20. Transmitter Control, Monitoring, and Interlocking

No matter how internally complex or how high-power a transmitter might be, its external or human-operator interface might be quite simple—as simple as turn it on, turn it off, or, if the transmitter stops working, to say what's wrong. The first two controls represent the operator or larger system communicating with the transmitter. The third is the transmitter communicating with the operator or larger system. In addition to providing a way for an operator to turn the system on and off, all transmitters must be capable of monitoring their internal status and automatically taking the proper action in case of anomalies. Such a capability is called a control circuit. In terms of computer logic, a transmitter control circuit will be one form or another of an AND gate. Often it will have literally hundreds of independent inputs, all of which must prove "true" before ultimate transmitter output will be obtained. We begin with the simplest control circuit of all.

20.1 The relay-logic ladder-network control circuit

The great preponderance of existing transmitter control circuits—and even some that are still on the drawing board (or CAD workstation)—are nothing more than expanded versions of the bare-bones illustrative circuit shown in Fig. 20-1. This is the classic "daisy-chain" ladder-logic-type AND gate with indicator lamps at every rung. As shown, it is capable of automatically sequencing transmitter turn-on from the very beginning to the point where the transmitter is ready for high-voltage dc to be applied to the transmitter output device. This is basically accomplished by applying control power at the beginning of the chain. (The circuit logic also assumes that low-voltage primary power is continuously connected to it.) Let's follow the daisy chain.

If the transmitter is water-cooled, a liquid-level switch in the water reservoir will be closed if the level is adequate. The switch closure applies power to the first indicator lamp (green) and to a power contactor that applies primary power to the cooling-system pump, which we do not want to turn on if there is not enough water in the system. Simultaneously, it powers the next rung in the ladder. (Any cooling-system air blower has no preconditions that need to be satisfied; it can be turned on coincident with control power.)

Assuming that the RF device is magnetically focused and that the focus coil is water-cooled, the next event of importance is water flow in the coil. If it is adequate, the water-flow interlock switch will be closed, applying control power to the next indicator lamp, to the coil of a contactor that applies primary power to the focus-coil dc power supply, and to the next rung.

If the RF device has a focus coil, it usually has a body and a collector that also need to be water-cooled. The next switches in line are the body- and collector-water-flow interlock switches, which are closed if flow is adequate. If these switches are closed, power is applied to their respective indicator lamps, to the contactor that applies filament power to the RF device, and to the next rung.

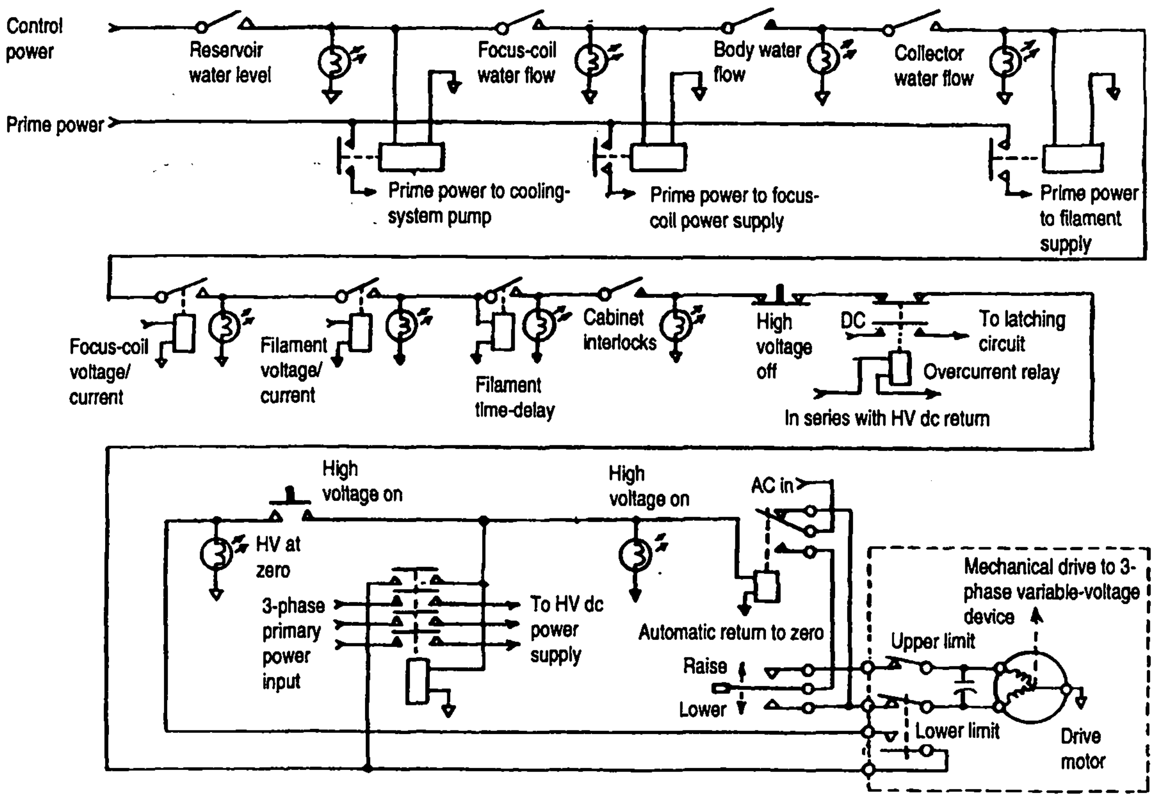


Figure 20-1. A simple series-string, or daisy-chain, type of transmitter control AND circuit.

The next two rungs are closed by relay contacts. The first is closed when its relay is energized with adequate measured current and focus-coil voltage. (In today's world of current-regulating dc power supplies, it is not enough to measure current alone, because even a short-circuited coil could draw the proper amount of current.) The second is closed by a relay that is energized by adequate filament voltage and current.

When both filament voltage and current are at their proper values, a cathode-heating time-delay relay is energized. For large-area, indirectly heated cathodes, this heating time can be as much as 15 minutes. Once the delay has elapsed, the relay contacts close, lighting the "delay-complete" indicator and sending control power to the next rung, which is the last fully automatic one in our simplified control ladder. Nevertheless, it is by far the most important. It is shown as a cabinet-interlock switch, but it represents all of the personnel-safety interlocks and interrupts that are not shown—including the kinds of switches described in Chapter 4, Section 6—and any mushroom-button emergency-off switches. Once this daisy chain has been activated, the cabinet-interlock indicator lamp is illuminated and the high-voltage on/off operator commands are enabled.

The high-voltage on/off controls illustrated are of the momentary-actuation type. The high-voltage off push button is normally closed and is in series with the normally closed contacts of a high-speed electro-mechanical overcurrent relay, whose actuating coil is in series with the high-voltage dc return conductor. The relay contacts will open if the high-voltage dc return current exceeds a preset threshold. Coupled to the normally closed contacts is a pair of normally open contacts in series with a latching circuit, which gives a more permanent indication that the overcurrent relay was momentarily tripped. The control path then leads through normally open contacts that are mechanically linked to the normally closed contacts of the lower-limit switch for the motor, which drives the variable-voltage device in series with the input to the high-voltage dc power supply. When the ac input to the high-voltage dc supply is at a minimum, these contacts will be closed, and the "high-voltage-at-zero" indicator (usually amber) will be on.

The closing of these contacts signals the operator that the on push button for the high voltage has been enabled. When it is momentarily depressed, control power will be applied to the three-phase contactor that connects primary power to the high-voltage dc power supply. An auxiliary set of contacts (the uppermost ones) short-circuits both the high-voltage-on push button and the high-voltage-at-zero contacts, latching the three-phase contactor in the closed position. The operator may now raise the output of the variable-voltage device, increasing the high-voltage dc applied to the transmitter output device until the desired operating conditions are reached. If nothing else happens between this event and the time when RF power is no longer needed and the high-voltage power supply can be shut off, the operator pushes the high-voltage-off push button, momentarily interrupting the coil power of the three-phase contactor, which opens the auxiliary contacts. Neither the high-voltage-on nor the high-voltage-at-zero contact is still closed, so the three-phase contactor remains open. If an automatic return-to-zero relay is used, as shown, it too will be de-energized, and its normally closed contacts will perform the same drive-motor function as the high-voltage "lower" switch position.

Suppose, however, that in the midst of normal operation an internal malfunction occurs, such as a leak in the water reservoir that drains all of the cooling water. On its way to empty, the water level will pass by the minimum amount required to keep the water-level switch closed. When its contacts open, the effect on the three-phase contactor is the same as pushing the high-voltage-off push button. The high-voltage dc will be dumped. The unambiguous indication given by the interlock ladder is that there is insufficient water level in the reservoir because the water-level indicator lamp is no longer illuminated. Unfortunately, none of the downstream indicators lights is on either. Here lies the basic shortcoming of the daisy-chain interlock string. Its indications are often ambiguous because it suggests that everything else in the circuit has failed too. (A simple burned-out bulb can give a false failure indication as well.) Until the leak is fixed and water-level is restored, there is no way of knowing if anything else has gone wrong. The opening of any interlock contact will dump the high-voltage dc and extinguish its indicator and all of the others downstream of it. (It is only fair to relate that large systems using this type of control sometimes make available to the operator "cheat" switches that permit the bypassing of an interlock switch to see how many of the downstream indications are still functioning. Although the cheat switch is supposed to be used only temporarily, it subverts the integrity of an otherwise fail-safe system. With respect to personnel-safety interlock systems, use of the cheat switch can also raise the issue of gross negligence.) This shortcoming, along with the availability of robust and affordable

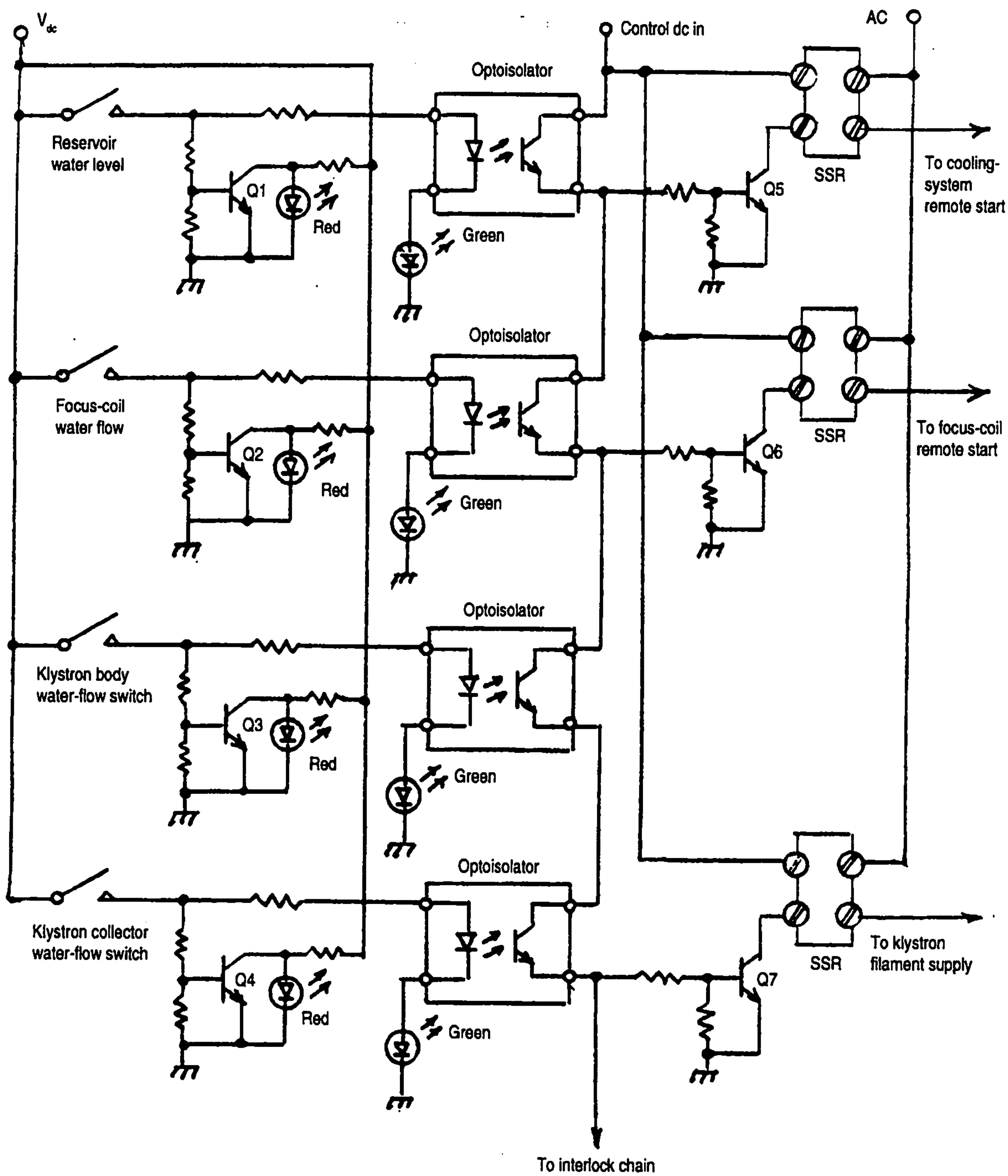


Figure 20-2. A solid-state transmitter-control circuit with series control but parallel status read-out (continued in Fig. 20-3).

solid-state components, has engendered advanced control-system topologies.

