Don't shun the shunt regulator. It may have lower efficiency than the series type, but its advantages in some applications outweigh its shortcomings.

More often than not, modern power supplies use series regulation, because of its inherently high efficiency. Yet the shunt regulator may fill the bill better in the long run. In some applications its desirable features outweigh its disadvantages. Consider these excellent qualities of the shunt regulator:

- Inherent short-circuit protection.
- Relative insensitivity to input transients.
- Automatic protection against overvoltage transients at the output.

That last feature is particularly important in large integrated-circuit systems.

All of the advantages can be attained, along with remarkable levels of performance, if you design the shunt regulator simply.

The basic object of any shunt regulator is to maintain a constant terminal voltage. To accomplish this, the regulator senses the terminal voltage and varies conduction of a shunt element accordingly. As shown in Fig. 1, the basic parts of such a regulator are a reference element, a comparison amplifier and the driven shunt element.

The amplifier, *A1*, compares the reference voltage on its negative input to the sample of the output voltage, provided by R_1 and R_2, on its positive input. The resulting amplified error signal drives *Q1*, which conducts to the degree necessary to maintain the output, E_o, at the level established by D_1 and R_1-R_2.

Aside from the obvious need for a proper *Q1*—one that can handle the required current and power—the performance of the regulator depends mainly on the reference diode and the input characteristics of the amplifier (assuming adequate loop gain). Good temperature stability and drift performance dictate a temperature-compensated element for D_1 and a differential amplifier with a good input drift characteristic for *A1*. A large forward transconductance will result in minimal load-voltage changes as the load current varies.

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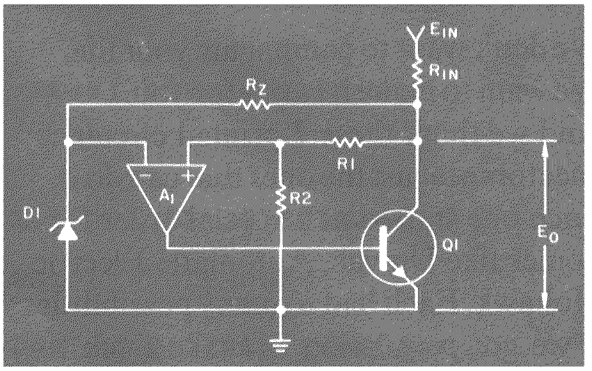


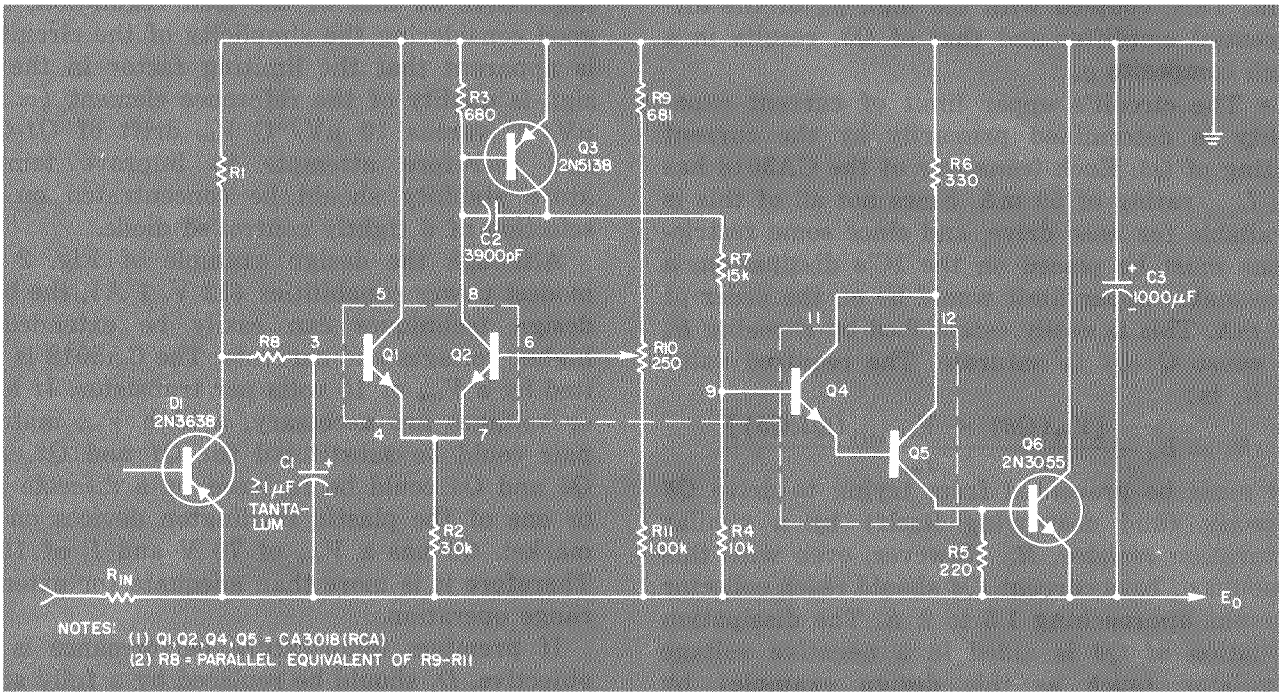
1. **Basic shunt regulator** consists of a reference element, D1, a comparison amplifer, A1, and a driven shunt element Q1.

