16. Electronic Voltage Regulators

Some microwave tubes require more precise regulation of beam voltage than can be obtained from the power supplies and regulators already discussed. This is especially true if the nature of the transmitter output waveform is continuously changing, which is often the case with modern sophisticated radar systems. High-frequency, switch-mode power supplies, which will be described later, can outperform the most precise variable-voltage device, which uses conduction-angle controlled rectifiers simply because of their usually higher self-generated operating frequency.

No regulating system can outperform the electronic variable-resistance voltage regulator, either in precision or in transient response. In the case where intrapulse voltage regulation is required, which is usual for a long-pulse system where the intrapulse voltage droop obtainable from a capacitor bank of practical size is excessive, there is virtually no alternative to the electronic variable-resistance regulator.

The variable resistance in such regulators, which are also sometimes called "linear" regulators, is supplied by a "series," or "pass," device that is connected in series with the high-voltage loop. This device must be "full-control," such as the ones previously discussed in regard to hard-tube pulse modulators. In large-power, high-voltage systems, the series device is customarily a large-power vacuum tube. The tetrode is often chosen for this application because adequate anode, or load, current can be obtained with a grid-cathode voltage that never needs to be positive. This situation greatly simplifies the design of the error amplifier that provides the beam voltage. (The power MOSFET has characteristics that are nearly ideal for such service, but its use is limited to low- or medium-power situations because it is considerably more difficult to harden against fault conditions in the high-voltage loop.)

16.1 Voltage regulators that handle system average current

Figure 16-1 shows a typical medium-power radar transmitter's high-voltage system that uses a variable-resistance electronic regulator. The peak-pulse current is supplied from a capacitor bank so large that intrapulse regulation is not required. The electronically controllable variable resistance in series with the high-voltage loop of the microwave tube comprises the parallel-connected tetrode vacuum tubes *V1-V3* and is connected between the high-voltage power supply and the capacitor bank. The microwave tube used in this system is a 35-GHz coupled-cavity TWT that produces 30-kW peak-power output. The optimum beam voltage is between 47 and 48 kVdc, but it must be precisely controlled and free from ripple-frequency components because of the high phase-pushing factor for this type of high-gain, broad-band tube. In addition, system specifications required that phase-modulation sidebands be -50 dBc.

To give some idea of what this means in terms of regulator performance, consider that for each phase-modulation component to be 50 dB below the carrier, its peak voltage must be 0.003 of the carrier. Assuming that the ripple lines

