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Photoresistor provides negative feedback to an op amp, producing a linear response

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AGC (automatic-gain-control) amplifiers use the nonlinear characteristics of control devices. The magnitude of the real component in some of their differential parameters changes depending on variations in their dc operating points. A typical example is the VA characteristic of a silicon PN junction, which results in the differential conductance directly proportional to the passing dc current (**Reference 1**). In this form of control, the main problem is the control element's nonlinear transfer characteristic, which causes a relatively large degree of nonlinear signal distortion once the processed voltage amplitude exceeds millivolts (**Reference 2**).

A photoresistor, which has a VA characteristic that's linear in a large range of voltages, is up to the task. Common photoresistors remain perfectly linear for signal amplitudes of 100V or more. Therefore, the amplification-control device can be an optocoupler whose controlled element is a photoresistor. The circuit in this Design Idea uses a radiation source whose spectral characteristic fits the spectral characteristic of the photoresistor, and its radiated power should, if possible, be a linear function of the drive signal. Such optocouplers are commercially available, but few have properties good enough for this purpose. Common photoresistors have spectral characteristics close to the spectral characteristics of the human eye, whose peak sensitivity has approximately a 500-nm wavelength. So a white or green LED (light-emitting diode) is a good alternative. To obtain the highest possible sensitivity, this circuit uses a white HB (high-brightness) LED.

Figure 1 shows the individual components of the optocoupler and the assembled device. The optocoupler comprises a cylindrical holder that accepts a standard 5-mm HB LED from one end and a photoresistor at the other end. An opaque nonconductive seal prevents external light from entering the device. The polished metallic inner wall of the holder results in minimum light loss between the LED and the photoresistor. Available off-the-shelf photoresistors include the LDR 05, the LDR 07, and a standard white, 5-mm HB LED type L-53MWC*E, with out-

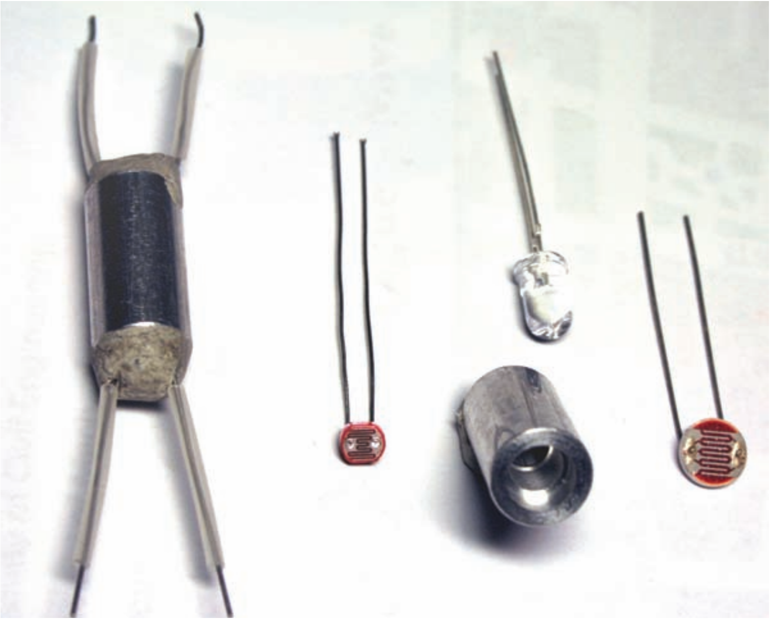

Figure 1 A metal tube with an HB LED and a photoresistor forms the optocoupler (left).

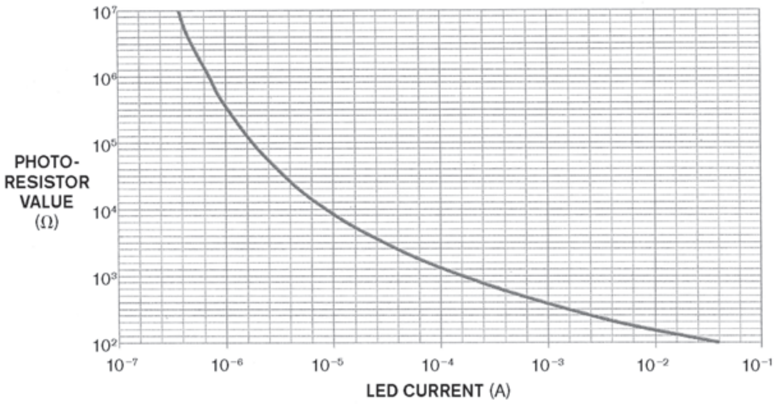

Figure 2 The optocoupler's logarithmic response in a feedback loop produces a linear amplifier response.

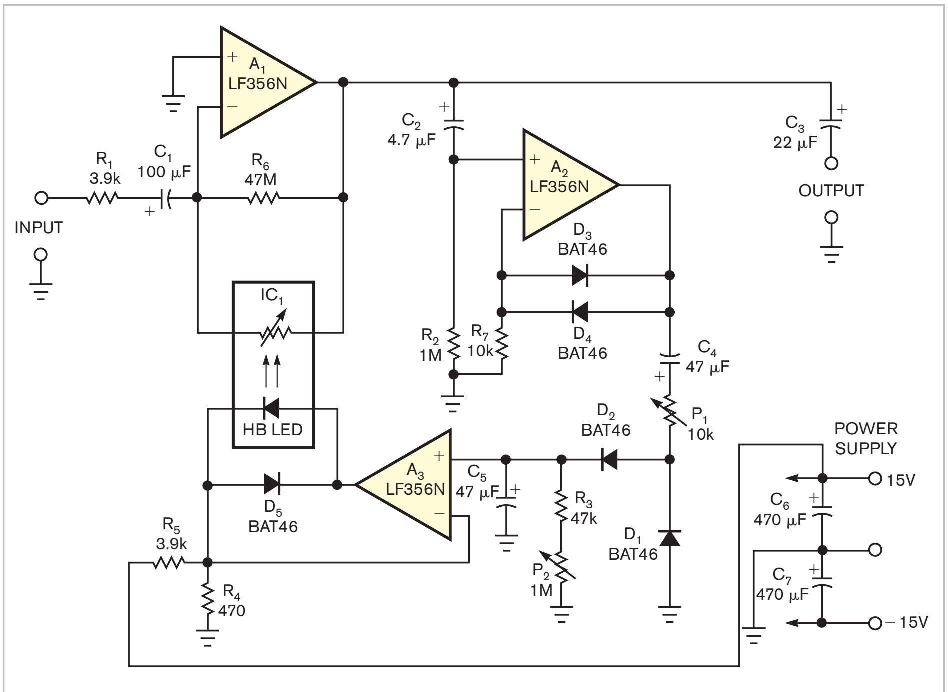

Figure 3 The adaptive-amplifier system has the optocoupler in a feedback loop.

put-light flux of 2500 mcd at a 20-mA drive current (**Reference 3**).

Figure 2 shows the transfer function of the optocoupler using the LDR 07-type photoresistor. The output resistance of the device can vary from 100Ω to 10 MΩ with LED-drive currents from 34 mA to 0.1 μA, respectively. The photoresistor's linear VA characteristic, even for large-amplitude signals, lets you use it as the control element even in situations that require a relatively large signal voltage, such as when the photoresistor is part of the feedback loop of an operational amplifier. **Figure 2** also shows that you can obtain a variation of linear output resistance over at least five decades with a maximum LED-drive current within the limits of permitted output current of common monolithic operational amplifiers.

Such an amplifier can control the overall amplification of the system in the same range without additional current amplification. Due to the photoresistor's linearity, the resulting degree of processed signal nonlinear distortion is almost solely due to the nonlinearity of the operational amplifier. Within the normal operating range, the overall linearity of the system improves with increasing input-signal amplitude because the amount of negative feedback increases with increasing signal amplitude.

Figure 3 shows the amplifier system. The basic signal-processing device is inverting op amp A_1. Its inverting connection lets you set the absolute value of the overall amplification from input to output to a value smaller than unity, permitting correct processing of an input-signal amplitude even larger than the regulated output value. Optocoupler IC_1 is the core component of the system, whose output, the photoresistor, serves as a variable part of A_1's negative-feedback network. At no-signal conditions, the LED does not illuminate the photoresistor. Thus, its resistance rises to a high value, which can cause dc runaway and the loss of the quiescent operating point of A_1. Such a condition is not harmful in principle because the signal path is ac-coupled, preventing the dc error value from getting any further. When a nonzero signal suddenly appears at the input, however, A_1's open-loop amplification would amplify it, causing a rapid rise in LED current. This action would drop the optocoupler's output resistance almost stepwise to a value sufficient to restore the dc operating point of A_1. The ac coupling transfers this transient to the output, and it may cause problems in signal-processing circuits following the adaptive amplifier. To prevent

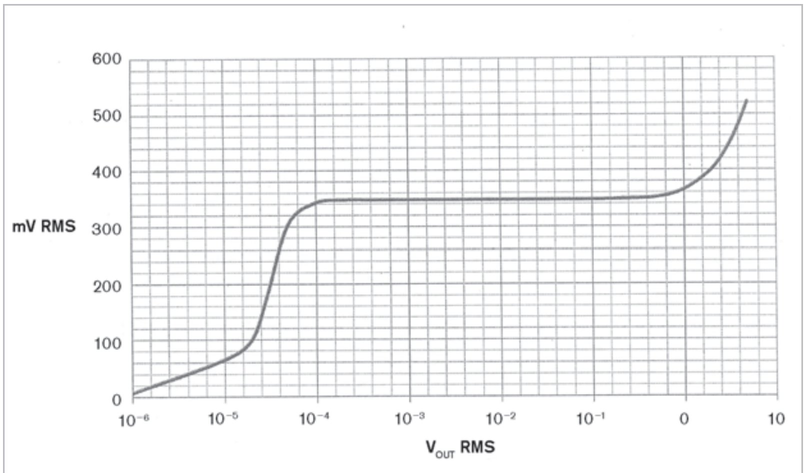

Figure 4 The amplifier system has a constant output from 0.1 mV to 1V-rms input.

this effect, you should limit the maximum value of the feedback resistance to a reasonable value, such as 47 MΩ, the value of R_6. Because the op amps have JFET inputs, the value of R_6 can be rather high. The value of 47 MΩ is a reasonable compromise, limiting the maximum absolute value of voltage amplification in A_1 to approximately 82 dB. The limiting factors for selecting a value for R_6 are the noise and the open-loop amplification of A_1.

Buffer A_2 separates the nonlinear load through the rectifying diodes from the output signal, thus preventing the nonlinear load from the rectifying diodes from distorting the output signal. Diodes D_3 and D_4 compensate the threshold voltage, including its temperature coefficient, of rectifying diodes D_1 and D_2. If you do not need to set the regulated output-voltage amplitude to a value smaller than the threshold value that the bias current in R_4 sets, you can replace D_3 and D_4 with a short circuit and omit R_7. You can set a larger-than-unity voltage amplification in A_2 to obtain a regulated output amplitude lower than the threshold that the bias in R_4 sets. Just insert an additional resistance in series with the D_3/D_4 pair.