Practical approach exploits advantages

A practical circuit that fully exploits the inherent advantages of the shunt regulator is shown in Fig. 2. In the circuit, R_9, R_{10} and R_{11} constitute the sampling element, while the reference element is a 2N3638, which is operated as a temperature-compensated zener diode (see reference). The differential input operational amplifier consists of *Q1* through *Q5*, with *Q1* and *Q2* being a matched monolithic pair having a typical offset temperature drift of 10 μV/°C. The discrete power transistor, *Q6*, functions as the shunt element. The configuration is ideally suited for responding to dynamic current changes because of the multiplication of transconductance that occurs within the loop. This results in high sensitivity to voltage changes.

A number of features contribute to the usefulness of this approach in shunt regulation:

- The active device complement consists of only 4 components: an integrated circuit, diode D_1 and transistors *Q3* and *Q6*.
- The operating current of D_1 (and the temperature coefficient) is quite predictable, since



2. **Only four active components** are required to implement the shunt regulator circuit: three transistors, one of which, D1, is operated as a zener diode, and a single integrated circuit.

the diode is fed from a constant voltage source through R_1. R_1 is given by

$R_1 = (E_o - E_{D1})/I_Z$,

where E_o is the required output voltage, E_{D1} is the reference voltage, and I_Z is the current for zero temperature coefficient. A typical value of E_{D1} is 6.6 V, and a good figure for I_Z is 5 mA. For ultimate stability, the operating current of D_1 should be trimmed for the temperature at which the regulator will operate.

A significant feature of the circuit is the fact that there is no current shift in the reference diode with variations in load. This is an advantage of a shunt regulator, which is basically a two-terminal device.

- The matched transistor pair, *Q1* and *Q2*, are ideally suited for differential connections. Typical input offset voltages for this pair of transistors is less than 1 mV. At the operating level of 1 mA, the offset drift is typically 10 μV/°C. To take full advantage of this stability, the dc base impedances must be matched. This necessitates the inclusion of resistor R_8, which has a value equal to the equivalent parallel resistance of the R_9-R_{11} divider network.

- Transistor *Q3* serves two purposes: First, it level-shifts the high collector voltage of *Q2* (by necessity $\geq E_{D1}$) down to the base requirement of *Q4* ($3V_{be}$'s). It is also a convenient point to frequency-compensate the amplifier with a Miller capacitor, C_2, from collector to base. *Q3* also contributes significant forward gain to the amplifier.

- The triple Darlington connection of *Q4*-*Q6* exhibits a phenomenally high h_{FE} characteristic. Thus large collector-current variations of *Q6* result in minimal variations in current through level-translation transistor *Q3*, and even smaller variations in the differential pair *Q1-Q2*. The differential amplifier therefore experiences a constantly balanced operating point, where *Q1* and *Q2* conduct equal current, the condition of maximum g_m.

