from Integrated Electronics by Millman and Halkias, see experiment "Regulated power Supply" p.66

The output voltage V_o can be changed by varying the feedback factor β. The emitter follower $Q1$ is used to provide current gain, because the current delivered by the amplifier A_V usually is not sufficient. The dc collector voltage required by the error amplifier A_V is obtained from the unregulated voltage.

Stabilization Since the output dc voltage V_o depends on the input unregulated dc voltage V_i, load current I_L, and temperature T, then the change ΔV_o in output voltage of a power supply can be expressed as follows:

$$\Delta V_o = \frac{\partial V_o}{\partial V_i}\Delta V_i + \frac{\partial V_o}{\partial I_L}\Delta I_L + \frac{\partial V_o}{\partial T}\Delta T$$

or

$$\Delta V_o = S_V\,\Delta V_i + R_o\,\Delta I_L + S_T\,\Delta T \qquad (18\text{-}50)$$

where the three coefficients are defined as

Input regulation factor: $$S_V = \frac{\Delta V_o}{\Delta V_i}\Big|_{\substack{\Delta I_L=0\\ \Delta T=0}} \qquad (18\text{-}51)$$

Output resistance: $$R_o = \frac{\Delta V_o}{\Delta I_L}\Big|_{\substack{\Delta V_i=0\\ \Delta T=0}} \qquad (18\text{-}52)$$

Temperature coefficient: $$S_T = \frac{\Delta V_o}{\Delta T}\Big|_{\substack{\Delta V_i=0\\ \Delta I_L=0}} \qquad (18\text{-}53)$$

The smaller the value of the three coefficients, the better the regulation of the power supply. The input-voltage change ΔV_i may be due to a change in ac line voltage or may be ripple because of inadequate filtering.

18-10 SERIES VOLTAGE REGULATOR

The physical reason for the improvement in voltage regulation with the circuit of Fig. 18-16 lies in the fact that a large fraction of the *increase* in input voltage appears across the pass element, so that the output voltage tries to remain constant. If the input increases, the output must also increase (but to a much smaller extent), because it is this increase in output that acts to bias the pass transistor toward less current. This stabilization is demonstrated with reference to Fig. 18-17 where $Q2$ is the comparison amplifier designated A_V in Fig. 18-16, and where the battery V_R is replaced by the breakdown diode D. Here a fraction of the output voltage βV_o is compared with the reference voltage V_R. The difference $\beta V_o - V_R$ is amplified by $Q2$. If the input voltage increases by ΔV_i (say, because the power-line voltage increases), then V_o need increase only slightly, and yet $Q2$ may cause a large current change in R_3. Thus it is possible for almost all of ΔV_i to appear across R_3 (and since the base-to-emitter voltage is small, also across $Q1$) and for V_o to remain essentially constant. These considerations are now made more quantitative.

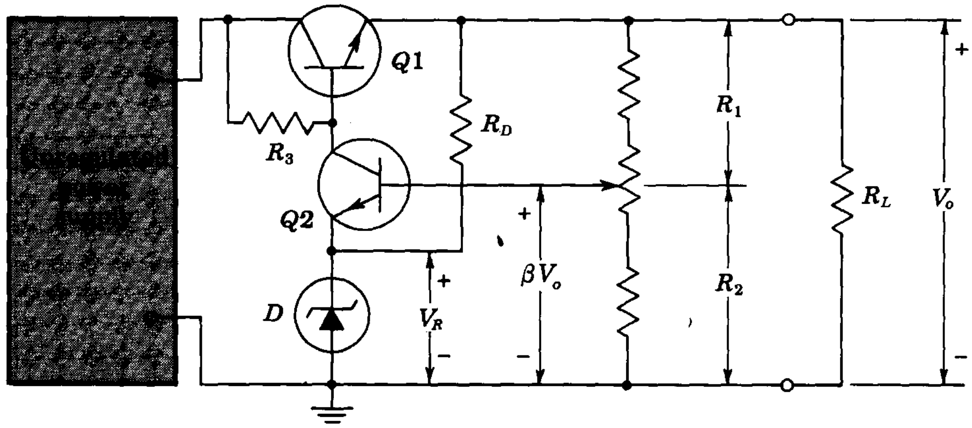

Fig. 18-17 A semiconductor-regulated power supply. The series pass element or series regulator is $Q1$, the difference amplifier is $Q2$, and the reference avalanche diode is D.