The rectifier uses Schottky diodes, which have a lower threshold voltage than conventional PN diodes. They also have a short recovery time, keeping the same rectification efficiency at high signal frequencies. The rectifier operates as a full-wave voltage doubler, providing peak-to-peak rectification even for signals with nonsymmetrical waveforms. The rectifier output feeds to A_3, a voltage-to-current converter, which drives the LED in the optocoupler. A rectification threshold-shifting bias-current source connects to current-sensing resistor R_4. In this case R_5 simulates a current source, setting the regulated output-voltage amplitude. If the 15V supply voltage isn't perfectly stable, obtain bias current from a separate stable source. An opposite-polarity diode connects across the optocoupler's input to protect the LED from reverse polarization at no-signal conditions.

This LED current-control circuit has an important advantage: It permits an almost-independent adjustment of the attack and release time. You can adjust the attack time through variable resistor P_1, using a higher value if necessary. You can also adjust the release time using P_2. The photoresistors used have a rather good response speed, and the introduced delay at a stepwise illumination variation is acceptable for most practical requirements.

Figure 4 shows the overall response of the adaptive amplifier system. The output signal remains constant at 350 mV rms ±1 dB for input-signal voltages of less than 70 μV rms to more than 1.2V rms—that is, over a more-than-85-dB range. The no-signal output noise is less than 6 mV rms, yielding an SNR (signal-to-noise ratio), or processed-signal dynamic range, better than 20 dB at the onset of regulation in the worst-case condition and improving proportionally with increasing input-signal level.

The key parameter this design follows is its linearity. Because of the photoresistor's linearity and the separation of the nonlinear rectifier load from the output, the gain control introduces negligible nonlinearity. Thus, A_1 alone, in principle, determines the overall linearity of the system.

Harmonic analysis of the output signal at 1 kHz yields higher harmonics with amplitudes lower than A_1's noise level for all input voltages to 200 μV rms and below −75 dB for input voltages to 1.5V rms. The nonlinear distortion becomes noticeable only at large input amplitudes exceeding the regulation range of the system, raising the second harmonic to −45 dB and the third harmonic to −40 dB at 2.5V-rms input.

Within the AGC's range limits, the overall transfer linearity improves with increasing input-signal amplitude due to the increasing degree of negative feedback to A_1 at increasing input-signal amplitudes. With a value of 10 kΩ for P_1 and 1 MΩ for P_2 and a stepwise input-signal variation between 100 μV and 50 mV rms, the attack and release times are approximately 0.2 and 2 seconds, respectively. The recovery time from a 1-kHz—more than 10V-rms input overdrive—to full no-signal sensitivity is less than 2 minutes. You can adjust all of these time intervals in a wide range by varying the values of C_4, C_5, P_1, and P_2, with P_1 setting the attack time and P_2 setting the release time. EDN

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Bootstrap circuit speeds solenoid actuation

By Ralf Kelz, Seefeld, Germany

The circuit in this Design Idea bootstraps a large capacitor in series with the solenoid to provide a large actuation voltage (**Figure 1**). This higher voltage provides substantially more current to operate the solenoid (**Figure 2**, which is available at www.edn.com/100610dib), speeding the operation of the solenoid. You can also choose operating voltages or solenoid specifications that result in lower continuous current through the solenoid, reducing dc power consumption and resulting in a cooler-running solenoid with better reliability.

When there is a 0V input to the circuit, both transistors are off. Resistor R_1 slowly charges the left side of capacitor C_1 to the 24V power-supply voltage. D_2 clamps the right side of capacitor C_1 to 0.6V. When the input signal goes high, both the Q_1 and the Q_2 transistors turn on. This action quickly drives the left side of C_1 to ground. Because voltage cannot change instantaneously across a capacitor, the right side of C_1 goes down to –23.4V. D_2 steers the solenoid current into the capacitor until it discharges, at which time the solenoid current conducts through D_2 to ground. D_1 prevents a voltage-overshoot spike when the circuit turns off, and current suddenly stops flowing in D_1. It clamps the bottom leg of the solenoid to 24.6V until the current decays in the solenoid.

THE TIME CONSTANT DEPENDS ON THE SOLENOID'S INDUCTANCE AND THE CAPACITOR'S VALUE.

The time constant of the circuit depends on the inductance of the solenoid and the value you choose for the capacitor, which you can calculate with the following **equations**:

$$I(t)=\frac{(2\,V_{IN}-V_D)\,e^{\left(-\frac{t}{\tau}\right)}\sinh(\omega t)}{\omega L};$$

$$\omega=\sqrt{\left(\frac{R}{2L}\right)^2-\frac{1}{LC}};\ \text{and}\ \tau=\frac{2L}{R}.$$

In these **equations**, e is the mathematical constant, ω is the radian angular frequency, and t is time in seconds. In addition, L is inductance and R is resistance. EDN

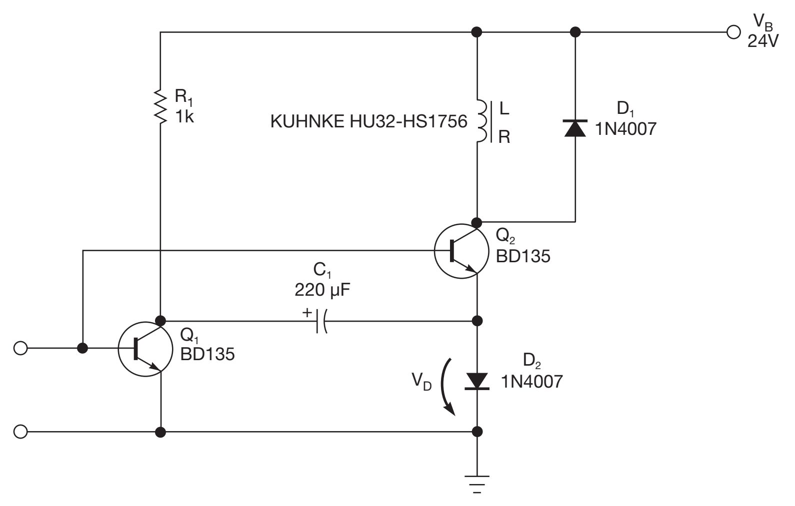

Figure 1 This circuit bootstraps the power supply voltage across the solenoid to temporarily double the actuation voltage.

Notch filter autotunes for audio applications

John R Ambrose, Mixed Signal Integration, San Jose, CA

Tracking notch filters find use in harmonic-distortion analyzers; they also can remove heterodyne noise from ham-radio systems. A conventional tracking switched-capacitor notch filter relies on a bandpass filter, a voltage-to-frequency converter, and a notch filter to track the incoming signal and remove undesired tones. The bandpass filter in these circuits sometimes adjusts to the wrong frequency, meaning that the undesired tone would have no attenuation.

The circuit in this Design Idea uses IC_1, a 74HC4046 PLL (phase-locked-loop) IC, which operates as fast as 1 MHz, to improve the noise immunity of the system (**Figure 1**). IC_2, an RDD104 IC from LSI Computer Systems Inc (www.lsicsi.com), provides a 1000-to-1 divider in an eight-pin package. IC_3, Mixed Signal Integration's (www.mix-sig.com) MSHN5 1000-to-1 clock-to-corner switched-capac-

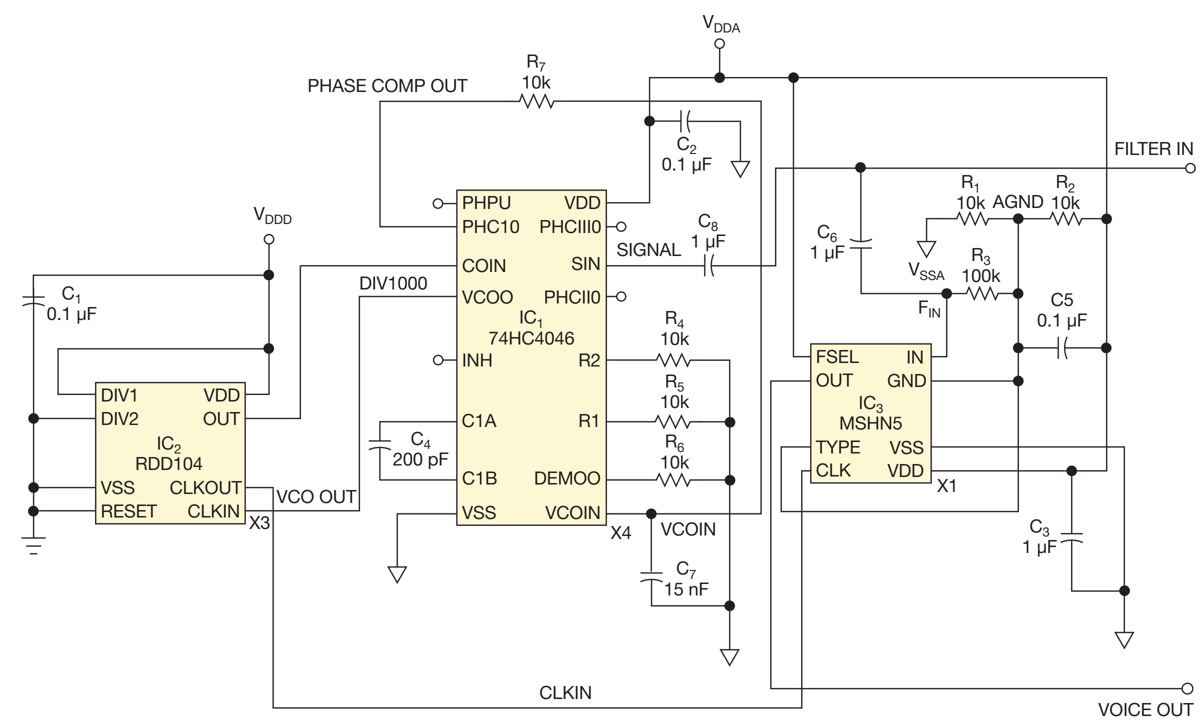

Figure 1 This audio notch-filter circuit uses a PLL to improve noise immunity.

itor highpass/notch filter, comes in an eight-pin package.

You feed IC_1's VCO (voltage-controlled oscillator) output into the clock input of IC_2. IC_2 can perform 10-, 100-, 1000-, and 10,000-to-1 divisions using the DIV1 and DIV2 pins. You tie the output of RDD104 to the COIN of IC_1. By using IC_1's EX/OR phase comparator, you can improve noise immunity. You apply the input signal to both IC_1 and the input of IC_3, whose clock you derive from the CLKOUT pin of IC_2.