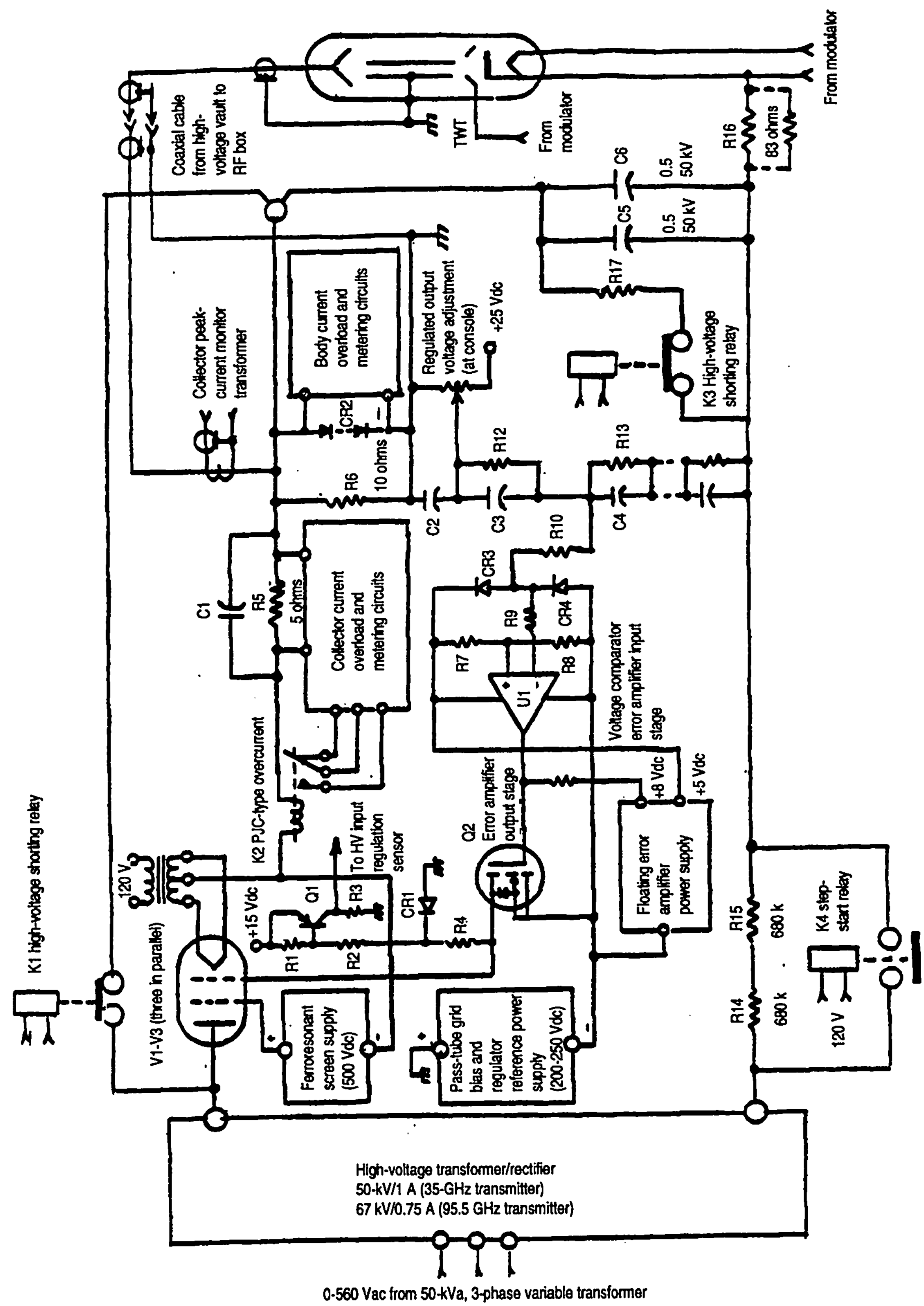

Figure 16-1. Example of electronic voltage regulator that handles microwave-tube average current.

are discrete and harmonically related to the 60-Hz primary-power frequency, we can expect there to be paired sidebands, with each sideband of a pair being above and below the carrier by the same frequency difference. The voltages of the two sidebands will add in quadrature with the carrier voltage so that the sine or tangent of the peak angular modulation will be 2 x 0.003, or 0.006. The corresponding phase angle is 0.36°. The phase-pushing factor of a TWT of this type is typically 50°/percent of beam-voltage change. The allowable peak ripple voltage, therefore, is 0.36° divided by 50°/percent, or 0.007% peak ripple voltage, which is not much.

Note that the anode-cathode circuit of the regulator series tubes is in the positive return of the high-voltage power supply. The cathode of the regulator series tube is connected to the collector lead of the TWT, through which passes the great majority of the TWT cathode current. The collector lead is connected to ground through resistor *R6*. The body of the TWT is also connected to ground. So are its waveguide input and output connections, which are an integral part of the TWT body structure. The voltage across *R6* is proportional to the current intercepted by the TWT body, which is normally a small fraction of the total beam current. However, in the case of an internal TWT gun arc, all of the current is body current, which is limited only by *R16* (83 ohms) to a value of 50 kV/83 ohms, or 600 A. The series-connected diodes, shown as *CR2*, shunt the fault current around *R6* while maintaining a total forward diode-drop great enough not to distort the voltage corresponding to normal body current.

The upshot of all of this is that the cathode of the regulator tube is very near ground potential, which is also the common point for the sample of the TWT cathode voltage obtained from the compensated voltage divider comprising *R13/C4* and *R12/C3*. Of course, the criterion for frequency compensation is that the product of *R13* and *C4* equals the product of *R12* and *C3*. The error amplifier topology can be relatively straightforward because there is no common-mode voltage difference to overcome. In this design, however, there is at least one unique feature. The common terminal of the error amplifier has been deliberately offset from ground by an amount equal to the negative voltage required by the error-amplifier output stage to assure current cutoff in the regulator series tube, which in this case is a minimum of -200 Vdc. This power supply, therefore, is a precision low-ripple dc source.

By offsetting the error-amplifier common terminal, the attenuation factor of the sample-voltage divider can be made many times smaller than with a grounded error amplifier. This offset is less important for the dc performance than it is for ripple rejection because the sampled ripple components are many times greater. The attenuation factor in this case is 50,000 V/200 V, or 250, whereas in a more conventional regulator the attenuation factor might be 50,000 V/5 V, or 10,000, if a 5-V reference is used for the error amplifier. The actual reference voltage used in this design is 2.5 V, but it floats atop the 200-V grid bias supply. The input stage of the error amplifier is a voltage comparator, *U1*, whose signal gate is protected by the resistor/diode network, *R8*, *R10*, *CR3*, and *CR4*. The error-amplifier output stage is a high-voltage MOSFET, *Q2*, which produces variable grid voltage to the *V1-V3* tetrode by varying the current through *R4* in response to the difference in voltage between the positive and negative gates of *U1*. The tetrode grid variable voltage produces the required variable resistance in the high-voltage loop to precisely regulate the TWT cathode voltage.

