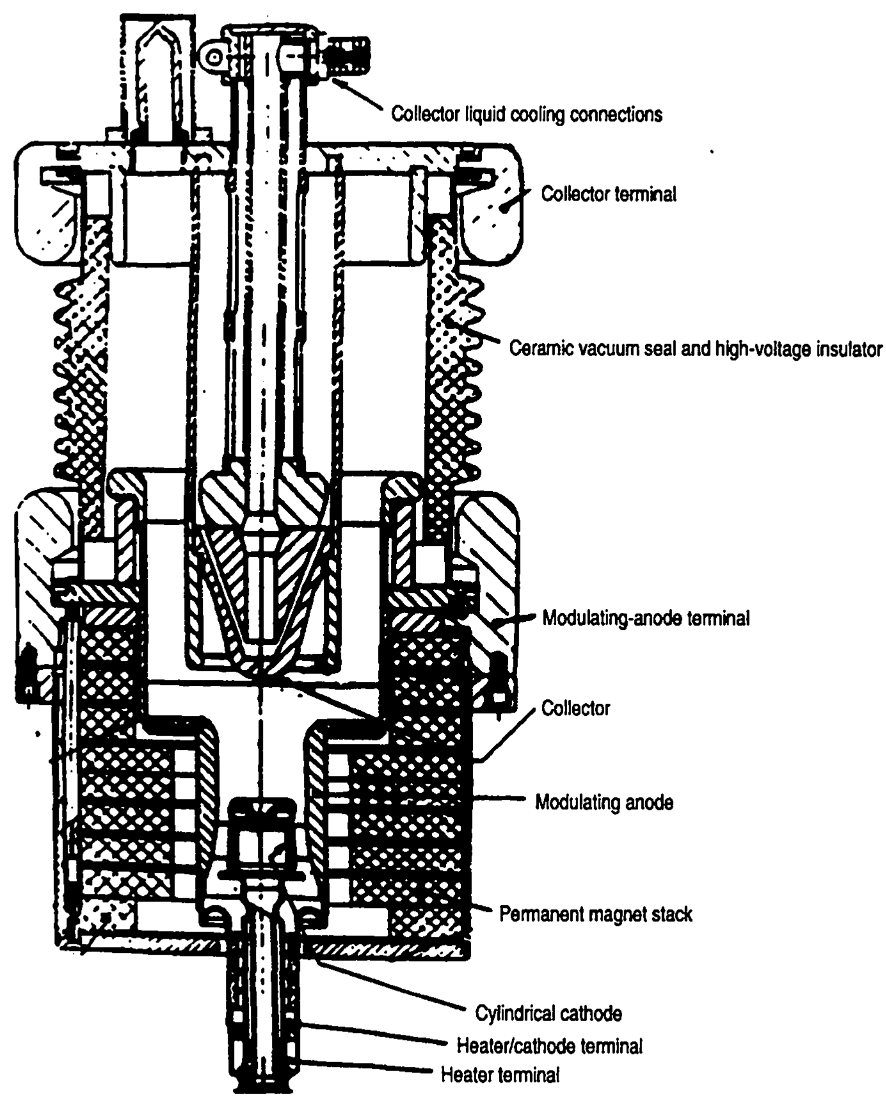


Figure 10-27. Cross-section view of L-5012 Injectron beam-switch tube.

rebuilt.

10.4.7 The L-5012 and L-5097 Injectron™ beam-switch tubes

We have seen how effective collinear electric and magnetic fields can be in minimizing grid current in the examples of the previous tubes. The L-5012 Injectron beam-switch tube (BST), the smaller of the two tubes of this type, was the first vacuum switch tube to make a complete break from the conventional wire-type grid structure. Instead, it used a completely solid control electrode. How can electrons get through a solid metal electrode? They can't, of course. But by using crossed rather than collinear magnetic and electric fields, they don't have to.

Figure 10-27 shows a cross-sectional view of the L-5012. Like the ML-6544, this switch tube uses an indirectly heated oxide-coated cathode that is cylindrical, which means that electrons leave it radially. The control electrode appears to the cathode as an external cylindrical anode, which it is. For this reason the control electrode is called a modulating anode rather than a control grid. When the voltage on the modulating anode is positive with respect to the cathode, electrons leave the cathode radially and are accelerated toward the modulating anode. Surrounding the modulating anode, however, is a stack of permanent magnets that are magnetized so as to produce a magnetic field along the axis of symmetry of the tube, which is at a right angle to the electric field between the modulating anode and cathode. This configuration is often referred to as a crossed-field, as in crossed-field amplifiers or oscillators, of which the magnetron is the most famous example.

The axially directed magnetic field acts to bend the electron trajectories to follow the magnetic-field lines (as in a magnetron), which is why this tube is sometimes called a magnetron-injection electron gun. Once headed in the right direction, the electrons fall under the accelerating influence of the electric field produced by the voltage on the beam collector, which is yet another anode, but one that is insulated from both the modulating anode and the cathode. The collector serves the role of the anode in a conventional triode or tetrode. The electric field produced by collector voltage is in the same direction as the magnetic field. The electron beam, therefore, is a cylindrical "hollow" beam. The electric fields produced by voltages on the modulating anode and the collector are at right angles to each other: the field produced by the modulating anode is perpendicular to the cathode surface, and that produced by the collector is parallel to the cathode surface. A designer would expect, therefore, that the collector voltage would have almost no effect on cathode current, or that the cut-off amplification factor, or μ, between modulating anode and collector would be infinite. A designer would also expect that above the diode line, the resistance of the incremental collector, or anode, would be nearly infinite as well. (This resistance is the amount of collector-voltage change for a given change in collector current, other things remaining constant.)

Figure 10-28 shows the voltage/current relationships of the L-5012 beam-switch tube. If one ignores the voltage scales, the shape of the characteristics is reminiscent of nothing more than the characteristics of a field-effect transistor. Indeed, the L-5012 is a true field-effect device. With no modulating-anode voltage, there