The MSHN5, IC_3, contains both selectable highpass filters and selectable notch filters. When you tie the FSEL pin high, it selects notch; tying TYPE to AGND selects the narrow notch filter. This step ensures the removal of only one tone from the input signal with little information loss. IC_3's 1000-to-1 clock-to-corner ratio reduces the chance that aliasing signals will affect the output. For voice applications, for example, no signals of 500 kHz or higher would be available to alias into the passband. A sample setup uses an input frequency of 789.13 Hz at a clock frequency of 789.13 kHz, 1000 times the input signal (**Figure 2**). The PLL tracks the input, moving the notch filter to 1.24 kHz. EDN

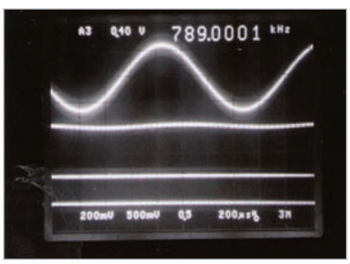

Figure 2 You feed the circuit an input tone (top trace) and get the output signal from the MSHN5 (second trace). The third and fourth traces represent the clock signal from the 74HC4046 PLL.

Tricolor LEDs create a flashing array

Jeff Tregre, www.BuildingUltimateModels.com, Dallas, TX

You can build a matrix of RGB (red/green/blue) LEDs using a simple and inexpensive circuit comprising the control logic and driver circuit in **Figure 1** and some LEDs (**Figure 2**). The center RGB LED is the first to come on, after which each sequential LED in the 8×8-LED matrix follows. This process gives the appearance that the display is alive and moving outward. This sequence repeats, producing a rainbow effect of colors.

You can adjust the frequency of each clock by changing the values of R_{17}, R_{19}, and R_{23}. Use different frequencies for each clock, which will display eight colors from the 65 tricolored LEDs, because using the same frequencies for all the clocks causes your display to appear white. The cost of building this circuit should be $25 to $30. You can purchase 100 5-mm RGB LEDs from eBay for a total of about $18. Be sure to use common-cathode LEDs.

This simple circuit comprises three clocks and three counters, one for each of the three LED colors. Setting each clock frequency to a different rate causes each color of each LED to appear to be random. All resistors are 0.25W, except for R_3, R_8, and R_{13}, which are 0.5W; R_4, R_9, and R_{14}, which are 1W; and R_5, R_{10}, and R_{15}, which are 1.5W resistors. These high-wattage resistors and the 12 NPN transistors are necessary because all LEDs in this matrix, ex-

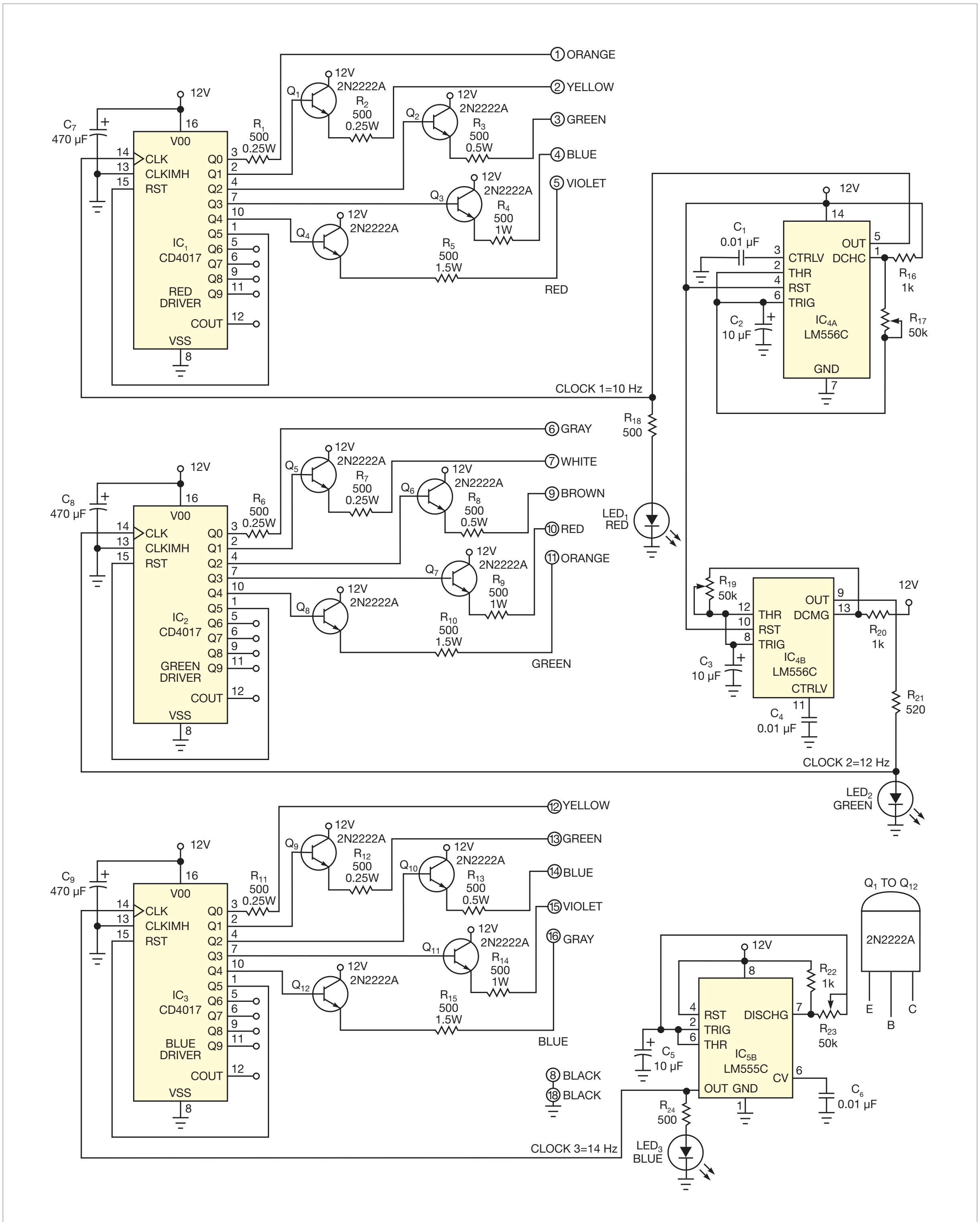

Figure 1 Three 555 timers generate clock signals, and CD4017 counters provide the drive signals for the transistors.

cept the center one, connect in parallel. Start by bending all of the ground leads flat and connecting them together. When wiring the LEDs, begin in the center and work outward. You can then mount the LED board onto the top of the PCB (printed-circuit board). See the online version of this Design Idea at www.edn.com/100624dic for photos, a parts list, and a video of this circuit in action.

To add the finishing touches to your project, use a small picture frame and install waxed paper onto the inside of the glass. Mount the LED board ¼ to 1 in. away. The magnifying lens of the LEDs will produce a beautiful effect when they shine through the waxed paper. EDN

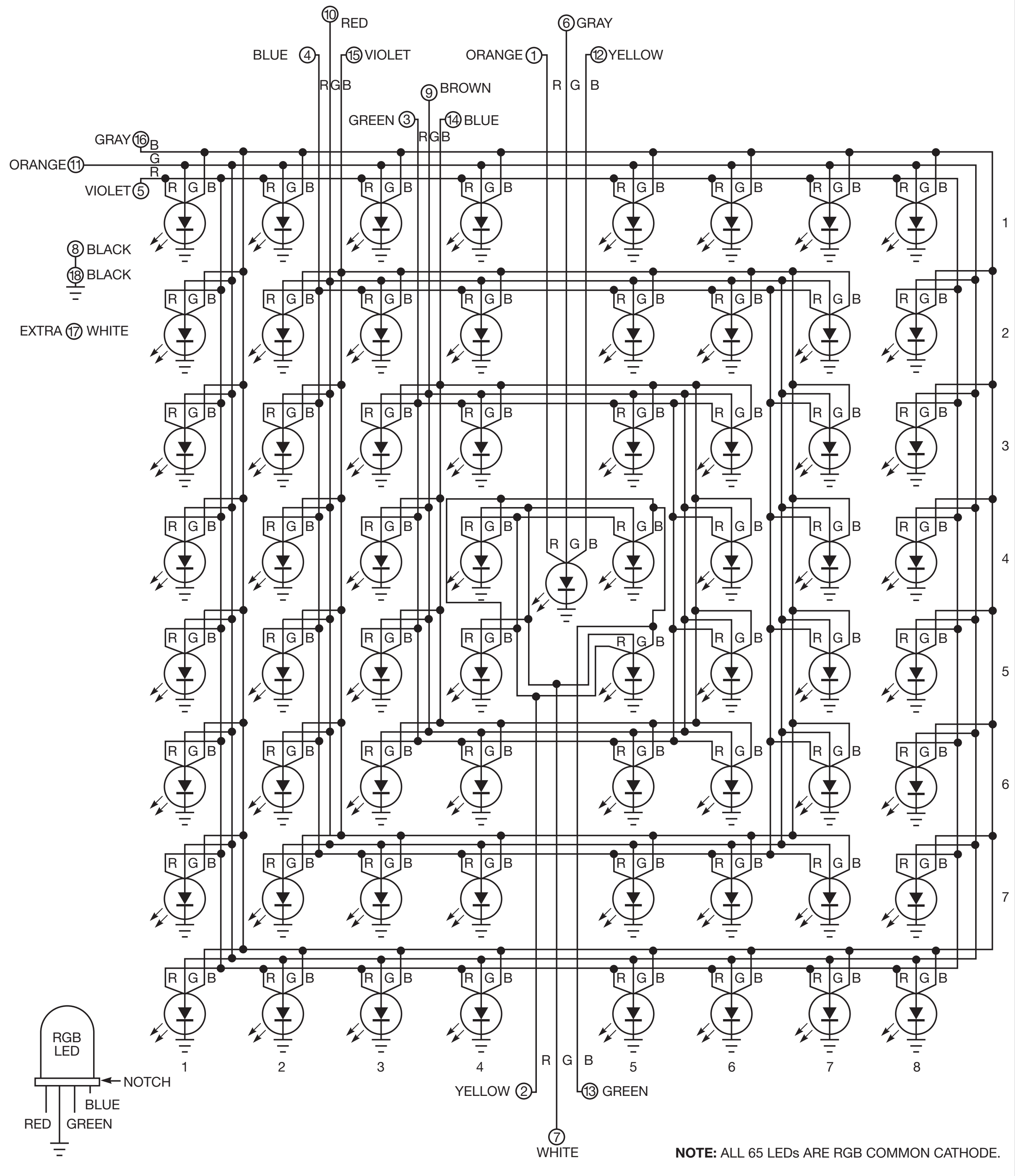

Figure 2 The LED in the center lights first, and the light then moves outward until the circuit products an 8×8-LED display.

