PROTECTING YOUR POWER SUPPLY

Components in modern power supplies cost too much to leave them unprotected from shorts, overloads, etc.

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THE COST of modern sophisticated power supplies is high enough to warrant as much consideration for their protection as that given to their rectification and regulation circuits. While fuses and circuit breakers have been the traditional means of protecting power supplies, they are often not fast enough to prevent solid-state devices in newer supplies from destructing due to overloads and the like. The devices most able to protect semiconductors are other semiconductors.

Here are ways to protect a power supply from the three primary causes of failure: shorted output; shorted filter capacitors; and excessive current through the load. There are two general methods of protection. The first is the control of the transformer's primary circuit. The second is the removal of base drive from a transistor in series with a load.

In either case, we will assume that the protection circuit is part of a more complex power supply. High-current power supplies are used in some examples simply to indicate that the methods of protection are not limited to low-current applications. Obviously, devices with lower current and power ratings can be used where possible.

Primary Circuit. The block diagram in Fig. 1 shows a basic method of power supply protection. Characteristic of this arrangement is the triac in series with the primary of the transformer. During normal operation, the trigger control allows the trigger circuit to apply a brief gate signal to the triac for every alternation of the ac line voltage. After the triac is turned on by the gate signal, it remains on for the complete half cycle until the zero-crossing point is reached at the end of the alternation.

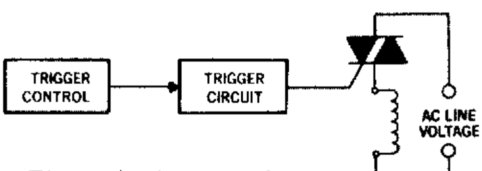


Fig. 1. A triac controls voltage across primary.

If the trigger control inhibits the trigger circuit while the triac is conducting, the triac cuts off when the line voltage approaches zero. It remains off until another gate signal is applied. Hence, the ac input to the transformer can be removed within a half cycle of the line voltage by designing the transformer's secondary circuit to inhibit the trigger circuit when a filter-capacitor short or supply output overload occurs.

The circuit shown in Fig. 2 is one type of control technique used in the primary circuit of a power supply. Under normal conditions, the gate of triac *Q1* receives a brief gate signal from the *IC2* zero-voltage switch at the beginning of each line alternation while the line voltage is near zero. Resistor *R1*, in series with the MT2 terminal of *Q1* and gate terminal of *Q2*, permits a continuous flow of alternating current through the gate of *Q2*. The primary of *T1*, in series with *Q2*, receives the full ac line voltage under these conditions.

Zero-voltage switch *IC2* can be used to provide pulses that are synchronized with the time of zero voltage in the ac cycle to the gate of a triac. Triac firing can be inhibited by the application of a positive (TTL-compatible) voltage to pin 1 of *IC2*.

The triple 3-input NAND gate used for *IC1* converts short-circuit logic-0 conditions to a logic-1 condition for inhibition of *IC2*. (A 5-volt dc supply was used for the IC's power and, consequently, for the inhibit signal.)

The inhibit signal appears at pin 1 of *IC2* when points A or B (at *Q3*) are shorted to point O (common). With *IC2* inhibited, *Q1* cannot provide ac to flow through the gate of *Q2*. When the line voltage falls to zero at the end of the alternation, during which the short occurs, *Q2* will cut off and remain off. After the short condition is removed, *Q1* turns on with the next gate signal from *IC2* and the system returns to normal operation.

If no filter capacitors were used (as in a simple battery-charging circuit), the self-resetting action would take place within one alternation of ac line voltage. Unfortunately, the inclusion of filter capacitors in the secondary circuit causes a resetting time lag on the order of one second for each 1000 μF used. If the resetting time is of no concern, no other consideration need be given this point. If you desire quick resetting time, you can do one of two things: First, include a dpst reset switch to momentarily break the connections between points A and B and *IC2*. Secondly, you can omit the connection between point A and *IC1* and include an isolation diode in the secondary circuit (Fig. 3); *IC2* will not, however, be inhibited by a shorted filter capacitor.

The circuit in Fig. 2 will not reset if a short occurs across the output terminals while a load is connected. In such a case, the load must be removed, or a reset switch must be used as explained above.