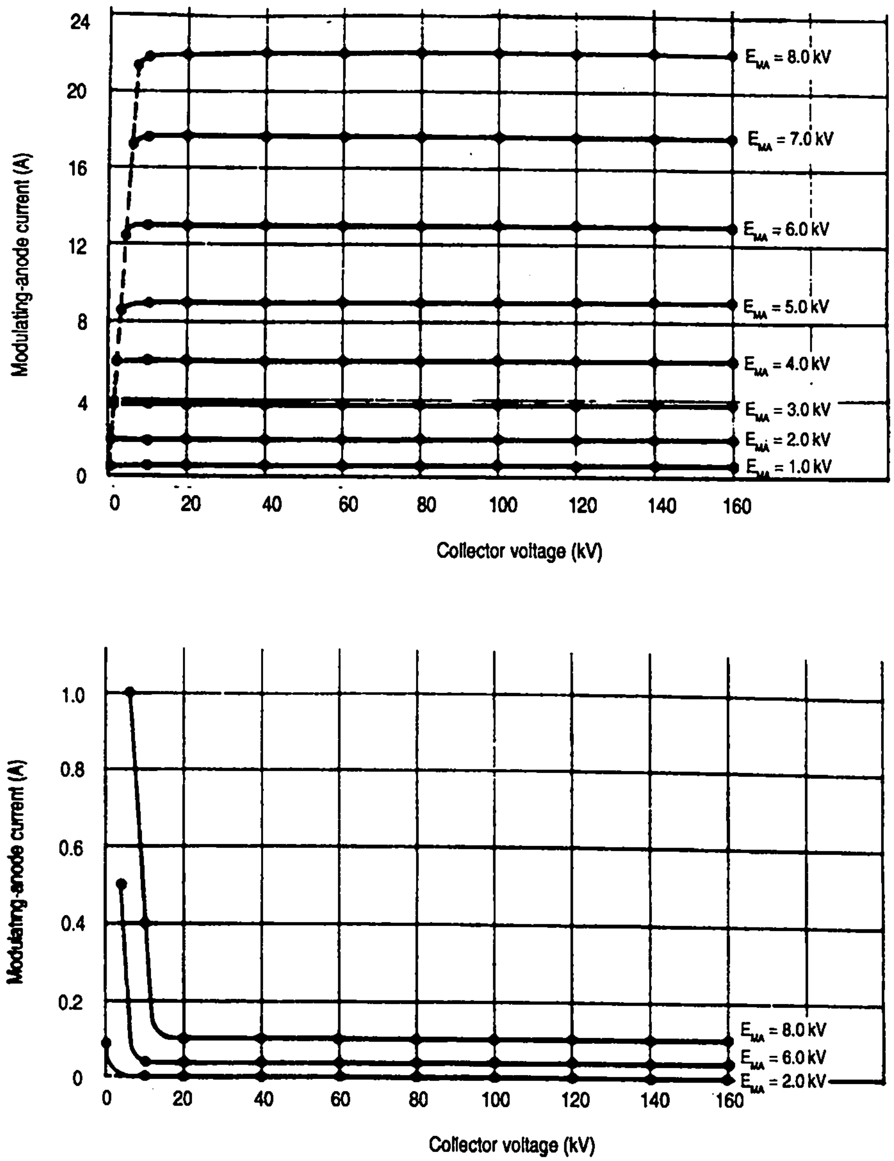


Figure 10-28. Voltage/current relationships of L-5012 beam-switch tube.

is virtually no collector current, regardless of collector voltage. The implies that μ is all but infinite. There is, however, competition for cathode-emitted electrons between the collector and modulating anode when their voltages approach one another, an example of the virtual-cathode effect. If we apply the 10% voltage criterion, we see that for collector current of 22 A and collector voltage of 15 kV (10% of the 150 kV hold-off rating), the modulating-anode voltage is 8 kV and its current is 0.1 A. The current ratio is 220—the highest so far. Above this "knee," the plotted line of the slope of the collector-current line corresponding to a given modulating-anode voltage is virtually horizontal, denoting nearly infinite collector (or anode) incremental resistance. The practical significance of this will be discussed later.

The L-5097 is very similar to the L-5012 in design and construction, but it is bigger and more powerful. It is rated at 180 kV maximum collector voltage and 70 A collector current, with 60-kW collector dissipation (compared with 150 kV, 30 A, and 12 kW for the L-5012). When operated at its full capability, or in parallel with other tubes in close physical proximity, the use of a solenoid electromagnet rather than a permanent-magnet stack has been found to give superior performance.

The type of construction used in both tube types has another advantage over conventional wire-grid construction that can be crucial. It has to do with what is likely to happen in the event of an internal arc. The cathode surface, for one thing, is not in the line-of-fire of a collector arc (or of back-streaming positive ions, either). The concave circular button on top of the cathode structure is the target for both. There are no fine wires to be fused by the i^2dt integral of the arc discharge. It is this susceptibility in conventional grid-control tubes that engendered the design of the electronic crowbar in the first place. (Crowbars themselves will be discussed in some detail later.) Ultimately, as the current-handling capability of a particular tube increases, it will be the amount of arc current (or the discharge action, $\int i^2dt$) that it can tolerate that will determine its ultimate internal impedance. We would like to be as close to zero as possible, as has been mentioned many times. Any linear resistance that must be added in order to limit arc or fault current is directly in series with the resistance represented by the tube conduction-voltage drop divided by the output current.

It would not be an unfair statement to say that the Injectron type of beam-switch tube represents the state-of-the-art in hard-tube modulator switches. It is sad to note that it is by no means a recent development. The first such tubes were used before 1960 (and the breakthrough that gave us the unit electron guns of the S94000E switch tube took place in the 1950s). There are, however, more recent design concepts that make use of the far more advanced technology of high-power microwave tubes. These devices promise the same type of performance as the Injectron without the use of magnetic fields. The devices, however, have yet to be built.

10.4.8 The 8454H Crossatron™ switch

Although this device is by no means a hard tube, it is not a thyratron either—although it is not coincidental that their names share the last five letters. The first five letters refer to the crossing of magnetic and electric fields, as with the Injectron. This device could be characterized as the "missing link" between the hard tube and the soft tube. Like the thyratron, it is filled with low-pressure hydrogen (0.05 torr) so it is a long way from being "hard" (10^{-8} torr). But it is not truly "soft" like a thyratron, whose hydrogen pressure is usually 0.5 torr. Its conduction characteristics are also intermediate. Whereas a hard tube may have conduction-voltage drop of approximately 10% of voltage hold-off (or 3 kV to 15 kV) and a thyratron may have 1% voltage drop (or 0.2 kV to 2 kV), the Crossatron voltage drop is more likely to be 2.5% to 3%.

The hard tube has proportional and reciprocal full control of output current (proportionality is retained whether current is being increased or decreased). The thyratron has half-control only. (Actually less than half-control, if you consider the fact that, once turned on, a thyratron has no ability to control current amplitude.) The 8454H Crossatron, on the other hand, can be gated into conduction and then gated off. A positive pulse applied to its control grid will turn it

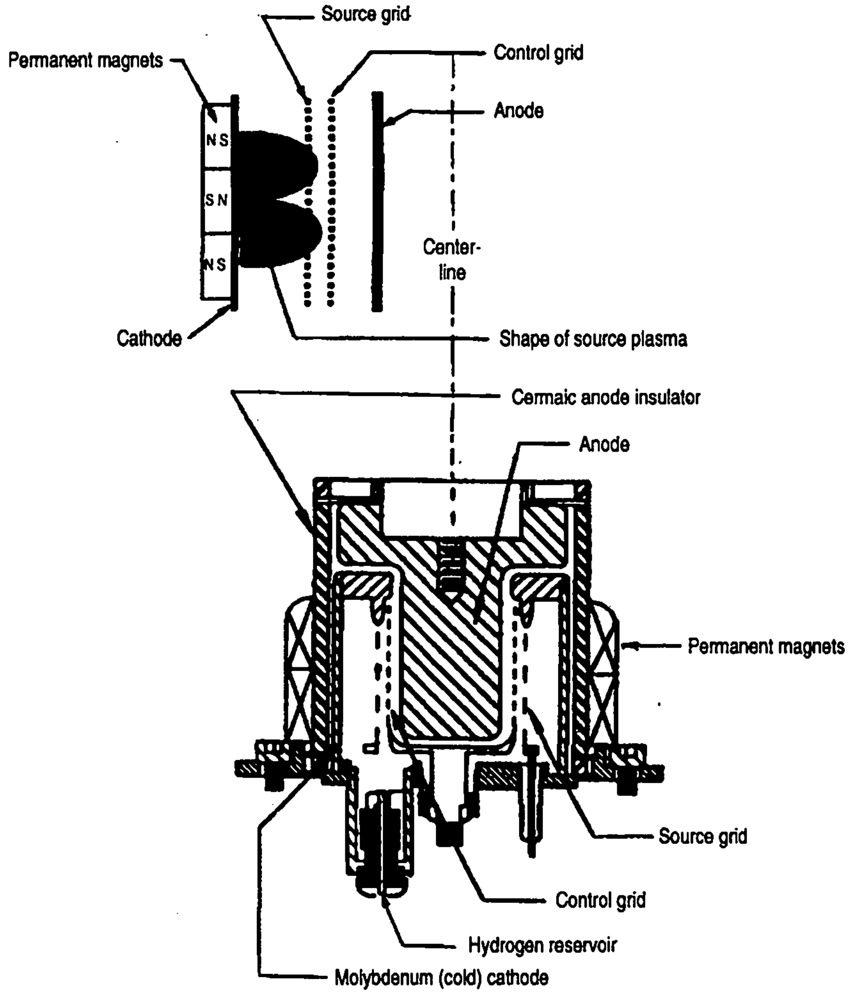


Figure 10-29. Cross-section view of Crossatron on/off control switch.