20.2 A solid-state, serial-control, parallel-indication interlock circuit

Figures 20-2, 20-3, and 20-4 show some of the features of a hard-wired, dedicated-circuit, solid-state transmitter-control system. It features a series-connected, ladder-network AND gate like that of the daisy chain and full-time status indication for all of the monitored functions. It uses dual indicators for each function—

green for "true" and red for "false"—to mitigate the burned-out-bulb ambiguity. If there were only green "true" indicators and one of them burned out while the interlock chain sequence was in some intermediate state, the light-out indication could be interpreted as either failure of the function or failure of the indicator. This is not the case with dual indicators. If neither the red nor green indicator is illuminated, the condition most likely suggests that one indicator or the other (or, more unlikely, both) is faulty, but not the function. Whatever ambiguity remains for this condition is lessened even further in the true daisy chain because if the downstream indicators are still on, it is most likely the indicator that has failed. To a large extent, the light-emitting-diode (LED) indicator has made the burnt-out-bulb syndrome a moot point because the LED tends to last as long as almost any other component in the system. On the other hand, incandescent, 24-Vdc bulbs, which are found in many older control systems, require almost continual replacement.

The building-block circuits are first shown in Fig. 20-2. The basic transducers and interlock switches are virtually the same as the ones used in the previous topology, starting with the water-reservoir level switch. Instead of being part of a series-connected string, however, its contacts are associated with an individual solid-state circuit. When reservoir level is too low, the contacts are open and control voltage V_{dc} produces current flow through the first red LED. When the contacts close, V_{dc} produces base current in *Q1*, whose collector short-circuits the red LED, thus turning it off. It also produces current through the photodiode of the first optically coupled isolator and the first green LED, which are connected in series. Note that the sum of these currents is not very large (in the milliampere range). The switches and sensors are operated in the dry-contact mode, so failure through contact welding is a virtual impossibility. (Failure of the switch to conduct when it is closed is the more likely failure mode, but it is fail-safe.)

The photo-sensitive transistor of the optoisolator is now turned on, allowing current from the control dc input to the base of *Q5*, turning on the first SSR and also enabling current flow to the next optoisolator phototransistor just below it. The turning on of the dc control input to the first SSR through *Q5* switches ac to the cooling system remote-start function.

The subsequent functions are the same as those shown in the daisy-chain system. The AND gate is formed by the series connection of all of the optoisolator phototransistors. Subordinate actions take place with the same prerequisites as before. Transistor *Q6* will not enable the focus-coil power supply SSR until both water-level and focus-coil water flow are "true." Transistor *Q7* will not enable the klystron filament supply SSR until klystron-collector water flow and all functions before it are "true."

The optoisolator chain continues on Fig. 20-3, but an important high-speed interrupt has been added to the sequence that responds to excessive klystron body current. The klystron illustrated has a collector that is electrically isolated from the body (without which there is no way to measure body current). The positive (low-voltage) return lead of the high-voltage dc power supply is connected to the collector through current meter *M3*, which indicates average collector current. The return is also connected to the klystron body (ground) through current meter *M4* and a series resistor. Any ground or klystron-body current

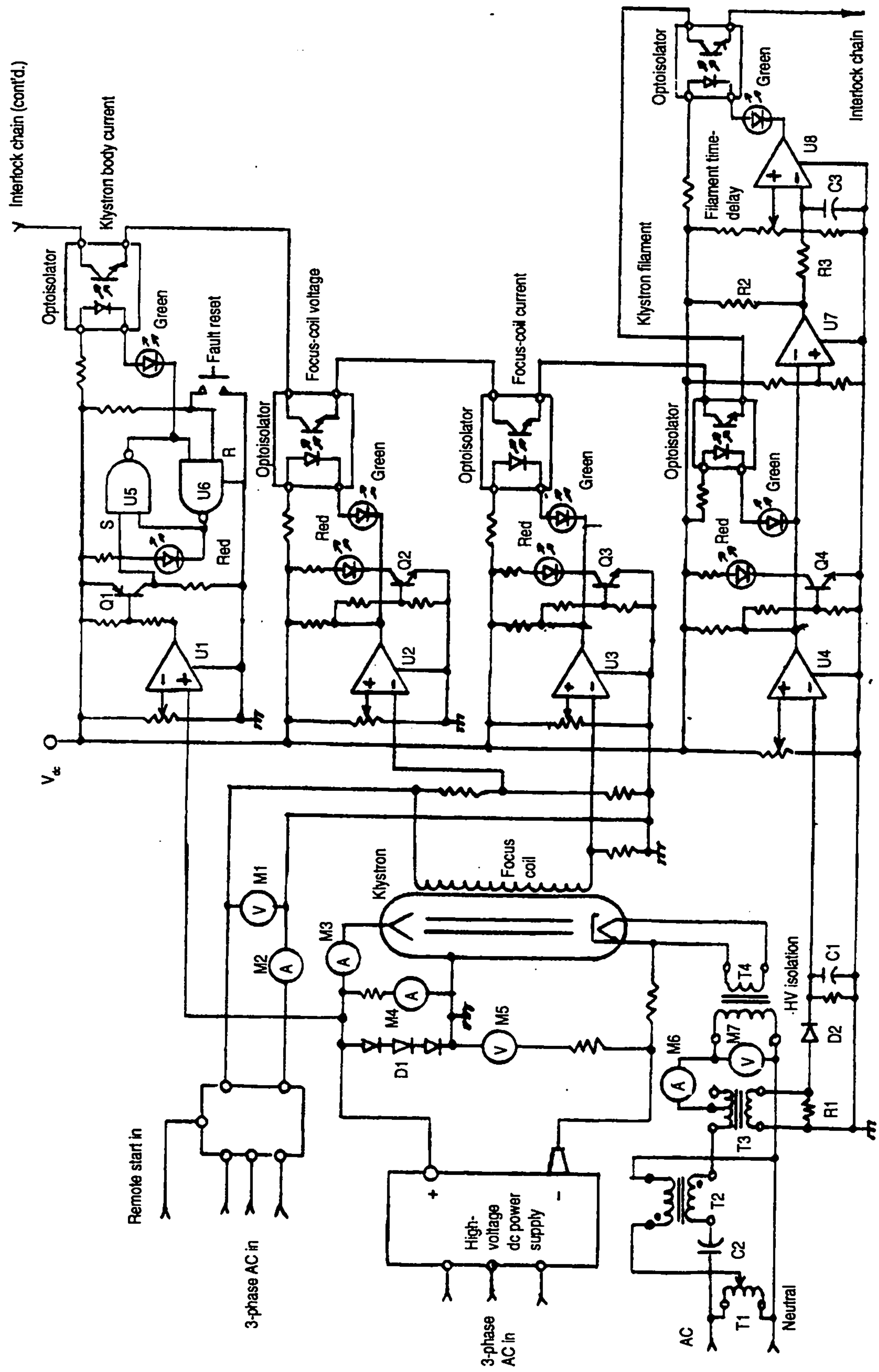


Figure 20-3. Continuation of solid-state, serial/parallel control circuit (continued in Fig. 20-4).

must return to the power supply through *M4*, which indicates its average value. The voltage across *M4* and the resistor in series with it is also proportional to instantaneous body current. This voltage is applied to the positive input of an integrated-circuit voltage comparator, *U1* (such as the ubiquitous LM339). An adjustable-threshold voltage is applied to the negative input. As long as the klystron-body current is below an acceptable upper limit, the negative input of the comparator will be more positive than the positive input and the output of the comparator will be low, meaning that its open-collector output device will be conducting. This produces base current in the PNP transistor *Q1*, turning it on and pulling the voltage at its collector positive to near the V_{dc} rail. This voltage is also connected to the "set" (*S*) input of the "set/reset" flip-flop (bi-stable multivibrator), which is created by interconnecting the two integrated-circuit, dual-input NAND gates, *U5* and *U6*, as shown. When both input gates of the NAND gate are high, the output will be low. If either input is low, the output will be high. In the reset (*R*), or normal, state of the flip-flop, the output of *U6* is high, which is one input to *U5*. The other input is the collector voltage of *Q1*, which is also high. So the output of *U5* is low, assuring that the output of *U6* will be high and permitting current through both the optoisolator photodiode and the green LED indicator.

Should the klystron-body current even momentarily exceed a value that makes the voltage at the positive input of *U1* exceed the threshold voltage at the negative input, the output of *U1* will go high, *Q1* will stop conducting, the *S* input of *U5* will go low, the output of *U5* will go high, current will stop in the photodiode and the green LED, and the interlock chain will be opened with microsecond speed. In addition, the output of *U6* will go low because both of its gate inputs are now high, turning on the red LED, which indicates a fault in this channel. The circuit is reset by pressing the "fault-reset" button, which momentarily grounds the *R* input to the flip-flop, making the output of *U6* high again. The klystron body current will be zero at this time because the high-voltage dc has been shut down, so this will reset the original conditions.

If the klystron-body-current overload was caused by a high-voltage arc in the klystron electron gun, all of the arc current will be body current, and it can have a peak current that is the operating beam voltage divided by the value of surge resistance in series with the cathode. The klystron illustrated has continuous beam current (it is not pulse-modulated), and might have a beam voltage of 30 kV or so. If the cathode surge resistor is 30 ohms—a value picked for an obvious reason, but it is by no means atypical—the peak instantaneous arc current could be 1000 A. This is the reason for the series diode string shown as *D1*, which shunts *M4* and its series resistor. The forward voltage drop of the diode string, even at low current, is chosen to be greater than the threshold voltage of *U1*. Diode *D1* is expected to clamp the voltage across it to a value safe for the positive input of *U1* while diverting nearly all of the 1000 A of fault current through itself.

The next two rungs in the optoisolator ladder are focus-coil voltage and current, both of which use identical electronic circuits. A sample of the focus-coil voltage that is taken from a voltage divider is applied to the negative input of comparator *U2*, and a voltage proportional to focus-coil current, which is obtained from a sampling resistor in series with the focus coil, is applied to the negative input of comparator *U3*. The positive inputs of both comparators are connected to adjustable threshold voltages. Note that both terminals of the focus coil float with respect to ground. If both the voltage and current samples exceed the threshold voltages, which represent acceptable minima rather than maxima, the outputs of both *U2* and *U3* will be low, enabling current flow in their respective optoisolator photodiodes (and their green indicator LEDs) and completing the conduction path through the series-connected optoisolator phototransistors. Should either sample signal fall below its threshold, its comparator output will go high, either *Q2* or *Q3* will be turned on (thus illuminating one or the other red "fault" LED), and conduction in the phototransistor chain will be stopped. Meters *M1* and *M2* indicate focus-coil voltage and current and are usually accessories of the focus-coil power supply.