Although *Q2* will remove power from *T1's* primary immediately when the output terminals of the supply are shorted, a spark will occur. The amplitude of the spark can be considerably reduced by incorporating the transistor stage shown in Fig. 3.

Removing Base Drive. If a transistor is placed in series with the output terminals of a power supply, an ar-

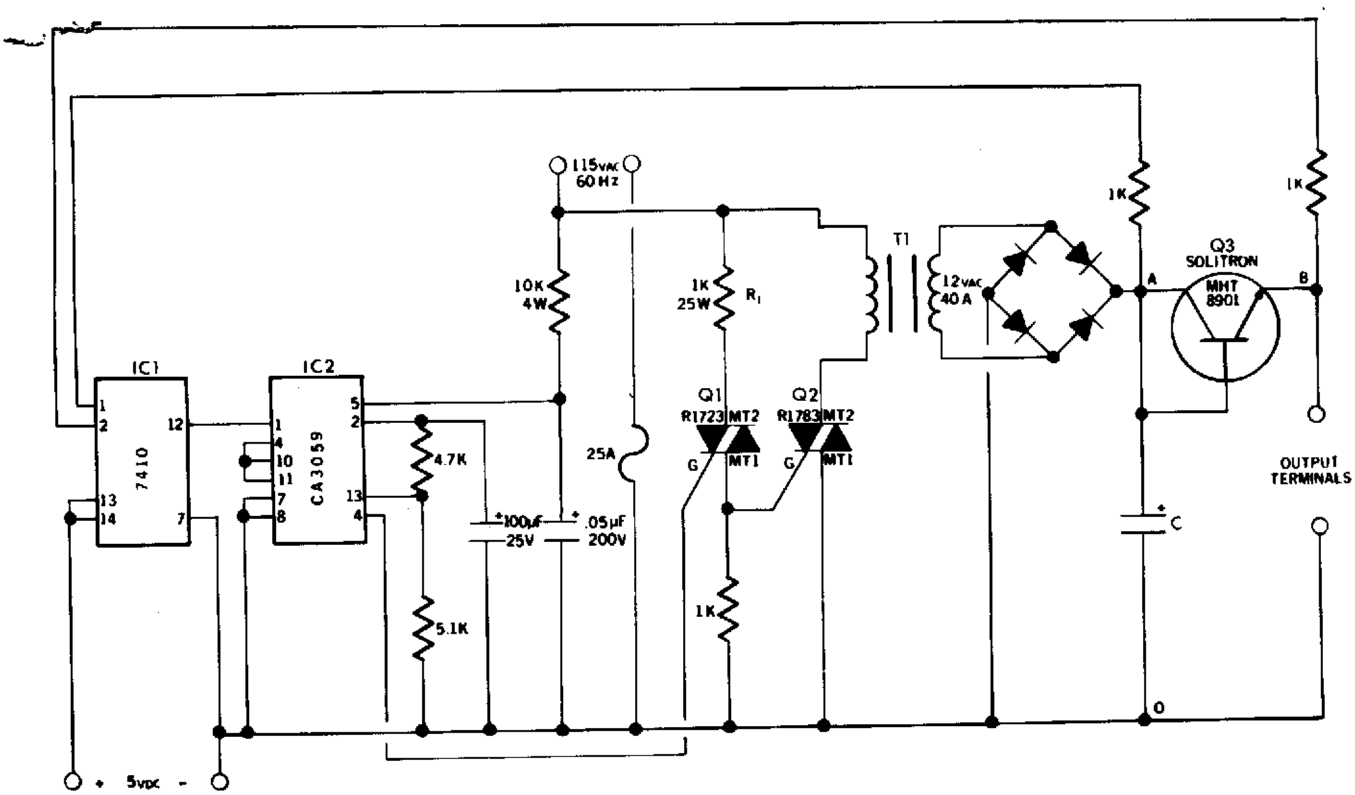


Fig. 2. One type of control technique used in the primary circuit.

rangement commonly used in series voltage regulators, the secondary circuit can be turned off by any action that removes base drive from this transistor. This can be done by shunting the base to ground with an SCR, optical coupler, or another transistor.

A method of removing base drive with an SCR when a desired maximum flow of current is exceeded is shown in Fig. 4. By varying R_x and R_s, the transistor can be cut off at any desired level of current flowing through the output of the supply.