The TWT beam current is in the form of 6-A peak-current pulses that do not exceed 50 μs in duration at duty factors up to 10%. The pulse current is obtained from a relatively small capacitor bank, *C5* and *C6*, totaling 1 μF. The intrapulse voltage droop is acceptably small. The voltage regulator, therefore, needs to pass only the maximum average beam current, 0.6 A, while maintaining constant a pulse-to-pulse voltage across the capacitor bank.

Like many high-voltage dc conditioning systems that have electronic regulators, this one has an ac variable-voltage device as well, a motor-driven variable autotransformer, that permits the dc output from the transformer/rectifier to be gradually applied. Or it can also be reduced to a lower value (including zero) for diagnostic purposes. In recovering from a high-voltage shutdown resulting from an internal TWT arc, the power supply can be "slapped on" with full ac input. The peak capacitor charging current and the rate-of-rise of TWT cathode voltage is determined by the "step-start" resistance, *R14* and *R16*. When the operating voltage is reached, this resistance is short-circuited by the high-voltage relay *K4*, and the operation of the electronic regulator is sensed by the flow of current through *R4*, which is in series with *R1* and *R2* as well. This current turns on transistor *Q1*, signaling the control circuits that the regulator is functioning.

However, before the regulator begins to function at dc input voltages lower than the normal operating value, the tetrode series tubes are zero-biased. There is also no anode-cathode voltage, which means that the screen grid will attempt to play the role of surrogate anode in the collection of current from the cathode. Screen current must be externally limited to prevent damage to the screen grids. A most effective way of doing this is to operate the screen supply from the output of a ferroresonant constant-voltage transformer, which is matched as closely as possible in VA rating to the normal full-load requirements of the screen supply. When such a transformer is short-circuited, its output current is typically only one-and-a-half to two times as great as the normal full-load current, and perhaps surprisingly it runs cooler under this condition than under full load. (The worst case is when it is open-circuited.)

A less expensive way of limiting screen current is to use ac capacitance in series with either the screen-supply ac input or ahead of the rectifier. The effect on screen-voltage regulation of capacitive reactance that is capable of limiting short-circuit current to two times full load can be quite modest. The ferroresonant transformer has the obvious technical advantage that it not only limits short-circuit current but it regulates source voltage in the normal operating range.

16.2 Voltage regulators that handle system peak current

When the transmitter waveform dynamics are such that an energy-storage capacitor bank would be impractically large in order to limit intrapulse voltage droop to acceptable levels, there is virtually no alternative to the pulse-current voltage regulator. The Haystack Hill LRIR is a classic example of such a transmitter (see Fig. 16-2). Its four parallel TWT high-power amplifiers draw a total intrapulse current of 40 A for pulse durations up to 50 ms. This is a per-pulse charge transfer of 2 C, which is large enough to completely drain capacitor banks