on. A negative pulse applied to the same grid will turn it off, much in the same fashion as a gate-turn-off (GTO) thyristor. The amount of current that can be gated off is unlikely to be as much as that which can be gated on, unless the grid construction of the tube is deliberately altered to optimize turn-off instead of turn-on.

The ability of the device to emulate full-control—or at least 2/3 control, if one considers intrapulse-current-amplitude control as being worth 1/3—derives from the action of the crossed magnetic and electric fields. Figure 10-29 shows the cross section of a typical Crossatron. Its design is coaxial-cylindrical. It has an internal anode, its current flow is radial, and it has a cold molybdenum cathode and two grids. A continuous voltage is applied to the "source" grid, which stimulates a continuous "source" plasma. The magnets, which surround the tube, are radially magnetized. They produce alternating north and south poles that progress vertically up the tube and a magnetic field that matches the shape of the source plasma. A positive-going trigger pulse applied to the control grid attracts plasma into the region between control grid and anode and initiates full conduction between cathode and anode. A negative-going pulse applied to the control grid will reverse the process.

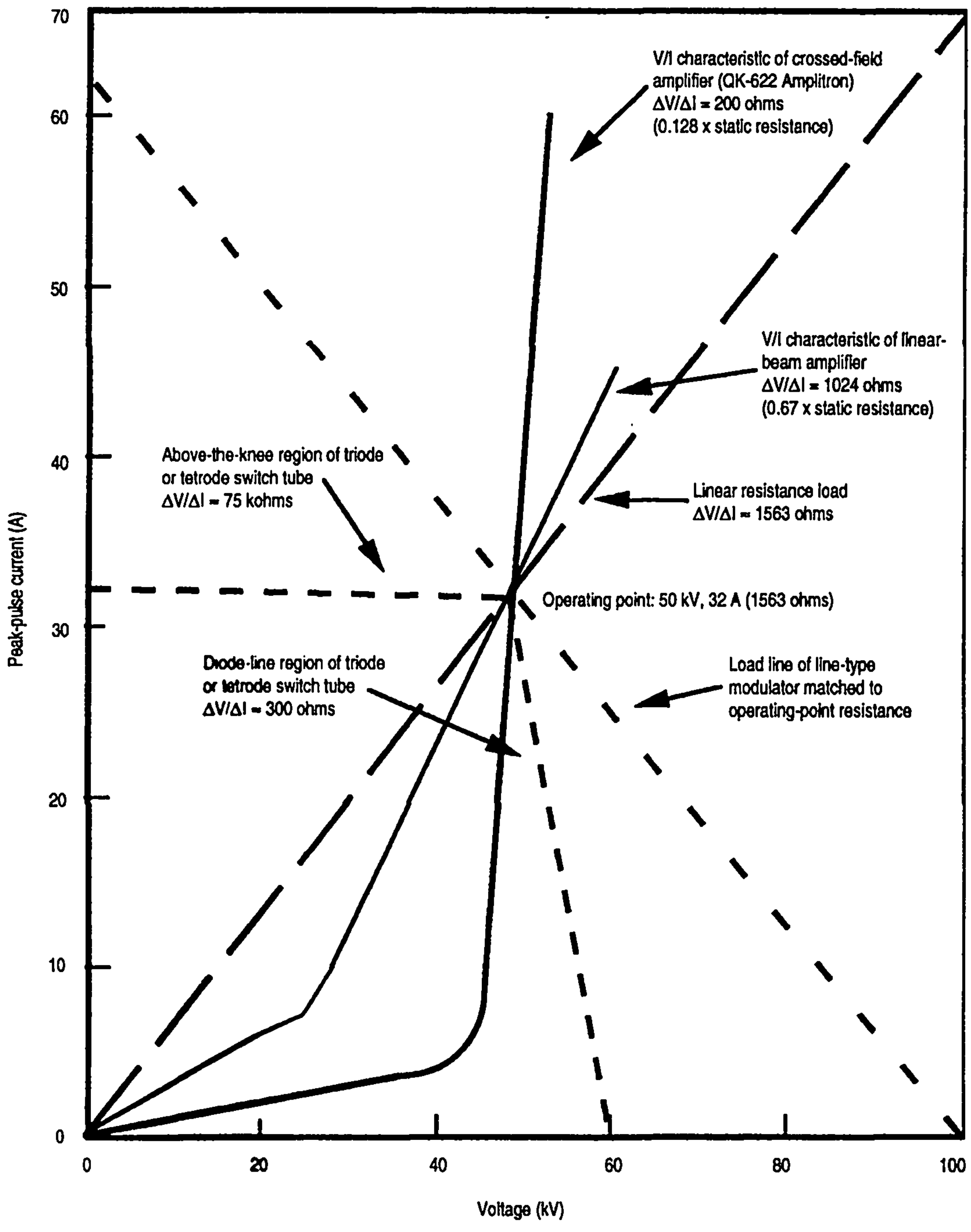


Figure 10-30. Load-line representation of modulator/load interaction.

If one of these devices is used to supplant a vacuum tube as the switch in a hard-tube modulator, external resistance must be provided to limit current in case of a shoot-through event. As with a gridded vacuum tube, the Crossatron is almost always more susceptible to damage than the diode-type microwave tube it is pulse modulating. For this reason, it determines the magnitude of the series resistance. The value of this resistance, in turn, often determines the usable output power from the switch because its resistance value can easily be greater than the ratio of conduction-tube drop and output current.

Crossatron tubes have been built with as much as 100 kV hold-off and closing currents as great as 12,000 A. When optimized as either a closing-only or opening-only switch, the 8454H is rated at 50-kV hold-off, 2500-A closing, and 1000-A opening. The Crossatron is one of the few full-control (or at least self-commutation) switches that is still in the process of being developed.

10.5 Load lines

Although we have discussed and illustrated the voltage/current characteristics of a number of triode and tetrode modulator switch tubes, we have yet to relate them to the complementary aspects of the non-linear load impedances that they will be expected to drive. Figure 10-30 shows how two types of modulators would interact with three types of loads, described as load lines. In this representation, load voltage/current relationships are plotted starting with the 0/0 points of the voltage/current axes. The modulator, or source, voltage/current relationships are plotted backwards, starting at the open-circuit-voltage, zero-current points.