DC-voltage doubler reaches 96% power efficiency

Marián Štofka, Slovak University of Technology, Bratislava, Slovakia

The voltage-doubler circuit in **Figure 1** can convert 2.5V dc to 5V dc or 1.8V to 3.3V. Most voltage doublers use an inductor, but this circuit doesn't need one. The circuit uses a capacitor, C, by charging it through serially connected switches. The charge switches let capacitor C charge, and the discharge switches are open. In the subsequent discharging phase, the charge switches are off, and the discharge switches close. The two discharge switches now connect capacitor C between the source of the input voltage, V_S, and the output capacitor, C_{OUT}. This connection scheme lets the applied voltages combine. Thus, the voltage at the output terminal has a value close to $2V_S$.

The two phases of operation repeat periodically at frequency f, which clock generator IC_2 determines. The duty cycle is about 50%, but the value isn't all that critical. One half of the Analog Devices (www.analog.com) high-performance ADG888 analog multiswitch provides the switching. The IC's two halves have independent control, so the other half occasionally shorts R_P, the 10Ω inrush-current-limiting resistor, which protects the charge switches from an initial overcurrent. That current occurs after power-on, before the output voltage reaches the predetermined percentage of the output's full voltage.

A micropower op amp, IC_{3A}, runs as a comparator with hysteresis. It compares input voltage to output voltage. Its output starts low and then goes high, which turns on paralleled switches S_3 and S_4. The comparator's action is ratiometric because the reference input voltage at the inverting input is the input-supply voltage, V_{IN}. This connection is possible because of the AD8617's rail-to-rail input/output operation. The circuit also provides overload protection for an excessive load, which connects to the circuit's output before power-on.

During soft start, the output voltage can't reach the threshold level for loads below a certain value. Consequently, the circuit remains in soft-start mode. The minimum value of R_L, which activates the protective subcircuit, is $R_L \leq m^2 \times (\alpha/(1-\alpha)) \times R_P$, where the multiplication factor

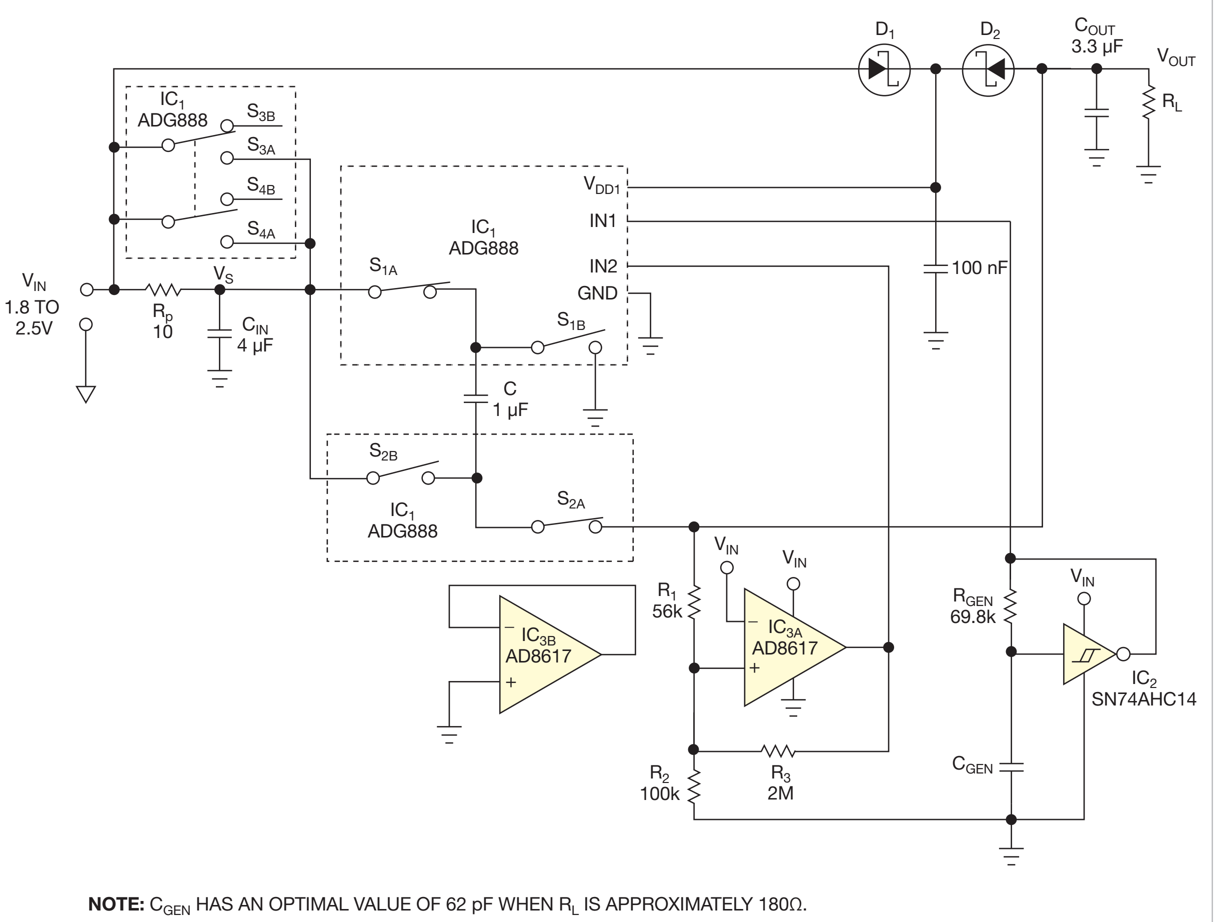

Figure 1 You can use this step-up dc/dc converter in applications in which power efficiency is a critical issue.

$m=(V_{OUT}/V_{IN})$ and α is a fraction of V_{OUT}, at which the soft start turns off. For $m=2$, $\alpha=0.8$, and $R_P=10\Omega$, R_L is 160Ω. Thus, loads of 160Ω or less will overload the circuit if you connect them to the circuit's output before power-on. IC_2 and IC_3 get their power from the input supply. IC_1, however, switches voltages of as much as $2V_{IN}$, and its V_{DD1} supply-voltage pin must remain at the same level. An analog OR switch comprising Schottky barrier diodes D_1 and D_2 provides that voltage. The higher of the input or output voltages appears at the V_{DD1} pin of IC_1. The high levels of output voltages for both IC_2 and IC_3 suffice for control of IC_1 because the ADG888's data sheet allows a $0.36V_{DD1}$ value for the high value at the control inputs. The circuit has been tested at an input voltage of 2.386V, R_L of 178.46Ω, a frequency of 200 kHz, a supply voltage of 2.377V, an input supply current of 51.285 mA, and an output voltage of 4.588V. Evaluating these data gives a multiplication factor of 1.929 and power efficiency of 96.39%.

This power efficiency remains more than 96% for frequencies of 150 to 350 kHz. The 9-mV drop at the switch-shorted R_P at the given input current indicates that the on-resistance of the paralleled switches has a value of approximately 0.175Ω. EDN

Methods measure power electronics' efficiency

Liping Zheng, Calnetix, Yorba Linda, CA

Validating the sytem efficiency of a power-electronics circuit is essential in evaluating the overall system performance, design optimization, and sizing of cooling systems. **Figure 1** shows the conventional method of performing efficiency measurement. The power-electronics system operates at the rated output-power level, and, by measuring the input power and output power, you can calculate the system's efficiency using the **equation** $\eta=(P_{OUT}/P_{IN})\times 100\%$, where P_{OUT} is output power and P_{IN} is input power. In other words, the measured input power is equal to the output power plus the power loss of the system.

However, measuring the efficiency of a high-power system that delivers power to loads such as motors, generators, or industrial-computer equipment requires a source that delivers the rated power. The infrastructue therefore should comprise a suitably rated source and an equivalent load that can support the rating of the power-electronics system you are evaluating. These requirements can drive up the facility's infrastructure cost; for one-time design-validation measurements, this cost is difficult to justify.

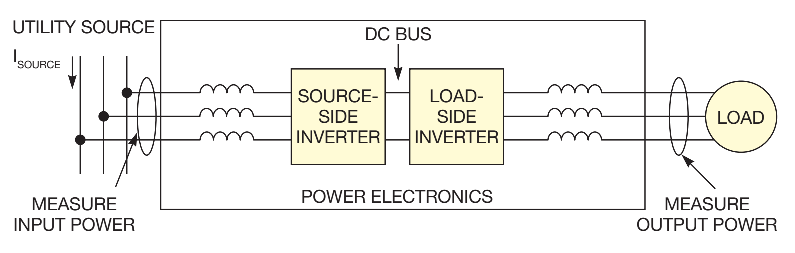

Figure 1 In a conventional method of performing efficiency measurement, the power-electronics system operates at the rated output-power level.

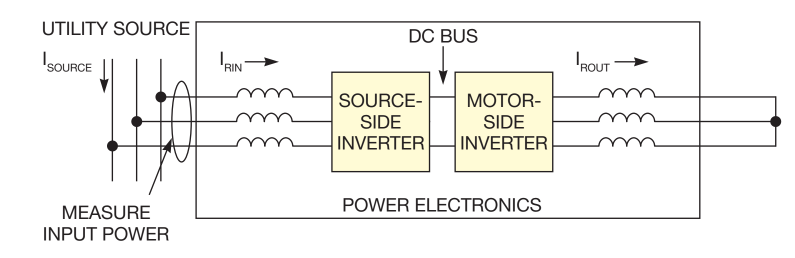

Figure 2 This method eliminates the test load by shorting the output-load terminals.

This Design Idea describes alternative methods of measuring the efficiency of a high-power power-electronics system that simplifies the test-infrastructure requirement by eliminating the test load and using a source that must support only the loss of the power-electronics system. **Figure 2** shows the proposed method, which eliminates the test load by shorting the output/load terminals. The system's control algorithm maintains the required input- and output-current amplitude and frequency by developing circulating reactive power. IGBTs (insulated-gate bipolar transistors) and magnetic components dominate the system's losses, which are functions of the amplitude and frequency of the input and output currents. The loss is also less sensitive to the power-factor and PWM (pulse-width-modulation) index.

To know the required input and output current, you must estimate the system's power factor, the motor's back EMF (electromotive force), and the systems's source voltage. This example uses a field-oriented control for both source- and load-side inverters, resulting in the following **equations**:

$$I_{ROUT}=I_{ROUT\_RE}+jI_{ROUT\_IM}=\frac{P_{OUT}}{\sqrt{3}V_{BEMF}};$$

$$I_{RIN}=I_{RIN\_RE}+jI_{RIN\_IM}=$$

$$\frac{P_{RIN}}{\sqrt{3}V_{GRID}}=\frac{P_{OUT}/\eta_E}{\sqrt{3}V_{GRID}},$$

where I_{ROUT} is the required output current, which comprises real current, I_{ROUT_RE}, and reactive current, I_{ROUT_IM}; I_{RIN} is the required input current, which comprises the real current, I_{IN_RE}, and the reactive current, I_{IN_IM}; P_{RIN} is the required input power; P_{OUT} is the output power at the test condition; V_{BEMF} is the motor's back EMF; V_{GRID} is the grid voltage; and η_E is the estimated efficiency of the circuit.