The last two rungs shown in Fig. 20-3 monitor klystron-filament current and cathode-heating time-delay. The actual filament current flows in the secondary of transformer *T4*, one side of which is connected to the klystron cathode. The cathode is normally 30 kV negative with respect to ground, thus making it awkward to directly monitor the filament current. (A useful, although not entirely accurate, indication of filament current can be obtained by monitoring the *T4* primary current, which is usually smaller by the *T4* voltage step-down ratio and referenced closer to ground. This method is not entirely accurate because of *T4* magnetizing current.) This is the reason for the presence of *T3*, which is often a low-voltage filament transformer that is connected with its low-voltage secondary in series with the primary current of *T3*. The secondary is center-tapped as shown to provide half the voltage in order to maximize the step-down ratio. A resistor, *R1*, is connected across its primary winding, which is now functioning as the secondary of a high-burden current transformer. The voltage across *R1* is rectified by *D2* to produce a usefully high voltage across *C1*, which will be proportional to *T4* primary current and approximately proportional to klystron-filament current. The value of *R1* will be reflected into the *T4* primary circuit as a resistance of $R1/n^2$, where n is the *T3* turns ratio. The voltage developed across the *T3* secondary, which serves as the current-transformer primary, will be $I_P \times R1/n^2$, and the voltage across *R1* will be

$$I_p \times \frac{R1}{n^2} \times n = I_p \times \frac{R1}{n},$$

where I_P is the *T4* primary current. If, for instance, the klystron heater power is approximately 500 W and the *T4* primary voltage is 120 V, the *T4* primary current will be around 4 A. If a voltage of 5 V across *R1* is adequate to operate the sensing circuit and T3 has a 120-V nominal primary and a 5-V center-tap secondary, the n = 120 V/2.5 V, or 48. The required value of *R1*, then, is 48 x 5 V/4 A, or 60 ohms. The voltage drop introduced into the primary circuit is 5 V/48, or 0.1 V, and the power dissipated is about 0.4 W.

The filament circuit also limits cold-filament inrush current. This feature is provided by the series ac-rated capacitor, *C2*. The cold resistance of a filament is typically one-fifth that of a fully heated filament. The starting inrush current, however, is usually specified as no more than twice the normal operating current.

Additional series impedance is required for start-up. For this purpose, designers often use a series resistor that is later short-circuited by relay contacts. An alternative that consumes no power is to use a series capacitor. A value can usually be determined that will limit inrush current to 1.5 times normal current. At turn-on, its capacitive reactance will be the current-limiting impedance. However, when the heater warms up, its resistance will increase by a factor of 5 and it will be the predominant impedance. The capacitive reactance, which is in quadrature with the resistance, will then have only a minor influence on total impedance. However, it will produce a small loss in total primary voltage, which is compensated for by the adjustable boost voltage provided by variable autotransformer *T1* and step-down transformer *T2*, whose secondary is connected, series-aiding, in the primary circuit. Meters *M6* and *M7* can be calibrated to produce quite accurate indications of actual klystron-filament current and voltage, especially near the nominal operating conditions.

Comparator circuit *U4* is the same as *U2* and *U3*. Its input is sensitive only to current in the klystron-filament circuit, which produces ambiguous information by itself because it cannot distinguish between a normal filament and a short-circuited one. In some cases, a separate circuit that is sensitive to *T4* primary voltage might be necessary. Open filaments, however, are a far more likely occurrence than short-circuited ones, making voltage sensing needlessly redundant in most cases.

Once *U4* has sensed proper filament current, its output goes low, pulling the negative input of comparator *U7* low with it and turning off its output transistor *Q4*. This allows current through *R2* and *R3* to charge C3. The time-constant (*R2* + *R3*) x C3 is made approximately equal to the desired cathode-heating time, which can be as great as 15 minutes, or 9000 seconds. The voltage across C3 is applied to the negative input of comparator *U8*. The adjustable voltage at the positive input of *U8* is set to equal the C3 voltage at the end of the desired heating interval, at which instant the output of *U8* goes low, again enabling current through the optoisolator photodiode and the green LED indicator. This action completes the path through the series-connected optoisolator phototransistors up to this point.

Resistor *R3* in the timer circuit provides some memory in case of a momentary interruption of filament power that would cause the *U7* output to conduct to ground again. If *R3* is the same value as *R2*, C3 will discharge at an initial rate that is only twice as great as the one at which it was charged. If the filament-power interruption is only a small portion of the total heating time, the cathode will not have cooled off that much and the timing circuit will not have completely discharged either.

The completion of this simplified control ladder is shown in Fig. 20-4. The last rung of the ladder represents the personnel-safety interlocks. As will be seen, the response to these very-important sensors is redundant in different ways. The first level of redundancy is not always permissible under some system-engineering guidelines. As shown, this method requires that the 120-Vac actuating power for the three-phase vacuum contactor, which makes and breaks the primary-power feed to the high-voltage dc power supply, pass through the contacts of the personnel-safety interlock switches. These interlocks monitor both

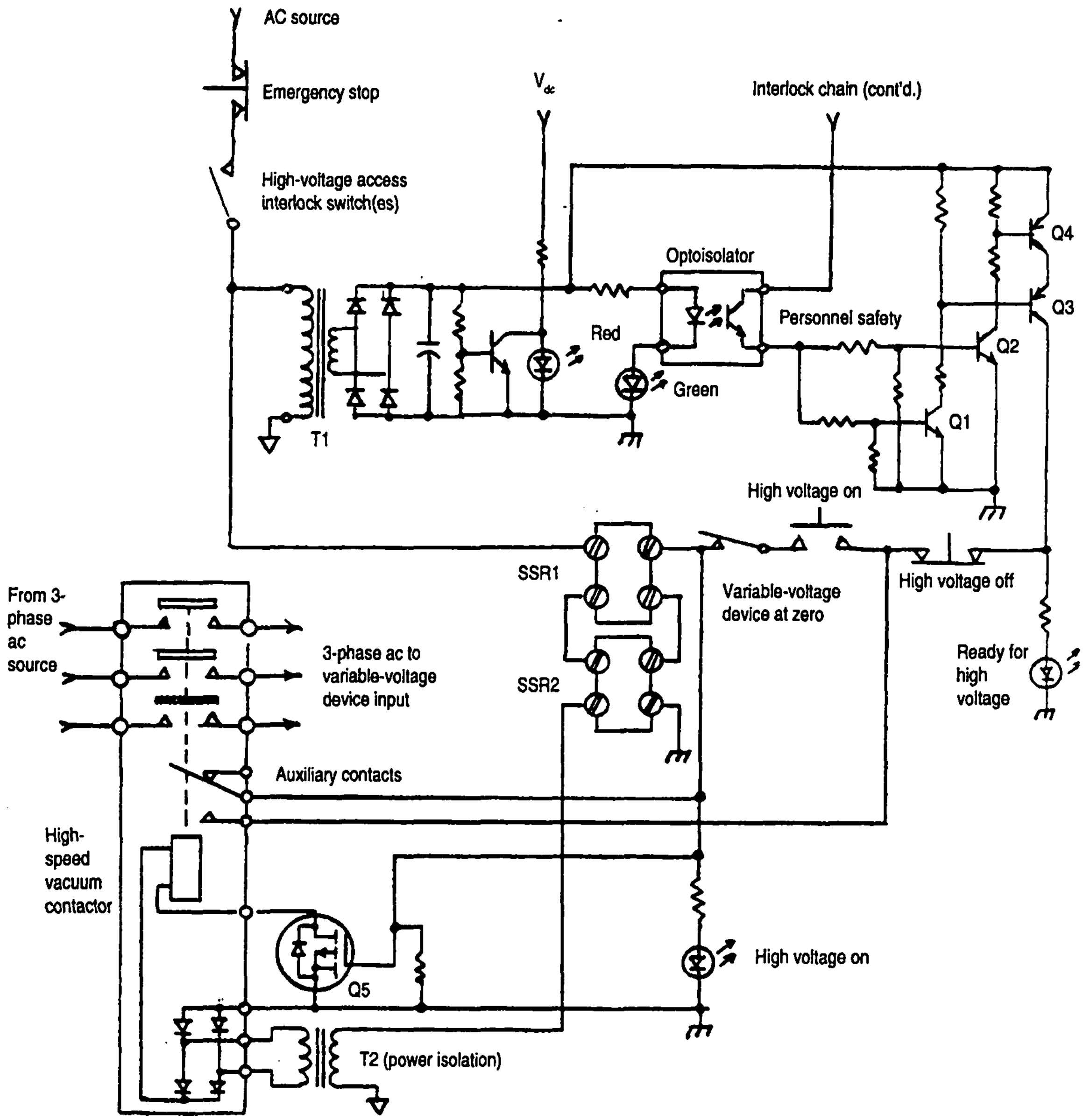


Figure 20-4. Completion of solid-state, serial/parallel control.

the status of panels and doors that permit access to hazardous voltages and the emergency-stop push button(s). This configuration is the most direct and positive means of assuring shutdown. When used, it is often augmented by running the same line to an undervoltage-trip coil on the three-phase-source main circuit breaker, making the primary-power break doubly redundant.

The objection to this approach is that the safety interlocks carry 120 Vac and therefore are potentially hazardous themselves. In addition, their closed contacts can carry significant load current, depending upon the closing and hold-in volt-ampere demands of the vacuum-contactor. Therefore, they are subject to arcing and possible contact welding. (The first objection is usually governed by some regulation, and not much can be done about it if 120-V interlocks are proscribed. The second objection can be surmounted by design. If cabinet interlocks like those illustrated in Fig. 4-2 are used, contact welding is not much of a threat, and positive action can easily be verified.)

If the use of 120-V interlock power is permitted and deemed advantageous, it can also be used to provide the last level of dc control power as well, after being stepped-down by *T1* and rectified and filtered. Before the interlocks are complete, or "true," control-power V_{dc} provides the current through the red LED that indicates the incomplete condition. When the interlocks are completed, rectified dc turns on the transistor shunting the red LED, diverting current from it. The transistor also produces current flow through the optoisolator photodiode and the green LED, which indicates that the interlocks are complete. When the last optoisolator phototransistor conducts, it completes the series-connected AND gate. This connection enables current flow into the bases of transistors *Q1* and *Q2*, which are part of a self-redundant output switch. Turning *Q1* on turns on *Q4*, and turning *Q2* on turns on *Q3*. Transistors *Q3* and *Q4* must both be on before the locally derived control dc reaches the high-voltage on/off circuit. This state is signaled when the green "ready for high-voltage" LED is illuminated. No single-point failure will cause the output switch to malfunction in an unsafe way.

If 120-Vac interlocks are not permitted for whatever reason, V_{dc} replaces the ac source, and step-down transformer *T1* and its output rectifier are not needed. The ac is connected directly to the ac line-side of *SSR1*. With all interlocks complete, high-voltage may be turned on by depressing the high-voltage-on push button, but only as long as the variable-voltage device is at zero (as before) and its lower-limit switch is closed, as shown. Also for the sake of redundancy, contactor-actuating ac is connected to power-isolation transformer *T2* through two series-connected solid-state relays, *SSR1* and *SSR2*. The *SSR* dc control voltage is also applied to the gate of a high-voltage MOSFET, *Q5*, which is in series with the return lead of the internal rectifier for the vacuum-contactor actuator. This not only effects redundancy, but it greatly speeds the opening time of the contactor by immediately interrupting the solenoid current. In so doing the MOSFET must withstand a severe voltage transient that will appear across the dc solenoid as its internal inductance attempts to sustain current flow to keep the contactor closed.

Like the daisy-chain interlock ladder, interruption of any of the monitored and interlocked functions of the circuit just described will cause the same direct interruption of the high-voltage power supply, except that now if only one function is to blame, only one green LED will be off and only one red LED will be on. If others functions share the blame, they will reveal themselves in the same way.

20.3 Integrated-circuit considerations

The designers of modern, hard-wired transmitter-control circuits have been able to take advantage of an ever-increasing repertoire of computer-logic integrated circuits to accomplish increasingly ambitious performance goals. The only such device used in the simple control circuit just described is a dual-input NAND gate. Two of them (*U5* and *U6* in Fig. 20-3) were interconnected to create a set/reset flip-flop. A negative-going voltage change at the *S* gate input to *U5* will shut down the transmitter. It is the intent of the design to limit such a voltage change only to excessive peak body current in the klystron RF amplifier tube. However, what about the "noise immunity" of that input? Are there other transient voltages coupled to that point that, one way or another, could also trip the circuit and shut down the transmitter? There are events that are completely systematic, such as the closing of the three-phase primary-power contactor, that are capable of generating large-amplitude transient voltages and currents. What assurance do we have that such a transient will not trip our flip-flop or any other similar circuit element that we have designed into our system? Unfortunately, the answer is that we really have no assurance until the transmitter is completed and tested.