Of the three load characteristics, only one represents an actual device. (It is also the least linear one.) It is a backward-wave, crossed-field amplifier, the QK-622 Amplitron. As voltage is increased across the QK-622, current builds up slowly, reaching only 4 A at 40 kV, which is near the knee of its characteristic. Between 40 kV and its operating voltage of 50 kV, current dramatically increases, from 4 A to 32 A. At its operating point, its equivalent linear-resistance impedance, its second characteristic, would be 1563 ohms, which is the slope of a straight line from the operating point back to the origin. However, its dynamic, or incremental, impedance is the slope of the characteristic, or $\Delta V/\Delta I$, as it passes through the operating point, which is 200 ohms, not 1563 ohms. The third load characteristic illustrates is a hypothetical linear-beam device whose operating point is the same as the QK-622 Amplitron, and it follows the 3/2-power voltage-current relationship. Its incremental impedance at the operating point would be $\Delta V/\Delta I$, or 1024 ohms.

Also shown are the source characteristics for two types of pulse modulators: line-type and hard-tube. The line-type modulator characteristic starts at 100 kV and zero current at the right side of the graph and ends at zero voltage and 64 A at the left side, passing through its matched-load operating point of 50 kV and 32 A. The open-circuit voltage is twice the matched-load voltage and the short-circuit current is twice the matched-load current, which is a sure-fire criterion for recognizing a matched source. (It is *not*, however, matched to the incremental impedance of either of our microwave-tube loads.) The hard-tube modulator characteristic starts at 60 kV and zero current at the right side. It is discontinuous, however, having a knee. In fact, it is much like the Amplitron characteristic only rotated by 90°. This means that the hard-tube modulator is the Amplitron's dual. As the voltage drop across the hard-tube switch increases beyond the knee, which occurs at 50 kV (10-kV tube drop, working from right to left), the current tends to be essentially constant, which is the exact opposite of what happened to the Amplitron. Therefore, the hard-tube switch has two distinct incremental impedances: one relatively low "below the knee" and another relatively high "above the knee." The most dramatic difference between the two was exhibited by the

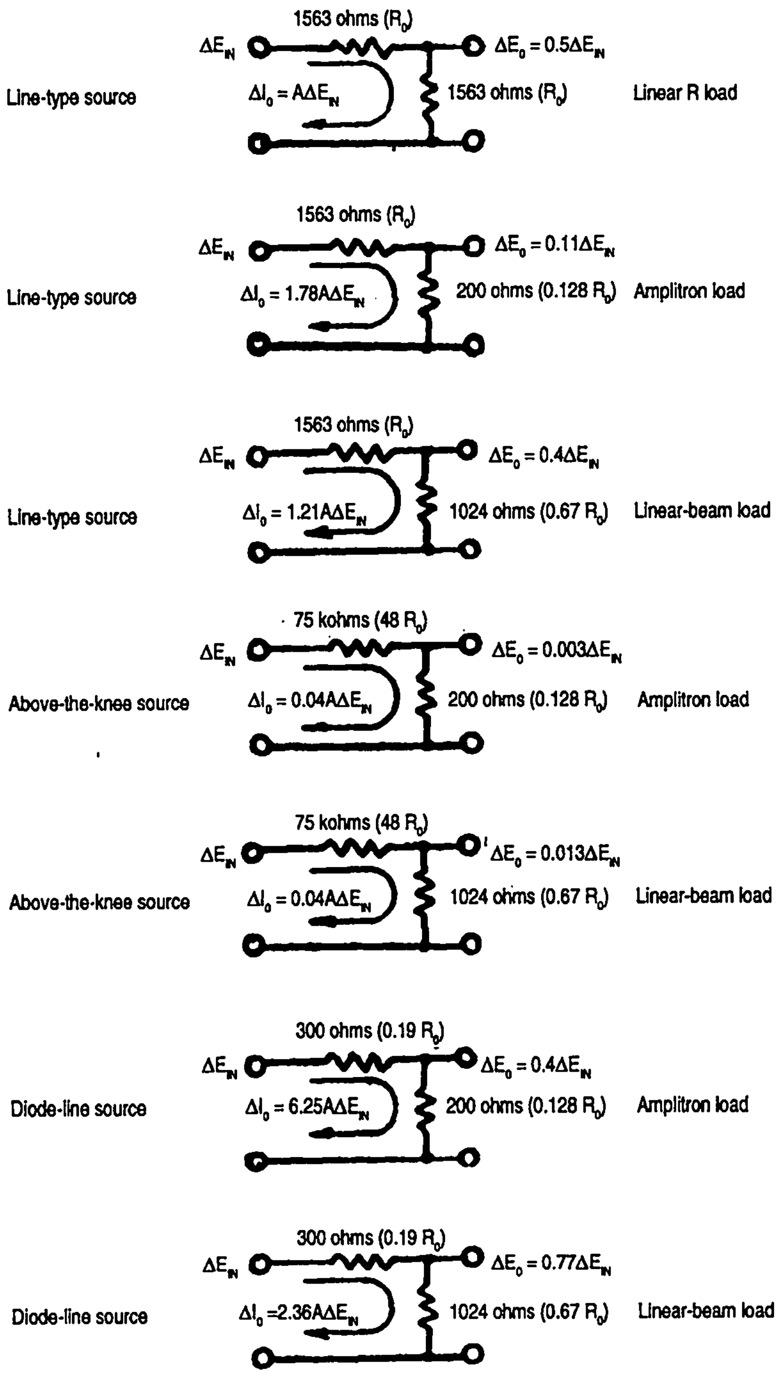


Figure 10-31. How modulator internal dynamic impedance interacts with load impedance to attenuate input voltage variations.

Injectron beam-switch tube. The incremental above-the-knee impedance of the L-5097, for instance, is approximately 75,000 ohms, which is what has been illustrated. The below-the-knee incremental impedance is 10,000 V/32 A , or a little more than 300 ohms. The ratio of the two is almost 250:1.

One of the functions of the load-line/source-line representation is to visualize what happens when one or another parameter varies. If, for instance, the characteristic impedance of the line-type modulator PFN were changed, the slope of its characteristic would change as would the point of intersection with a load characteristic, thus establishing a new operating point. Lower impedance would mean higher short-circuit current, tilting the line upward from its 100-kV starting point. Higher impedance would tilt it downward. What is more likely to happen, however, is a change in the supply voltage, which can take the form of imperfect regulation, intrapulse droop, or power-supply ripple. The effect of supply-voltage change, or modulation, is to shift the starting points of the source characteristics either to the right or to the left, moving the lines with them. For ripplelike modulation of source voltage that has no average value, the average operating point does not change either.

The change in load voltage will always be some fraction of the change in source voltage because of the voltage-divider action of the incremental source and load impedances in series. For example, take the case of a line-type source matched to the linear resistor. In this case, the voltage-attenuation factor is two because source and load resistances are equal. Less obvious is the case with the Amplitron load, which has an incremental resistance (200 ohms) that differs from its zero-intercept impedance (1563 ohms). The attenuation factor is the sum of 1563 ohms plus 200 ohms divided by 200 ohms, or about 10. A 1-V change in source voltage produces only a 0.11-V change in load voltage. With the Amplitron driven by a hard-tube source, an increase in source voltage of 1 V—which is in the above-the-knee direction for the hard tube—is attenuated by a factor of (75,000 ohms + 200 ohms)/200 ohms, or 376. So the Amplitron voltage increases by only 0.003 V. (This is why constant-current hard-tube modulators are used with crossed-field amplifiers for maximum performance.) If, however, the source voltage had been reduced by 1 V, the Amplitron voltage change would have been attenuated by the below-the-knee source resistance of 300 ohms, giving a factor of (300 ohms + 200 ohms)/200 ohms, or 2.5, which is a long way from 375. Needless to say, the actual knee of the hard-tube characteristic is never this abrupt, but the trend is the same. Figure 10-31 shows the transfer characteristics for a number of source and load combinations.