Simplified Analysis From Fig. 18-17 the output dc voltage V_o is given by

$$V_o = V_R + V_{BE2} + \frac{R_1}{R_1 + R_2} V_o$$

or using Eq. (18-48) for β

$$V_o = (V_R + V_{BE2})\left(1 + \frac{R_1}{R_2}\right) = (V_R + V_{BE2})/\beta \tag{18-54}$$

Hence a convenient method for changing the output is to adjust the ratio R_1/R_2 by means of a resistance divider as indicated in Fig. 18-17.

An approximate expression for S_V (sufficiently accurate for most applications) is obtained as follows: The input-voltage change v_i is very much larger than the output change v_o. Also, by the definition of Eq. (18-51), $\Delta I_L = 0$, and to a first approximation we can neglect the ac voltage drop across r_o. Hence $\Delta V_i = v_i$ appears as shown in Fig. 18-18. Neglecting the

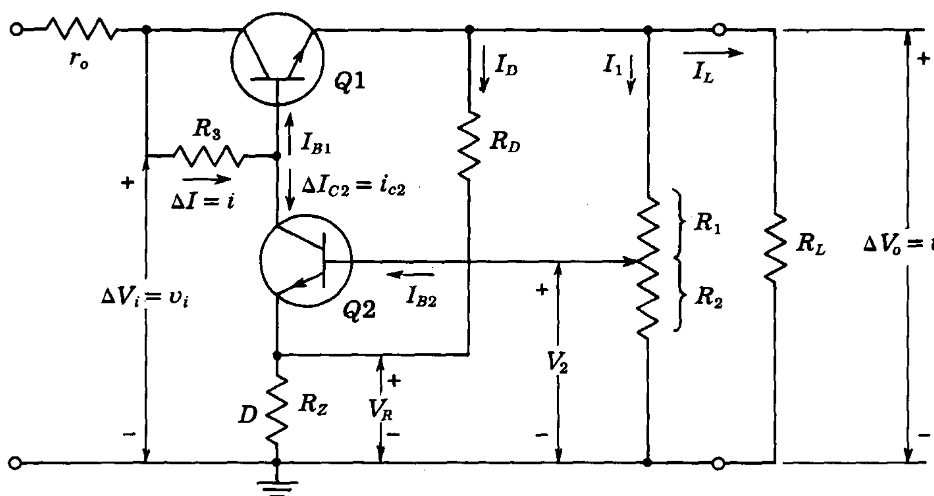

Fig. 18-18 Analysis of the series-regulated power supply.

small change in base-to-emitter voltage of $Q1$, the current change $\Delta I = i$ in R_3 is given by

$$i = \frac{v_i - v_o}{R_3} \approx \frac{v_i}{R_3} \tag{18-55}$$

Since R_L is fixed, constant output voltage requires that I_L, and hence I_{B1}, remain constant. Hence, for constant I_{B1},

$$i = \Delta I_{C2} = i_{c2} \tag{18-56}$$

In Prob. 18-23 we find, for small values of R_3, that $i_{c2} = G_m v_o$ where

$$G_m = h_{fe2} \frac{R_2}{R_1 + R_2} \frac{1}{(R_1 \| R_2) + h_{ie2} + (1 + h_{fe2})R_Z} \tag{18-57}$$

where R_Z is the dynamic resistance of the Zener diode. Using Eqs. (18-55) to (18-57), we find, since $v_i \approx i_{c2}R_3$,

$$S_V = \frac{v_o}{v_i} = \frac{1}{G_m R_3} \tag{18-58}$$

In Prob. 18-24 the output resistance R_o of the circuit of Fig. 18-18 is found to be

$$R_o \approx \frac{r_o + (R_3 + h_{ie1})/(1 + h_{fe1})}{1 + G_m(R_3 + r_o)} \tag{18-59}$$

where $G_m \equiv i_{c2}/v_o$ is obtained from Eq. (18-57). A design procedure is indicated in the following illustrative example.

EXAMPLE (*a*) Design a series-regulated power supply to provide a nominal output voltage of 25 V and supply load current $I_L \leq 1$ A. The unregulated power supply has the following specifications: $V_i = 50 \pm 5$ V and $r_o = 10\ \Omega$. (*b*) Find the input regulation factor S_V. (*c*) Find the output resistance R_o. (*d*) Compute the change in output voltage ΔV_o due to input-voltage changes of ± 5 V and load current I_L variation from zero to 1 A.

