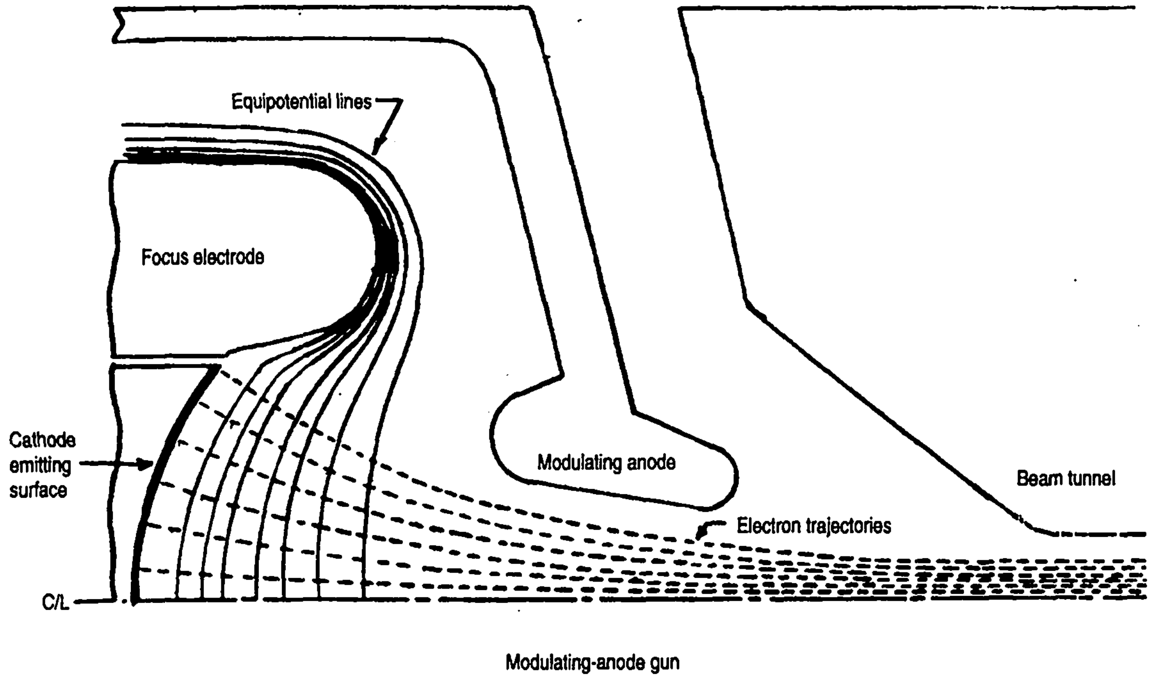
12. Control-Grid Modulators

Of course, the size, weight, complexity, and expense of a pulse modulator can be minimized if the voltage and power required by the current-control electrode of the microwave tube can also be minimized. At the present point of evolution of the electron tube, the high-μ, high-transconductance grid is the control electrode that best achieves these goals. Regardless of the current intercepted by the control electrode, the switching losses, or the energy that must be dissipated in charging and discharging the capacitance associated with the electrode, for the control-grid tube, each pulse will vary as the square of the voltage swing. All major parameters being approximately equal, the capacitance associated with a control grid is about the same as that for a modulating anode. But the voltage swing for a grid may be only a small fraction of that required by a modulating anode, and the switching losses will be lower by the square of that ratio. Therefore, for high-repetition-rate operation—typically above 5000 pps—a control grid is almost mandatory, regardless of duty factor.

To be effective, however, the grid must be as close to the cathode as possible. But this presents problems. The cathode is hot and so it will radiantly heat the grid. The cathode also emits electrons, which are intercepted by the grid. Both of these factors will limit the average-power capability of the control-grid tube. But just as triodes and tetrodes have grid geometries that can minimize beam interception, so does the control-grid tube. The most popular geometry is the "shadow grid." The shadow grid is not the grid itself but is a masking electrode that is located near, on, or even embedded in the surface of the cathode. It inhibits emission from the cathode in those regions where electrons could directly strike the control grid, thus casting an electron "shadow" on the control grid.

Even if the shadow-grid concept were completely effective in keeping electrons from hitting the grid, and even if the grid itself were capable of dissipating unlimited power, there would still be an average-power penalty that the gridded-tube must pay. This is illustrated in Fig. 12-1, which contrasts computer-derived electron trajectories for an electron gun using a modulating anode with trajectories for a shadow-gridded gun. The electrostatic lens produced by the modulating-anode geometry is at least theoretically capable of generating a perfectly collimated, area-convergent electron beam; there are no undulations or "scalloping" tendencies of the beam as it enters the narrow beam tunnel and comes under the influence of the magnetic focusing field. (This is not to say, however, that the magnetic field cannot penetrate into the gun region. When it does, and if its strength is tapered so as to be congruent with the electrostatic lens, what results is called "confined-convergent flow," which results in the highest-quality electron beams achievable.) The gridded gun, however, produces an electrostatic aberration almost immediately; it introduces beam scalloping that no amount of magnetic focusing can completely mitigate.

Beam interception by the RF interaction circuitry of the microwave tube will always be higher for the gridded gun than for the diode or modulating-anode gun. This problem is often compounded by designs that use periodic perma-



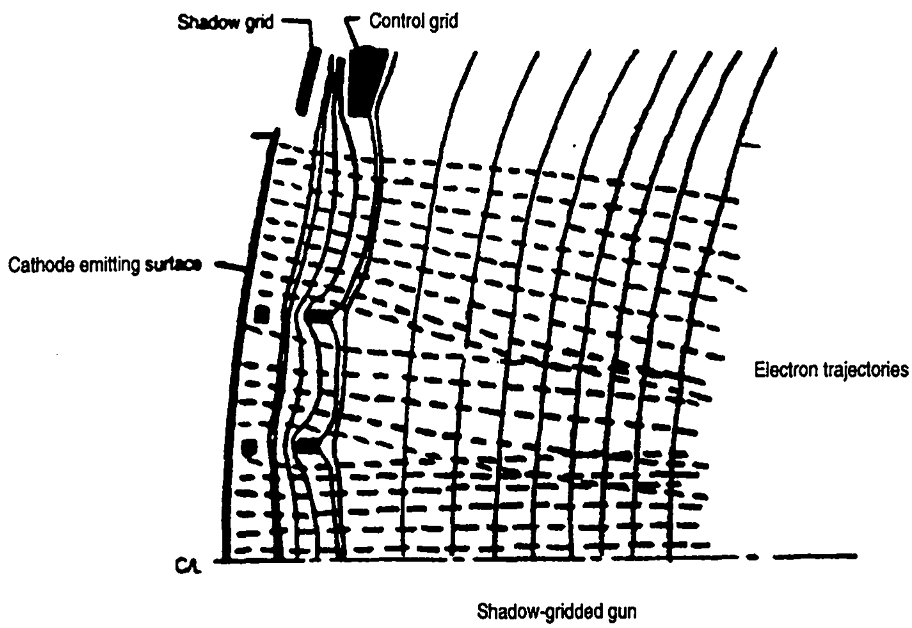


Figure 12-1. Electron-beam formation in modulating-anode and gridded electron guns.

nent-magnet focusing, which also has imperfect polarity transitions along the beam. Added to this problem is the fact that the electrostatic lens is even less perfect for voltage transitions between grid and cathode, such as during the rise-and-fall times of the control-grid voltage pulse. This phenomenon is shown in Fig. 12-2 for a gun with a typical grid base. (The grid voltage for beam cutoff is -1% of beam voltage; grid voltage for full beam current is +1.8% of beam voltage, therefore the total grid voltage swing is 2.8% of beam voltage.) Note how the total cathode current splits between the collector and body as the grid voltage