10.6 Pulse fall time of hard-tube modulator

The current available from a hard tube used as a series modulator switch (and illustrated again in Fig. 10-32) determines the rise time of the voltage pulse applied to the parallel combination of diode-electron-gun load and stray capacitance, which includes that of the switch tube itself. If the switch tube is a power triode, there will likely be greater anode current available when there is full supply voltage across it than when it has reached its intrapulse operating point (typically a 10% tube drop). The higher the amplification factor of the tube, the smaller the difference in currents is likely to be, and in the case of the Injectron-type beam-switch tube, there is virtually no difference at all. The switch tube in this case is a constant-current source, regardless of the voltage across it. Some of its current is required to charge up the stray, or distributed, load capacitance, but at any state of charge, the remainder of the current, which is proportional to the 3/2 power of the voltage that we are attempting to build up, will flow into the diode-electron-gun load. Therefore, there is no simple circuit model such as an

RC time-constant, because R varies continuously with voltage.

The rise-and-fall waveforms for a constant-current source are identical but inverted, as shown in Fig. 10-32. They have variable slopes like RC time-constant exponentials, except that the closer the exponentials come to the asymptotes, the greater are the instantaneous values of R and the more slowly they approach asymptotes, much like exaggerated RC time-constants. It is possible, however, to evaluate the length of time required for the pulse voltage to change from one normalized value to another by using the formula shown in the equation

$$T_F = t2 - t1 = \frac{2C_S}{k\sqrt{V_k}}\left(\frac{1}{\sqrt{a2}} - \frac{1}{\sqrt{a1}}\right),$$

where $a1$ and $a2$ are two such normalized values, typically 0.1 and 0.9. The three important parameters are the perveance of the diode gun, k; the flat-top pulse cathode voltage, V_k, and the value of the stray capacitance, C_S. In the example shown, a tube that operates at 50-kV beam voltage and 1-A beam current and has a perveance of 0.9×10^{-6}, has a stray capacitance of 200 pF. Given these parameters, the time between the 10% and 90% points of the rise-and-fall waveforms would be 38 μs if there was a constant-current source of 1 A. This would be unacceptably long for many, if not all, applications.

However, this is a rare case because most cathode-pulsed tubes have higher perveances and operating voltages. Moreover, switch tubes are usually overdriven during the rise time to provide an excess of current in order to charge the stray

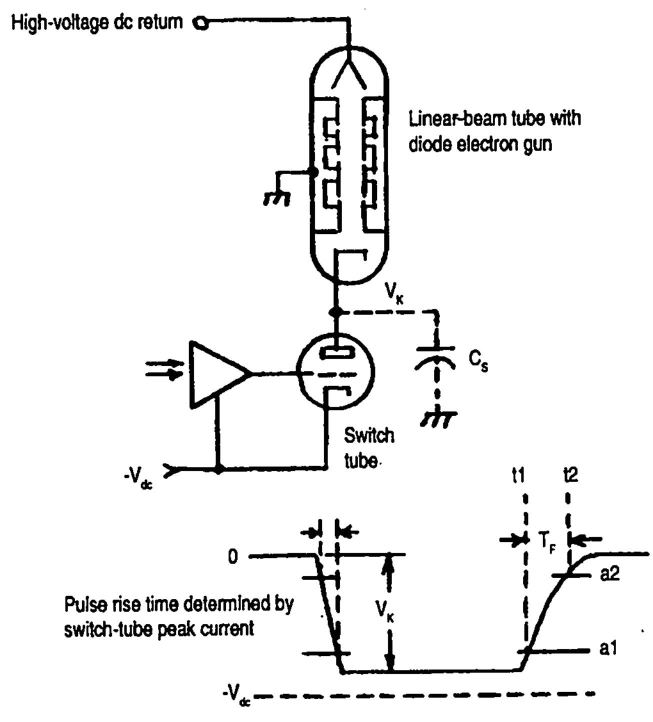


Figure 10-32. Determination of pulse fall time of a hard-tube modulator.

capacitance. The fall time, however, cannot be reduced by anything that can be done to or with the switch tube. When shorter fall time is required, a "tail-biter" or "active pull-up" stage is usually employed. (This feature was illustrated in Fig. 2-1, the transmitter block diagram.) This device, usually another switch tube, is connected between the diode-gun cathode and ground and is pulsed into conduction only during the desired fall time of the pulse. Its current adds to the ever-diminishing current of the diode load to facilitate discharging of the stray capacitance. A constant-current tail-biter of current I will produce a fall time, Δt, that is no longer than $\Delta V \times C_S/I$, where ΔV is the pulse voltage amplitude, even if the diode-load current is negligibly small.

In the case of grid-driven triode switch tubes, which include the modulating-anode type of tube as well, there is another aspect to rise-and-fall time: the effect of "Miller" capacitance, which is related to the capacitance between anode and control grid. Miller capacitance and the capacitance between grid and cathode make up the static input capacitance of a triode. With the anode connected to a fixed voltage source, the rate at which the voltage between grid and cathode can be varied depends upon the driver stage's source resistance, or current-delivery capability, and the sum of the two input capacitance components. The anode, however, is usually not connected to a fixed voltage. It is connected to the load, and the load-voltage excursion is the same as the anode-voltage excursion. The change in charge, or ΔQ, experienced by the anode-grid capacitance is not the product of the grid-voltage change times anode-grid capacitance but is the anode-voltage change times the anode-grid capacitance. In effect, the anode-grid capacitance component of total triode input capacitance is multiplied by the voltage gain of the triode. This fact must be considered in the design of a grid driver. Moreover, the internal impedance of the driver stage in both directions must be considered. This is because all amplifiers that have a single active device have source impedances that are different, depending on whether the device is a source or a sink for output current. For this reason, active pull-up/active pull-down pairs connected in a "totem-pole" configuration are the preferred grid driver stages in critical rise-and-fall-time applications. This is because they can be designed to deliver equal output currents in both directions.

With high-voltage and high-current transistors now available—especially the metal-oxide semiconductor field-effect transistor (MOSFET)—triodes do not have to be grid-driven. They can be cathode driven by what is often called the grounded-grid, or grid-separation, connection. The control grid can be connected to a fixed source of positive grid bias, if necessary. A typical cathode-drive connection is shown in Fig. 10-33. During the switch tube's interpulse (or off) state, the MOSFET is in the off state. To reduce cathode current to the same value as the MOSFET's drain-leakage current, which is negligibly small, the drain of the MOSFET will automatically assume whatever positive voltage is required, including an amount equal to whatever positive voltage is applied to the grid. No negative grid bias supply is needed at all. To cause output current and load voltage, the cathode MOSFET is turned on, pulling the switch-tube cathode to within a few volts of ground (or deck) potential. The effective grid-cathode voltage at this time is whatever positive voltage source is connected between grid and ground. The "Miller" charge and discharge currents flow either into the ground-reference (or deck-reference) conductor or into the output capacitance of the positive grid-voltage source. This grid-voltage source can be made sufficiently large so that ΔQ through the anode-grid capacitance produces negligible power-supply output-capacitance voltage change.

"Miller" capacitance affects a tetrode-type switch tube about the same way it affects a cathode-driven triode: the electrode closest to the anode is also connected to a fixed source of voltage that is the high-frequency equivalent of ground (deck). The control-grid, or input, capacitance comprises both grid-cathode and grid-screen components, but the screen appears to the grid as an anode connected to a fixed voltage source. The screen-voltage supply, however, must act as both a sink and source for the ΔQ produced by the anode-screen capacitance charge and discharge without generating significant terminal-voltage change.