Solution *a*. Select a silicon reference diode with $V_R \approx V_o/2$. Two 1N755 diodes in series provide $V_R = 7.5 + 7.5 = 15$ V and $R_Z = 12\ \Omega$ at $I_Z = 20$ mA. Refer to Figs. 18-18 and 18-19. Choose $I_{C2} \approx I_{E2} = 10$ mA. The Texas Instruments 2N930 silicon transistor can provide the collector current of 10 mA. For this transistor the manufacturer specifies $I_{C,\max} = 30$ mA and $V_{CE,\max} = 45$ V.

At $I_{C2} = 10$ mA, the following parameters were measured:

$$h_{FE2} = 220 \qquad h_{fe2} = 200 \qquad h_{ie2} = 800\ \Omega$$

Choose $I_D = 10$ mA, so that $D1$, $D2$, operate at $I_z = 10 + 10 = 20$ mA. Then

$$R_D = \frac{V_o - V_R}{I_D} = \frac{25 - 15}{10} = 1\text{ K}$$

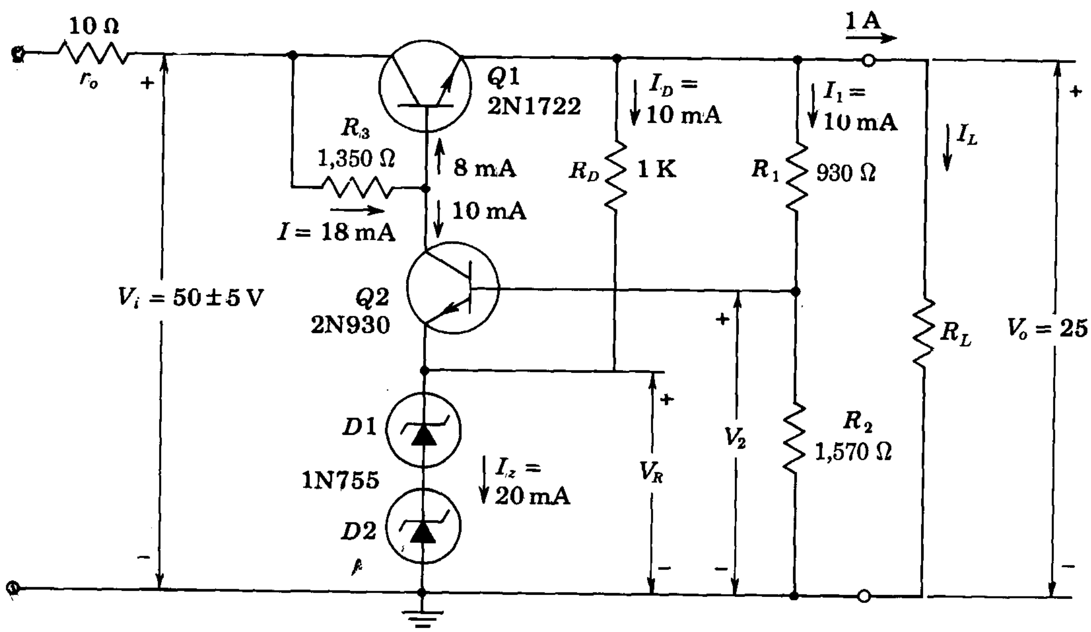

Fig. 18-19 The series regulator discussed in the example.

The ratio R_1/R_2 may be found from Eq. (18-54). Each resistor is determined as follows:

$$I_{B2} = \frac{I_{C2}}{h_{FE2}} = \frac{10 \text{ mA}}{220} = 45\ \mu\text{A}$$

Since we require $I_1 \gg I_{B2}$, we select $I_1 = 10$ mA; then, since $V_{BE} = 0.7$ V,

$$V_2 = V_{BE2} + V_R = 15.7 \text{ V}$$

$$R_1 = \frac{V_o - V_2}{I_1} = \frac{25 - 15.7}{10 \times 10^{-3}} = 930\ \Omega$$

$$R_2 \approx \frac{V_2}{I_1} = \frac{15.7}{10 \times 10^{-3}} = 1{,}570\ \Omega$$

If we select the Texas Instruments 2N1722 silicon power transistor for $Q1$, we measure at $I_{C1} = 1$ A the following parameters:

$$h_{FE1} = 125 \qquad h_{fe1} = 100 \qquad h_{ie1} = 20\ \Omega$$

We thus have

$$I_{B1} = \frac{I_L + I_1 + I_D}{h_{FE1}} = \frac{1{,}000 + 10 + 10}{125} \approx 8 \text{ mA}$$

The current I through resistor R_3 is $I = I_{B1} + I_{C2} = 8 + 10 = 18$ mA. The value for R_3 corresponding to $V_i = 45$ and to $I_L = 1$ A is given by

$$R_3 = \frac{V_i - (V_{BE1} + V_o)}{I} = \frac{50 - 25.7}{18 \times 10^{-3}} = 1{,}350\ \Omega$$