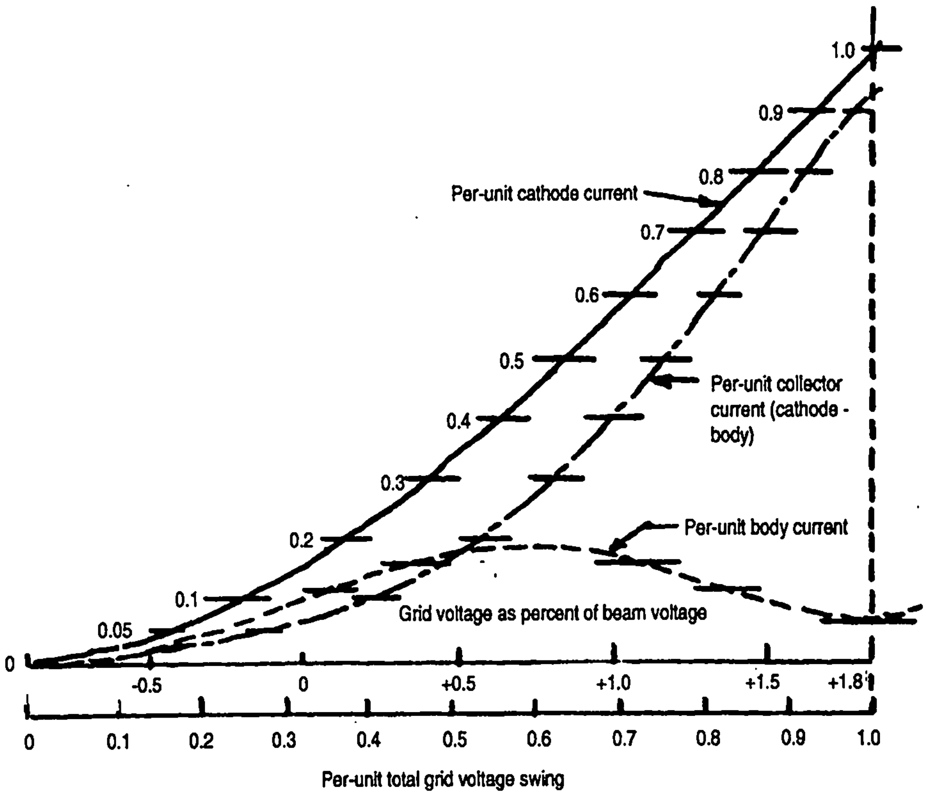


Figure 12-2. Typical transfer characteristic of shadow-gridded gun.

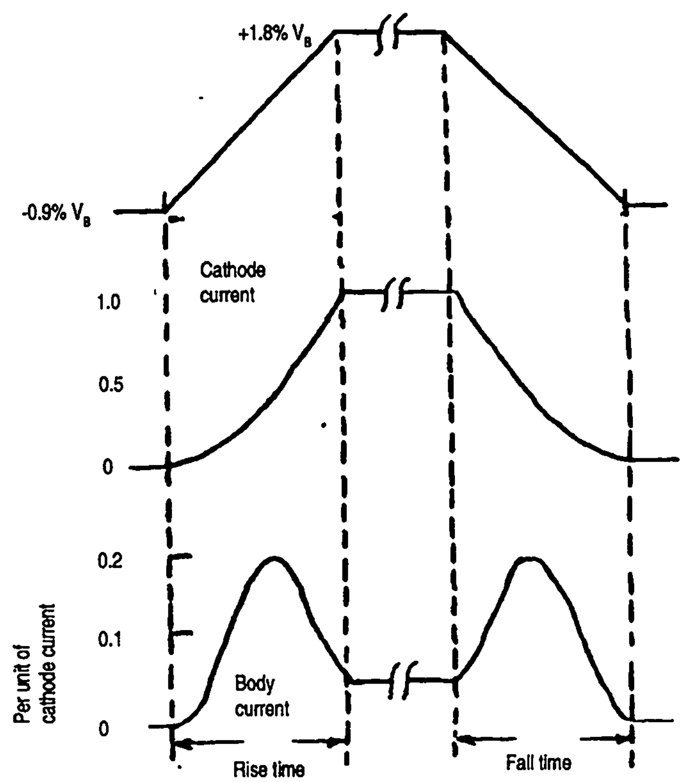


Figure 12-3. Time-domain voltage and current waveforms for tube with shadow-gridded electron gun.

swings from full-off to full-on. At full current, the portion intercepted by the body is only 5%. At about 60% of the grid-voltage swing, however, the body current rises to a value that is 20% of the eventual full-cathode current, which, at that point, is almost equal to the current that arrives at the collector. Beam transmission at that point is only slightly greater than 50%, whereas it will reach 95% at full current.

This is of no great consequence in long-pulse, low-repetition-rate service, where the percentage of the total "on" time devoted to the rise-and-fall portions of the pulse is small. But one of the reasons for using a control-grid geometry in the first place is to facilitate operation at a high repetition rate. Given the average-power constraints, the duty factor is usually only modest (typically less than 1%), which means that at high repetition rates—and rates up to 100,000 pps are not unheard of—pulses will tend to be short, and rise-and-fall intervals will be significant fractions of the flat-top durations. This situation gives rise to waveforms similar to those shown in Fig. 12-3. If we assume a linear rise and fall of grid voltage, then cathode current, varying as the 3/2 power of instantaneous voltage, will approximately follow the shape shown. More important, however, is the shape of the intercepted body current in the form of "rabbit-ears." As the flat-top duration gets smaller, the ratio of the average value of body current to the average value of cathode current can be 2 to 3 times as great as it would be for the same useful (flat-top) duty factor, if the ratio was obtained with longer pulses at lower repetition rate. This problem, mitigated by faster grid-voltage rise-and-fall performance, theoretically disappears altogether with vertical-sided pulses. Vertical-sided pulses, however, cause infinite *di/dt*, subjecting any inductance in the cathode-current loop to infinite transient voltage, which has also been known to cause problems.

12.1 Grid-modulator topologies

Using the rule of thumb that grid-voltage base (or the required grid voltage swing from fully cut-off to fully turned-on) is less than 3% of the operating beam voltage, we can see that voltage swings of less than 1000 V will suffice for tubes operating at less than 30 kV of beam voltage, which includes the vast majority of tubes. (We will later discuss a notable exception.)

Today's transistor technology supports the design of single-device circuits that will switch 1-kV pulses—with not much safety factor, to be sure. This situation means that solid-state designs dominate the field of grid modulators. The availability of reliable and inexpensive fiber-optically coupled signal-links that are immune to noise and interference has made direct-coupled modulator designs far more practical than they were not so long ago. Nevertheless, many grid modulators use pulse-transformer signal coupling of the entire grid pulse, which is generated by a ground-referenced pulse generator. Some have even used capacitive coupling. (When duty factor is low, as it usually is, dc restoration of the coupled pulse is not required.) With capacitive coupling to the grid from a ground-referenced pulse generator, there will be a natural tendency for an intercepting-type grid to clamp the pulse voltage at zero bias because the grid-cathode circuit of the microwave tube will function as a forward-biased clamping diode. The average value of this rectified pulse current will develop a voltage

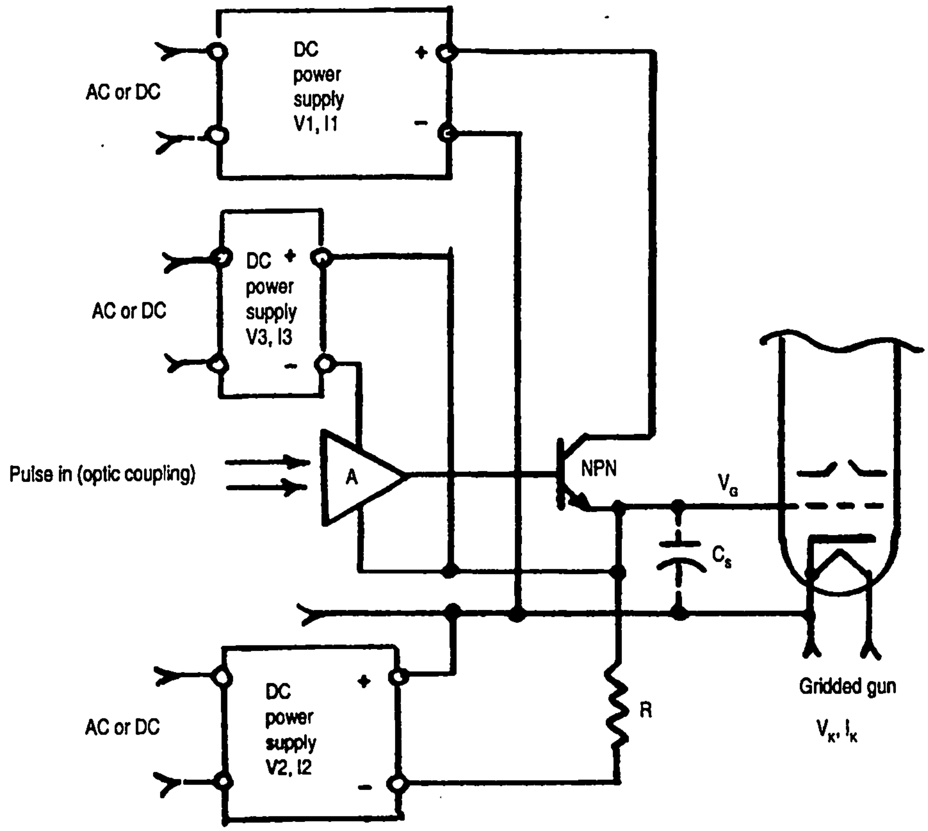


Figure 12-4. Grid modulator with bootstrap active pull-up, passive pull-down.