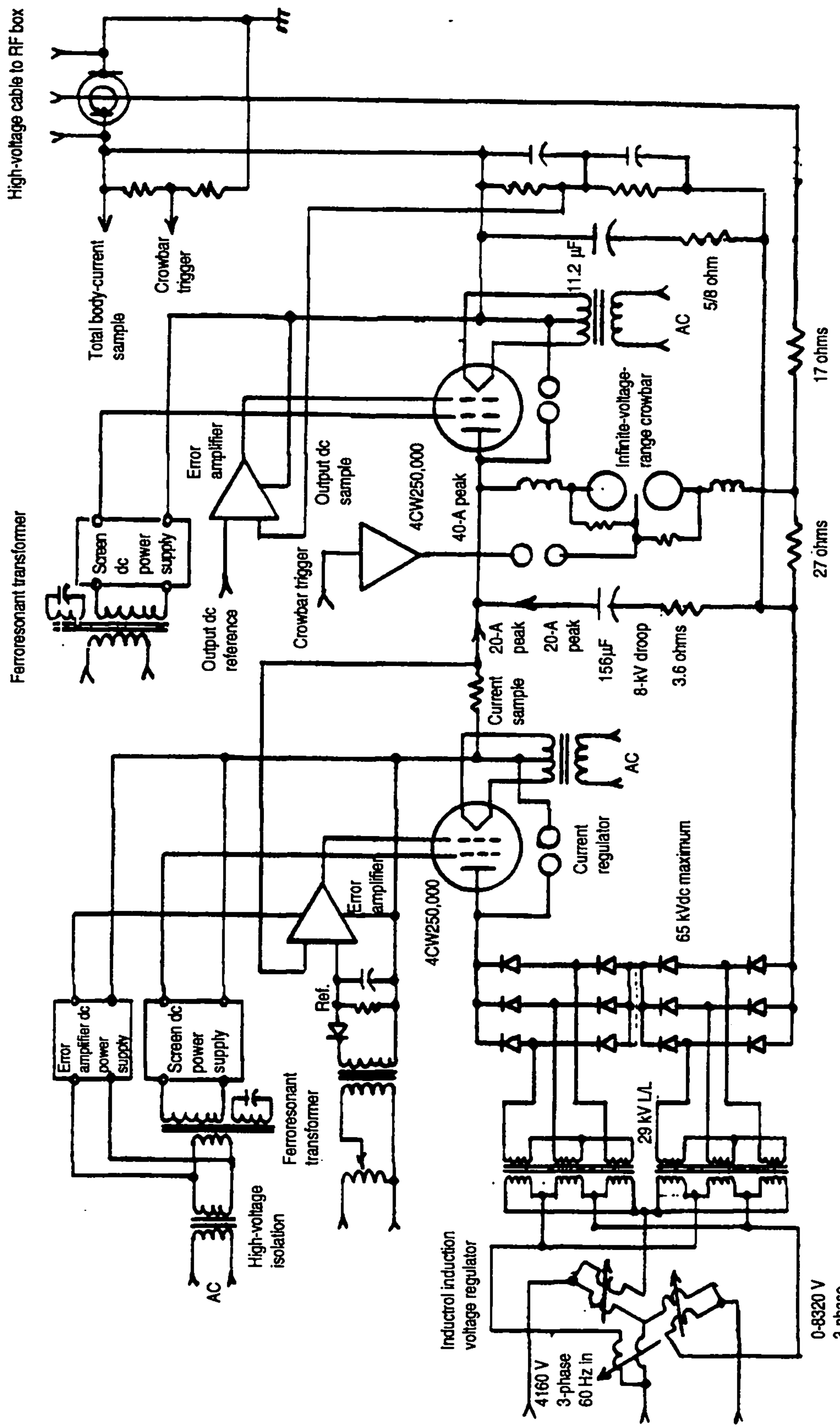

Figure 16-2. Example of high-voltage system having an electronic voltage regulator handling microwave-tube peak-pulse current and an electronic regulator handling average current.

of common size. Even though the LRIR transmitter main capacitor bank of 156 µF could hardly be considered commonplace, the removal of 2 C from it would cause its voltage to droop by almost 13 kV. However, the dc power supply is capable of putting out an average current of 20 A, which it did whenever the transmitter was operated at its maximum duty factor of 50%. Therefore, the entire 40-A peak-pulse current did not have to come from the capacitor bank, because half of it would come from the dc power supply. Even so, the capacitor-bank droop was over 6 kV per pulse.

A voltage regulator was provided that would remove the 6-kV voltage change while passing the full 40-A peak-beam current. The output voltage applied to the cathodes of the TWTs remained within approximately 30 V of the desired operating voltage, approximately 47 kV. (Tetrodes are inherently good source-voltage regulators because of their high incremental anode resistance or constant-current nature.) The anode voltage of the 4CW250,000 tetrode series regulator tube started each pulse 6 kV higher than at the end. The total source voltage at the beginning of each pulse was 47 kV + 6 kV + 3 kV for a total of 56 kV. This value takes into consideration the headroom required for the tube to pass 40-A peak without excessive screen-grid current.

Apart from the obvious requirement that the series-connected device in a pulse-current regulator must be capable of passing the full-peak-pulse current of the microwave tube(s)—a conventional voltage regulator must merely pass the time-averaged value of the total beam current—the major difference between the two types of regulators is transient response. Whereas the series tube of a conventional regulator in the linear mode is in a state of almost-continuous conduction—the degree of which changes only slowly in response to changes in transmitter duty factor—the conduction of a pulse-current regulator must change from full-off to full-on during each pulse. In addition, it is precisely the intrapulse grid-cathode voltage of the regulator tube that keeps the microwave-tube beam voltage constant. The error amplifier, therefore, must be both precise and swift.

Fortunately, the timing of the load-current pulses is not unknown to the designer. So, often the slew rate of the error amplifier is assisted by means of an open-loop synchronous switch that receives its timing information from the same source as the pulse modulator. Other than that, there is no reason why the error-amplifier topology of the pulse-current regulator would be significantly different than that shown in Fig. 16-1. In that figure, the error-amplifier output resistor *R4* and the current-handling capability of the output stage, *Q2*, must be sized so that the input capacitance of the regulator tube will be charged and discharged so rapidly that the buffer capacitor bank, which is at the output of the voltage regulator, is not excessively discharged before the regulator tube can take over the delivery of the full-load current. Needless to say, due to the higher gain and speed and lower internal impedance of the pulse-current regulator, this design is more difficult to stabilize against oscillatory instability than a conventional regulator. It is also essential to provide for high-frequency bypassing of the filament leads of a directly heated cathode tube, which most high-power tetrodes tend to be. (Although this feature is not shown in Fig. 16-2.)