The complete circuit is shown in Fig. 18-19.

b. From Eq. (18-58) we find

$$S_V = \frac{2.50}{1.57} \times \frac{584 + 800 + (201)(12)}{(200)(1{,}350)} = 0.022$$

c. The output resistance is found from Eqs. (18-58) and (18-59). Since

$$G_m = \frac{1}{S_V R_3} = \frac{1}{0.022 \times 1{,}350} = 0.033$$

$$R_o = \frac{10 + (1{,}350 + 20)/101}{1 + (0.033)(1{,}350 + 10)} = 0.51\ \Omega$$

d. The net change in output voltage, assuming constant temperature, is obtained using Eq. (18-50):

$$\Delta V_o = S_V\,\Delta V_i + R_o\,\Delta I_L = 0.022 \times 10 + 0.51 \times 1 = 0.22 + 0.51 = 0.73\ \text{V}$$

The circuit designed in this example was built in the laboratory, and excellent agreement between measured and calculated values was obtained.

Very often it is necessary to design a power supply with much smaller value for S_V. From Eq. (18-58) we see that S_V can be improved if R_3 is increased. Since $R_3 \approx (V_i - V_o)/I$, we can increase R_3 by decreasing I. The current I can be decreased by using a Darlington pair (Fig. 8-29) for $Q1$. For even greater improvement in S_V, R_3 is replaced by a constant-current source (so that $R_3 \to \infty$), as shown in Fig. 18-20 (see also Sec. 15-3). For this circuit, which incorporates a Darlington pair, values of $S_V = 0.00014$ and $R_o = 0.1\ \Omega$ have been obtained.[9] The constant-current source in Fig. 18-20 is often called a *transistor preregulator*. Other types of preregulators (Prob. 18-28) are possible.[9] The 0.01-μF capacitor in Fig. 18-20 is added to prevent high-frequency oscillation.

Practical Considerations The maximum dc load current of the power supply shown in Fig. 18-18 is restricted by the maximum allowable collector current of the series transistor. The difference between the output and input voltages of the regulator is applied across $Q1$, and thus the maximum allowable V_{CE} for a given $Q1$ and specified output voltage determines the maximum input voltage to the regulator. The product of the load current and V_{CE} is approximately equal to the power dissipated in the pass transistor. Consequently, the maximum allowable power dissipated in the series transistor further limits the combination of load current and input voltage of the regulator.

The reverse saturation current I_{CO1} of $Q1$ in Fig. 18-18 plays an important role in determining the minimum load of the regulator. If $I_{B1} = 0$, then