By maintaining the input current to be I_{RIN} and the output current to be I_{ROUT}, the measured input real power will be close to the power loss, P_{LOSS}, at the actual output-power level, P_{OUT}. Therefore, you can calculate the efficiency as follows: $\eta=(P_{OUT})/(P_{OUT}+P_{LOSS})\times100\%$.

If the measured efficiency, which you calculate using this **equation**, does not quite match the estimated efficiency, η_E, update the second **equation** using the measured efficiency, η, and repeat the measurement until they are close. Calnetix (www.calnetix.com) has used this method to evaluate the efficiency of a 125-kW power-electronics system, compared the results with the conventional measurements, and found them to be closely matching.

Most high-power power-electronics systems have high efficiency, which means that the real current is much less than the reactive current. To reduce the required current from the grid, you can use the method in **Figure 3**, which uses another identical system to offset the input reactive current that the test system creates. By providing a path for circulating reactive power, the utility sources the lost power only, not the total power. In **Figure 3**, the input current of the second power-electronics circuit is $I_{RIN}=I_{RIN\_RE}+jI_{RIN\_IM}$. By setting the first circuit to have an input current of $I_{RIN1}\approx I_{RIN\_RE}-jI_{RIN\_IM}$, the power from the source is only $I_{SOURCE}=I_{RIN1}+I_{RIN}\approx I_{RIN\_RE}+I_{RIN\_RE}+j(I_{RIN\_IM}-I_{RIN\_IM})=2I_{RIN\_RE}$. The circuit uses the input current from the source only to overcome the power losses of the two circuits, thereby eliminating the need for a high-power infrastructure. EDN

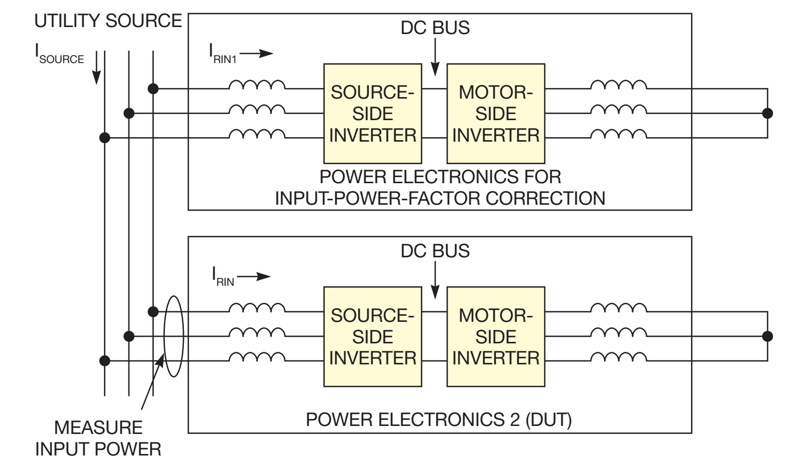

Figure 3 This method uses two identical power-electronics systems. The second system offsets the input reactive current that the test system creates.

Simple battery-status indicator uses two LEDs

Abhijeet Deshpande, People's Education Society Institute of Technology, Bangalore, India

Properly maintained rechargeable batteries can provide good service and long life. Maintenance involves regular monitoring of battery voltage. The circuit in **Figure 1** works in most rechargeable batteries. It comprises a reference LED, LED_{REF}, which operates at a constant current of 1 mA and provides reference light of constant intensity regardless of battery voltage. It accomplishes this task by connecting resistor R_1 in series with the diode. Therefore, even if the battery voltage changes from a charged state to a discharged state, the change in current is only 10%. Thus, the intensity of LED_{REF} remains constant for a battery state from a fully charged state to a fully discharged state.

The light output of the variable LED changes with respect to changes in battery voltage. The side-by-side-mounted LEDs let you easily compare light intensities and, thus, battery status.

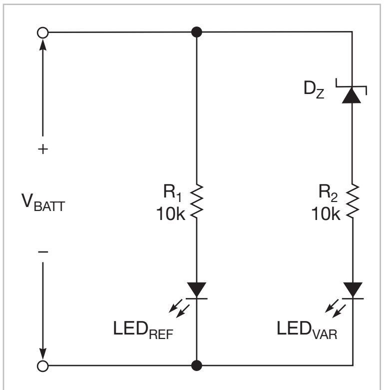

Figure 1 This circuit works in most rechargeable batteries. It comprises a reference LED, LED_{REF}, which operates at a constant current of 1 mA and provides reference light of constant intensity regardless of battery voltage.

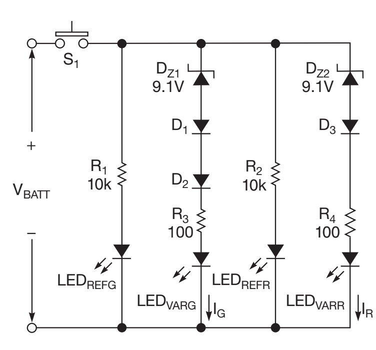

Figure 2 This circuit can withstand 13V because it has a 10-mA margin. If the LEDs are bright, quickly release pushbutton switch S_1.

TABLE 1 LED INTENSITY

Light output of LED_{VARG}	Light output of LED_{VARR}	Battery status (%)
Much brighter than LED_{REFG}	Much brighter than LED_{REFR}	70 to 100
Equally as bright as LED_{REFG}	Much brighter than LED_{REFR}	60
Off	Brighter than LED_{REFR}	50 to 30
Off	Equally as bright as LED_{REFR}	20
Off	Off	0 to 10

Using diffused LEDs as crystal-clear LEDs can damage your eyes. Instead, mount the LEDs with sufficient optical isolation so that the light from one LED does not affect the intensity of the other LEDs.

The variable LED operates from 10 mA to less than 1 mA as the battery voltage changes from fully charged to fully discharged. Zener diode D_Z in series with resistor R_2 causes the current to change with battery voltage. The sum of the zener voltage and the drop across the LED should be slightly less than the lowest battery voltage. This voltage appears across R_2. As the battery voltage varies, it produces a large variation of current in R_2. If the voltage is approximately 1V, then 10 mA will flow through LED_{VAR}, which is much brighter than LED_{REF}. If the voltage is less than 0.1V, then the light intensity of LED_{VAR} will be less than LED_{REF}, indicating that the battery has discharged.

Immediately after the battery has charged, the battery voltage is more than 13V. The circuit can withstand this voltage because it has a 10-mA margin. If the LEDs are bright, quickly release pushbutton switch S_1 to avoid damage to the LEDs (**Figure 2**).

The **figure** uses a 12V lead-acid battery indicator as an example, but you can extend the design to accommodate other types of chargeable batteries. You can also use it for voltage monitoring. It uses two green LEDs to indicate whether the battery has charged above 60%. A set of red LEDs indicates whether the battery charge drops below 20%. LED_{REFG} and LED_{REFR} feed through 10-kΩ resistors R_1 and R_2. For the variable-intensity LEDs, a zener diode works in series with 100Ω resistors R_3 and R_4. Diodes D_1, D_2, and D_3 provide the required clamping voltages. **Table 1** shows how LED intensity indicates battery charge.

The following **equation** calculates the variable intensity for the green LED: $V_{BATT}=I_G\times100+V_{D1}+V_{D2}+V_{LEDG}+V_{DZ1}$. For a green-LED current of 1 mA, $V_{BATT}=10^{-3}\times100+0.6+0.6+1.85+9.1=12.25$V. The selected LEDs have a drop of 1.85V at 1 mA.

If the LED has different characteristics, then you must recalculate the resistor values. At this voltage, the LEDs have the same intensity, and the battery is 60% charged. See **Reference 1** for lead-acid-battery voltages.

The following **equation** calculates the variable intensity for the red LED: $V_{BATT}=I_R\times100+V_{D3}+V_{LEDR}+V_{ZD2}$. For a green-LED current of 1 mA, $V_{BATT}=10^{-3}\times100+0.6+1.85+9.1=11.65$V.

At this voltage, both red LEDs have equal intensities, and the battery is 20% charged. LED_{VARG} is off. **Figure 3** shows that both variable-intensity LEDs are brighter than the reference LEDs, indicating that the battery is 100% charged. **EDN**

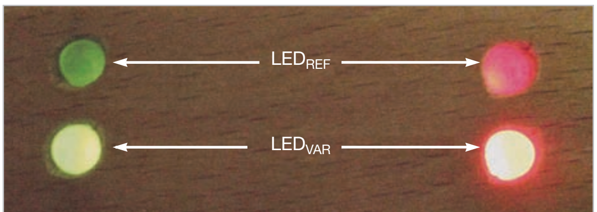

Figure 3 Both variable-intensity LEDs are brighter than the reference LEDs, indicating that the battery is 100% charged.

REFERENCE

1 "Lead Acid Battery Charging," *Solar Navigator*, 2005, www.solarnavigator.net/battery_charging.htm.

Get four colors from 2 bits

Marián Štofka, Slovak University of Technology, Bratislava, Slovakia

Three-color LEDs contain red, green, and blue LEDs in one package. Using two digital control signals, you can drive these LEDs to produce four colors. The circuit in **Figure 1** uses an Analog Devices (www.analog.com) ADG854 dual analog 1-to-2 demultiplexer that lets you select the current through each LED.

The circuit uses a distinct current, I or 2I, to drive each LED. The demultiplexers determine the routes of the currents through transistors Q_1, Q_2, and Q_3 in transistor array IC_2 to the LEDs. These transistors act as both current sources and summing elements.

The following **equation** yields the value of the current: $I=(V_{REF}-V_{BE})/R_E$, where V_{BE} is the base-emitter voltage of bipolar transistors Q_1, Q_2, and Q_3. The base-emitter-voltage value varies slightly depending on the total collector current, but you can neglect this variation. Refer to the data sheet of your transistor array for this information.

One unit of current constantly flows through the green LED. Demultiplexer D_1 routes another unit of current to either the red LED or the blue LED, and D_2 routes the third unit of current to either the green LED (2I total) or the red LED.

Table 1 shows the states and colors that this circuit produces. The sum of currents flowing through all LEDs is 3I at one time for all four combinations of control variables. Thus, the generated light is approximately of the same intensity regardless of color.

The decreasing value of the base-emitter voltage with temperature, which is approximately –1.42 mV/°C, causes an increase in current through the LEDs by approximately 0.33%/°C. It has a beneficial effect because it compensates for the decreasing radiance of the LEDs as temperature increases.

Drops in radiance are approximately –0.27%/°C for the blue LED and about –0.35%/°C for the green LED. The radiance of these two LEDs, which are both indium-gallium-nitride types, thus remains almost constant over ambient temperature. The red LED is an aluminium-indium-gallium-phosphorus type, having a radiance drop of approximately –0.77%/°C, and the current source roughly halves this drop.