across whatever shunt resistance is connected in the circuit between grid and cathode. This voltage is called "grid-leak" bias. The bias voltage automatically adjusts itself so that the positive excursion of the capacitively coupled pulse just reaches zero grid-cathode voltage. In order to drive the grid positive, it is necessary to provide a low-impedance sink at the desired fixed bias to absorb the rectified grid current.

Common-mode rejection is not a great problem, because the microwave-tube cathode voltage, which is the return reference for the grid pulse, is nominally non-variant with respect to ground. (However, it will experience a step-voltage drop due to pulse-cathode current flowing through the fault-current-limiting resistance and a linear droop due to charge removal from the energy-storage capacitance.) Fast rise is made difficult by the leakage inductance of the pulse transformer, which can be physical large, too. The higher the microwave-tube beam voltage, the greater will be the insulation required between the primary and secondary of the pulse transformer—and the more difficult it becomes to minimize leakage inductance between windings. The coupling of long pulses depends upon the magnetizing inductance and core-saturation characteristics of a pulse transformer and/or the magnitude of the capacitance value of a coupling capacitor.

Transformer-coupled grid pulsers do have an advantage in that greater grid voltage can be obtained at the expense of proportionately greater pulse-generator output current by using the appropriate value of the secondary-to-primary turns

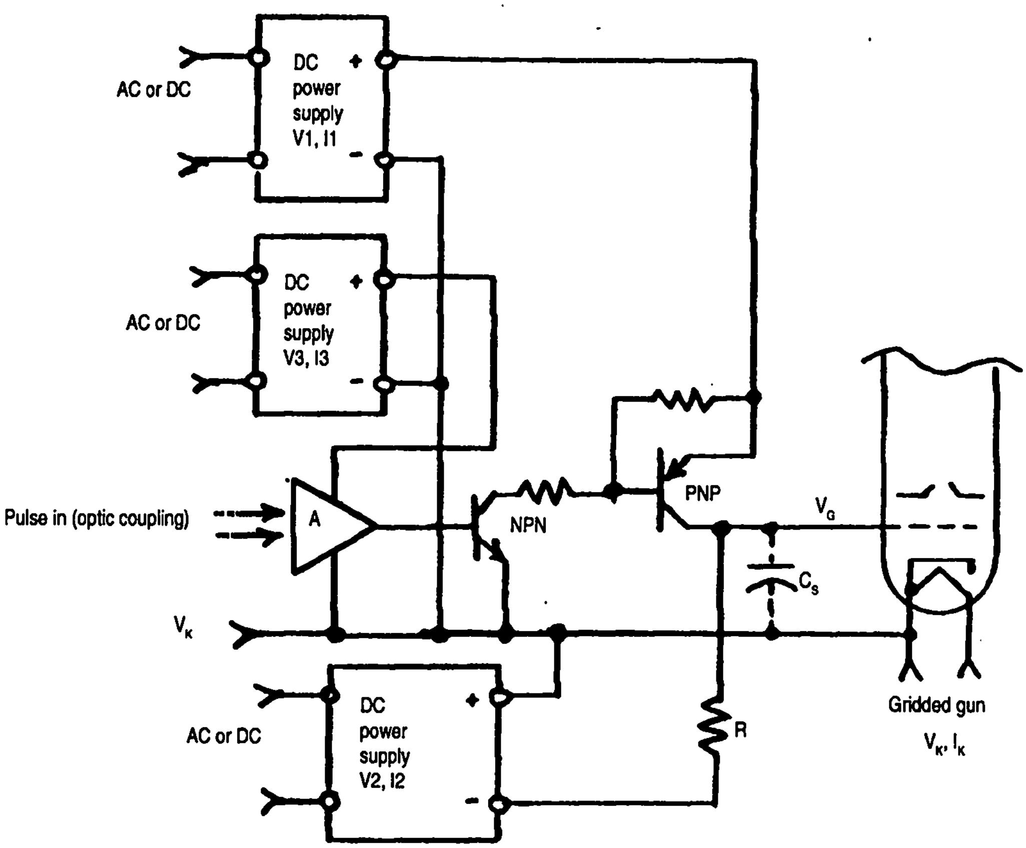


Figure 12-5. Grid modulator with open collector active pull-up, passive pull-down.

ratio. Fixed dc interpulse grid bias must also be generated by a power supply referenced to the microwave-tube cathode. The input to this supply can usually be obtained from the microwave-tube filament supply, which is already isolated from ground for the full dc cathode voltage.

Direct-coupled grid-modulator circuits do not suffer from duty-factor-induced level shifts or pulse-top duration limitations. The specific topologies of these modulators will now be described.

12.1.1 *Active pull-up, passive pull-down*

As before, we will begin our discussion with the simplest type of topology: the single active switch, as shown in Fig. 12-4. However, unlike the modulating-anode pulsers, passive pull-up is not a practical option for grid electrodes because of the finite intrapulse grid current that the grid pulser must supply. This current must flow through the pull-up resistor. The passive pull-down resistance, as before, determines the fall time-constant of the modulator pulse, RC_S, where C_S is the stray capacitance of the grid to cathode and focus electrode. The active pull-up switch is shown schematically as an NPN bipolar transistor, but is more likely to be a high-voltage N-channel MOSFET. The output is taken from the emitter, or source, but the transistor is not an emitter or source follower. It is a true bootstrap connection. The optic receiver and pre-driver, lumped together as amplifier *A*, and their housekeeping dc power supply (*V3, I3*), have common returns that are connected to the grid-pulse output. They should have low self-