In addition to the electronic voltage regulator in the LRIR transmitter, some means was also required to keep the dc power supply from being virtually short-circuited by the linearly discharging capacitor bank. In which case it would attempt to supply the entire 40-A peak current by itself. The solution, as shown, was another electronic current regulator set to pass no more than 20 A, which it did continuously at a 50%-duty factor. It required anode-voltage headroom of approximately 2 kV to pass 20 A, so the total dc-power-supply output voltage was 58 kV. The anode drop of the current regulator increased throughout each pulse as the voltage-regulator anode-drop and capacitor-bank voltage both decreased.

Unlike the voltage regulator, the cathode of the current regulator was not at virtual ground potential. All of the circuits referenced to its cathode had to be isolated for at least 10 kV and had to tolerate the same rate-of-change of voltage as the main capacitor bank, which was I/C = 20 A/156 μF, or approximately 105 V/s. The error-amplifier inputs, however, required no elaborate level-shifting or common-mode rejection, because circuit current could be sampled by means of a series resistance wherever it was most convenient. As shown, this point was in series with the cathode of the current-regulator tube.

Not all systems that require pulse-current voltage regulators also need current regulators. If capacitor-bank droop can be maintained at 1-2% of TWT beam voltage, a pulse-current regulator will not be required. If the actual capacitor-bank droop is only 1-2 kV or so, it is unlikely that the dc power supply will hog the pulse current. A dc inductor in series with the power supply can be made large enough to keep its output current constant. In so doing, however, the inductor can also store a great deal of energy that must be dealt with one way or another during fault conditions. Often this is a difficult problem.

16.3 Voltage regulators that do not handle the full microwave-tube beam current

The critical voltage for all microwave tubes is the voltage between cathode and ground because the RF circuit for all practical tubes is grounded. The only microwave-tube current component that must flow in ground is that which is intercepted by the body because the electron beam has been imperfectly focused. If the collector of the microwave tube is isolated or insulated from ground, then the collector current can be conducted in a path that is separate from the body-current path. (Microwave tubes, of course, can have multiple collectors, in which case there will be multiple collector-current paths.) Therefore, the series tube of a voltage regulator that is regulating the voltage between cathode and ground needs to handle only the current intercepted by the body of the microwave tube, which in a well-designed tube will be less than 10% of the total beam current.

A good example of such a regulator is shown in Fig. 16-3, which is a simplified schematic diagram of the high-voltage circuits for high-power RF amplifier groups. Each group has eight high-power (200-kW peak power) TWTs operating in parallel with respect to the high-voltage system. The nominal peak-pulse cathode current is 16 A for each TWT, 128 A for all of them. A pulse-current linear regulator for this much current would be awesome, indeed. So would any capacitor bank used to replace it because its maximum pulse duration is 2000 μs.

However, the TWT—which is specifically designed for long life, a high degree of phase linearity, and high efficiency—does not have one isolated collector but