The R_0 resistors force the logic inputs to logic zero at manual control by

TABLE 1 DISTRIBUTION OF CURRENT AND COLORS

IN_1	IN_2	I_R	I_G	I_B	Color
0	0	I	I	I	White
0	1	Off	2I	I	Aqua
1	0	2I	I	Off	Red-orange
1	1	I	2I	Off	Yellow

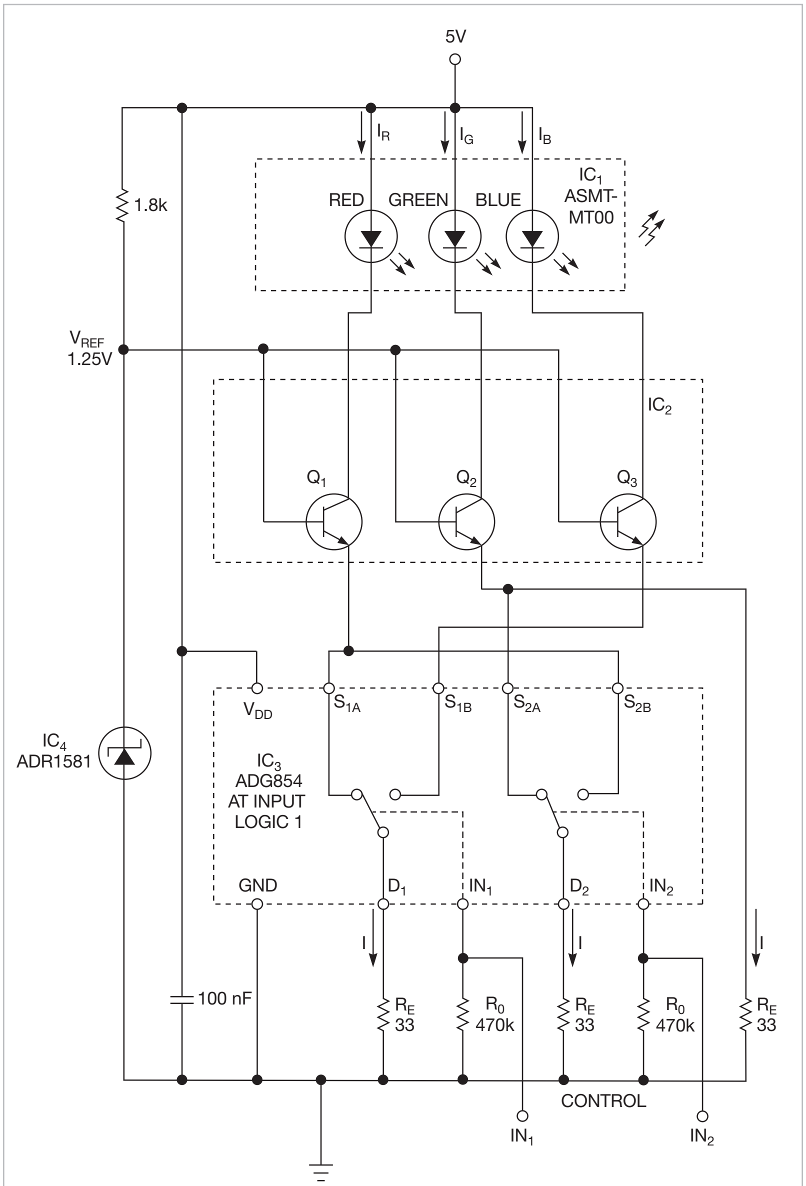

Figure 1 A three-color LED IC emits mixtures of two or three spectrally "pure" colors. The human eye perceives the mixtures as special colors.

connecting or not connecting the IN_1 and IN_2 control leads to V_{DD}, the power-supply voltage. The maximum current flowing through the LEDs, about 26 mA, is far below the nominal current of 350 mA that Avago Technologies (www.avagotech.com) rates for the ASMT-MT00 power RGB (red/green/blue) LED that this circuit uses.

The radiance is sufficient, yet the junction temperature of the LEDs is low. Junction-to-pin thermal resistance for the green LED is 20°C/W. IC_1 dissipates approximately 0.1W. Therefore, you can estimate the junction temperature to be higher than the ambient temperature by less than 2°C (**Reference 1**). Consequently, you increase the LED's expected lifetime well beyond thousands of hours. EDN

REFERENCE

1 Oon, Siang Ling, "The Latest LED Technology Improvement in Thermal Characteristics and Reliability: Avago's Moonstone 3-in-1 RGB High Power LED," White Paper AV02-1752EN, Avago Technologies, Jan 20, 2009, www.avagotech.com/docs/AV02-1752-EN.

Control a dc motor with your PC

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Polytechnic Higher Institute of Yefren, Yefren, Libya

The circuit in this Design Idea controls the speed of a 5V permanent-magnet dc motor through the PC's parallel port (**Figure 1**). You use the C++ computer program, available at www.edn.com/100826dia, to run the motor at three speeds. The circuit uses PWM (pulse-width modulation) to change the average value of the voltage to the dc motor. You connect the motor to the PC's parallel port with an interface circuit. The design comprises IC_1, a 74LS244 buffer; IC_2, a ULN2003 driver; relay switches S_1, S_2, and S_3; IC_3, a 555 astable multivibrator circuit; and Q_1, a 2N2222 driving transistor. The 555 timer operates as a variable-pulse-width generator. You change the pulse width by using relays to insert or split resistors in the 555 circuit.

The computer program controls these resistors. When S_1 is on and both S_2 and S_3 are off, the timer output is set to logic one, thereby driving the motor with its maximum speed. When S_1 and S_2 are on, the 555 timer generates a pulse signal with a 50% duty cycle. In this case, the charging resistor, R_{A1}, is equal to the discharging resistor, R_B. In the third case, S_1 and S_3 are on, and the charging resistor is R_{A2},where R_{A2}=0.1×R_B, reducing the on time of the pulse signal and, consequently, the speed of the motor to the lower limit. **Table 1** summarizes the on/off-operation conditions of the relays and the corresponding dc-motor speeds.

The code prompts you to select a certain speed, stores your selection as

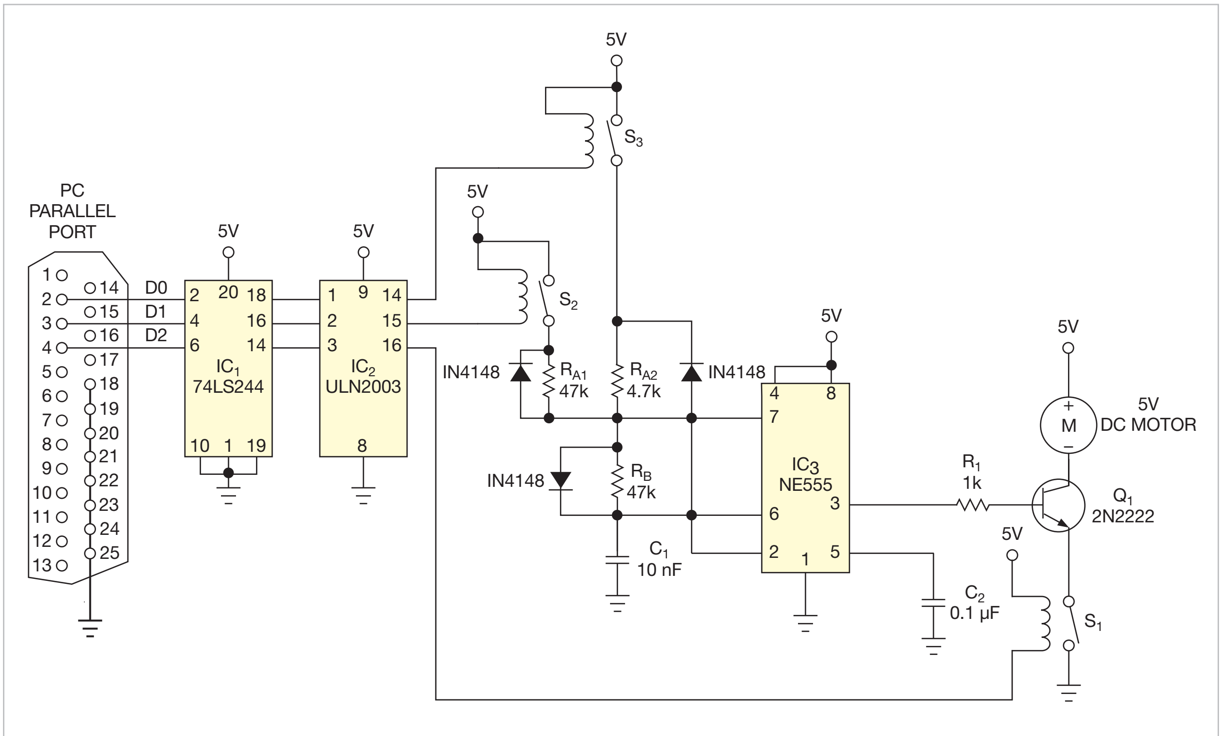

Figure 1 This circuit controls the speed of a 5V permanent-magnet motor through the PC's parallel port.

an integer variable choice, generates the proper digital sequence, and stores it at another integer variable. You place the value of the integer variable data at a PC's parallel port using the outportb function. The program uses the kbhit function to stop the motor when you hit any key on the keyboard. EDN

TABLE 1 SWITCH STATES AND GENERATED PC SEQUENCES

S_3	S_2	S_1	Equivalent digital sequence	Motor speed
Off	Off	Off	000	Stop
Off	Off	On	001	Maximum
Off	On	On	011	Medium
On	Off	On	101	Minimum

Current monitor compensates for errors

Chau Tran and Paul Mullins, Analog Devices, Wilmington, MA

You sometimes need to measure load currents as large as 5A in the presence of a common-mode voltage as high as 500V. To do so, you can use Ana-log Devices' (www.analog.com) AD8212 high-voltage current-shunt monitor to measure the voltage across a shunt resistor. You can use this circuit in high-current solenoid or motor-control applications. **Figure 1** shows the circuit, which uses an external resistor and a PNP transistor to convert the AD8212's output current into a ground-referenced output voltage proportional to the IC's differential input voltage. The PNP transistor handles most of the supply voltage, extending the common-mode-voltage range to several hundred volts.

An external resistor, R_{BIAS}, safely limits the circuit voltage to a small fraction of the supply voltage. The internal bias circuit and 5V regulator provide an output voltage that's stable over the operating temperature range, yet it minimizes the required number of external components. Base-current compensation lets you use a low-cost PNP pass transistor, recycling its base current, I_B, and mirroring it back into the signal path to maintain system precision. The common-emitter breakdown voltage of this PNP transistor becomes the operating common-mode range of the circuit.

The internal regulator sets the voltage on COM to 5V below the power-supply voltage, so the supply voltage for the measurement circuit is also 5V. Choose a value for the bias resistor, R_{BIAS}, to allow enough current to flow to turn on and continue the operation of the regulator. For high-voltage operation, set I_{BIAS} at 200 μA to 1 mA. The low end ensures the turn-on of the bias circuit; the high end is limited, depending on the device you use.