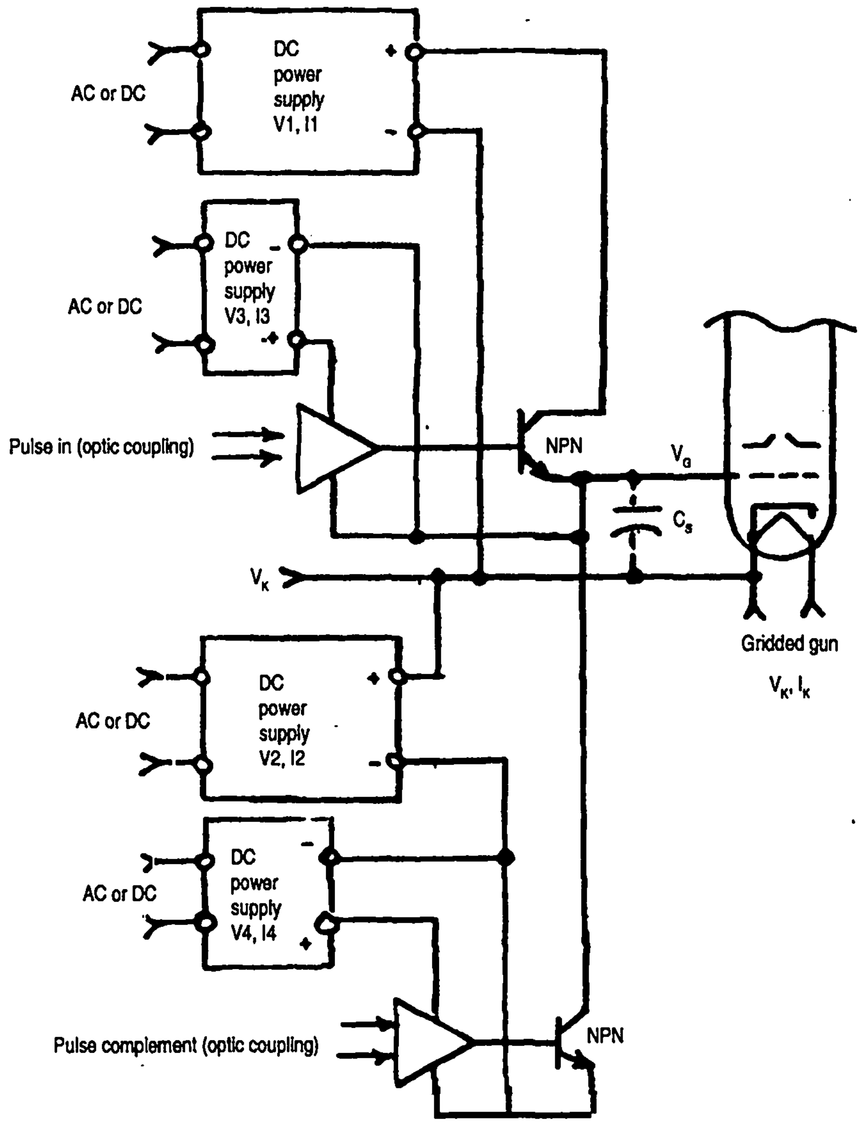


Figure 12-6. Solid-state grid modulator with active pull-up (bootstrap type) and active pull-down.

capacitances because they add to C_S in determining rise-and-fall times.

There are two other dc power supplies whose return leads are connected to the microwave-tube cathode. Supply *V1* must be equal to the sum of the voltage drop of the pull-up switch (usually negligibly small), which is approximately 1.8% of the beam voltage, V_K. And it is equal to the amount by which V_G is pulled positive. Supply *V2* is equal to the interpulse cut-off bias voltage, approximately 1% of V_K. The voltage across the transistor switch during the interpulse interval is *V1* + *V2*, or approximately 0.028 V_K, which is also the voltage across *R* during the pulse. The average currents, *I1* and *I2*, have two components: 0.028 V_K/R x duty-factor, and 0.028 V_K x C_S x repetition rate. Assuming the conduction-voltage drop of the transistor switch is negligibly small, its average-power dissipation is 1/2 C_S x $(0.028V_K)^2$ x repetition-rate. The average-power dissipation in the pull-down resistor is

$$\frac{(0.028V_R)^2}{R \times duty\ factor} + \frac{1}{2}C_S(0.028V_R)^2 \times repetition\ rate.$$

If the duty factor is small, as it often must be for a gridded-microwave tube, this simple topology is often the best.

An internal gun arc in the microwave tube can result in almost-instantaneous destruction of the transistor switch if it occurs between tube body and grid. Such an arc can also destroy the grid. So microwave tubes with gridded guns are designed to minimize the possibility of an internal arc path that terminates directly on the grid itself. The focus electrode, usually electrically connected to the cathode, is positioned within the gun and shaped so as to function as a lightning arrester, or arc shield, so that an internal arc will form between the body and focus electrode.

Figure 12-5 shows a grid pulser with the same basic topology, but it makes use of either a PNP bipolar or P-channel MOSFET output stage. This refinement gives the open-collector output connection a slightly more robust nature, thereby improving, at least marginally, the circuit's tolerance for microwave-tube internal arcs.

12.1.2 Active pull-up and pull-down grid pulser

When duty factor becomes too great or pulse fall time becomes very short, or both, pull-up and pull-down must both be active. The power supplies for *V1* and *V2* have similar roles as before. Only *I1* is duty-factor dependent, having two components: (peak-pulse grid current) x duty factor; and ($V1$ + $V2$) x C_S x repetition rate. Current *I2* is ($V1$ + $V2$) x C_S x repetition rate. Average-power dissipation in both transistor switches, assuming small voltage drop, is 1/2 C_S $(V1 + V2)^2$ x repetition rate, and average microwave-tube grid dissipation is *V1* x (peak-pulse grid current) x duty-factor.

As shown in Fig. 12-6, an additional housekeeping power supply, *V4* and *I4*, is required for the optic receiver and pre-driver for the pull-down switch, but self-capacitance is of no concern because it is not a bootstrap circuit and there is no common-mode pulse voltage.

12.1.3 Single-switch active pull-up and pull-down grid pulser

How can you have active pull-up and pull-down with only a single electronic switch and no cutoff bias power supply? The answer is shown in Fig. 12-7. It is yet another variant of the cascode-connected cathode drive that has already shown up here and there. The active pull-up is literally active pull-down because, in order to initiate pulse-current flow, it is the cathode of the microwave tube that will be pulled down by driving the NPN transistor (or N-channel MOSFET) into conduction. The grid of the microwave tube is connected to the positive terminal of a dc source whose output voltage is equal to the positive grid voltage required for nominal peak beam current (V_G = 0.018 V_{BM}). During the interpulse interval ,the voltage at the junction of the transistor collector (or drain) and microwave-tube cathode automatically rises to a value equal to the cathode-grid voltage required for beam cutoff (1% of V_{BM}) plus the grid power-supply voltage (1.8%

V_{BM}). This amounts to the same total of 2.8% V_{BM} that was switched by the previous grid modulators. When the transistor switch is turned on, current through it discharges the stray capacitance C_S, pulling the tube cathode voltage toward the V_{BM} bus. The rate at which this happens is determined by the total available switch current or conduction-state resistance. Note that the capacitance that must be charged and discharged is not the same as when the grid is pulsed. This includes the capacitance of the filament supply V_F, which should be designed for minimum value.

During the flat-top intrapulse interval, the transistor switch must conduct the entire beam current of the microwave tube. However, it is this current *after* the switch is turned off that provides the "active" pull-down, which is literally pull-up, by recharging the stray capacitance to its initial interpulse value. As discussed earlier, neither this current nor an equivalent series resistance is constant during the recharge, or trailing-edge, portion of the pulse, so that the waveform will be neither linear nor exponential.

The fault tolerance of this circuit connection depends upon the transient overvoltage capability of the transistor switch, which must have external transient suppression, shown as an MOV. If the microwave-tube electron gun has its grid shielded from internal arcing, the arc will terminate on a cathode-potential electrode, and the arc-current outlet is through either the transistor switch or the transient suppressor. If the circuit is to survive, the arc current had better flow through the transient suppressor, whose clamp voltage, even for short-duration residual overvoltage due to series-inductance effects, must be coordinated with the transistor safe-operating-area restrictions.