Under normal operating conditions, the transistor is biased on by R_b. The current flowing through the output develops a voltage drop across R_s. Because a very low amplitude gate voltage is needed to trigger on the SCR, the resistance and power rating of R_s can be relatively small for high-current applications. The voltage drop across R_s is used to provide a gate signal for the SCR that is proportional to the level of the current flowing through the load. If the resistance of R_s is high enough, the level of load current at which the SCR shunts the transistor's base drive to ground can be varied by R_g. When the voltage drop across R_s is sufficient to trigger on the SCR, the transistor cuts off within microseconds. (A reset switch must be provided as shown to return the circuit to normal operation.)

Because of R_b, the transistor must operate in the active region. If the resistance of R_b is too low, the SCR will be required to handle a large current. Conversely, if the resistance of R_b is too high, the transistor will be forced to dissipate considerable power. Usually, a value for R_b must be chosen to keep the transistor's power dissipation and the current through the SCR at reasonable levels. The necessary current rating of the SCR can be determined (after R_b is chosen to provide the desired transistor power dissipation) by dividing the input voltage by the value of R_b.

Assume you're working with the following components and conditions: V_{in} = 34 volts dc, C = 18,300 µF (40 V), R_b = 30 ohms (50 W), R_g = 1780 ohms, R_s = 2.2 ohms (220 W), SCR = 2N682, and Q = HEP S7000. Here, the SCR will trigger on when the current reaches 10 A. You can also measure the following parameters: V_{CE} = 11.5 V dc, V_{BE} = 1.5 V dc, I_B = 350 mA, and I_{SCR} = 1.1 A. And the power dissipation of the transistor can be found by using the formula $P_D \approx V_{CE}I_C$, which would yield 115 watts.

Light-emitting diodes can be switched on and off in nanoseconds, and optical couplers with transistor detectors can switch at speeds of 2 to 5 µs. It is logical, therefore, to consider a protection system based on these high-speed devices. A typical optical coupler protection circuit is shown in Fig. 5.

It is not necessary for the series transistor in Fig. 5 to dissipate large amounts of power because this transistor (*Q1*) can be operated in or near the saturation region. Heavy base drive is applied to *Q1* through *Q2* according to the formula $I_{BQ1min} = I_{CQ1}/h_{FEQ1}$, where I_{BQ1min} is the minimum base current that assures saturation of *Q1*, I_{CQ1} is the maximum expected collector current, and h_{FEQ1} is the minimum expected h_{FE}. Transistor *Q2* is used to supply base drive for *Q1* so that only *Q2*'s relatively small base current need be shunted to ground to turn off *Q1*.

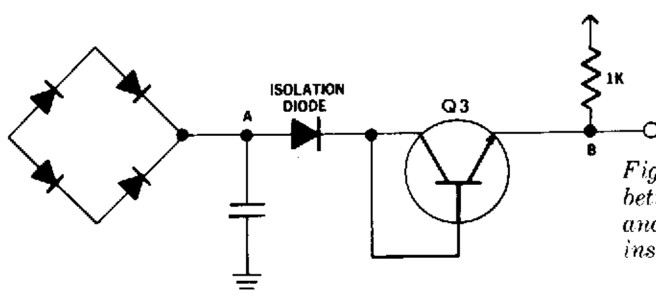


Fig. 3. Isolation diode between filter capacitor and transistor permits instantaneous reset.

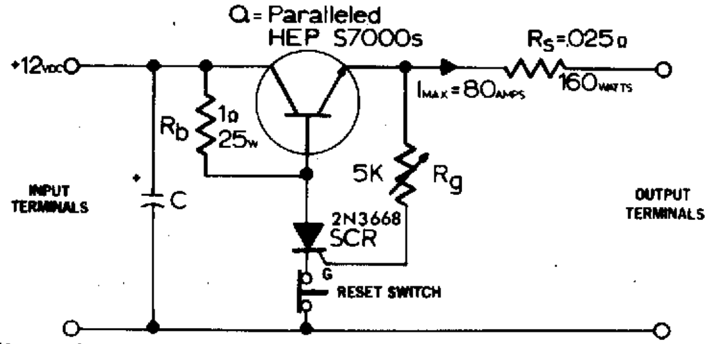


Fig. 4. A gate signal for the SCR is developed by the current through R_S.