Typical gain figures for *Q4* and *Q5* show the natural h_{FE} to be 7000-8000 at an I_c of 5 to 10 mA. The V_{be} of *Q6* (about 1 volt), together with R_5, force *Q4* and *Q5* to operate near this gain peak. In addition *Q6* exhibits a typical h_{FE} of 100 or more at an I_c of 300-500 mA. The resultant combination exhibits large forward transfer gain over all reasonable operating levels of current. This, coupled with the high g_m of the differential amplifier and that of *Q3*, results in a high composite g_m.

▪ The circuit's upper limit of current capability is determined primarily by the current rating of *Q5*. Each transistor of the CA3018 has an $I_{c_{max}}$ rating of 50 mA. Since not all of this is available for base drive, and since some restrictions must be placed on the IC's dissipation, a reasonable upper limit would be on the order of 30 mA. This is easily established by choosing R_6 to cause *Q4-Q5* to saturate. The required value of R_6 is:

$$R_6 = E_o - \frac{[V_{be}(Q6) + V_{ce_{(sat)}}(Q4,Q5)]}{I_{max}}.$$

Q3 must be prevented from trying to drive *Q6* directly (and destroying itself) by a similar saturation resistor, R_7. However, even with this amount of base current, *Q6* should reach collector current approaching 1.5 to 2 A. The dissipation situation of *Q6* is aided in a negative voltage regulator (such as this design example) by clamping the collector shell directly to a grounded heat sink and avoiding the additional thermal interface of an insulating washer.

High-level performance achieved

The design approach of Fig. 2 results in a regulator that achieves some remarkable levels of performance. At the design level of 12 volts, the circuit exhibits a current threshold, or "knee," of 18 mA, beyond which little or no change in voltage is measurable. Within the current design limit of 1 ampere, the voltage change is less than 1 mV, which is better than 0.01%.

A better figure of merit for the degree of regulation can be arrived at by an ac test. If an ac ripple current is superimposed in series with the dc input current (sometimes called "purring"), the resultant voltage swing attributable to the regulation impedance can be observed with a high-gain oscilloscope. The equivalent regulation impedance, R_o, is

$$R_o = \Delta E_o/\Delta I_o\,,$$

where ΔE_o = the peak-to-peak ac voltage observed on the scope, and ΔI_o is the peak-to-peak input ripple. For the circuit of Fig. 2, ΔI_o is 0.4 A and ΔE_o is 200 μV, giving a regulation impedance of

$$R_o = 2 \times 10^{-4}/4 \times 10^{-1} \doteq 0.5 \text{ milliohms.}$$

The figure of 200 μV for a 0.4-A change in current tends to justify the assumption of the dc case—namely, that the ΔE_o is less than 1 mV per 1 A load change.

Temperature tests on the circuit bear out the claims for the 2N3638 temperature-compensated zener (see reference). The measured temperature coefficient of the circuit of Fig. 2 is less than 0.01%/°C. Although this percentage is not as impressive as that of the load regulation, it is good considering the simplicity of the circuit. It is apparent that the limiting factor in the design is quality of the reference element ($\simeq$ 700 μV/°C versus 10 μV/°C V_{be} drift of *Q1-Q2*), and therefore attempts to improve temperature stability should be concentrated on the selection of a tightly controlled diode.

Although the design example of Fig. 2 has modest power capabilities (12 V, 1 A), the basic design techniques can easily be extended to higher voltages and currents. The CA3018 is limited by a V_{ceo} of 15 volts per transistor. If higher voltages are necessary, a high V_{ceo} matched pair could be substituted for *Q1* and *Q2*. Also *Q4* and *Q5* could be replaced by a discrete pair or one of the plastic Darlington devices on the market. *Q6* has a V_{cer} of 70 V and I_c of 15 A. Therefore it is more than adequate for extended range operation.

If premium temperature performance is the objective, D_1 should be replaced by a fully specified temperature-compensated diode, such as the 1N829. And the zener current should be carefully maintained at the test level. ▪▪

Reference:

E. J. Kennedy, "Inexpensive 6-V Reference Is Also Temperature Stable," ELECTRONIC DESIGN, ED 23, Nov. 8, 1967, p. 112.

Test your retention

Here are questions based on the main points of this article. Their purpose is to help you make sure you have not overlooked any imporant ideas. You'll find the answers in the article.

1. What is the basic disadvantage of a shunt regulator?

2. What are its advantages?

3. Why is a high h_{FE} and g_m desirable in a shunt regulator?