There are, however, choices that will either increase or decrease the likelihood of false alarms. Perhaps the most popular of all logic-element families is TTL, or transistor-transistor logic. At its most basic, TTL uses two transistors. The collector of one transistor is connected to the base of the next. When the first transistor is turned on, its collector voltage is lower than the base voltage required to turn on the next transistor. The collector of the first transistor is connected to the supply rail, nominally 5 Vdc, through a pull-up resistor. When the first transistor is turned off, its collector voltage is pulled up to a voltage that exceeds the base voltage required to turn on the next transistor. Thus, transitions are made from logical 0 to logical 1.

In Fig. 20-5 these voltage relationships are shown in terms of specification limits for TTL logic. Throughout the family of devices, the collector voltage of no conducting transistor will exceed 0.4 V and the base voltage required to turn on any transistor will not be less than 0.8 Volts. There is thus a noise-voltage margin of the difference between the two, or 0.4 V. At the other end, the maximum base voltage required to turn on any transistor is 2 V and the minimum collector voltage of any non-conducting transistor is 2.4 V. The noise-voltage margin, or noise-immunity, is again the difference between the two, or 0.4 V. Many interfaces between transistors will have higher actual margins, but they cannot be guaranteed by specification.

Integrated-circuit manufacturers have not been entirely unaware of the relative "hair-trigger" thresholds inherent in TTL logic because this problem can become intolerable when their devices must be used in threatening electromagnetic environments. Therefore, they have developed families of integrated circuits called high-threshold-logic (HTL) and high-noise-immunity logic (HINIL). As shown in Fig. 20-5, an additional noise-voltage margin was introduced by inserting a series Zener diode between input and transistor base, and by operating the circuit at higher rail voltage, either 12 Vdc or 15 Vdc, as illustrated by the simple dual-input NAND gate. Operating at 15 Vdc, the noise immunity was increased to 3.2 V for the low-to-high transition, and 6.5 V for the high-to-low transition.

These families of integrated circuits are not as fast as TTL to be sure, but what's the big hurry? In the case of the transmitter control previously discussed, the fastest fault-corrective event—the opening of the three-phase contactor—will require on the order of one-half cycle at the primary-power frequency (typically 60 Hz), which is 8.3 ms, or 8,300 μs. It has been suggested that there is no good reason to have an individual response time that is less than one-tenth that of the ultimate response, which means that the toggle time of the flip-flop could be 830

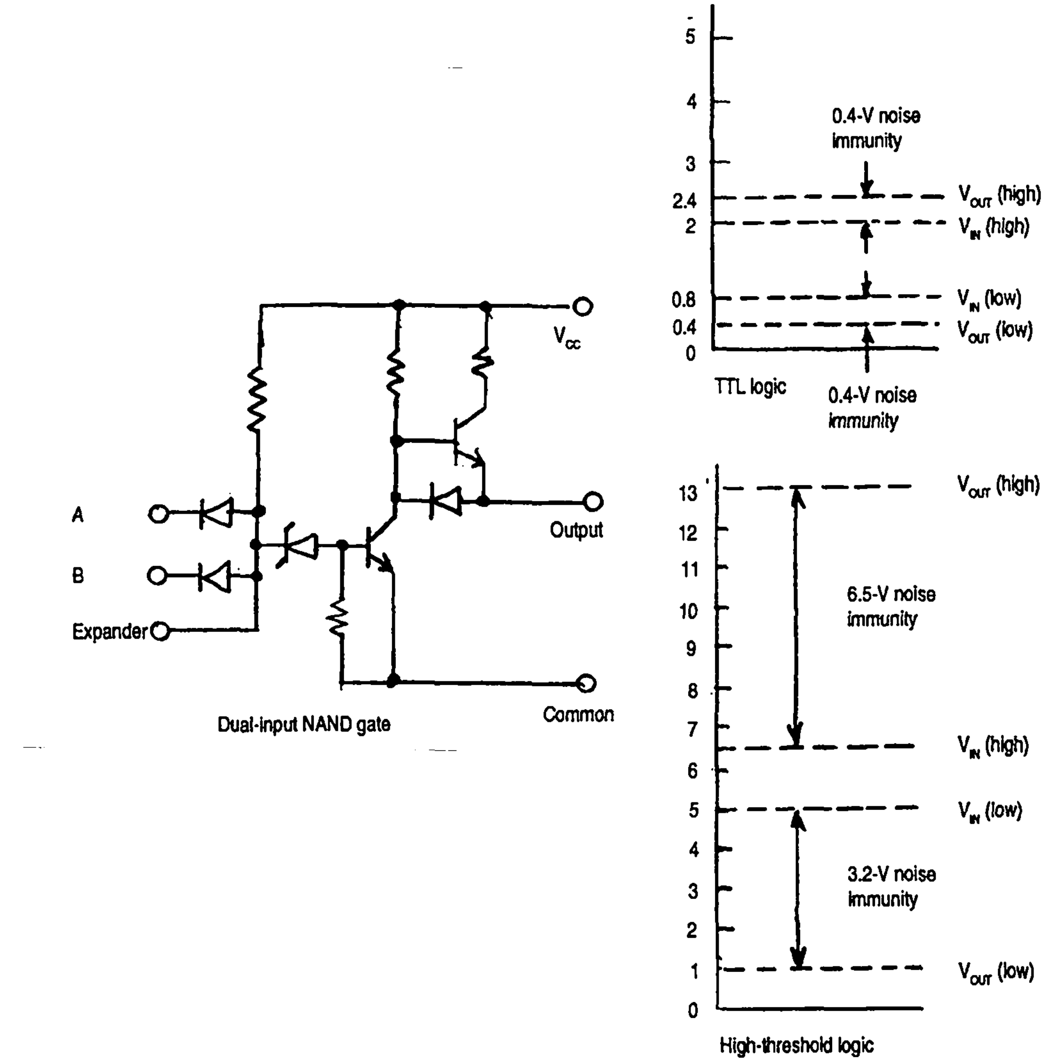


Figure 20-5. High-noise-immunity (high-threshold) integrated circuits.

μs without noticeable effect on the overall response time. Making the response faster only increases the bandwidth of the transient spectrum to which the system might respond. With this in mind, designers have been known to deliberately slow down a NAND gate like the one shown by connecting an integrating capacitor between the "expander" input and ground. This allows the input to be present for a longer period of time before the decision is made that a real fault has occurred. (Transmitter-control logic must be slow to anger, but it should shoot to kill.)

Even where considerably faster response is required—such as when an electronic crowbar is used to discharge an energy-storage capacitor bank—submicrosecond response time is still not required. So, you might ask, why don't all hard-wired transmitter-control circuits use HTL or HINIL logic elements? Many successful ones have, but the product lines are dwindling and there are no guarantees that replacement devices will always be available. Some designers are turning to the voltage-comparator IC, from which numerous logic elements can be constructed. These devices have even greater noise-immunity levels than

HTL or HINIL, and they offer a greater promise of being around for the foreseeable future.

Regardless of what logic devices are used, their gate inputs need external protection if they are to survive for long. Figure 20-6 shows some of the protection techniques. A series resistor is always required, either to limit the fault current that a Zener-diode clamp will have to handle or to provide its share of the time-constant when an integrator or low-pass-filter capacitor is used. The diode, used as a supply-rail clamp, will prevent the gate from ever being above the supply rail by more than a diode forward-voltage drop.

A final note has to do with what could be called "connectivity safety." If we review the circuit topology in Figs. 20-3 through 20-5, we should notice that some of the functions could be accomplished with fewer active components. However, this circuit was designed so that if *any* of the removable components like the ICs or transistors were not in place, a "true" indication could not be obtained. (In the case of the *U5-U6* flip-flop in Fig. 20-3, both were part of the same IC.) Designing for connectivity safety is a good idea. At least one control circuit that did not follow the concept of connectivity safety was capable of closing the prime-power contactor whenever a circuit card was pulled from its nest. Nevertheless, connectivity safety is not the same as "fail-safe," which can never be absolutely assured.

20.4. The programmable-logic controller

The more complex a transmitter becomes and the greater the desire for its fully automatic operation, the more sense it makes to base a control system on a programmable-logic controller (PLC), its associated input and output modules, and its display and operator-input components—to say nothing of its ability to interact with other computer-based systems. Components of PLC systems have been engineered for industrial applications. They have built-in transient-voltage tolerance, hard-wired and digital filters, and with many other hardening features.

To the transmitter hardware designer, there should be nothing intimidating or

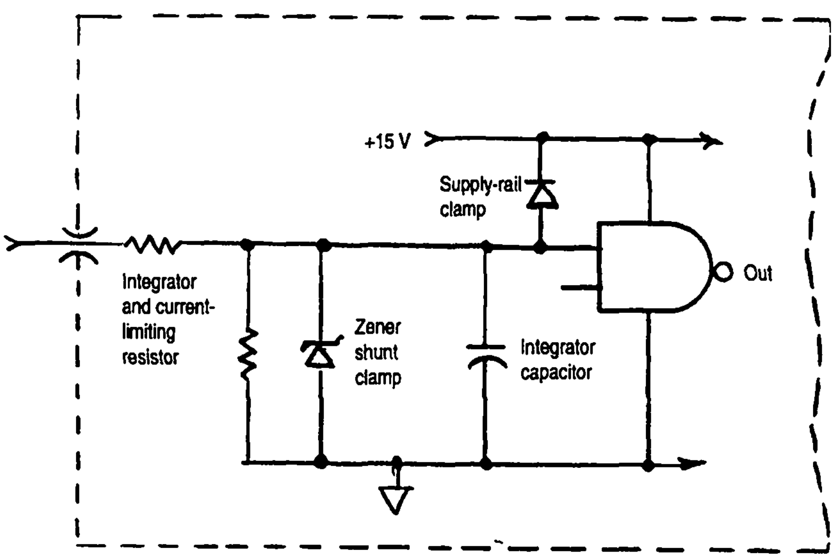


Figure 20-6. Methods of protecting integrated-circuit inputs.

off-putting about the use of a control system that will ultimately be driven by a computer program. The hard-wired interfaces are much like those that would be encountered in a conventional, dedicated transmitter-control system, such as the types already discussed. Figure 20-7 shows the layout drawing of a controller crate that is configured to handle a dual-klystron transmitter. The far-left crate location, which has no slot number, is reserved for the processor itself. A typical processor can handle over 500 input/output channels and is programmed to emulate the way that a well-designed hard-wired control circuit would respond. A major advantage of the processor, however, is that its behavior can be improved by merely changing a program. The next slot in the crate, Slot 0, is a dedicated fiber-optic converter. Two such converters can communicate with each other over either two fiber-optic channels or two twin-axial electrical cables. Slots 1 through 11 are used by different input and output modules, as shown. In this specific example, not all of the available slots are used—nor are all of the different module types used.