There are many variations of the circuit shown in Fig. 5, but we will limit our discussion to this specific circuit configuration. It should be noted that R_S can have a value much lower than 1 ohm, which results in a lower wattage rating for this resistor. The optical coupler should consist of an infrared LED and a silicon transistor detector.

After assembling the circuit as shown, disconnect the anode lead of the LED from point A. Power up the supply, and monitor $I_{C(Q1)}$ and $V_{CE(Q1)}$ while decreasing R_B until *Q1* goes into saturation. Then adjust R_B until *Q1* is operating in the active region, just short of saturation. (This speeds up the cutoff action of *Q1* and keeps the photodetector current low.) When adjusting R_B, monitor $I_{B(Q2)}$ to make sure you don't exceed the current rating of the photodetector.

Connect the LED's anode back to point A and short the output of the supply. Adjust the R_{LED} control for an $I_{C(Q1)}$ short-circuit current of 15 mA. (Actually, the short-circuit current can be set to about 1 mA, but the adjustment of R_{LED} becomes critical for currents below 15 mA.)

With the adjustments performed as described and a 5-ohm value for R_L, $I_{C(Q1)}$ would be 640 mA. Reducing R_L to 3.33 ohms would drop $I_{C(Q1)}$ to 27 mA and $V_{CE(Q1)}$ to 4.4 V. This yields a 0.195-W $P_{D(Q1)}$. This current-limiting circuit produces an I_C versus R_L curve with a very steep slope, which results in very little overshoot of the desired maximum current.

With R_B and R_{LED} properly adjusted, *Q1* will operate in or near the saturation region with a heavy base drive supplied by *Q2*. If the current through R_S exceeds the maximum for which the circuit is adjusted, determined by the resistance of R_S, R_B, and R_{LED}, the LED will emit enough light to reduce the resistance of the photodetector. The result is that *Q2*'s base drive will be shunted to ground and the transistor will be cut off.

Because this circuit is very temperature sensitive, a reset switch must be provided as shown. When power is first applied to the system, no current flows through *Q2* until the reset switch is operated to momentarily disconnect R_P from ground. After a few minutes warm-up, the system is self-resetting.

A Current Limiter. The self-resetting circuit shown in Fig. 6 lacks some of the advantages of the previous circuits. Transistor *Q1* is again in series with the load, while *Q2* supplies sufficient base current to keep it operating in the saturation region for a significant range of loads.

In the following discussion, we will assume that a wide range of loads will be applied to the output of an unregulated power supply rated at 40 amperes. (If the load is to be fixed, circuit components can easily be chosen so that *Q3* provides a sharp turn-off of *Q1* if $I_{C(Q1)}$ increases beyond a chosen maximum. However, if the load is variable, R_C must be selected so that *Q1* passes the desired range of currents, with R_B and R_E chosen to provide rapid turn-off for the loads that will cause excessive current to pass through *Q1*.)

The graphs shown in Fig. 7 are plots of $I_{C(Q1)}$ versus R_L for the circuit shown in Fig. 6. From plot A, it can be seen that as R_L is decreased (increased load), the current through *Q1* increases to a maximum of 14.5 A when R_L is 0.3 ohm. Decreasing R_L further yields a reduction in $I_{C(Q1)}$ instead of an increase. This $I_{C(Q1)}$ decrease with increasing load continues until R_L is 0.25 ohm, at which point the circuit becomes unstable. When a load of 0.25 ohm is connected to the supply's output, $I_{C(Q1)}$ momentarily goes to 12 A, after which *Q1* turns off and $I_{C(Q1)}$ reduces to zero. For load resistances less than 0.25 ohms, *Q1* is in the cutoff region and $I_{C(Q1)}$ is zero.

Plot B, an expanded view of $I_{C(Q1)}$ for small values of R_L, shows how the *Q1* collector current varies when load resistance approaches zero. When R_L is 0.3 ohm, $I_{C(Q1)}$ is at the maximum 14.5-A value. For loads between 0.3 and 0.25 ohm, $I_{C(Q1)}$ decreases almost linearly. Load resistances of less than 0.25 ohm are a virtual short circuit at the output terminals of the supply and cause *Q3* to keep *Q2* at cutoff.