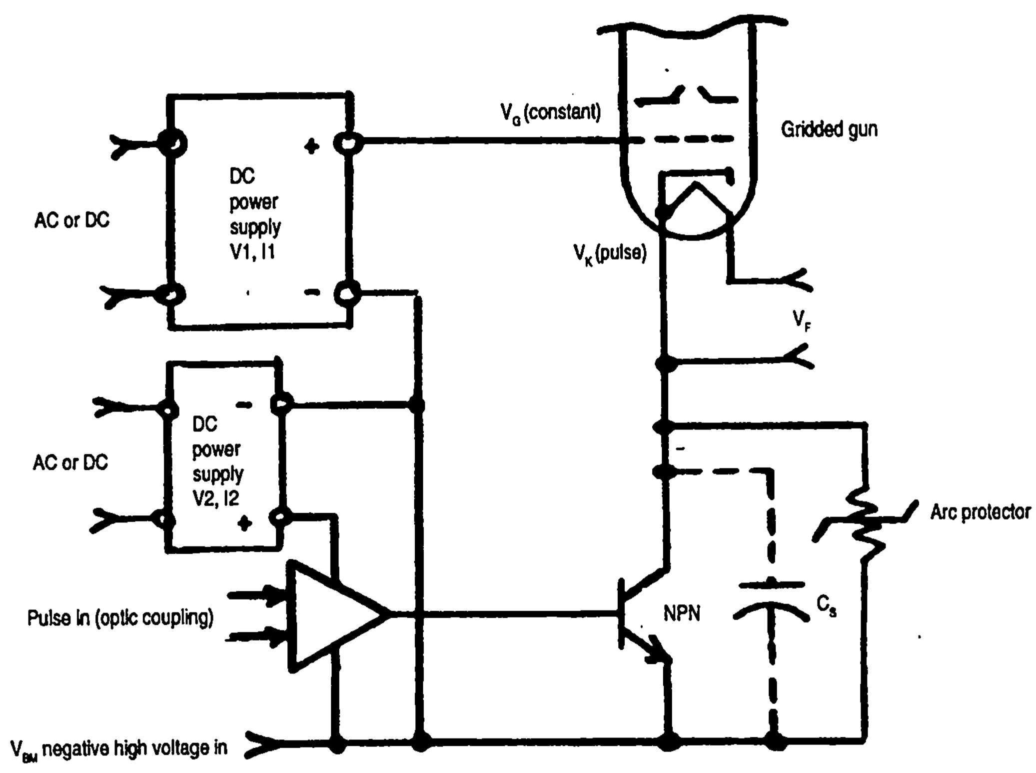


Figure 12-7. Another solid-state grid modulator with active pull-up and pull-down, but only one control switch.

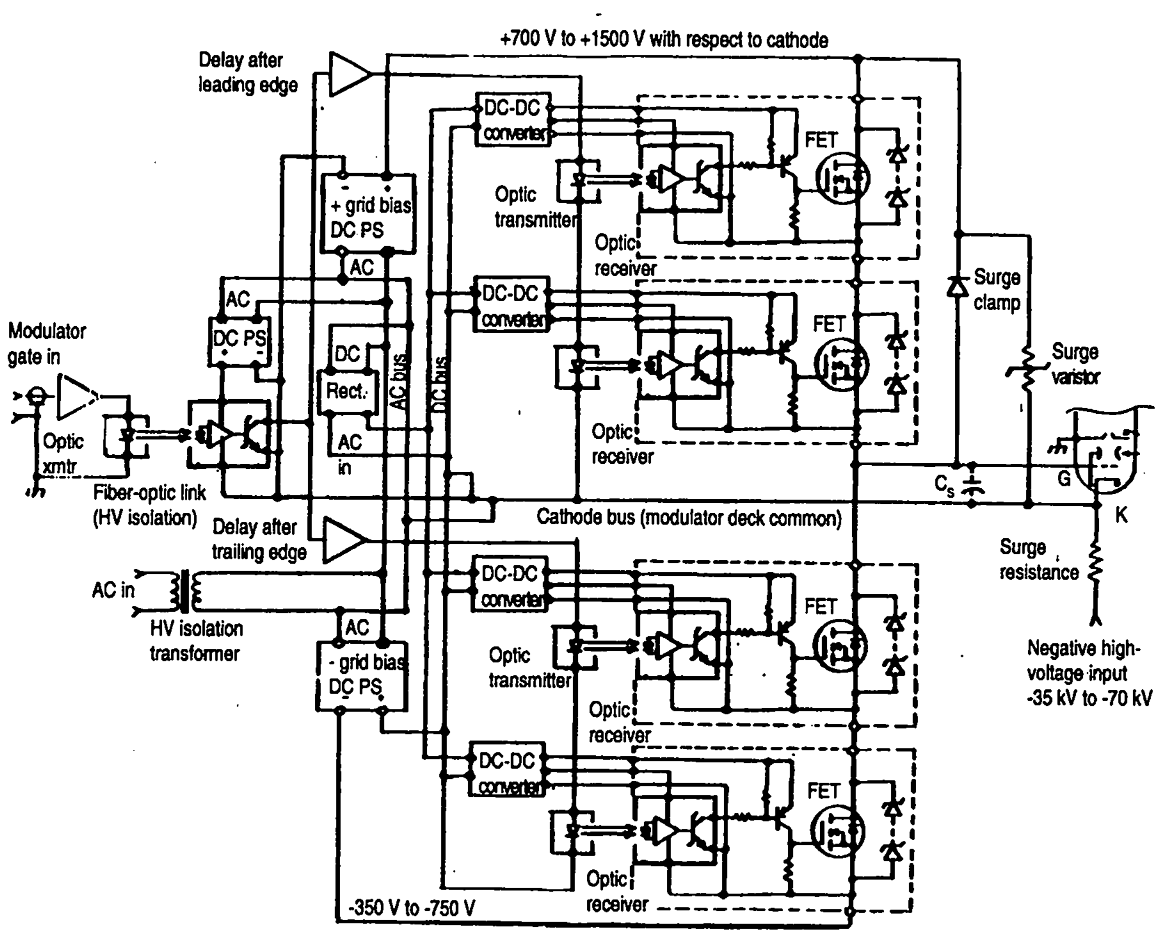


Figure 12-8. Cascaded solid-state modulator with active pull-up and pull-down for gridded guns operating at higher grid bases.

12.1.4 Cascaded solid-state active pull-up and pull down

Not all gridded-gun microwave tubes operate below 30-kV beam voltage. Those that don't may have grid bases that exceed the safe-operating conditions of single transistor switches. There is nothing, except circuit complexity and expense, that prevents the use of series-cascading transistor-switch modules to accommodate the switching voltage requirement of virtually any gridded electron gun. An example of such a stacked-module topology is shown in Fig. 12-8, which is but one of many practical implementations.

The circuit shown is expanded to only two modules each for the pull-up and pull-down segments of the pulser. A single optically coupled signal link is used to inject a low-level modulator gate to an electronics deck whose common plane is connected to the cathode of the microwave tube. In the absence of an optical signal, the quiescent, or interpulse, conditions are obtained. During this period, the current is driven through the series-connected optical transmitters associated with the pull-down modules, thus maintaining the FET switches in the conducting state. (Note that this is not a "fail-safe" implementation, because active pull-down throughout the entire interpulse, or stand-by, interval requires the presence of current in the optic-transmitter loop, fiber-optic continuity, and fiber-optic connection to each pull-down module optic receiver. It would be preferable if all modules in the complete pulser were identical because using modules with complementary logic would not be "fail-safe" during the active pull-up. However, pull-down resistance connected between microwave-tube grid and the negative-bias source would mitigate the situation, especially for the stand-by condition.)

When the leading edge of the modulator gate is received at the pulser gate, current in the optic-transmitter loop of the pull-down portion is terminated without intentional delay. The start of current in the optic-transmitter loop for the pull-up modules is intentionally delayed by a few microseconds to assure the recovery of the FET pull-down switches to their non-conducting states. This is done to preclude simultaneous conduction in pull-up and pull-down paths, or shoot-through.

At the end of the modulator gate, the process is reversed. Current in the pull-up optic-transmitter loop is terminated with no delay, while intentional delay is inserted in the start-up of current in the pull-down optic-transmitter loop. The circuit then toggles back and forth between full-conduction and full-cutoff grid conditions, using as many series-connected switch modules as safety-factor considerations dictate for grid base and switch-voltage. The classic problem of series-connected electronic switches is the lack of simultaneity in turn-on and turn-off, which results in unequal voltage-sharing during the leading-edge and trailing-edge transient conditions. This is why each transistor switch is shunted by a Zener-diode string whose transition voltage is coordinated with the voltage hold-off capability of its associated switch. The reluctance of a switch to either turn on or turn off manifests itself in a stair-step discontinuity in the rise or fall wave shape. It does not result in destructive transient overvoltage. Each switch module requires an isolated, low-capacitance source of housekeeping voltage and current located between isolated input and output terminals In Fig. 12-8 this is shown as a dc-dc converter. (Later, a type of converter that is nearly ideal for such service will be discussed.)