The dc-input modules used in Slots 1, 3, and 9 have a logic function similar to that of the relay coils in relay-coupled control logic. Each module has 16 inputs. A minimum input of 10 V is required for a low-to-high transition, a maximum of 6 V will cause a high-to-low transition, and a maximum of 30 V can be tolerated, making them even more noise-tolerant than the high-level-logic ICs previously mentioned. In addition, an input filter delays "on" response by 6 ms, and "off" response by 20 ms, and the actual coupling to the internal circuitry is by means of

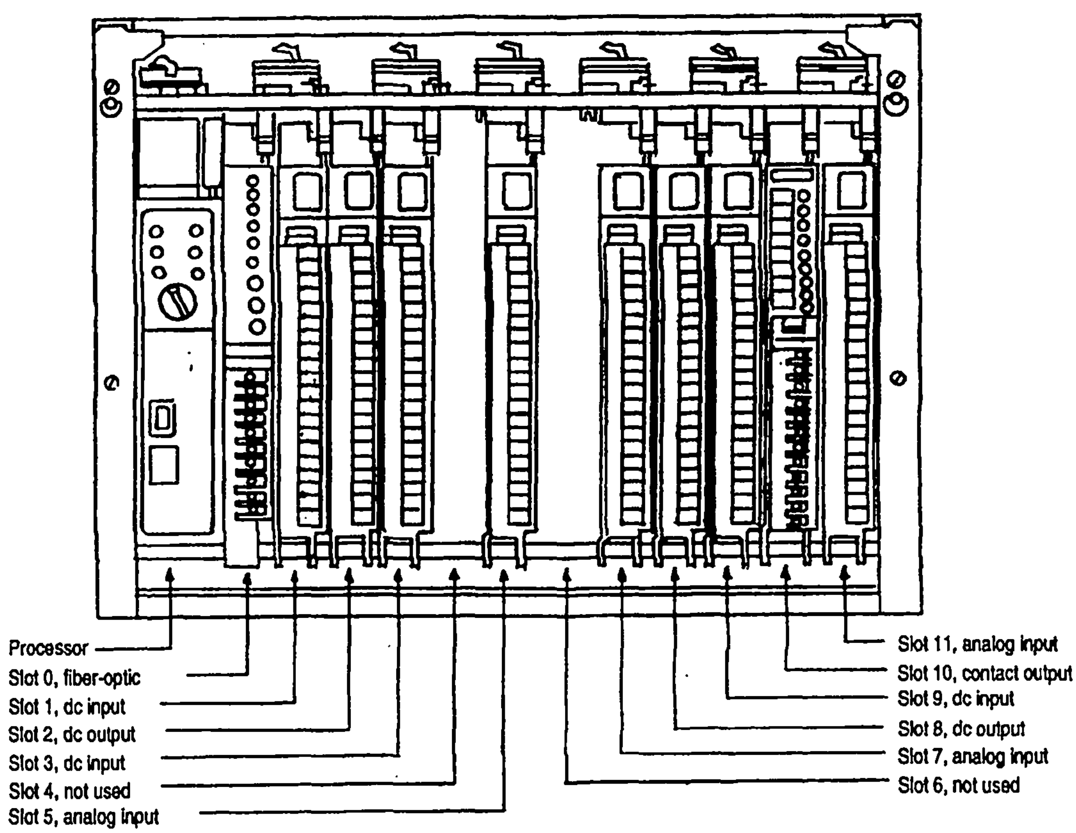


Figure 20-7. Typical arrangement of modules in programmable-logic-controller input/output chassis.

optically coupled isolators with 1.5-kV isolation.

The analog input modules used in Slots 5, 7, and 11 serve the same roles as analog or digital meters or adjustable-threshold comparators. Each will accommodate either 16 single-ended signals or eight differentially driven signals. The inputs, having a maximum range of -10 V to +10 V, are converted into either a four-digit binary-coded decimal or a 12-bit binary value, which gives the signals a resolution of 1 part in 4095. The signals are then passed on to the processor.

The dc-output modules used in Slots 2 and 8 have 16 outputs each. The input to each module is a dc power supply, which ranges from 10 V to 60 V. As commanded by the processor, each module will switch as much as 2 A per output (8 A total per module) to a load connected to an output.

The selectable-contact output module used in Slot 10 has four pairs of relay contacts, each pair sharing a common lead. Each relay contact is jumper-selectable as either normally closed or normally open. It will switch ac or dc in the range from 24 V to 125 V at up to 1 A, with 5-ms operate-and-release times. Its maximum actuation rate is 10 Hz.

Each module has its own set of diagnostic LED indicator lamps to locally indicate activity. The signal lines that the module inputs and outputs are connected to are shown in Fig. 20-8. Most of the signal designations should be self-explanatory. Some, however, may be more scrutable than others. Together they are sufficient to control the operation of a dual-klystron, pulse-modulated, high-power RF source.

Programming the processor to deal with the input and output signals in the same fashion as a hard-wired control circuit is only part of what can be accomplished. The fact that the operator/transmitter interface can be ergonomically designed (or programmed) is equally, if not more, important. Figure 20-9 shows a typical (and now almost obsolete) operator interface. It has a 12-in. color-video display, through which the control system can communicate with the operator, and 16 push buttons (or "programmable function keys"), through which the operator can communicate with the control system. Illustrated in the figure is but one of many display screens on which transmitter status is alpha-numerically presented. In addition to standard boxes—"go" (green, "OK"), "no-go" (red, "NOK")—to the right of each function, analog information is digitally presented,

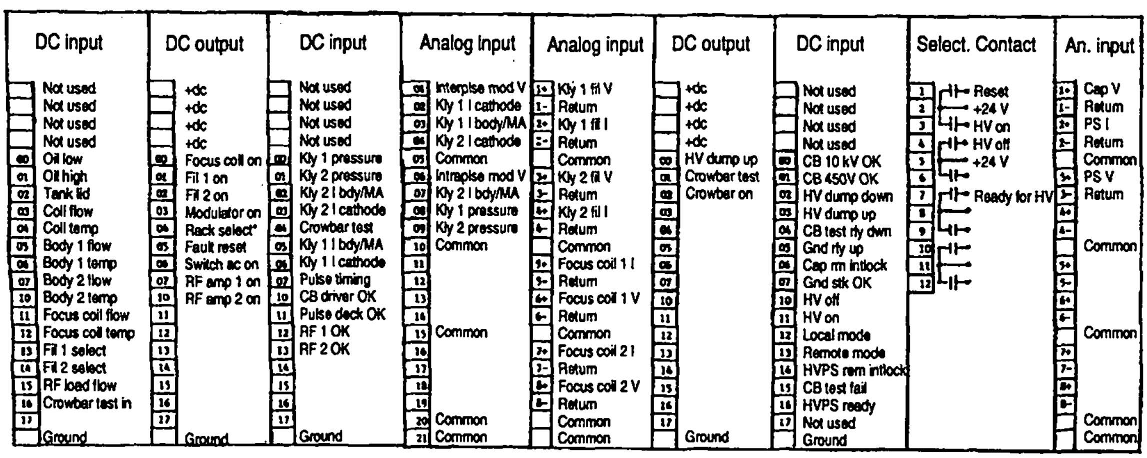
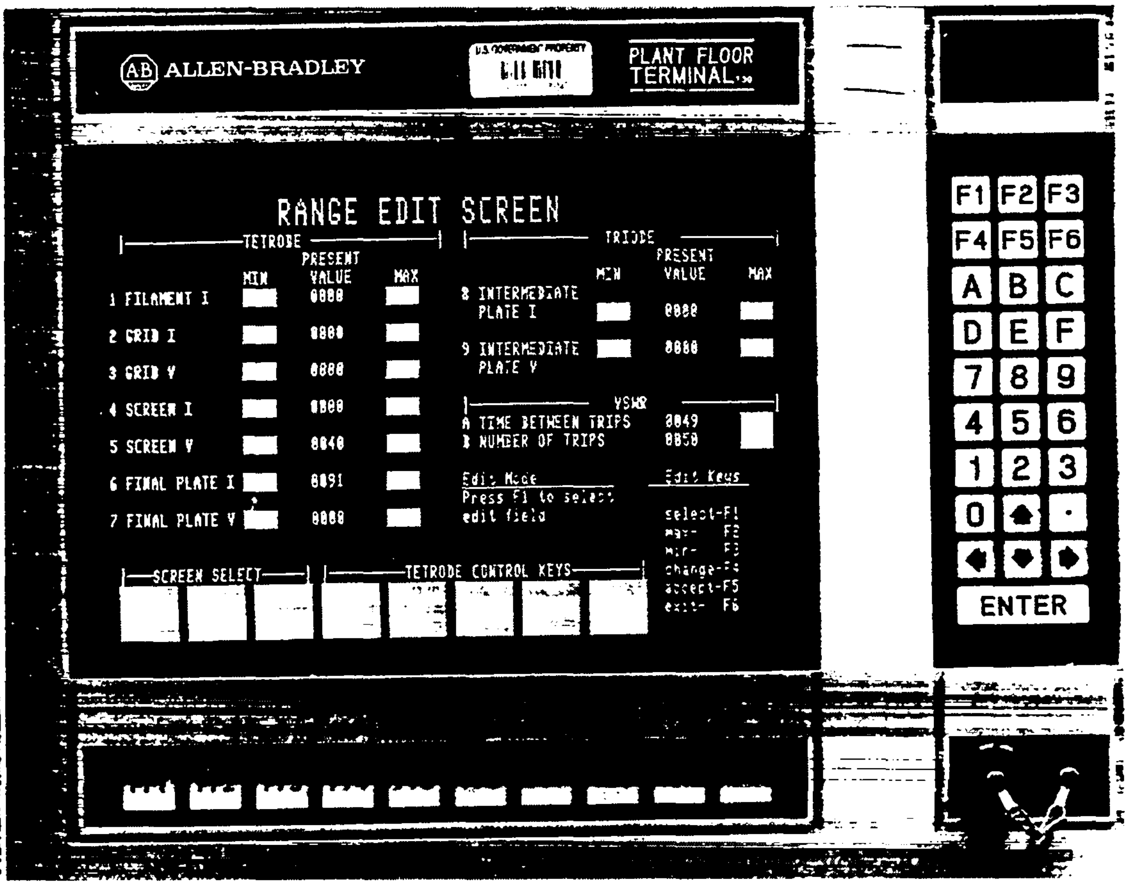


Figure 20-8. Actual inputs and outputs of programmable-controller modules for dual-klystron transmitter.



LANL Photo No. RN91-30.8027

Figure 20-9. Typical operator interface for transmitter control using programmable logic.

as well as the limits to which it may be compared. The klystron-beam voltage, for instance, is presented both as a value from 0 to 100 kV and as a solid horizontal bar graph. Filament voltage and current are compared with maxima and minima, which are automatically time-compensated for the 150% cold-filament-inrush current. The cathode-heating timer can be viewed as it runs down from 900 seconds to zero. Water-flow interlock switches have digital delays associated with them, so that a momentary drop-out will not dump the entire system until approximately 5 seconds has elapsed.

More modern implementations of PLC transmitter-control systems use special-purpose software to allow even more elaborate ergonomics, including a touch-screen interface and almost cartoonlike identification of functions. But the PLC approach offers a level of dynamism that can, if not carefully controlled, get out of hand. As easy as it is to program a function and its display response, it is just as easy to change it. And almost anyone can do it if they have a computer and a keypad. Configuration control for PLCs can become a nightmare.

The system illustrated was eventually married to a host computer that was programmed to dictate its behavior. Under fully automated computer control, the PLC changed beam voltage, peak beam current, RF power level and frequency, and even the phase and amplitude of RF load mismatch, making possible the fully automatic creation of RF bandwidth plots and Reike diagrams.

To many engineers, PLC transmitter control is the ultimate, and the concept gains proselytes easily (especially among younger technologists). The PLC will not do everything, however. In the first place it, is relatively slow. The processor does not simultaneously look at the inputs all of the time; it looks at them sequentially. It updates its outputs sequentially as well, and has a cycle time of approximately 10 ms. This is fast enough for many stimuli and responses but not for all. If the transmitter has an electronic crowbar, then it must be fired by dedicated and relatively high-speed circuits. The same is true of RF arc detectors, which should be capable of interrupting RF drive with a submicrosecond response time. Personnel-safety interlocks must also be dedicated and hard-wired in the fashion of direct interrupts. All of this additional information can be sent to the PLC, and it can respond to it redundantly, which is all the better.

20.5 High-speed instrumentation for transmitter monitoring

In order to generate transmitter-control decisions in the microsecond time domain and view on an oscilloscope pulsed currents and rapidly varying voltages that may have pulse durations and rise-and-fall intervals measured in nanoseconds, special components and assemblies are required. The two most important are the wide-band current-monitor transformer and the frequency-compensated voltage divider.