The operation of the circuit is quite

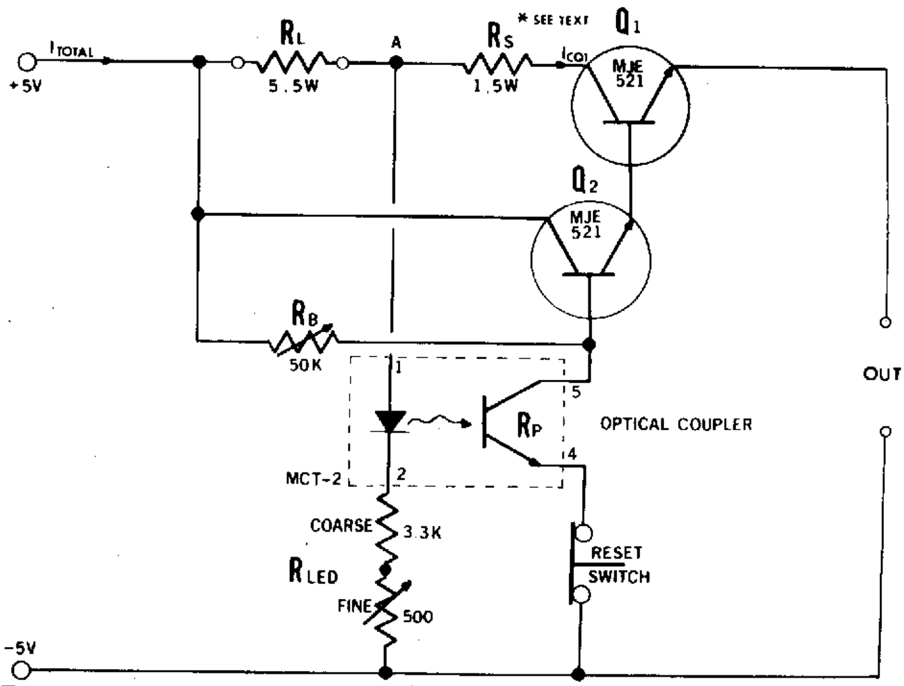


Fig. 5. Opto-coupler controls Q2 on basis of current in Q1.

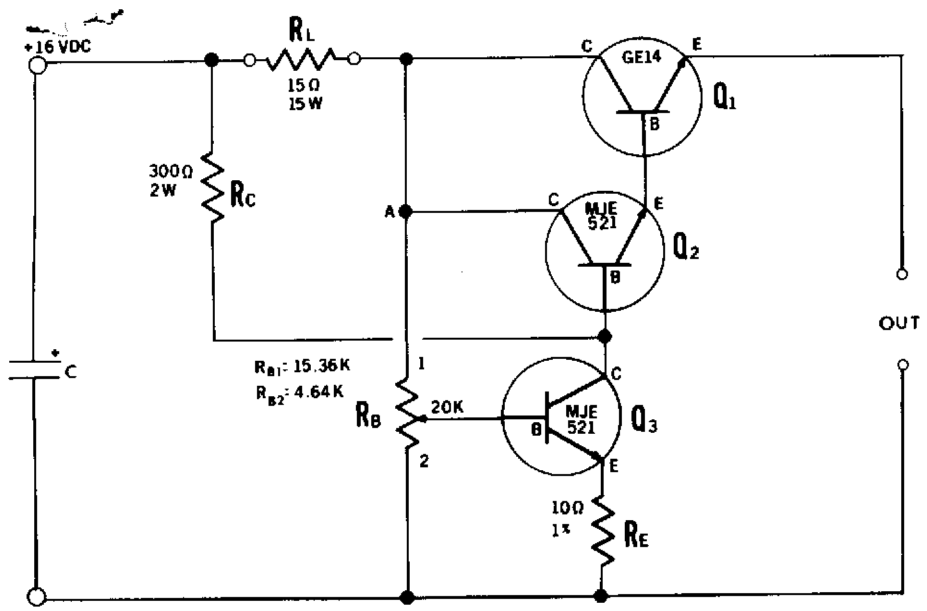


Fig. 6. A self-resetting current limiter with short-circuit protection.