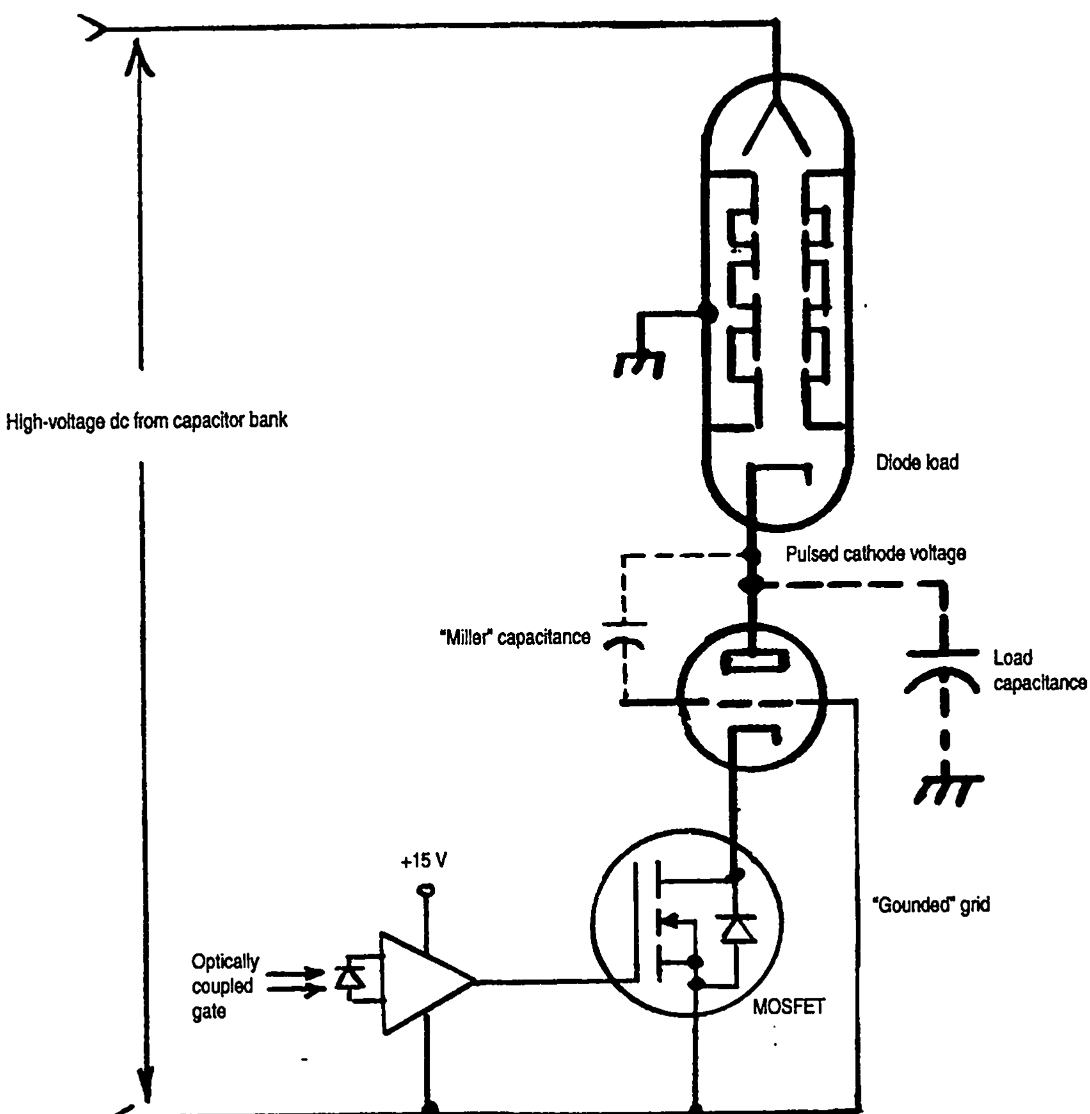


Figure 10-33. Cathode drive or "grounded-grid" operation of triode modulator switch tube.

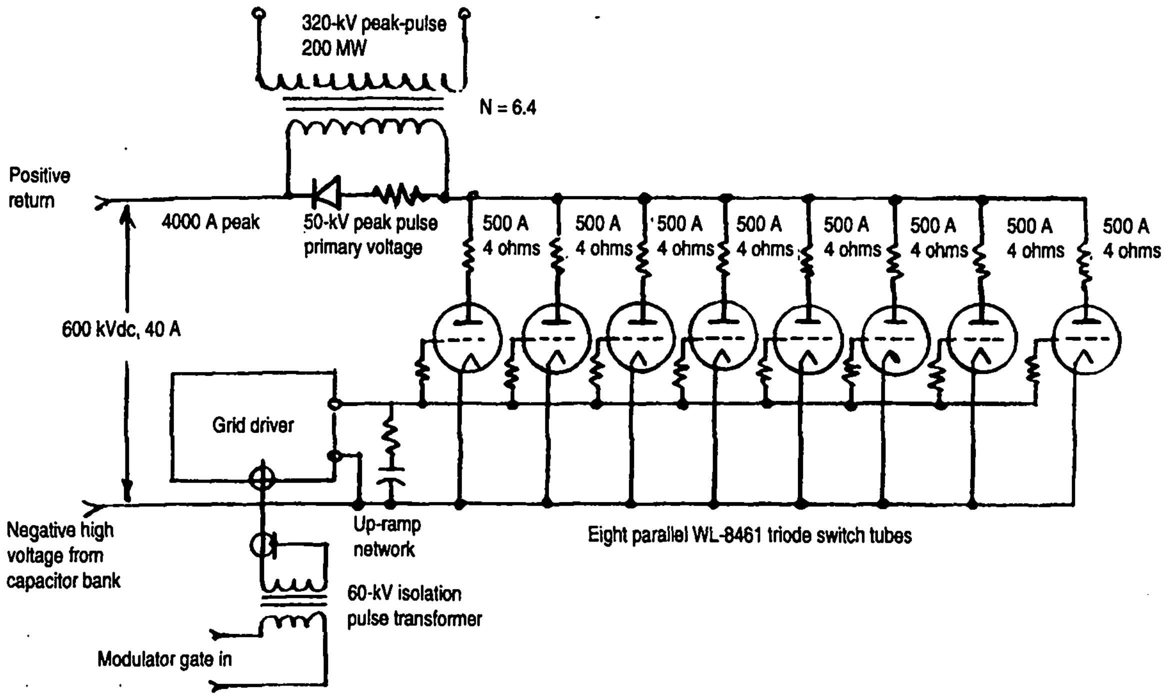


Figure 10-34. Simplified schematic diagram of "200-MW" modulator using WL-8461 triodes.

10.7 Some practical examples of hard-tube modulators

Now let's look at some applications of a couple of these switches.

10.7.1 The 200-MW triode (WL-8461) hard-tube modulator

A hard-tube pulse modulator was actually built using eight parallel-connected WL-8461 triode switch tubes coupled to a super-power klystron load through a 320-kV output-pulse transformer. Its simplified schematic diagram is shown in Fig. 10-34. Each of the triodes was to be operated with peak plate current of 500 A (for a total primary-loop pulse current of 4000 A) and a supply voltage of 65 kVdc, which is adequate to support a 50-kV primary voltage output pulse with as much as 15-kV switch-tube anode-voltage drop.