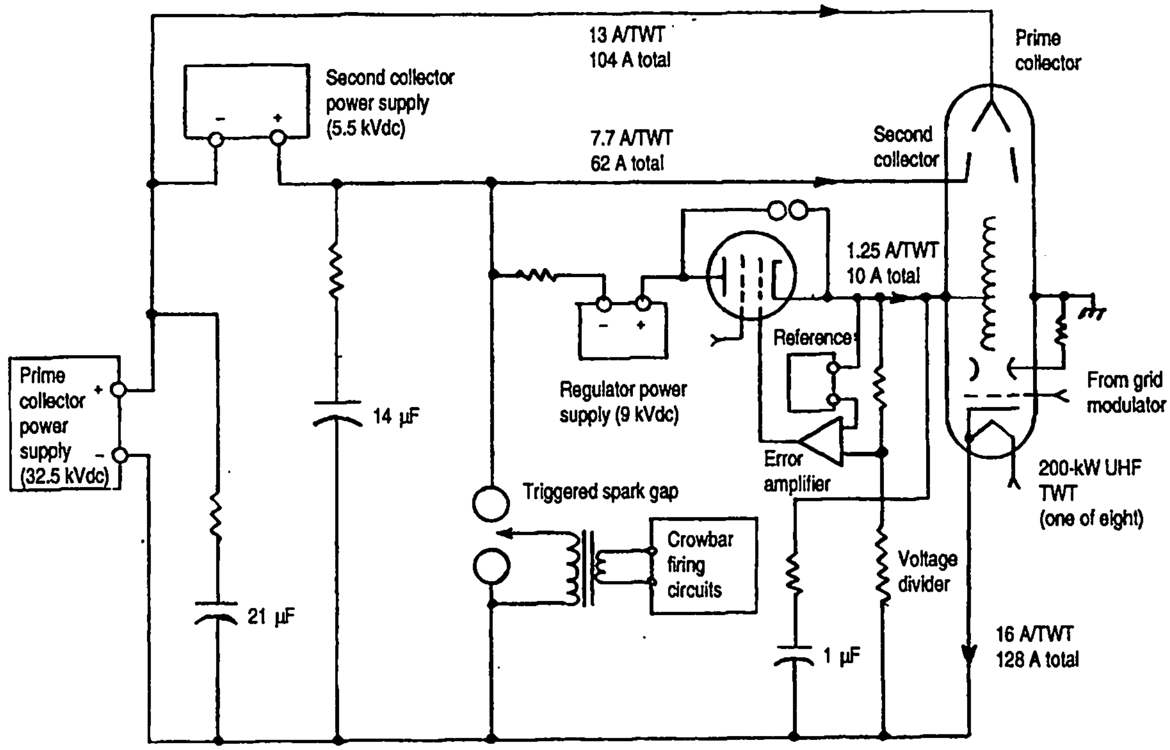

Figure 16-3. Electronic linear voltage regulator that handles only TWT body current.

two, both of which are depressed in voltage with respect to the grounded helix-type circuit. Three dc power supplies are required: one at 32 kVdc, which is connected between the cathode and prime collector; a second at 5.5 kVdc, which is connected between the prime and second collectors; and a third at 9 kVdc, which is connected between the second collector and the anode of the tetrode series tube of the linear voltage regulator. The resistance of the voltage regulator is electronically varied to maintain the voltage between cathode and ground of the TWT at its nominal value of 41 kVdc. However, the current that passes through the regulator tube is only about 8% of the total peak-cathode current, or 10 A. A modest-sized tetrode (4CX5000 class) is adequate for this purpose.

16.4 Regulators that must also be modulators

The regulator circuits discussed so far are used in transmitters having separate pulse modulators, either modulating-anode or control-grid pulsers. But such regulators are also used in transmitters that have microwave tubes with diode-type guns. In order to pulse-modulate them, either a line-type pulser or a hard-tube switch is required that is capable of handling the full beam current and voltage of the diode electron gun. If active pulse-top regulation is a requirement, the line-type pulser is not an option. However, we have already discussed voltage regulators that must handle the full peak-pulse current of the microwave tube. If, in addition, the series regulator tube is capable of holding-off the full diode beam voltage as well, the roles of pulse modulation and pulse-top regulation can be integrated into the same circuit.

Figure 16-4 shows a simplified schematic of a practical pulse-current voltage regulator in which the regulator tube is in series with the positive return bus of

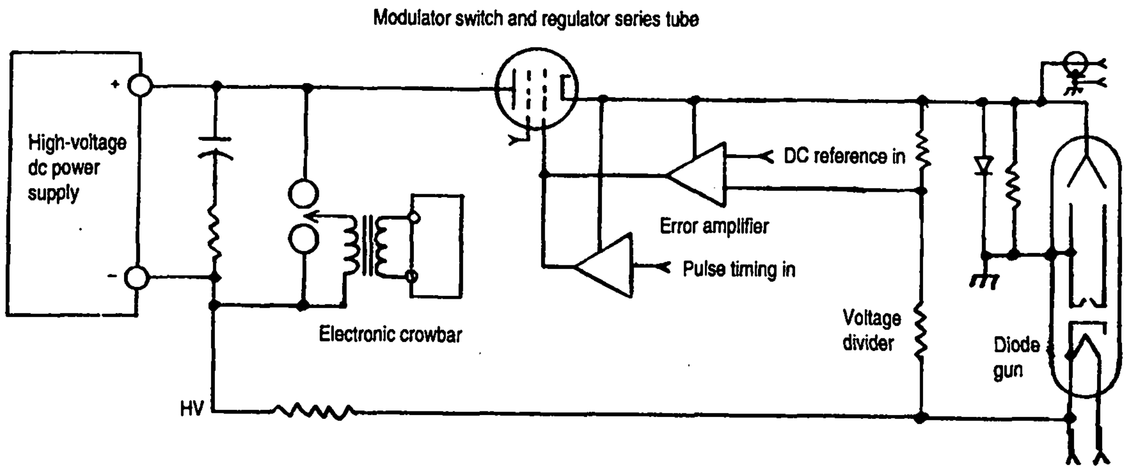

Figure 16-4. Modulator switch and regulator series tube in positive return of high-voltage power supply.

the high-voltage circuit. The advantage of this location in the circuit is that the common nodes for the error amplifier and the voltage sample are both virtually at ground potential, so that direct-coupled, high-speed circuits can be used throughout. Presumably, if the voltage hold-off capability of the regulator tube is adequate, it can be pulsed or biased to the non-conducting state during the interpulse interval by means of an amplifier channel that overrides the error-amplifier channel in response to low-level pulse-timing information, as shown. The combination modulator switch and regulator tube is shown as a tetrode because it is highly advantageous that the control grid operate in the negative-grid regime under all conditions.