Figure 20-10 shows a simplified high-voltage circuit for a transmitter module using eight parallel-connected TWTs, each with dual-depressed collectors. To accurately view the pulse currents in the cathode and collector leads of an individual TWT requires the use of current-monitor transformers, which not only have a high degree of waveform fidelity but are compatible with the high volt-

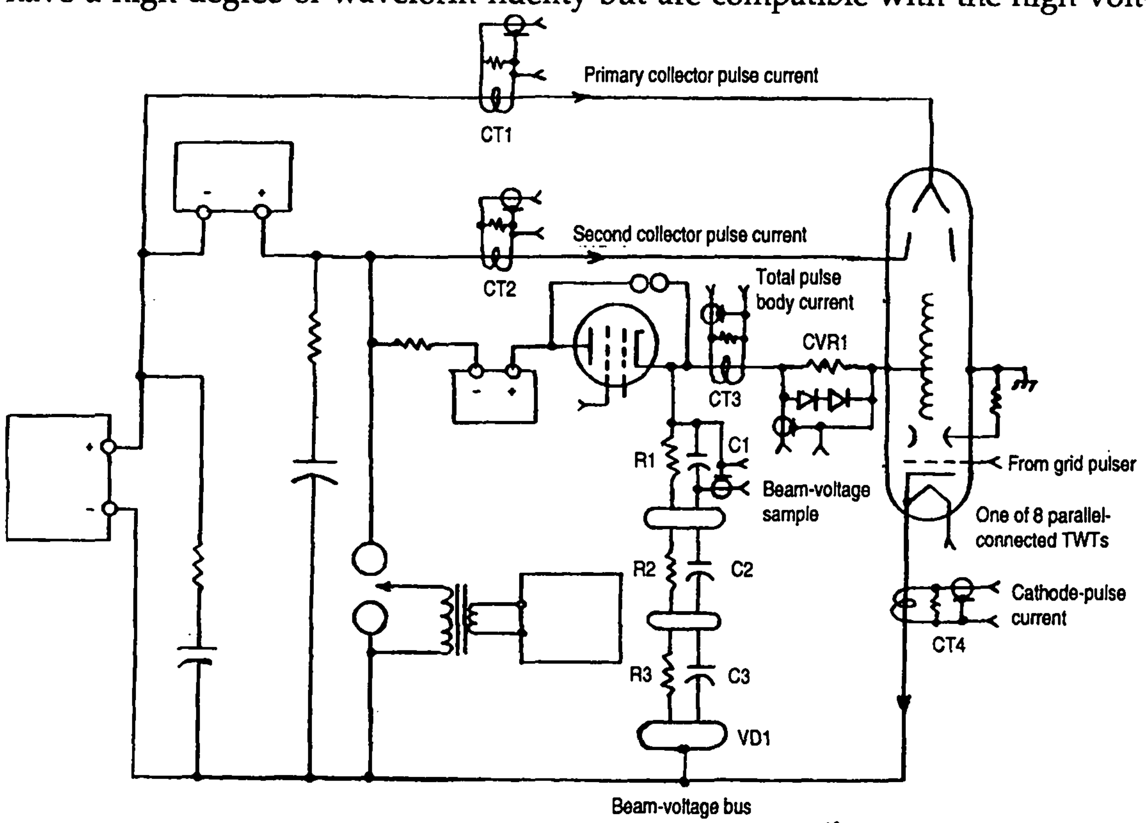


Figure 20-10. Current and voltage instrumentation for dual-collector TWT in multiple-tube transmitter.

ages applied to these electrodes. The ring-type current-monitor-transformer family—the members of which either look like doughnuts, flat-sided doughnuts, or a square metal box of modest depth with a large hole through it—is ideally suited to perform this task.

Inside the doughnut shape is an electrostatically shielded toroidal winding, one end of which is connected to the outer shield of a coaxial cable and the other end to a terminating resistor and the center conductor of a coaxial output connector. The shield, which is the outside of the doughnut, is actually in two parts. This is because it cannot be continuous, or else it would function as a short-circuited turn. The two half-shields overlap but do not touch. The conductor carrying the pulse current to be monitored is passed through the hole in the doughnut. A time-varying magnetic field surrounds the conductor when pulsed current passes through it. The rise time of the output from the toroidal winding is shorter than the delay time of the helical transmission line that the winding forms with the surrounding shield. This is due to the fact that voltage is induced into all of the toroidal turns simultaneously because they surround the current-carrying wire. The highest-performance versions of these transformers have distributed loading around the periphery of the winding to maximize the bandwidth. The low-frequency, or pulse-droop, performance is determined by the ringlike magnetic core, around which the toroid is wound. Such transformers not only have pulse droop but a volt-time product (or more correctly they have a primary current-time product limit beyond which they saturate and suddenly force the output to drop to zero). The most broad band of the transformers, those with the lowest product of useful pulse rise-time and flat-top droop rate, tend to have the smallest hole diameters and the lowest sensitivity. Useful rise times are as low as 10 ns and droop rates as low as 0.5% per ms. Typical voltage/current sensitivities range from 0.01 V/A up to 1 V/A, although very large transformers have been built to handle pulse currents in the 100,000-A range. Although these huge transformers have droop rates of 0.1%/ms, their sensitivity is only 0.1 mV/A.

The size of the hole in the doughnut is usually dictated by the voltage on the conductor, or "primary," that passes through it, and there is a wide range of hole sizes, up to a standard maximum of 10-3/4 in. Because of the coaxial geometry between wire and hole, the maximum voltage gradient will be at the outer surface of the wire. There is an optimum diameter for this conductor to minimize that gradient. It is the hole diameter divided by *e* (the base of natural logarithms). For the 10-3/4-in. hole, the optimum conductor outer diameter is 4 in. and the flashover voltage for sea-level air is 150 kV.

Most current-monitor transformers have 50-ohm internal impedances. When feeding a standard 1-Mohm oscilloscope-input resistance through a cable of 50-ohm characteristic impedance, the voltage sensitivity is as rated. The scope impedance produces a 100% voltage-aiding reflection, which is totally absorbed by the transformer internal impedance without re-reflection. When the scope end of the cable is terminated in a matching resistance, the sensitivity is cut in half. There is no practical advantage to terminating the scope or load end. Not all transformers have 50-ohm internal impedance. It is possible to combine higher sensitivity and lower droop rate if the internal impedance can be made greater.

For long-duration pulses of modest rise time, the mismatch will produce pulse distortion that is usually not noticeable. Values up to 500 ohms have been used successfully. The transformer output voltage, which is produced across an actual internal impedance, represents finite power that must be supplied by the primary circuit. This means that there is a finite resistance coupled into the primary conductor. It is usually of negligible magnitude, but it can be as great as 0.2 ohms.

Although the output of a high-quality current-monitor transformer can be a highly accurate replica of the sampled current, it can never have all of the properties of that current. In the first place, the monitor is a transformer and, therefore, can have no average value of voltage across its output. In a rectangular-pulse system, it is restricted to duty factors less than 50%. Furthermore, even in low-duty pulse systems, the baseline can never be zero volts. The area of the volt-time product above zero must be the same as that below zero. In a 10%-duty-factor system, for instance, the pulse itself may be reproduced with great fidelity, but if the transformer droop rate is small compared with the interpulse interval, it will start at a voltage that is negative with respect to ground by 10% of the pulse amplitude and return to that level when the pulse ends. This offset will increase as duty factor increases until it reaches a maximum value of 50% at 50%-duty factor. If the transformer droop rate is small compared with the interpulse interval, each output pulse will start at zero, the pulse top will droop (even if imperceptibly), and the pulse will under-shoot the baseline at the end of the pulse, exponentially recovering to zero before the next pulse. The requirement for equal volt-time areas will be met, but with an exponentially decaying negative voltage.

Returning to Fig. 20-10, note that the "body," or ground, current is monitored by both a current-viewing resistor, *CVR1*, and a current-monitor transformer, *CT3*. However, in this example there are eight tubes in parallel whose bodies must all be grounded. The current monitored, therefore, is the total body current of all of them.

The broad-band voltage monitor shown is a frequency-compensated voltage divider, *VD1*. In this example, it is used to view the TWT beam-voltage bus, nominally -40 kVdc. The output sample across *R1* and *C1* is typically in the range between 1/10,000 and 1/1000 of the bus voltage. The compensated divider is not only a resistive divider comprising *R1*, *R2*, and *R3* (and having a frequency response down to zero frequency) but a capacitive divider comprising *C1*, *C2*, and *C3* (and having theoretical frequency response of everything but zero frequency). If the two are connected in parallel and have the same division ratio, their combination has a theoretical total frequency response from dc to infinity. The equal division ratio is the same as the equal *R-C* product: $R1 \times C1 = R2 \times C2 = R3 \times C3$. The upper-frequency limit is determined by series inductance in both the resistive and capacitive legs. Clever component design can extend frequency response to the megahertz region. Some designs even use printed parallel-plate capacitance with the resistance mounted to the printed wiring in order to minimize inductance. These designs are more successful when insulated with dielectric oil because the printed-capacitance edge effects tend to produce very high electric-field enhancement. When these devices are air-insulated, it is customary

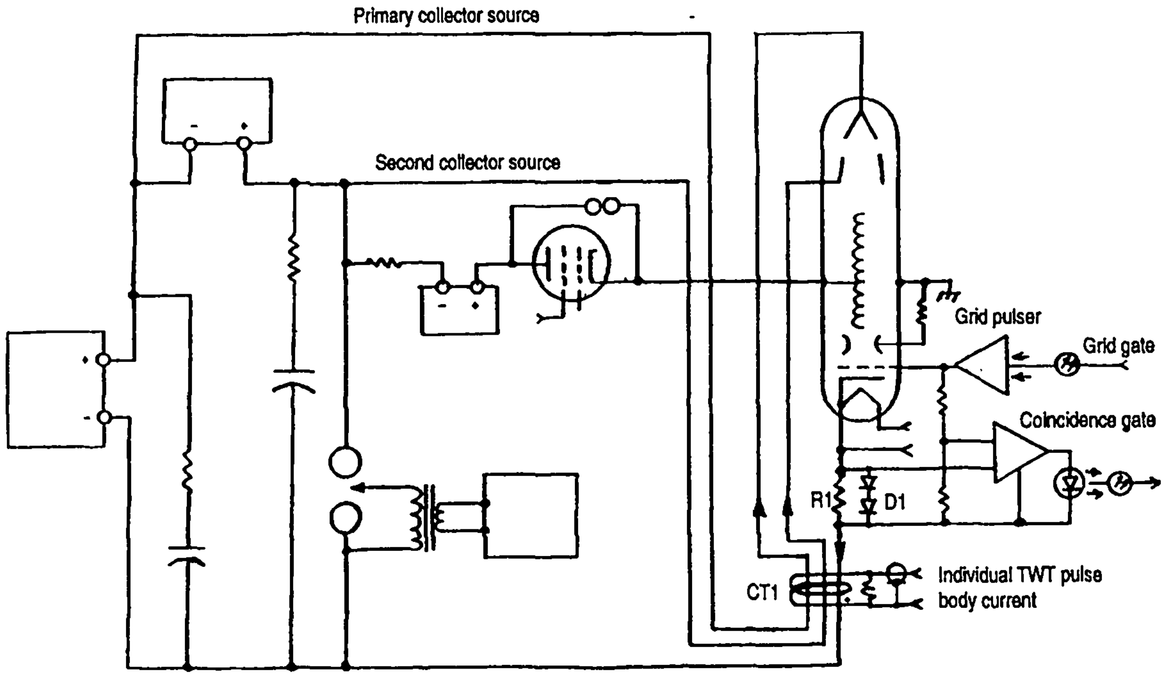


Figure 20-11. TWT of multiple-tube transmitter instrumented for individual-tube body current and interpulse cathode current.

to break up the divider into a number of layers with some corona-reducing metal shape, such as the doughnut, separating them. Two layers are schematically represented in the illustration.