Fault tolerance for this type of pulser is greater than that for the single-switch pulser only because this modulator is used in situations where more grid-voltage swing is required, so it is inherently a higher-voltage device. On a per-unit voltage basis, however, its susceptibility is the same (higher grid voltage, higher beam voltage). Just as in the single-switch topology discussed above, a MOV surge-voltage clamp is used as the primary path for fault current resulting from a body-grid internal arc. The MOVs can be effective, but they add considerable load capacitance to the circuit. Where blazing speed is required, their loading effect must be considered at all points of the design. (They certainly cannot be tacked on as an afterthought without measurable performance penalty.) Spark gaps, which have much lower shunt capacitance, are often considered for fault protection. They have a shortcoming, too. Their internal plasma development has inertia, which means that the gap's sparkover voltage increases as the rate-of-rise of voltage increases. In any case, even the lowest-inertia spark gaps have a ratio of static voltage hold-off to transient-voltage breakdown of at least a 1.5:1. This means that for guaranteed protection the pulser must have at least a 1.5:1 voltage safety factor.

An alternative to direct shunt voltage-limiting is the surge diode connected between tube grid and the positive grid-voltage source, as shown. The only capacitance added is the reverse-biased diode capacitance. An internal arc to the tube grid, which will tend to pull the grid toward ground (forcefully), will forward-bias the diode as soon as the grid becomes more positive than the positive-source power-supply terminal, forcing fault current through the power supply (and reverse-biasing its internal rectifier diodes). If, however, the storage capacitance that shunts this supply is large enough and the total charge transfer from the beam power supply is not too large, the fault will be absorbed without damage to anything. The power-supply terminal voltage will rise by an amount $\Delta V = \Delta Q/C$, where ΔQ is the charge transfer from the beam supply energy-storage capacitor and C is the capacitance shunting the pulser positive-bias power supply. The peak current through the surge diode will be limited by the surge resistance in series with the microwave-tube cathode. The surge diode must be rated for at least this much peak current. In addition, the diode must be rated to handle the action integral

$$\int i^2 dt\,.$$

This integral, having the dimension of ampere^2-second, or joules/ohm (which are the same), can be evaluated either by doing the time integral of the exponentially decaying fault current or by finding the total energy stored in the beam-supply system and dividing by the total series surge resistance.

12.1.5 Hybrid grid pulser

Even though today's "technologically correct" design strategy is to use solid-state components wherever possible—and the grid pulser is the only high-level modulator where their exclusive use is even remotely practicable—there are still applications where a combination of solid-state and vacuum-tube devices can yield an optimum design solution. For instance, there is nothing simple or inherently inexpensive about the cascaded-module pulser just discussed—especially as the module count becomes large. There are small, planar-geometry triodes such as the Y-540 that can hold off kilovolts (6 kV for the Y-540) and deliver short-pulse currents that are more than adequate for microwave-tube grid interception. Therefore, they could supplant a number of transistor switch modules.

The circuit of Fig. 12-9, an obvious adaptation of a previous modulating-anode pulser topology, shows such a hybrid. The pull-up, or "on," portion of the circuit uses a triode vacuum-tube high-voltage switch in a cascode connection with two MOSFET switches, which are optimized for separate functions. The upper FET, designated "switch," turns the triode on and off in response to the optically coupled modulator-gate timing signal. The lower FET, designated "clamp," is connected in a gate-catcher circuit in which its gate electrode is diode-coupled to an adjustable dc power supply that permits grid-pulse amplitude control from a ground-referenced operator adjustment. The anode supply for the pull-up triode has an output voltage that is approximately 1 kV greater than the maximum output-pulse amplitude, up to the anode-voltage capability of the triode. The low-current amplitude-control supply has a minimum output volt-

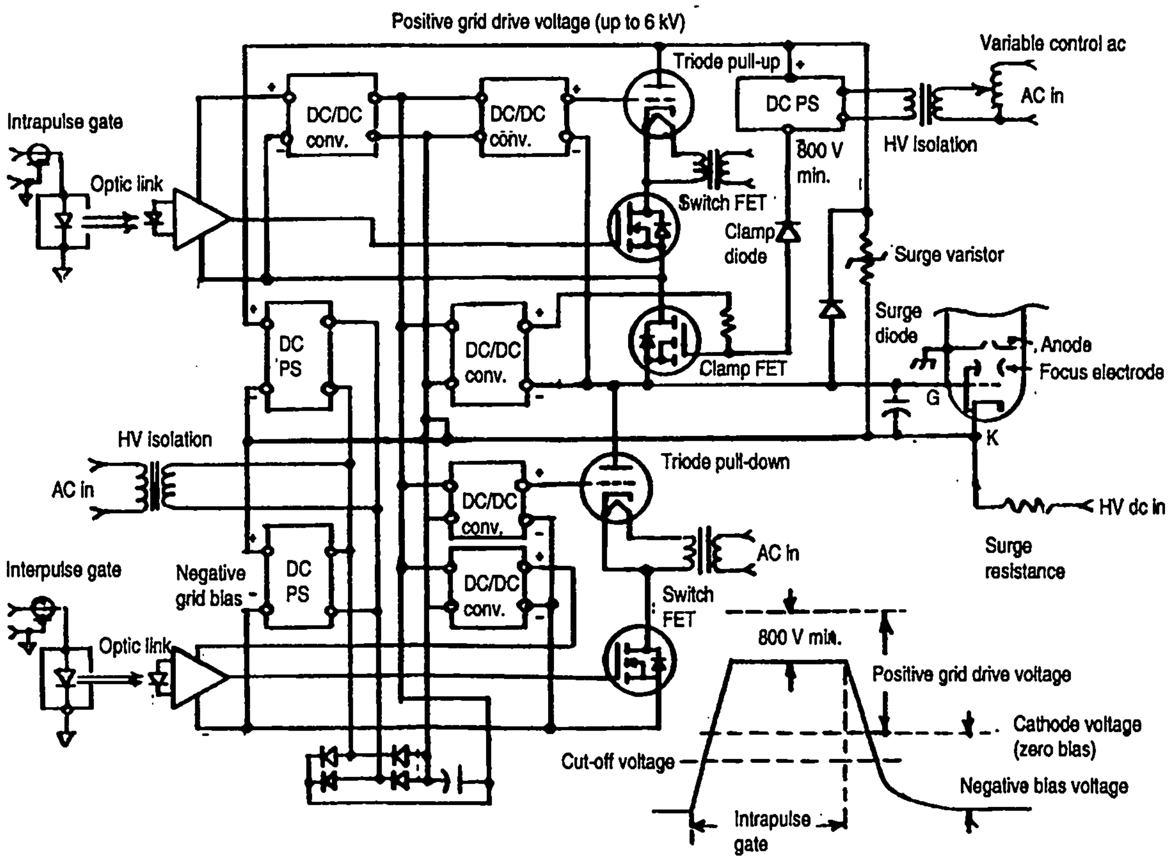


Figure 12-9. Hybrid solid-state/vacuum-tube grid modulator for high-voltage guns.

age of 1 kV because it bucks the anode supply. (The greater the amplitude-control supply voltage, the smaller the output-pulse amplitude.)

As in the comparable modulating-anode pulser topology, the timing-switch FET and the clamp-switch FET are shown as separate, isolated entities that are connected in series as a logical AND gate. With some compromise in overall performance (particularly affecting rise-and-fall times and reference-supply average current), the "clamp" FET can be also used to perform the timing function by connecting a small-signal NPN transistor switch between the FET gate and source. The NPN transistor conducts during the interpulse interval, removing FET gate drive, and is driven off for the duration of the modulator-gate signal, allowing the FET gate to rise until the clamp circuit pulls it back down again when the desired pulse amplitude has been reached (or slightly exceeded). Although this design results in a simpler topology, it is not "fail-safe," because the low-level transistor must conduct in order to prevent attempted pull-up. If it fails in the short-circuit mode, nothing potentially hazardous will occur. If, on the other hand, it is simply removed from the circuit, continuous pull-up current will flow. Unless the situation is sensed and corrected by status-monitor circuitry, it could damage the microwave tube, the modulator, or both.

12.1.6 *"Inductive-kick" single-switch grid pulser*

One answer to the question "what is the voltage across an inductor?" is "whatever voltage is required to maintain constant current flow through it." The operation of the high-voltage grid pulser shown in Fig. 12-10 is based on that answer. The figure's single electronic switch, which is made up of a number of series-connected transistors, conducts throughout the interpulse interval. In so doing, it connects the grids of the multiple microwave tubes (eight, in its most significant actual applications) to a -900-V grid-bias power supply. At the same time it conducts current from a 600-V, 300-mA constant-current power supply through a 16-H inductor.