TABLE I—CIRCUIT PARAMETERS FOR DIFFERENT LOAD RESISTORS						
	(A) R_L = 15 ohms			**(B)** R_L = 0		
Parameter	**Q1**	**Q2**	**Q3**	**Q1**	**Q2**	**Q3**
V_{CE}	0.61 V	0 V	1.49 V	16 V	15.8 V	0.25 V
V_{CB}	−0.143 V	−0.74 V	1.3 V	15.8 V	15 V	−0.35 V
V_{BE}	0.75 V	0.74 V	0.17 V	0.42 V	0.385 V	0.65 V
I_C	1.09 A	38 mA	0	23 mA	14 μA	48 mA
I_B	9 mA	46 mA	0	15 μA	9 μA	0.7 mA
I_{RB1}	32 μA	—	—	1 mA	—	—
I_{RB2}	32 μA	—	—	0.3 mA	—	—
V_{in}	16 V	—	—	16 V	—	—
V_L	15.9 V	—	—	0 V	—	—
P_D	0.7 W	0 W	0 W	0.37 W	2×10^{-4} W	0.012 W

straightforward. However, the adjustment of R_B and value of R_E that allow self-resetting are critical. Before R_B and R_E can be adjusted to the proper values, R_C must be chosen to allow the desired maximum $I_{C(Q1)}$. To select R_C, disconnect point 1 of R_B from junction A. The minimum value of $I_{B(Q1)}$ to keep *Q1* in saturation can be estimated from the formula $I_{Bmin} \approx I_C/h_{FE}$. For a maximum of 15 A for $I_{C(Q1)}$ using a GE14 transistor (h_{FE} = 45) for *Q1*, I_{Bmin} would be 0.333 A. The minimum value of $I_{B(Q2)}$ is found by using the $I_{B(Q1)min}$ value for $I_{C(Q2)}$. Hence, using an MJE521 transistor whose h_{FE} is 40 for *Q2*, we obtain $I_{B(Q2)min}$ = 0.33/40 = 8 mA. Then, determining the value for R_C we have $R_C = V_{in}/I_{B(Q2)}$ = 15/0.008 = 1875 ohms.

You can select a potentiometer for R_B by using the formulas given in "A Simple Method For Biasing Transistors" (June 1975). Let the base voltage of *Q3* be approximately 0.7 V, $I_{C(Q3)} = I_{B(Q2)}$ = 8 mA, and $I_{bias} \approx 0.1 \times I_{C(Q3)}$ = 0.8 mA. Assume that R_B consists of two resistors, *R1* for the portion of the pot above the wiper and *R2* for the lower portion. Now, $R2 = V_{base}/I_{bias}$ = 0.7/0.0008 = 875 ohms. Then for *R1*, consider the output of the power supply to be shorted and determine *R1* from the formula $R1 = (V_{in} - V_{base})/I_{bias}$ = (15 − 0.7)/0.0008 = 17,875 ohms. Adding the results obtained, we end up with a total resistance of 18,750 ohms. A