A solution to the problem of monitoring body current for each individual TWT in the eight-tube transmitter is shown in Fig. 20-11. Body current in a microwave tube is current that does not go somewhere else. The "somewhere else" in the TWT illustrated is the two collectors. If the total collector current is subtracted from the cathode current, the remainder is body current (or body-plus-grid current). This subtraction can be accomplished vectorially by passing the cathode lead of each TWT through a current-monitor transformer window in one direction and the two collector leads through the same window in the opposite direction, as shown. The net magnetic field in the window will be proportional to the vector difference between the cathode and the two collector currents, which is the body current. (This is the same way that many household ground-fault-interrupter circuits work. Ground current is inferred from the difference between high-side supply current and low-side, or neutral, return current as subtracted by a current transformer.) However, the ground current indicated will also include the charging and discharging currents of the stray capacitances associated with the grid or any other modulating electrode supplied by the pulse modulator.

Although this technique of inferring body current is effective in most situations, sometimes reality is not what it appears to be. Consider a medium-power gridded TWT. It operates at very short pulse duration, typically 100 ns, and has rise-and-fall times of less than 10 ns. The beam voltage is typically 10 kV or so. The electron beam will be travelling at approximately 20% the speed of light. If the beam traverses one foot of interaction region between cathode and collector, it will take about 5 ns for the electron beam to make the trip. The collector-current waveform will be delayed from the cathode current waveform by 5 ns, and the subtraction process will give very large errors during the rise-and-fall intervals of the pulse because two large numbers are being subtracted to get a small one. Fortunately, in most cases it is the mid-pulse body current that is the most important, and its replica will not be distorted.

Not only does vector subtraction work in a current transformer but vector addition does too. The sensitivity of a transformer can be doubled by passing the "primary" lead through the transformer window twice in the same direction, or it can be multiplied by n by passing the lead through *n* times. However, the resistance coupled into the primary circuit will increase by the same factor.

Sometimes in a high-power gridded microwave tube it is crucial to be able to measure even small values of cathode current, down to zero frequency, and compare them with the timing of the grid pulse. In Fig. 20-11, that is the function of *R1*. Diode *D1* shunts the major portion of the intended pulsed beam current. The voltage across *R1* is compared with a portion of the grid-drive pulse. If it overlaps by too much, the system is shut down.

A major threat to the longevity of such a tube is the open grid, a situation in which everything is on while the grid is floating, not connected to anything. This is not just a hypothetical condition, especially in a transmitter that has 128 such tubes. What happens if the grid is unconnected is that it will be self-biased to a point near beam-current cutoff, called grid-leak biasing. Electrons striking the grid will produce a negative potential on it. The current will be infinitesimal because the circuit impedance is infinite. Beam current will be very small but continuous—and worst of all, super-focused. It will pass down the center of the beam tunnel, much like a laser beam, and will disperse only slightly when it leaves the beam tunnel, or interaction region. This beam current is capable of boring a hole in the center of the collector, a fatal occurrence for the tube.

20.6. High-speed crowbar-firing circuits

There are nearly as many low-level crowbar-firing circuit designs as there are circuit designers. Nevertheless, the circuit shown in Fig. 20-12 illustrates some of the features that such a circuit should have. The inputs to the circuit are outputs from current-monitor transformers. Three monitor points are shown for a single microwave tube, in this case one with a modulating anode. The pulse currents sampled are cathode, body, and modulating anode. A typical 1-2 MW peak-power modulating-anode klystron will operate at a beam voltage of less than 100 kVdc with peak-pulse cathode current of 30 A, body current of 0.3 A (or 1%), and modulating-anode current of 0.03 A (0.1%). The customary transformer sensitivities used are 0.1 V/A for cathode current (3 V peak, nominal), and 1 V/A for body and modulating-anode currents (0.3 and 0.03 V peak, nominal).

A 15-V, high-surge-current Zener diode directly shunts the output of each current-monitor transformer with no intentional series resistance at the input to each channel. Its purpose is to clamp the output voltage of each monitor transformer to 15 V peak. Peak instantaneous fault current due to an electron-gun arc in a transmitter like this is typically between 1000 and 2000 A (100 kV limited by 50 to 100 ohms). It will flow through both the cathode- and body-current moni-

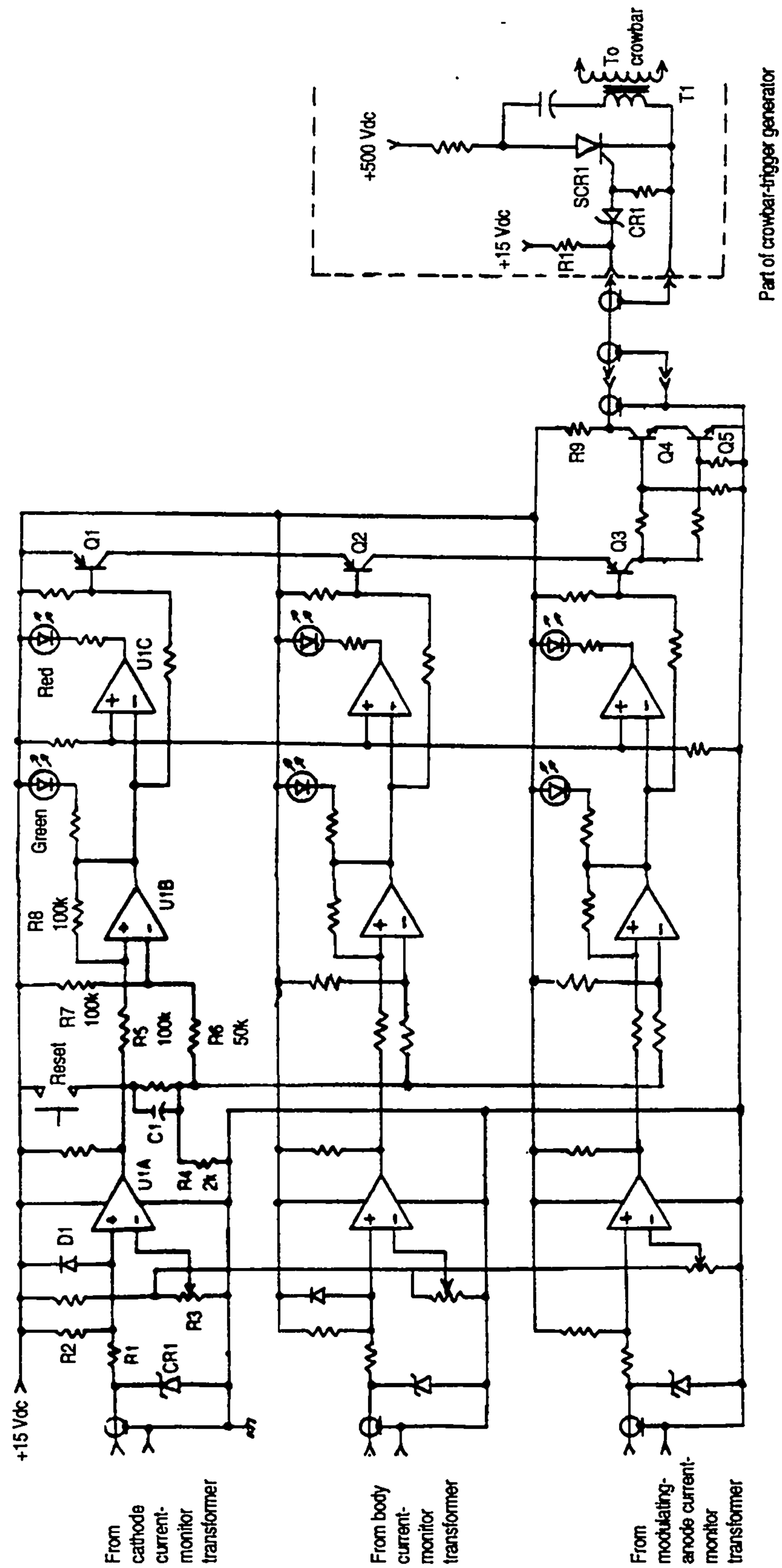


Figure 20-12. Crowbar-firing logic circuit using voltage-comparator integrated circuits.

tors. With no clamping, the peak output voltage of the cathode monitor would be between 100 and 200 V and the body-current monitor would be between 1000 and 2000 V. Neither output pulse would last very long, because both transformers would almost instantaneously saturate due to excessive $V \times t$ product. The desired input to the fault channel is no more than 15 V for a time duration that is at least long enough to insure that the circuit reacts. (The maximum peak current through the cathode-current-channel Zener diode is 200 V/50 ohms, or 4 A. The maximum peak through the body-current-channel Zener diode is 2000 V/500 ohms, or 4 A because the transformer is a non-standard high-impedance type.)

Each of the three channels is identical. The first stage is a voltage comparator, typified by *U1A*, that has an adjustable-voltage threshold set by potentiometer *R3*. Resistor *R1* limits current in the signal channel. If, for any reason, input voltage exceeds the dc rail voltage, the excess will forward-bias diode *D1* (one of the protective ploys first shown in Fig. 20-6). Resistor *R2* normally does nothing. It has a high value (100 kohm or more) compared with the very low internal impedance of the current-monitor transformer that is normally connected to each input. Should a transformer be disconnected, however, or never connected in the first place, *R2* will pull up the gate input to the rail. This will have the same effect as a massive fault. This strategy is consistent with the connectivity-safe philosophy.

Comparator *U1B* is connected as a set/reset flip-flop by means of resistors *R5* through *R8*. This use of a comparator is another high-threshold alternative to TTL logic. In the "reset," or normal, state of the flip-flop, the voltage at the negative input of *U1B* is the voltage across *R6*, which is about + 5V. The voltage at the left end of *R5* is a small fraction of a volt because *U1A* is normally turned on. The output of *U1B* in the reset state is also conducting, as we will see. The left end of *R5* is low and the right end of *R8* is low also, so the voltage at their junction, which is the positive input of *U1B*, must also be low. This is a stable state.

Any input voltage exceeding the threshold voltage of *U1A* will cause its output to go from low to high. When this happens, the left end of *R5* will be pulled high (the output pull-up resistor is small compared to 100 kohm). The right end of *R8* is still low at this moment, but the voltage at the resistor junction will rise to about 7.5 V. When it passes +5 V (the voltage at the negative input of *U1B*), the output will toggle from low to high, raising the voltage at the positive input momentarily to +15 V. As soon as the input fault voltage recovers to zero, the output of *U1A* will go low again, and the voltage at the positive input of *U1B* will stabilize at +7.5 V with its output high. This is also a stable state. Manual reset is accomplished by pushing the reset switch, which momentarily pulls up the voltage at the left end of *R6* through *C1*. The voltage at the negative input of *U1B* is pulled to +15 V, toggling its output from high to low, which is where the process began.

When the output of *U1B* is low, it enables current to flow in the base of *Q1*, part of an AND gate that includes *Q2* and *Q3*. When all three are conducting, series-redundant transistors *Q4* and *Q5* conduct as well, providing a low-voltage-drop pull-down, or sink, for current in the output circuit. The output of the crowbar-firing circuit is connected to *SCR1*, the first capacitor-discharge stage of

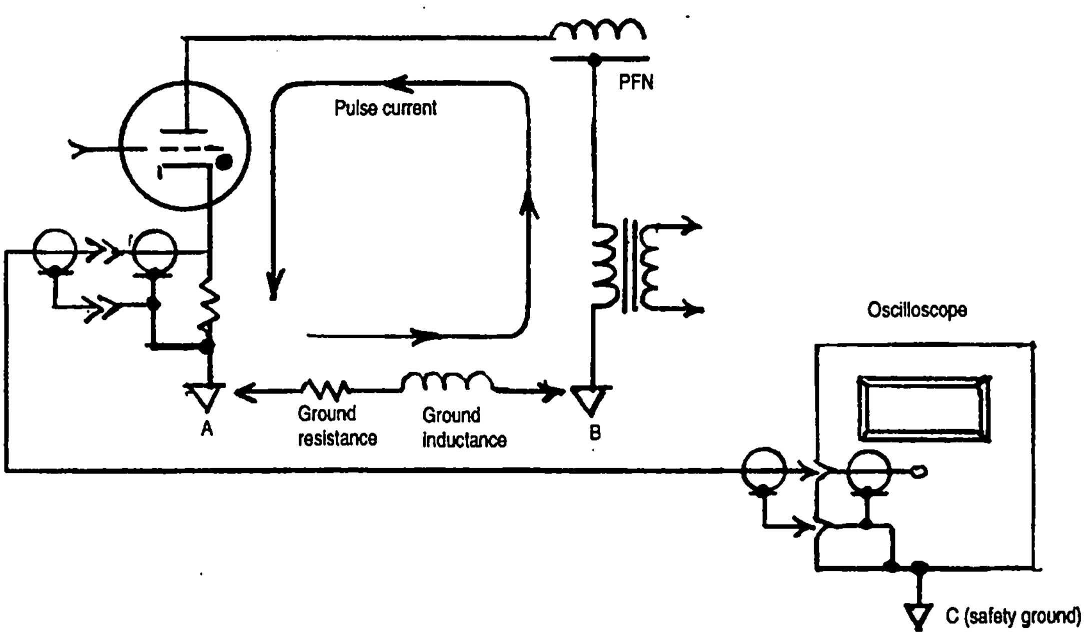


Figure 20-13. A typical source of ground-loop, or "ground-bounce," spurious signal.

the crowbar-trigger generator. The output of the capacitor-discharge stage initiates breakdown of the high-voltage electronic crowbar. A local power supply in the trigger generator will cause gate current to *SCR1* to flow through *R1* and *CR1*, unless their junction point is pulled down by conduction of *Q4* and *Q5*. The cable between the crowbar-firing logic and the trigger-generator must be connected, or the crowbar will fire (or the discharge capacitor will fail to charge to +500 V, a situation which is interlocked). This is more connectivity safety.