When the trigger-amplifier/light-receiver receives an optically coupled signal corresponding to the beginning of a pulse, the electronic-switch transistors are turned off, opening the path for the 300-mA inductor current. With no place for the current to go (yet), the voltage rapidly builds up across the inductor, such that the voltage at the junction of the inductor and the electronic switch increases in the positive-going direction. This voltage becomes the leading edge of an output-voltage pulse that keeps rising until it exceeds the +2-kV output voltage of the clamp power supply, the conduction-voltage drop of the seven series-connected Zener diodes, and the bias-blocking diode in series with the clamp

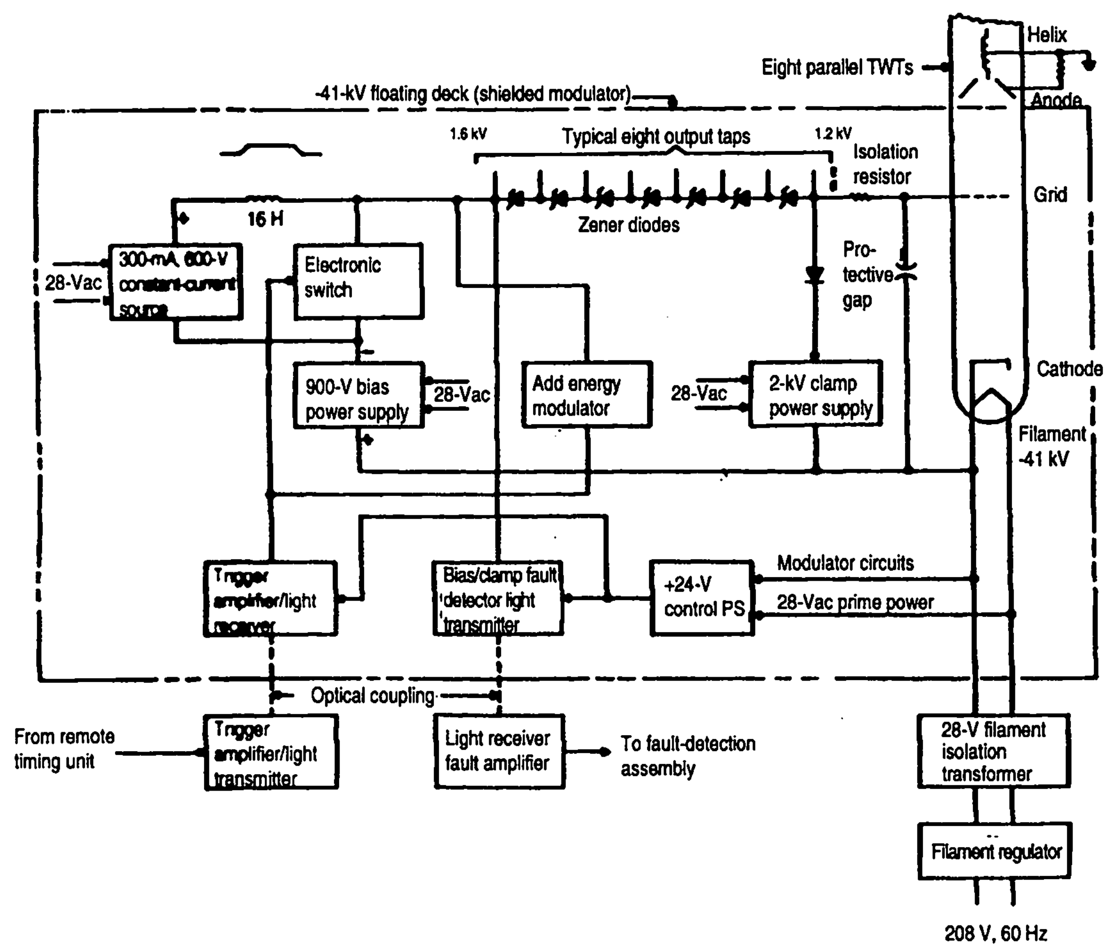


Figure 12-10. Inductive ring-up grid modulator for high-voltage gridded gun.

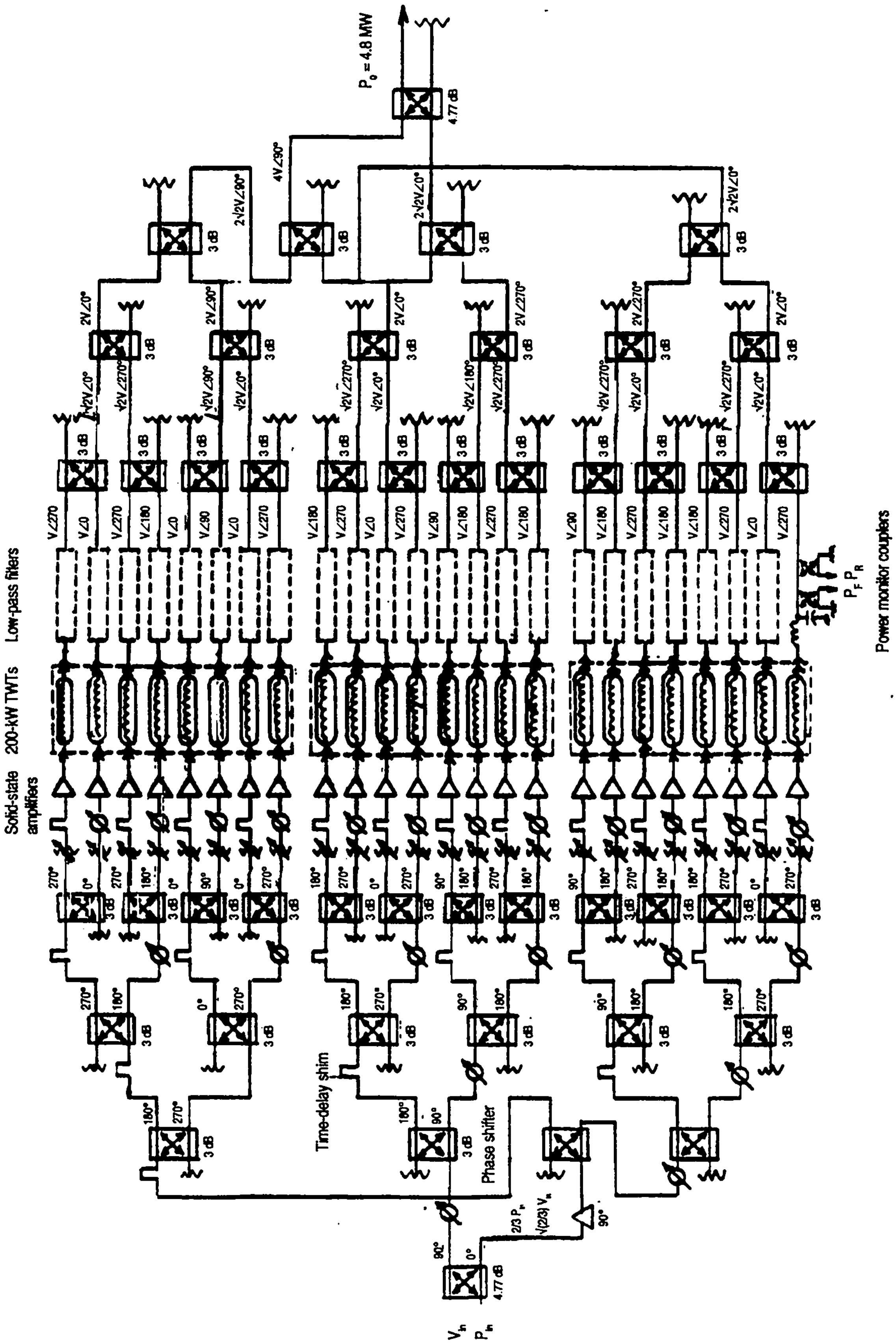


Figure 12-11. The 24-channel, single-output transmitter of the ALTAIR UHF radar.

supply. With a path thus re-established for the inductor current, the voltage no longer rises, and a flat-top pulse continues until either the volt-time product of the inductor core is exceeded or the electronic switch is turned back on, terminating the pulse and restoring the original negative-bias conditions at the pulser output. The inductor is large enough and has sufficient internal distributed capacitance so that the rate-of-rise of voltage is not quite as fast as desired. To increase the speed, additional current in the form of a sharp pulse is coupled to the output node from the add-energy modulator.