Although several identifiable pulses at the 200-MW peak-power level were actually obtained, the modulator was never capable of continuous, repetitive operation at this level. The operation of the electronic crowbar in response to individual switch-tube arcs was far too frequent an occurrence, even at output levels below 100-MW peak. This was an experimental modulator, however, and much was learned from it. One empirical lesson, which has not added much to an analytically satisfying data base, was that the performance demonstrated by a single tube operating alone could not be repeated when the tube was operated in parallel with others trying to do the same thing. It might be expected, for instance, that if an individual tube had an average period between internal arc events called T, an ensemble of n such tubes in parallel might have an average spark-down interval of T/n. The observed interval was much shorter than that.

Internal switch-tube arcs are presumably random, non-synchronous events. Extensive diagnostic checks were used, including broad-band delay lines to de-

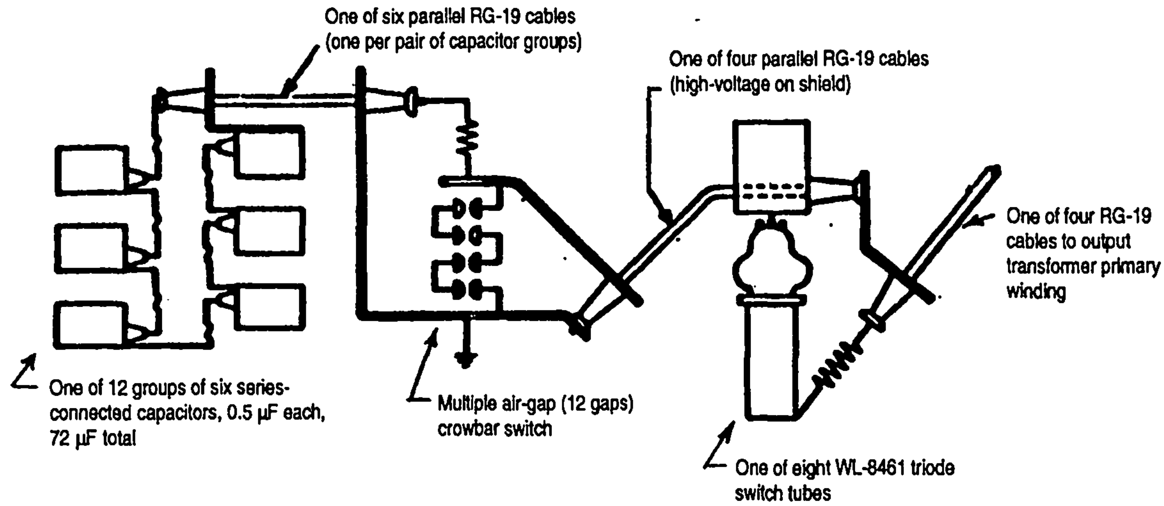


Figure 10-35. Physical arrangement of components of 200-MW hard-tube modulator (end view: parallel-connected components extend into the page).

lay the pulse output information so that the conditions immediately preceding an arc could be captured on a conventional storage oscilloscope that was triggered by the arc event itself. (This is a data-collection problem ideally suited to today's continuous-sampling digital storage oscilloscopes.) Unfortunately, the results showed that the vast majority of arcs occurred precisely at the end of a current pulse, when the switch tubes were being switched to the voltage-blocking state. So common were these that the acronym EOPSTA was used as log-book shorthand for "end-of-pulse switch-tube arc."

The first experimental objective to become a casualty of this problem was pulse fall time, which was originally intended to be 0.1 μs. Of course, the culprit

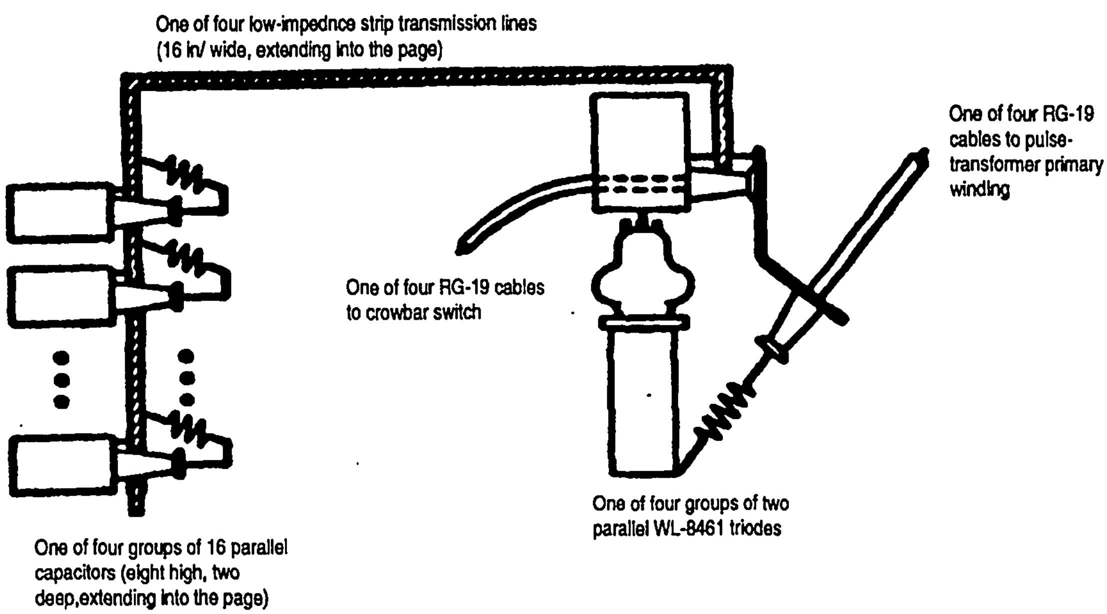


Figure 10-36. Experimental use of low-impedance strip transmission line to reduce inductance of primary-current loop of 200-MW hard-tube modulator.

in such a low-impedance primary loop was series inductance. The inductance of the as-built primary loop—which included the capacitor bank, the eight switch tubes, the electronic crowbar connection, and the pulse-transformer primary connections (see Fig. 10-35)—was 1.7 μH. When the primary-loop current of 4000 A is brought to zero in 0.1 μs, or 10^{-7} seconds, the rate-of-change of current is 4×10^{10} A/s, and the impulse, or *L di/dt*, voltage across 1.7 μH would be 68 kV, which more than doubled the instantaneous anode voltage. This factor, of course, came in addition to the voltage spike produced by the pulse-transformer back-swing clipper circuit, which shunted the primary winding. Substantial *dv/dt* snubbing was required before this component of stress was brought to more reasonable levels.