$$I_{C1} = -I_{E1} = I_{CO1}(1 + h_{FE1})$$

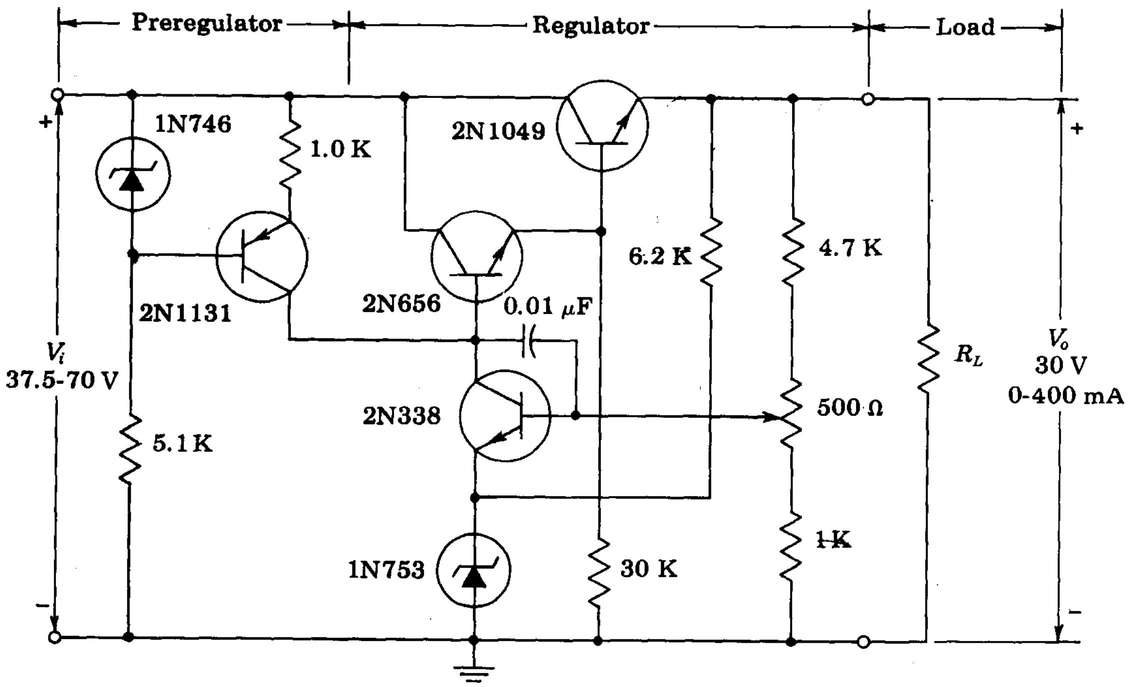

Fig. 18-20 Typical series regulator using preregulator and Darlington pair. (Courtesy of Texas Instruments, Inc.)

Hence, if the emitter current of $Q1$ $(I_L + I_D + I_1)$ falls below $I_{CO1}(1 + h_{FE1})$, then V_{CE1} cannot be controlled by I_{B1}, and the regulator cannot function properly. We thus see that, at high temperatures, where I_{CO} and h_{FE} are high, the regulator may fail when the load current falls below a certain minimum level. Various techniques have been proposed[10] to reduce this minimum-load restriction due to I_{CO}. The 30-K resistor in Fig. 18-20 is added to allow operation at low load currents.

A power supply must be protected further from the possibility of damage through overload. In simple circuits protection is provided by using a fusible element in series with r_o. In more sophisticated equipment the series transistor is such that it can permit operation at any voltage from zero to the maximum output voltage. In case of an overload or short circuit, the circuit of Fig. 18-21 can provide protection. Here the diodes $D1$, $D2$ are nonconducting until the voltage drop across the sensing resistor R_S exceeds their forward threshold voltage V_γ. Thus, in the case of a short circuit, the current I_S would increase only up to a limiting point determined by

$$I_S = \frac{V_{\gamma 1} + V_{\gamma 2} - V_{BE1}}{R_S}$$

Under short-circuit conditions the load current would be, approximately,

$$I_L \approx \frac{V_i}{R_3} + \frac{V_{\gamma 1} + V_{\gamma 2} - V_{BE1}}{R_S} \tag{18-60}$$

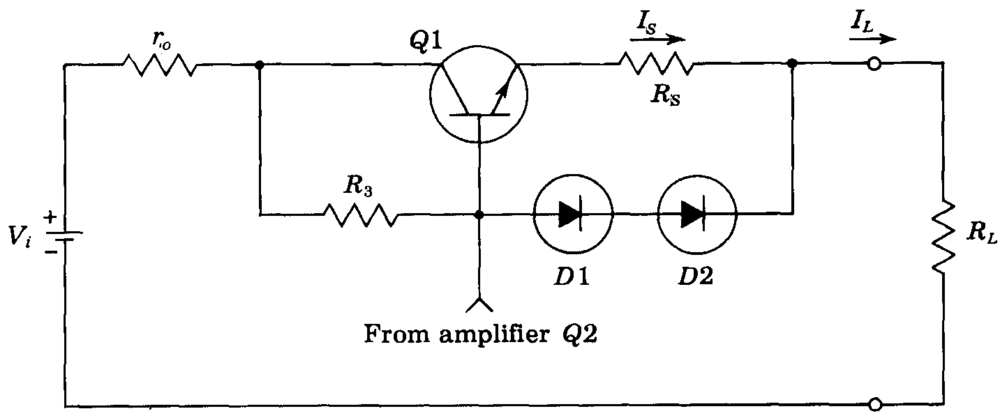

Fig. 18-21 Short-circuit overload-protection circuit.

Finally, an important practical consideration is the variation in output voltage with temperature. From Eq. (18-54) we see that, approximately,

$$\frac{\Delta V_o}{\Delta T} \approx \left(\frac{\Delta V_R}{\Delta T} + \frac{\Delta V_{BE2}}{\Delta T}\right)\left(1 + \frac{R_1}{R_2}\right) \tag{18-61}$$

Thus cancellation of temperature coefficients between the reference diode $D1$ and the transistor $Q2$ can result in a very low $\Delta V_o/\Delta T$. The GE reference amplifiers RA-1, RA-2, and RA-3 have been designed for this purpose. They are integrated devices composed of a reference diode and *n-p-n* transistor in a single chip. Typical temperature coefficients for these units are better than ± 0.002 percent/°C.