To satisfy any differing grid-drive requirements of the electron guns connected to it, the pulser has eight separate output voltages at 50-V intervals over a 400-V range. These can be tapped at the junction points between Zener-diodes. Note that the filaments of the electron guns operate from 28 Vac, coupled to the -41-kVdc floating deck through a filament isolation transformer, which also provides the prime-power input to the dc power supplies used by the pulser.

12.2 Transmitter applications of grid pulsers

As mentioned earlier, gridded-gun microwave tubes—primarily the helix-type TWTs— whose peak power is usually less than 5 kW are almost universally used in systems demanding a high degree of waveform versatility. One use for such a transmitter is as a radar jammer for electronic-countermeasures applications. The pulse trains produced by this transmitter may have pulse recurrence intervals of only a few microseconds and repetition rates up to 100,000 pps, while their flat-top pulse durations may be on the order of fractions of a microsecond. Pulse Doppler radar systems use the frequency domain in much the same way that long-range search radars use the time domain. To achieve low Doppler ambiguity, a transmitter must have a high repetition rate (for greater frequency deviation between PRF lines in the frequency domain), whereas to achieve low range ambiguity, a transmitter may have more time between successive pulses and a low repetition rate.

There are literally hundreds of designs of small, solid-state, high-speed grid modulators for the host of compact, light-weight transmitters used in military aircraft. On the other hand, there are only a handful of truly high-power transmitters that use RF power amplifiers with gridded electron guns. And in this group are some of the most powerful transmitters in the world.

The last class of grid pulser to be discussed is the one used in the most impressive of these transmitters: the ALTAIR (ARPA Long-Range Tracking and Instrumentation Radar). The ALTAIR UHF transmitter uses 24 TWTs rated at 200-kW peak power, 6% duty factor, and 2-ms pulse duration. The tubes are organized into three groups of eight tubes each. The transmitter is of interest not only because it uses three of the grid modulators described above, but because it is a multi-channel (24) transmitter that starts with a common RF drive signal in one RF transmission line and ends up with the outputs of all of the TWTs combined in a single-output waveguide. (This waveguide, in turn, feeds a 150-ft-diameter parabolic-reflector antenna system through a multi-mode tracking feed and Cassegrain optics.) Why is the number 24 of particular interest? Because it is not part of a binary progression, like 2, 4, 8, 16, and 32.

Because the transmitter's microwave-tube organization does not follow a binary pattern, it follows that power splitting and power combination cannot be accomplished by relying exclusively on networks comprising 3-dB couplers. In fact, as shown in Fig. 12-11, which depicts the simplified RF circuit of the 24-channel transmitter, the very first power-splitter hybrid has a coupling factor of 4.77 dB. This value was chosen because it produces a 2:1 power split, with 1/3 of the input power, P_{in}, going to the quadrature port, and 2/3 P_{in} to the in-phase port. After that, the remaining breakdown of the input signal into 24 equi-amplitude channels is accomplished with 3-dB hybrid junctions. The same hybrid junctions recombine high-power outputs of each tube of a group into one channel that is the sum of the group. A second similar junction sums the output of the other two tube groups. The last step of combining is accomplished in another 4.77-dB hybrid junction implemented in WR-2100 waveguide. The progression of 90°-phase-shift increments produced by the quadrature hybrids leads to the final in-phase condition of all of the voltage vectors at the output. This progression is listed in the figure as are the voltage amplitudes throughout the combiner tree. The final power output of the system can be determined by the output voltage, V_0, as expressed in the following equation:

$$V_0 = 4V\sqrt{\frac{2}{3}} + 2\sqrt{2}V\sqrt{\frac{1}{3}} = 6\sqrt{\frac{2}{3}}V.$$

Power output, P_0, can be defined as

$$P_0 = kV_0^2 = kV^2 \times 36 \times \frac{2}{3} = 24kV^2,$$

where kV^2 is the TWT output power of 200 kW. Final output power is 24 times 200 kW, or 4.8 MW.

Just as there can be three-cylinder engines, there can be non-binary combiners. Yet, how do we determine the required coupling value for the needed hybrid junctions after all of the binary-progression numbers have been factored out? The quadrature-hybrid criterion for zero waster-load power, shown in Fig. 12-12, can supply the answer.

Assume that two arbitrary input voltages, V_1 and V_2, already have the requisite 90° phase differential at the two input ports of a quadrature hybrid, which has a voltage coupling factor of A such that the voltage coupled from an input port to the quadrature output port is 1 divided by A, as shown. If all of the power represented by the two input voltage vectors is to combine at the desired output port, then none of it can be coupled to the opposite, or waster-load port. The two voltage components incident on the waster load are already in phase opposition. If they are of equal amplitude, they will cancel out. To be equal,

$$V_1\sqrt{1 - \frac{1}{A^2}} = \frac{V_2}{A},$$

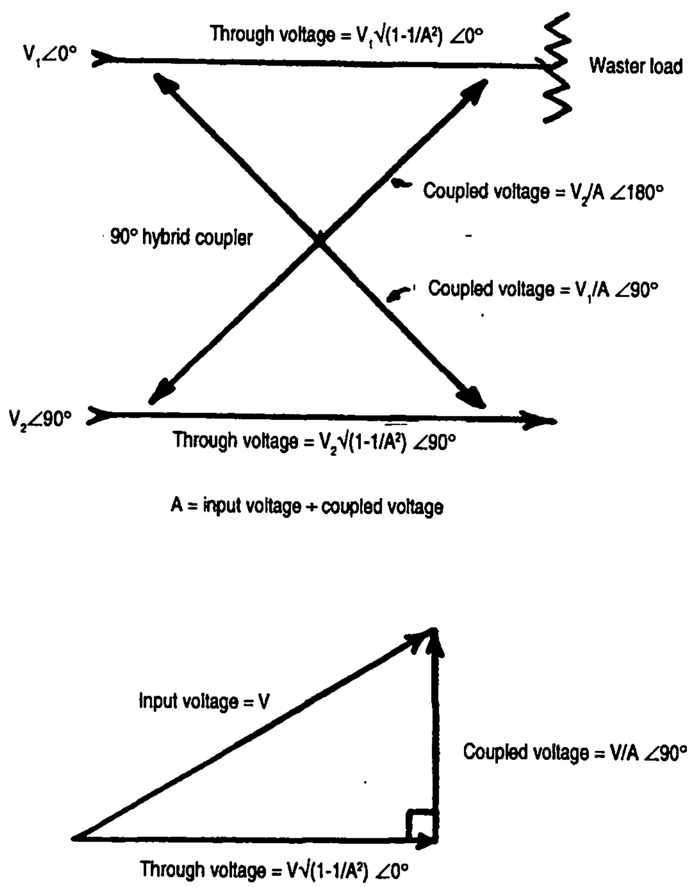


Figure 12-12. General relationships in quadrature combiner-coupler.

which gives

$$A = \sqrt{\frac{V_2^2}{V_1^2} + 1} \ .$$

$V2^2$ and $V1^2$ are proportional to the corresponding power levels P_2 and P_1, so A can be expressed as

$$\sqrt{\frac{P_2}{P_1} + 1} \ .$$

Because A is a voltage ratio (presumably in a constant-impedance system), the coupling factor can also be expressed as

$$20 \log \sqrt{\frac{P_2}{P_1} + 1} \text{ dB} \ ,$$

or

$$10\log\left(\frac{P_2}{P_1}+1\right)\ \mathrm{dB}\ .$$

For the case where P_2 is twice P_1, the coupling factor will be

$$10\log(2+1)=4.77\ \mathrm{dB}\ .$$

For the more common equal-power case, $P_1 = P_2$, the coupling factor is

$$10\log(1+1)=3\ \mathrm{dB}\ .$$