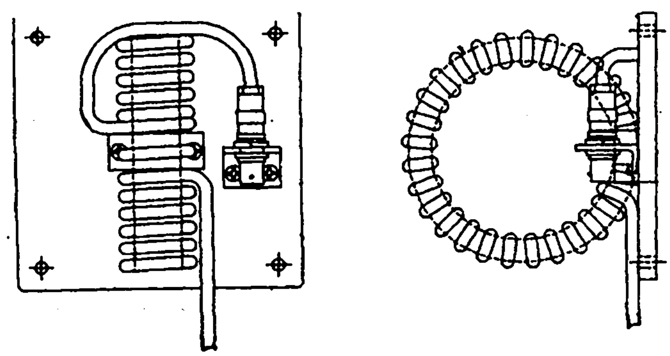
The third voltage comparator, *U1C*, is used as in inverter to illuminate the red LED that indicates a fault in that channel. The same information is conveyed to a PLC input module or comparable interface of a generic control system.

20.7 The pulse "balun"

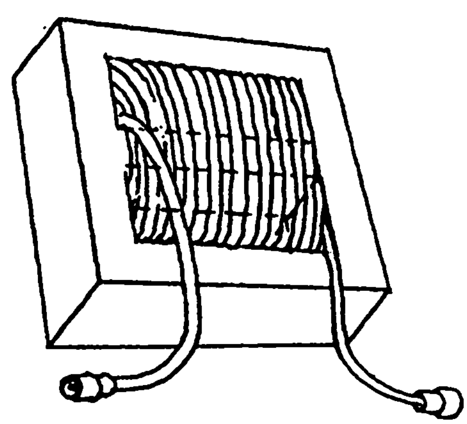
The last component to be discussed is the pulse "balun." The generic balun is a circuit element that allows a differential, or balanced, circuit to be connected to single-ended, or unbalanced circuit, or vice-versa. (The "bal" in its name is for balanced, and the "un" is for unbalanced.) The pulse balun is a circuit element that permits this to happen over the frequency domain associated with impulses. Its most common use is not to couple signals but to permit unwanted signals to be suppressed.

A classic example of a situation that will generate an unwanted common-mode signal is shown in Fig. 20-13. The offending circuit is part of a line-type modulator. The designer's mistake, whether intentional or not, is to use ground as the return conductor for the low-side of the high-current discharge circuit. By doing this, the designer subverts the ground; it is no longer ground. Ground, no matter how high its conductivity is intended to be—a copper-sheet floor notwithstanding—has resistance and, more importantly, inductance. Current will produce an $I \times R$ voltage drop, and change-of-current will produce an $L \times di/dt$ voltage drop between points *A* and *B*. If the discharge current is sampled by means of a CVR in the cathode of the discharge switch, referenced to point *A*, and the oscilloscope used to monitor the CVR voltage is connected to safety ground, point C, there will be shield current in the interconnecting cable that will superimpose all or part of the "ground-bounce" voltage on the composite signal that the scope thinks it is looking at. (An acid test for this situation is to touch just the outer conductor of the cable to the outer surface of the scope input jack. If there is a vertical scope trace, it is time for a pulse balun.)

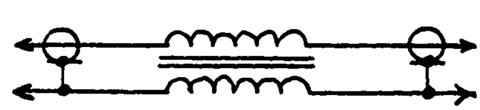
A pulse balun can take many shapes, but it is usually nothing more than a length of coaxial signal cable wound around a magnetic core. Figure 20-14 shows two common types. In Fig. 20-14a, the cable has been wound toroidally around a ring-like core. In Fig. 20-14b, a longer length of cable has been wound around a standard transformer iron core for greater low-frequency performance. The equiva-



A. Toroidally wound pulse balun



B. Pulse balun wound on conventional transformer core



C. Equivalent circuit of pulse balun

Figure 20-14. Aspects of the pulse balun.

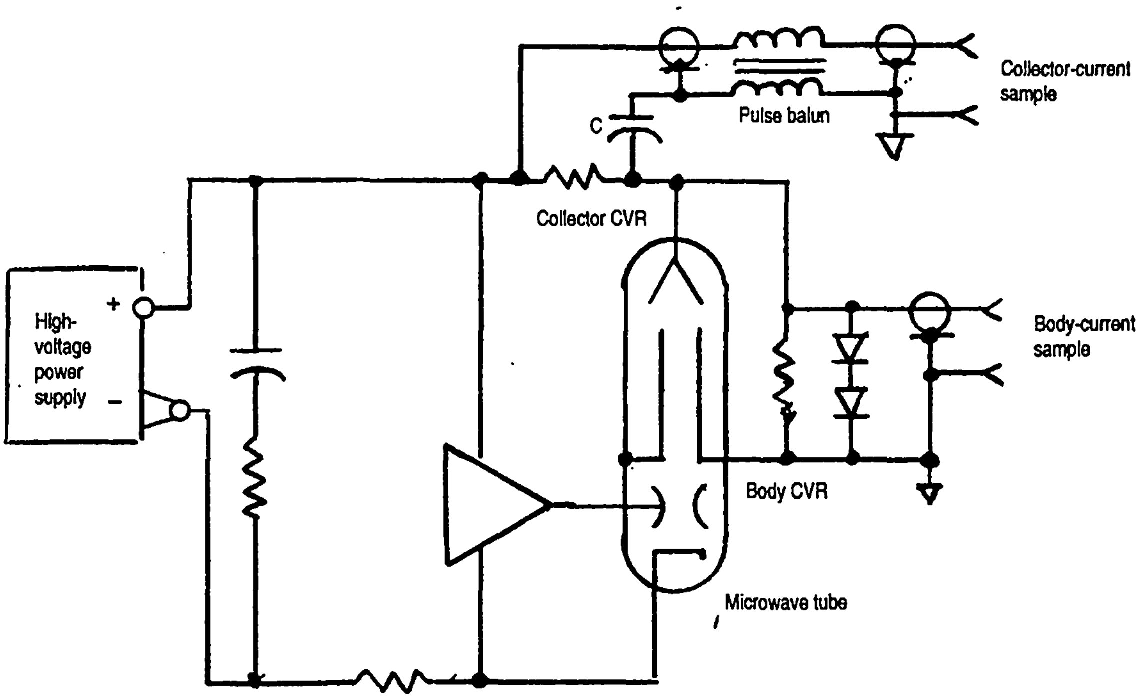


Figure 20-15. Pulse balun used to translate differential collector-current sample to a single-sided signal referenced to ground.

lent circuit is shown in Fig. 20-14c. (The pulse balun is actually the same thing as a bifilar transformer winding, but without the other winding or a bifilar inductor.) The two conductors of the cable—the inner conductor and shield—have unity coupling. Whatever voltage was between them at one end is there at the other end. The shield, however, now has series impedance. A varying voltage can exist between the shield at one end and the shield at the other end with current limited by the winding inductance.

If the cable of Fig. 20-13 had a pulse balun in series with it, the ground loop would be open-circuited by it. Shield current resulting from the differential ground voltage between A and B would be limited by the winding inductance of the balun, and the scope would respond only to the voltage between cable shield and inner conductor. There was a time in virtually all high-power installations, especially where current levels were in the multiple kiloampere range, when all coaxial-cable interconnections were made via pulse baluns.

Figure 20-15 shows another practical application of the pulse balun. It is in an application where a direct-coupled sample of collector current is required to guard against "collector boring" should the modulator fail to completely cut off beam current in the interpulse interval. Both ends of a CVR must float, however. One end is connected to the body-current CVR, and the other end is the insulated low-voltage return of the high-voltage dc power supply. The only ground reference is at the body of the microwave tube, as usual. A pulse balun can be used to develop outer-conductor winding inductance that limits the current flow that would shunt the body-current CVR. The body-current pulse-top droop rate would be determined by L/R, where L is the pulse-balun winding inductance and R is the body-current CVR value. An important detail is the coupling and dc-block-

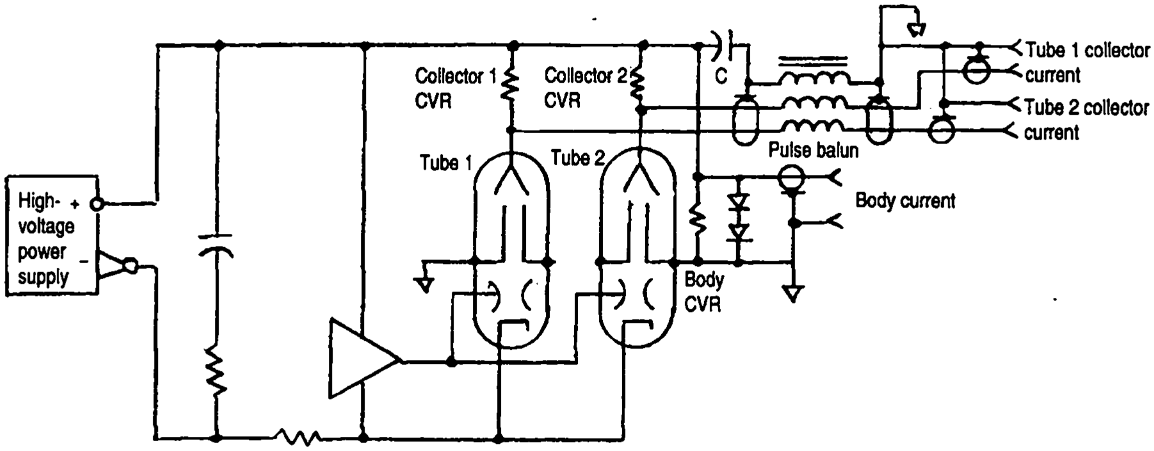


Figure 20-16. Use of trifilar pulse balun to translate differential collector-current signals of dual microwave tubes to single-sided signals referenced to ground.

ing capacitor C, which is in series with the outer-conductor of the balun. It does two things. First, it blocks dc magnetization of the balun core. In a high-duty-factor application this step increases the volt-time product obtainable before the core is saturated. More importantly, it prevents the balun inductance from short-circuiting the dc component of the body-current CVR voltage. The dc component appears both across the CVR and, in reverse-polarity, across C. Note that the balun center conductor can be connected to either end of the collector-current CVR to produce a ground-referenced output that is either positive- or negative-going. The connection illustrated gives a positive-going sample.

Figure 20-16 shows the use of a single trifilar pulse balun that uses a shielded, twisted-pair cable wound on a magnetic core to isolate the collector-current CVRs of two parallel-connected microwave tubes. The shield of the balun winding is common to the junction of both CVRs, so there is no alternative to a negative-going ground-referenced output. If positive-going output is required, two independent pulse baluns can be used. The winding inductance of a pulse balun for such an application can easily be more than 10 H. Assuming a 10-ohm body-current CVR, the flat-top body-current sample pulse will have a 1-s time-constant.