Although such a connection can presumably be made to work, it has one usually crucial disadvantage. When the switch tube is off, the full power-supply voltage appears across it and not across the diode electron gun. Thus, the negative terminal of the power supply (and capacitor bank) is at ground potential, and the positive terminal is above ground by the full system voltage. When the switch tube is turned back on, the voltage across it drops to the minimum headroom requirement, pulling the positive terminal of the high-voltage system back to near ground and the negative terminal down to the operating cathode voltage of the diode gun. This is very stressful for the high-voltage power supply, the capacitor bank, and even the electronic-crowbar circuits. It should be avoided.

Figure 16-5 shows the combination of modulator switch and regulator tube in the circuit location most common for a hard-tube modulator switch: in series with cathode of the microwave-tube's diode electron gun. In this location there is no common terminal shared by the beam voltage sample, the error amplifier, and the grid-cathode circuit of the regulator tube. In fact, the error amplifier and voltage-sample common are separated from the regulator-tube grid-cathode circuit by the entire high-voltage power supply output. Modern high-speed data links using fiber-optic coupling are capable of bridging this voltage gap, as shown. Voltage telemetering is commonly performed by complementary voltage-to-frequency and frequency-to-voltage converters. Usually the variable-frequency pulse train is optically coupled.

However, the disadvantage of this topology is that the proportional-voltage

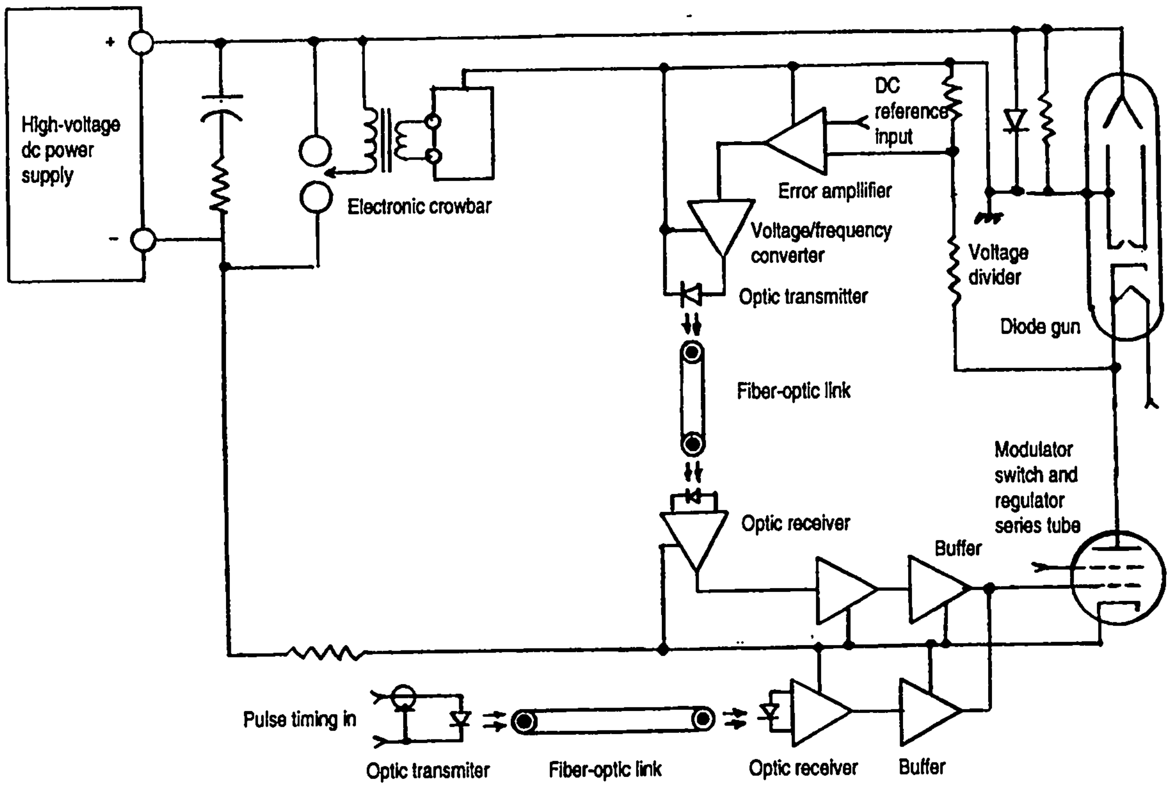

Figure 16-5. Modulator switch and pulse-voltage-regulator tube in series with high-voltage bus.