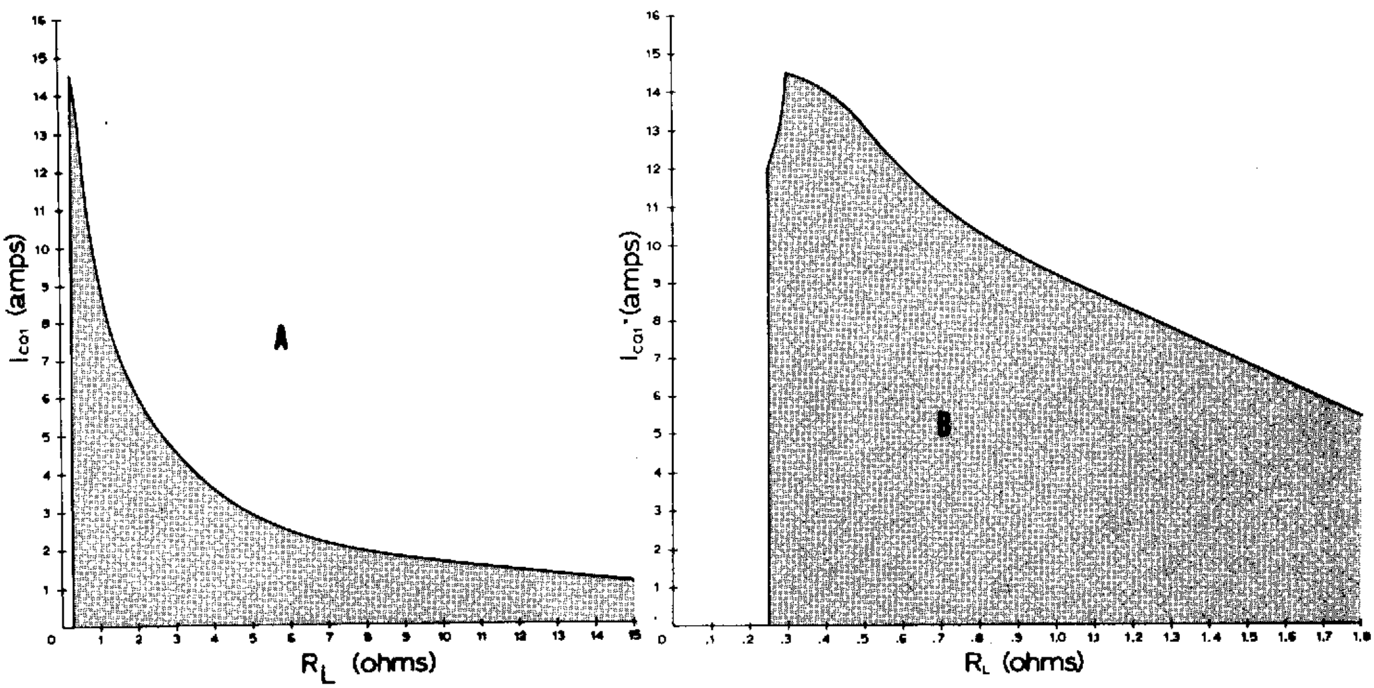


Fig. 7. In (A) load varies from 15 to 0 ohms; (B) load approaches zero.

standard 20,000-ohm potentiometer can, therefore, be used for R_B.

The adjustment of R_B is accomplished by connecting point 1 of the pot to junction A and installing a resistance decade box set to zero ohms as R_E. With a 15-ohm load connected to the output of the supply, $I_{C(Q1)}$ should measure 1 A. (The collector current for *Q1* should be monitored during all adjustments. Also, the power supply should be shut off when installing and removing ammeters.) Start with a 0-to1-A ammeter and adjust R_B until $I_{B(Q3)}$ is nearly zero. Continue in this manner, using a more sensitive ammeter, until $I_{B(Q3)}$ is exactly zero.

Remove the ammeter from the supply and connect the base of *Q3* as shown in Fig. 6. The decade box in the emitter leg of *Q3* should still be set to zero. Short the supply's output terminals; $I_{C(Q1)}$ should drop to zero. If the current through the collector of *Q1* doesn't drop to zero, $I_{B(Q3)}$ or R_E has not been set to zero. Remove the short from across the supply's output, leaving only the 15-ohm load. The collector current of *Q1* should remain at zero; if it doesn't and the circuit self-resets, no R_E is needed. (This is not likely to occur.) With no self-reset, increase R_E in 1-ohm steps until $I_{C(Q1)}$ goes back to 1 A. Short the output of the supply again; $I_{C(Q1)}$ should go to zero and the circuit should restart when the short is removed.

The higher the resistance of R_E, the greater will be the off current of *Q1*. With repeated trimming of R_B and R_E, the collector current of *Q1* when the supply's output is shorted can be brought down to 5.4 mA. The circuit can be made much less dependent on the setting of R_B, and R_E can be zero, if a reset switch is used to return the circuit to normal operation after an overload. In this case, R_B would be adjusted as before, and a reset switch would be operated to momentarily break the R_B connection to junction A. The pot could then be trimmed to yield a minimum collector current in *Q1*.

The capacitor shown across the input of the circuit in Fig. 6 is not part of the protection system. It is simply representative of the filter capacitor in the power supply. Under normal conditions, *Q2* supplies the base current to *Q1*. Both *Q1* and *Q2* operate in the saturation region to assure that full power is delivered to the load. The A section of Table I shows the measured and calculated parameters for *Q1, Q2, and Q3* for the 15-ohm load, R_L.

If R_L is reduced to zero (shorted output terminals), *Q3* will conduct and *Q1* and *Q2* will be driven into cutoff. The B section of Table I shows the parameters for the transistors when $R_L = 0$.