Before giving up on fall time, however, engineers made another attempt to further reduce the primary-loop inductance. A low-impedance, parallel-plate transmission line was used. It had a very large width-to-thickness ratio to make the interconnections, as illustrated in Fig. 10-36. The intent was good but the implementation was doomed to failure because of the electric-field enhancement

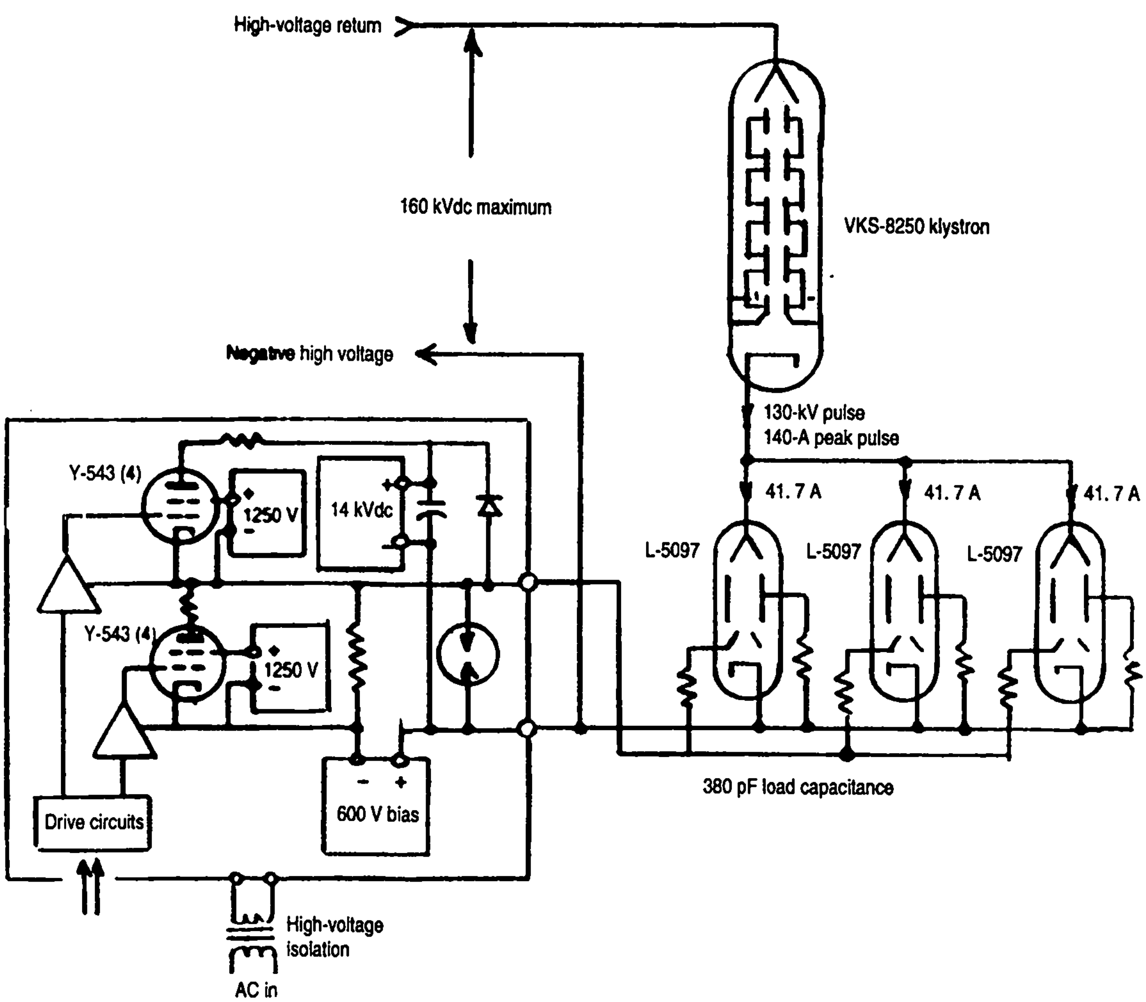


Figure 10-37. Simplified schematic diagram of 18-MW peak-power, high-precision, hard-tube modulator using L-5097 BSTs.

Table 10-1. Operating parameters of 18-MW hard-tube modulator shown in Fig. 10-37.

Parameter	Value
Pulse voltage to klystron	130 kV
Pulse output current	140 A
Peak-power output	18 MW
Pulse duration	3-28 μs
Burst mode	32 pulses, max., 4-μs duration, 7-28-μs spacing
Pulse-pair mode	10-μs duration, 13-μs spacing
Rise time	1.2 μs
Duty factor	0.03, max.
Pulse-top ripple	0.1% (during RF)
Pulse-to-pulse variations	0.02%

at the side-to-side endings of the parallel plates. Merely extending the dielectric material between the plates far beyond the ends of the plates did nothing to prevent corona discharges, which rather quickly punctured the dielectric material.

Reducing fall-time *di/dt* did not completely cure EOPSTA, however. Other effects were also at work. The temperature of the control grids, which were subject to power dissipation throughout each output pulse, reached transient maxima at the end of each pulse. This made them more likely to emit secondary electrons and less likely to establish zero-current conditions within the tubes when their voltages were brought to the negative voltage that should have been adequate for current cutoff. Experiments were performed where grid drive voltage and conduction-anode-voltage drop were interchanged between so-called hard-drive and soft-drive conditions. Unfortunately, reducing intrapulse grid drive voltage and increasing anode supply voltage to compensate for it produced a statistically insignificant difference in spark-down rate. The only thing that really worked was operating the circuit at lower voltage and lower power.

The suggestion that the problem may have been caused by the tubes' "hot grids" gave rise to an experiment that replaced the eight triodes with eight tetrodes of roughly comparable capability (type 4CW250,000). It turned out, however, that "hot screens" produced much the same effect as "hot grids." Virtually no improvement was noted.

This modulator experiment did, however, lead to a successful 60-MW pulse modulator, which used six parallel-connected WL-8461 triodes that pulsed a 20-MW-peak-power UHF klystron in the transmitter of an instrumentation radar. The system worked well for many years.

10.7.2 An 18-MW hard-tube modulator using the L-5097 beam-switch tube

Waveform versatility and both pulse-top and pulse-to-pulse precision are the salient features of the hard-tube modulator shown in Fig. 10-37. A broad-band high-performance S-band klystron, the VKS-8250, is directly cathode pulsed by a group of three paralleled L-5097 BSTs. The pulsed cathode voltage, which is 130 kV, is driven by a power supply with maximum dc output of 160 kV. The total klystron beam current of 130 A peak is shared equally among the three L-5097 switch tubes, 41.7 A each. They share current without external isolation components in quite the same fashion as MOSFETs.

The modulating anodes of the three switch tubes are driven from a common pulse source that is configured as a "floating-deck" modulator (which is the next general category of modulator to be considered). This design satisfies a previously mentioned requirement: it behaves as both a source and a sink for the BST's modulating anode displacement currents, which are magnified by the "Miller" effect. Both driver stages use four parallel-connected, high-performance, pulse-rated tetrodes (type Y-543), operating with 1250-V screen supplies. The "pull-up" stage is a "bootstrap" circuit, which means that the common connection for screen and filament supplies and for the low-level drive circuit is the pulse-output bus, which is, so to speak, pulled up by its bootstraps. The driver output is capable of pulsing the modulating anodes of the BSTs to 14-kV positive and returning them to an interpulse negative bias level of -600 V (which is not an absolute necessity for zero BST current).

As shown in Table 10-1 by the listed ensemble of pulse waveforms, including doublets and multi-pulse bursts, advantage is clearly being taken of the electronic agility of the full-control hard-tube modulator switch. Pulse-top variations are 0.1% and pulse-to-pulse variations are 0.02%. Both of these performance achievements are facilitated through the attenuation of power-supply variations by the relatively high incremental-source resistance of the BSTs. The source resistance of each one is 75,000 ohms, giving a parallel-combination resistance of 25,000 ohms for the three tubes. The operating-point incremental resistance of the klystron, $\Delta E_k/\Delta I_k$, is 640 ohms. Therefore, the voltage-attenuation factor is 25,000 ohms divided by 640 ohms, or 40, which is 32 dB.

This is but one of a number of highly successful, highly precise pulse modulators that share very similar circuitry but use different numbers of parallel BSTs.