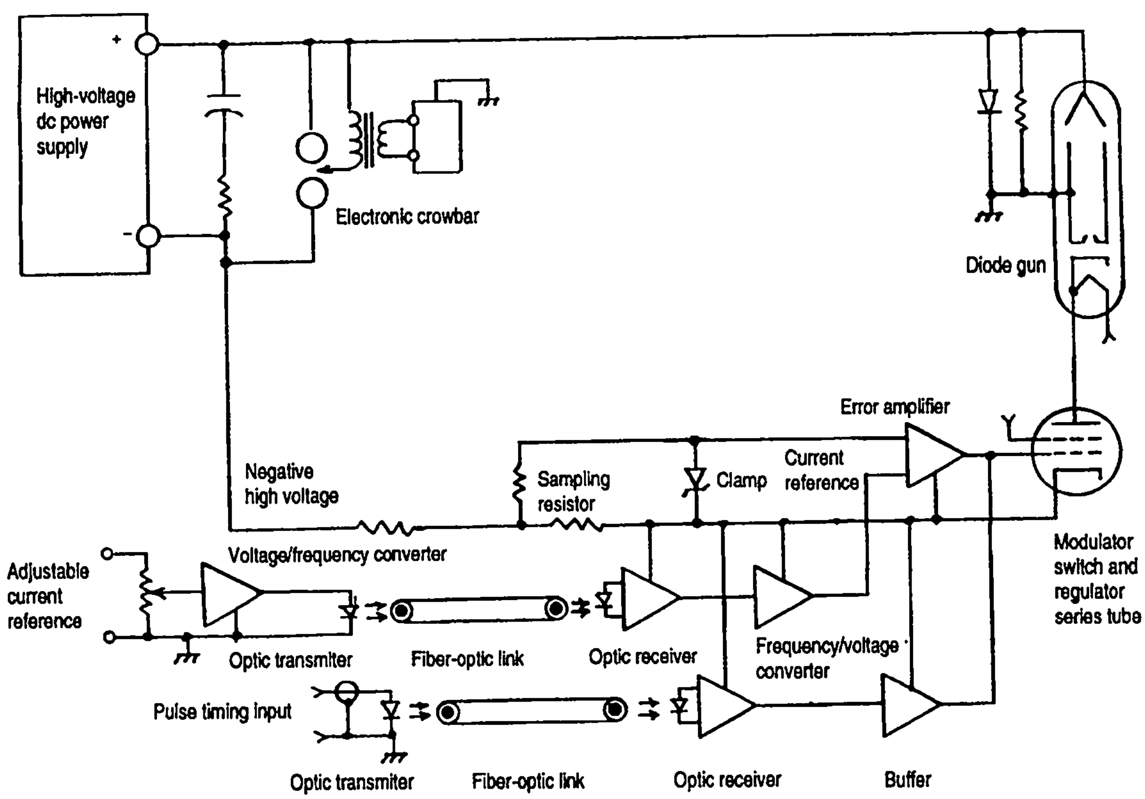

Figure 16-6. Modulator switch and pulse-current-regulator tube in series with high-voltage bus.

data link, no matter how high its speed, lies within the closed loop of the voltage-regulating servo system, thus limiting the speed of its response. To get around this problem, it is not unusual to use a proportional light-link whose optic-receiver output voltage is at least monotonic with optic-transmitter input current. (Absolute linearity is not a requirement. Monotonicity is.)

A permissible question at this point is, why try to regulate beam voltage in the first place? It may be better to attempt to regulate pulse-beam current instead, as shown in Fig. 16-6.

In a properly functioning diode electron gun, beam current is always related to the 3/2 power of beam voltage. Regulating beam current to a given degree of precision will maintain beam voltage to a value that is one-and-a-half times as precise. Moreover, the beam current can be sampled by means of a series resistor, as shown. This sampling resistor restores the desirable common-terminal situation of the circuit in Fig. 16-4, only at the high side of the power supply rather than at ground. Both the pulse-timing information and a voltage proportional to the desired peak-beam current can be optically coupled to the high-voltage common node without impinging on regulator-response speed or open-loop gain. During the rise time of each pulse, the error amplifier will be saturated, permitting the switch tube to deliver all of the capacitance-charging output current at its disposal. As beam current approaches the desired value, the voltage across the sampling resistor will approach the current-reference voltage, and the regulator will attempt to maintain that condition until the pulse-timing information once again overrides the error-amplifier output and shuts off the regulator/modulator switch.