With a 500V battery and an R_{BIAS}

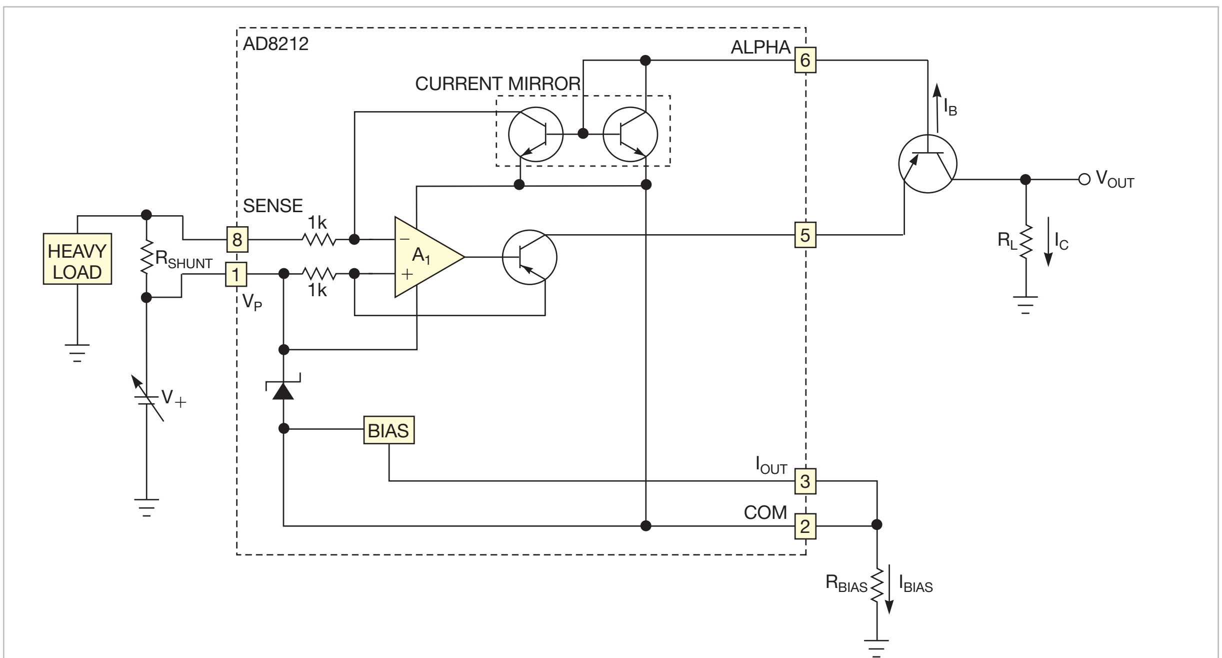

Figure 1 An external PNP transistor lets you operate the circuit at high voltages.

value of 1000 kΩ, for example, $I_{BIAS}=(V_+-5V)/R_{BIAS}=495V/1000$ kΩ=495 μA.

The circuit creates a voltage on the output current approximately equal to the voltage on COM plus two times the V_{BE} (base-to-emitter voltage), or $V_+-5V+2V_{BE}$. The external PNP transistor withstands two times the base-to-emitter voltage of more than 495V, and all the internal transistors withstand voltages of less than 5V, well below their breakdown capability.

Current loss through the base of the PNP transistor reduces the output current of the AD8212 to form the collector current, I_C. This reduction leads to an error in the output voltage. You can use a FET in place of the PNP transistor, eliminating the base-current error but increasing the cost. This circuit uses base-current compensation, allowing use of a low-cost PNP transistor and maintaining circuit accuracy. In this case, current-mirror transistors, the AD8212's internal resistors, and amplifier A_1 combine to recycle the base current.

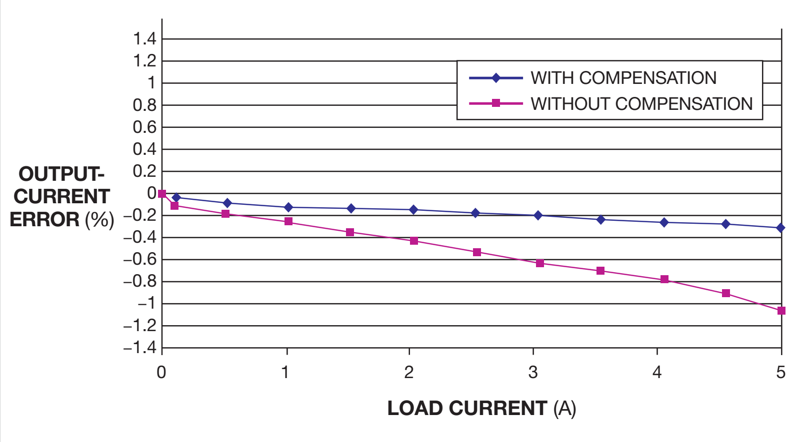

Figure 2 Internal base-current compensation reduces error.

Figure 2 shows a plot of output-current error versus load current with and without the base-current-compensation circuit. Using the compensation circuit reduces the total error from 1 to 0.4%. You should choose the gain of the load resistor, R_L, to match the input voltage range of an ADC. With a 500-mV maximum differential-input voltage, the maximum output current would be 500 μA. With a load resistance of 10 kΩ, the ADC would see a maximum output voltage of 5V. EDN

Amplifiers deliver accurate complementary voltages

Marián Štofka, Slovak University of Technology, Bratislava, Slovakia

The circuit in **Figure 1** generates two analog voltages, which you can vary in a complementary manner. When the straight output voltage rises, the complementary output voltage decreases, and vice versa. The sum of both output voltages is a constant: $V_{OUT}+V_{OUTC}=V_{REF}$, where V_{OUT} is the straight output voltage, V_{OUTC} is the complementary output, and V_{REF} is a reference voltage you derive from bandgap cell IC_1. You choose the ratio of the resistor divider that connects to the output of IC_1 so that the reference voltage is approximately 400 mV. Potentiometer R_P sets the desired analog voltage, which connects to the noninverting input of voltage follower IC_{2A}. The output of IC_{2A} provides the straight output voltage, which connects to the inverting input of unity-gain inverter IC_{2B}. The noninverting input of IC_{2B} has a gain of two and connects to the

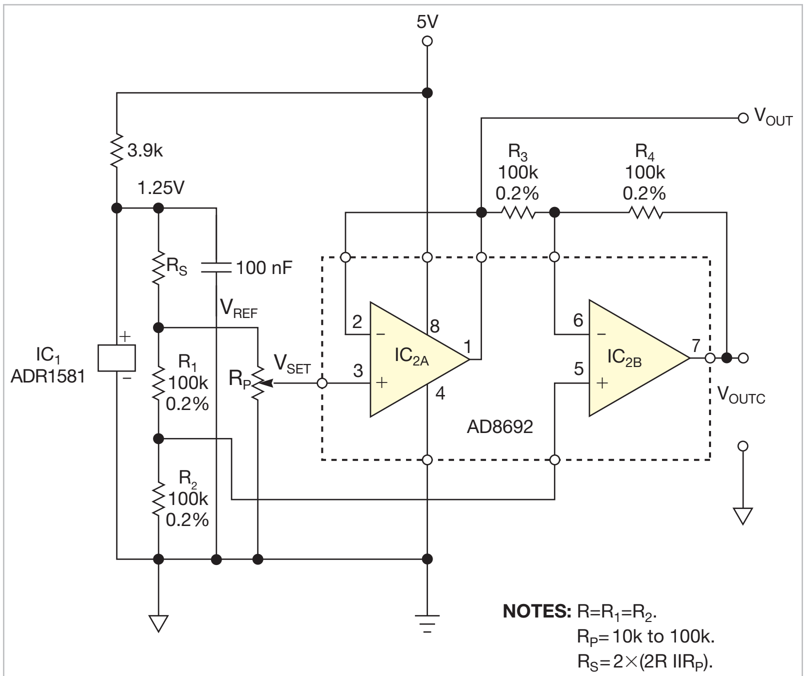

Figure 1 The circuit outputs two analog voltages whose sum always equals the reference voltage.

TABLE 1 COMPLEMENTARY VOLTAGES FOR THREE INPUT SETTINGS

V_{SET}	V_{OUT} (mV)	V_{OUTC} (mV)
V_{REF}	411.45	0.15
0	0.45	410.45
$V_{REF}/2$	205.8	205.1

middle of the high-precision resistive divider comprising R_1 and R_2, which halves the reference voltage. The following **equation** calculates the output voltage of IC_{2B} with respect to ground: $V_{OUTC} = -V_{OUT} + 2 \times (V_{REF}/2) = V_{REF} - V_{OUT}$. Thus, the straight output voltage plus the complementary output voltage give the desired constant value equal to the reference voltage.

You should use either a quad resistor or two pairs of matched resistors for precision resistors R_1 through R_4. Resistors R_3 and R_4 form the negative feedback in IC_{2B}, and the other pair of resistors halves the reference voltage. You can omit these four resistors if you use an instrumentation amplifier instead of IC_{2B}. In this case, you must use an RRIO (rail-to-rail-input/output) type of instrumentation amplifier. The output of a contemporary RRIO instrumentation amplifier approaches the low side by a margin of approximately 60 mV, and it would severely degrade the circuit's accuracy. The output of the Analog Devices (www.analog.com) AD8692 op amp, however, typically approaches the lower rail by 0.75 mV at a 10-μA load current. The guaranteed value of the margin is 1 mV at this current.