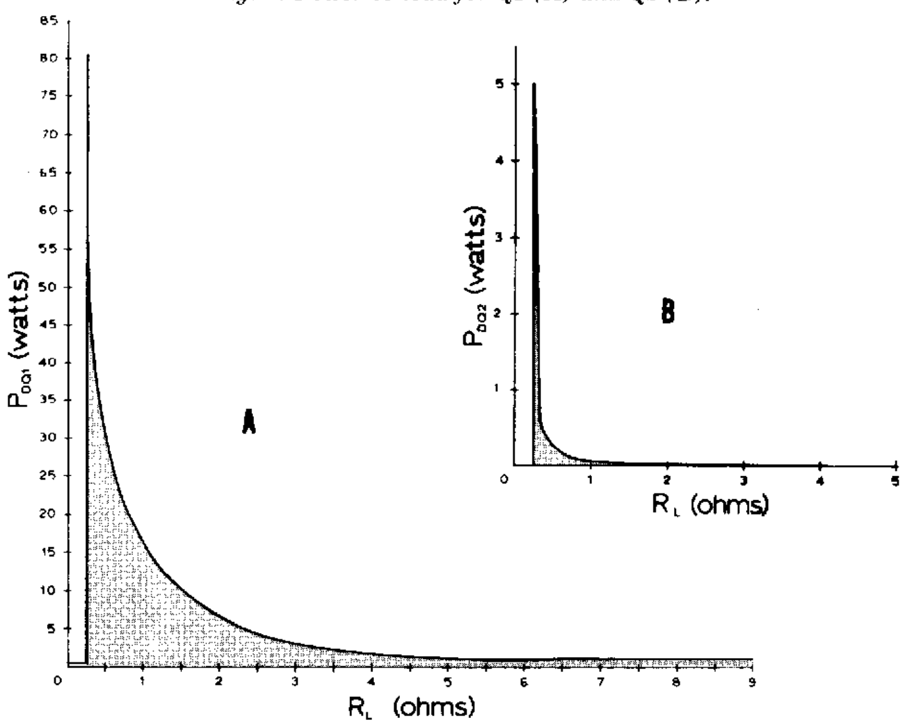
For intermediate values of R_L, the transistors pass through all three regions of operations. These regions and the loads that cause the transitions are listed in Table II. When R_L is reduced to 1.66 ohms, enough forward bias is applied to the base of *Q3* to bring it out of cutoff. As R_L is further reduced, $I_{B(Q3)}$ increases until, finally, when R_L is 0.296 ohm, *Q3* is shunting a large enough portion of *Q2*'s base drive to ground to cause both *Q2* and *Q1* to come out of saturation and begin operating in the active region. Eventually, when R_L is reduced to less than 0.25 ohm, *Q3* is driven into saturation and *Q2* and *Q1* go into cutoff.

TABLE II—OPERATING REGIONS FOR DIFFERENT LOAD RESISTORS

R_L (Ohms)	Region Of Operation		
	Q1	Q2	Q3
15	saturation	saturation	cutoff
1.66	saturation	saturation	active
0.296	active	active	active
0.25	unstable	unstable	unstable
0	cutoff	cutoff	saturation

Because *Q1* and *Q2* must operate in the active region, even for so narrow a range of loads as from 0.296 to 0.25 ohm, the voltage drop across these transistors over this range causes the power dissipation of the devices to increase tremendously while they are operating in the active region. Plots of P_D versus R_L for *Q1* and *Q2* are shown in Fig. 8. Plot A shows that, for maximum protection, *Q1*'s P_D rating should be greater than 80 watts at the desired operating temperature. Plot B shows that *Q2*'s P_D rating should be greater than 5 watts.

Fig. 8. Power vs load for Q1 (A) and Q2 (B).



Although the P_D ratings of *Q1* and *Q2* must be much greater than is necessary while the transistors are operating strictly at saturation, they need not handle the power dissipation that would be necessary in an unprotected series voltage regulator. For example, with a 14.5-A $I_{C(Q1)}$ and a 16-volt V_{in}, the P_D rating of a transistor used as an unprotected series regulator would have to be 232 watts at the operating temperature under shorted conditions. This would require a very expensive transistor.

Conclusion. We have proposed only a few of the many possible ways of protecting the more expensive and fragile components found in modern power supplies. Proper utilization of the proposed circuits, individually or in combination, will produce protection systems that are relatively inexpensive and reliable. ◈