The circuit has undergone testing for three values of test voltages: the reference voltage, which represents a full-scale; half the reference voltage; and 0V. **Table 1** lists the measured voltages at both outputs. Any of the output voltages can approach the lower supply rail with an error of less than 0.25% at 400 mV full-scale.EDN

Set LEDs' hue from red to green

Marián Štofka, Slovak University of Technology, Bratislava, Slovakia

The circuit in **Figure 1**, which lets you create light of 32° of hues, uses red and green LEDs. A constant current divides into two components. One component flows through a red LED, and another one flows through a green LED. You can vary the current from 0 to 100% through the red LED, and thus you simultaneously

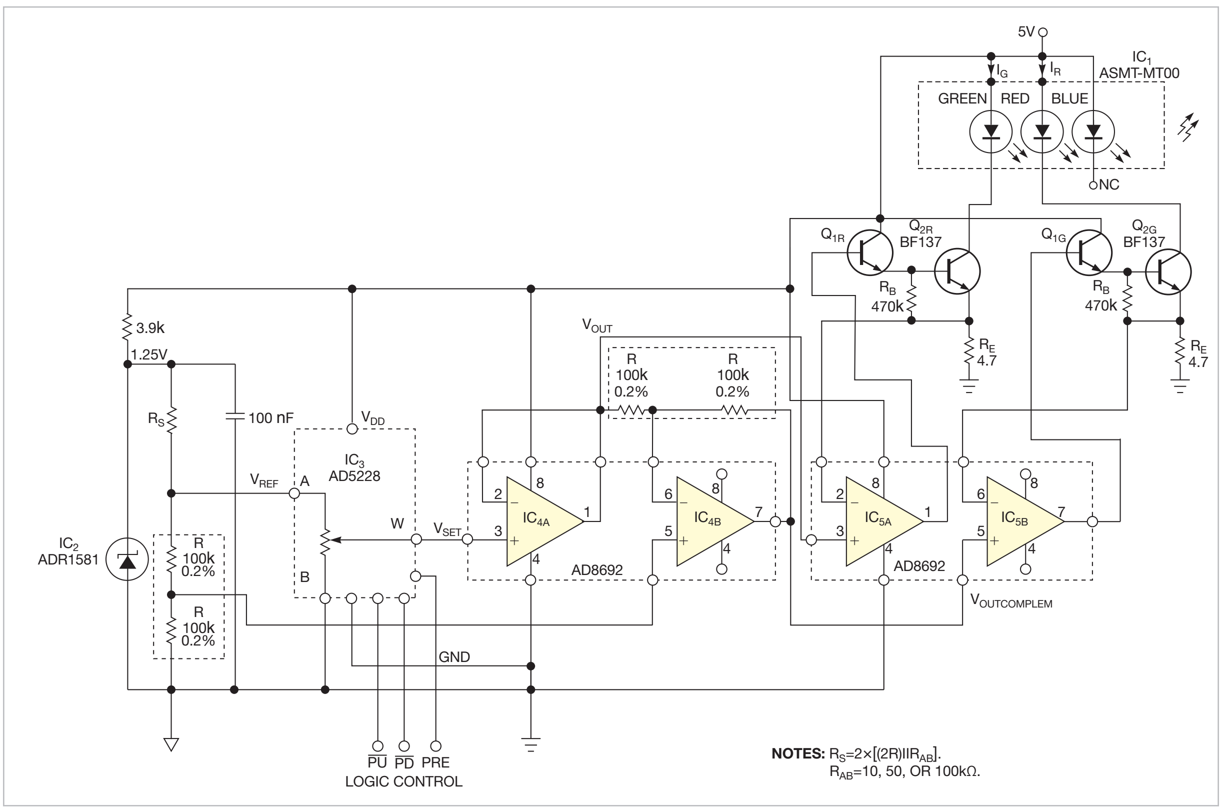

Figure 1 This circuit lets you set one of 32° of hues between red and green using a short-term grounding of the pullup or pull-down control pins.

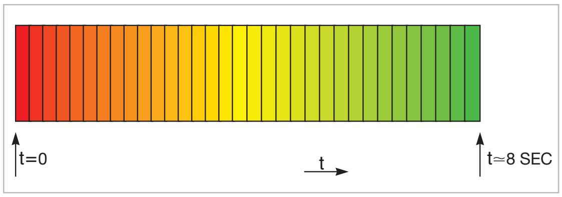

Figure 2 The light output changes quasicontinually from red to green within approximately 8 seconds, using long-term grounding of the pullup pin or a continuous grounding of the pin at power-on.

vary the current through the green LED as a slave-type complement to 100%. When this scenario happens, your eye perceives the resulting light mixture as any hue between red and green. Roughly speaking, the transition from red to green passes through orange, amber, and yellow. You can set any of the 32 hues between red and green, passing through orange, amber, and yellow.

IC_3, an Analog Devices (www.analog.com) AD5228 resistive DAC, has one-in-32 resolution, and it thus sets the resolution of this circuit. In this application, the resistive DAC functions as a digital potentiometer. You can manually set its wiper position through short-term grounding of its $\overline{PU}$ pullup and $\overline{PD}$ pulldown control pins. The resistive DAC has no memory, so you have to make this setting after each power-on.

Holding the $\overline{PU}$ and $\overline{PD}$ pins to a logic low, the wiper position increments or decrements with an increased speed of one step per 0.25 sec, so the output light's color varies stepwise for a low pullup (**Figure 2**). You can also preset the hue of the LED, which appears at power-on. For a high Preset, the color is 100% red when you apply power. At a low Preset, a midposition is preset at the resistive DAC and thus the color at power-on is 50% red and 50% green; you perceive it as yellow.

The circuit uses two LEDs in IC_1, a high-performance, tricolor ASMT-MT00 LED from Avago Technologies (www.avago.com). The blue LED remains unused. You can, however, connect any of the remaining five red/green, red/blue, blue/red, green/blue, or blue/green combinations instead of the green/red combination this circuit uses.

Although the sum of currents flowing through the red and green LEDs is approximately one-fourth of the nominal per-LED current, the radiance is high, and you should not look directly at the lid of IC_1 when it is on from a distance of less than approximately 1 foot.

IC_2, IC_3, and IC_4 comprise a low-side source of two complementary analog voltages (**Reference 1**). The resistive DAC replaces the classic potentiometer in the earlier Design Idea. These complementary analog voltages are the input voltages for the two power stages comprising transistor Q_1 and midrange-power transistor Q_2.

The power stage—voltage-to-current converters you make by cascading two bipolar transistors and an op amp—drives each of the two LEDs. The circuit senses output current at resistor R_E. The R_B resistors eliminate the leakage currents of both bipolar transistors in the cascaded series. These power stages would be functional even with one bipolar transistor instead of two. The cascaded bipolar transistors provide precision in the voltage-to-current converter. With a single power transistor, the relative error would be approximately $1/\beta$, whereas using the cascaded series, the error is approximately $1/(\beta_1\beta_2)$, where β_1 and β_2, the current gains of the bipolar transistors, are approximately 300 and 100, respectively. The error results from the current flowing through resistor R_E, which is the sum of the output current and the base current of transistor Q_1.

You can use this circuit in industries ranging from entertainment to toys; it may eventually find use in experimental psychology and in modern fine arts, which involves the use of optoelectronics.

Holding $\overline{PD}$ low and feeding a 50% duty cycle, 0.05-Hz-frequency logic waveform to the $\overline{PU}$ pin produces a slow, periodic, quasicontinuous "waving" of the color from red to green and back. EDN

REFERENCE

1 Stofka, Marian, "Amplifiers deliver accurate complementary voltages," *EDN*, Sept 23, 2010, pg 44, http://bit.ly/cljzKQ.

Accurately simulate an LED

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National Semiconductor Corp, Santa Clara, CA

Solid-state-lighting applications are quickly moving into the mainstream. Although they are more efficient, the LEDs that produce the low-cost light often require a complicated driver circuit.

Testing the driver circuit using LEDs, although easy, yields only typical results because the tests don't factor in worst-case LED parameters and often generate undesirable light and heat during driver debugging. Although using a constant resistance might seem to be an appropriate approach, a resistor approximates an LED load at only one point on the current/voltage curve. An electronic load may prove to be a more useful approach. The control loops of the driver circuit and the electronic load, however, often result in system instability and oscillations.

Figure 1 illustrates a typical LED-driver circuit using a low-cost simulated-LED circuit. The simulated LED accurately mimics a real LED at a user-programmable threshold voltage. A simple Darlington current sink, Q_1, provides a wide range of LED threshold voltages. The size of the heat sink attached to Q_1 and the power capability of Q_1 are the only limits on the amount of power the simulated LED can dissipate.

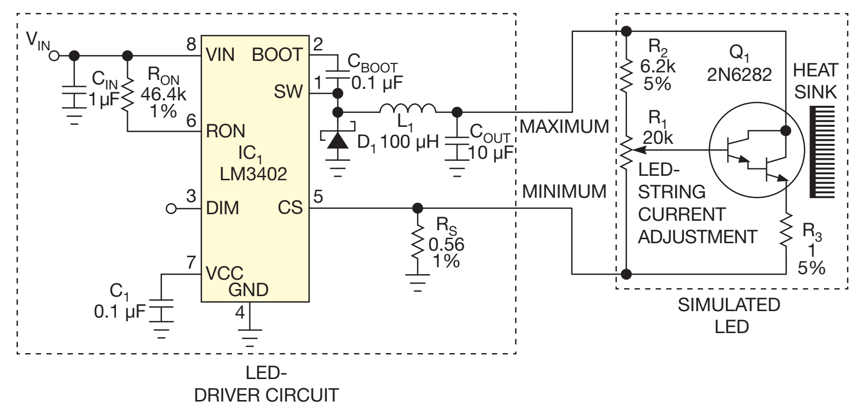

Figure 1 Use this circuit for quick testing of an LED-driver circuit over minimum, typical, and maximum LED parameters.

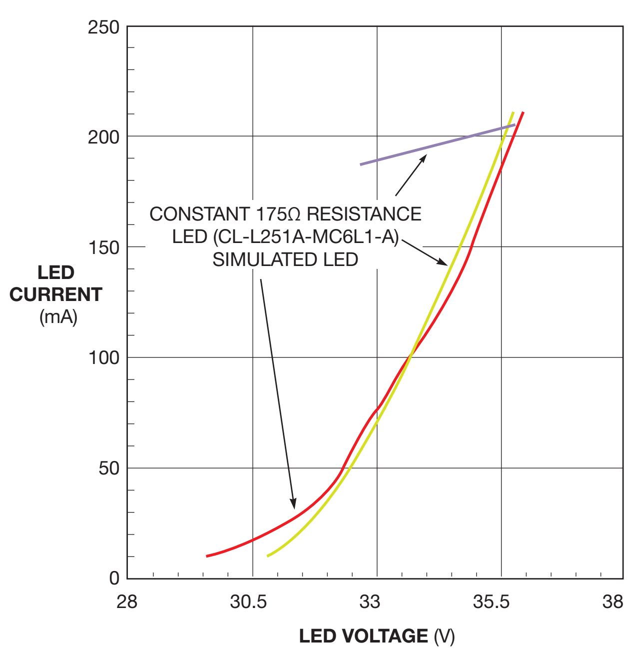

Figure 2 The simulated LED approximates the turn-on characteristics of a real LED. A constant-resistance load approximates a real LED load at only one point on the curve.

You can easily tune the circuit for any LED voltage. Place a constant voltage across the simulated LED. Tune the circuit by adjusting resistor R_1 until the circuit draws the desired current. You can adjust the shape of the voltage knee by making small changes to resistor R_3, although this step is not usually necessary.

Figure 2 compares the simulated LED's current and voltage characteristics to those of a real LED and a constant resistance. The soft turn-on of the simulated LED accurately mimics that of a real LED. Furthermore, the simulated LED quickly retunes to test minimum and maximum LED characteristics, thus giving you confidence that the circuit will work over all load conditions.EDN

Power USB devices from a vehicle

Fons Janssen, Maxim Integrated Products Inc, Bilthoven, the Netherlands

Automotive accessories such as PNDs (portable navigation devices) usually receive their power or charge using a simple adapter that a user plugs into a cigarette lighter. Sometimes, however, you may want to power or charge two devices at once. The circuit in **Figure 1** can handle that task.

IC_1 generates 5V from any 7.5 to 76V input—a wide enough range to include the complete range of car-battery voltage plus the 40V spike that can occur during a load dump. The IC is simple to use because it has an internal power switch and requires no compensation circuit.

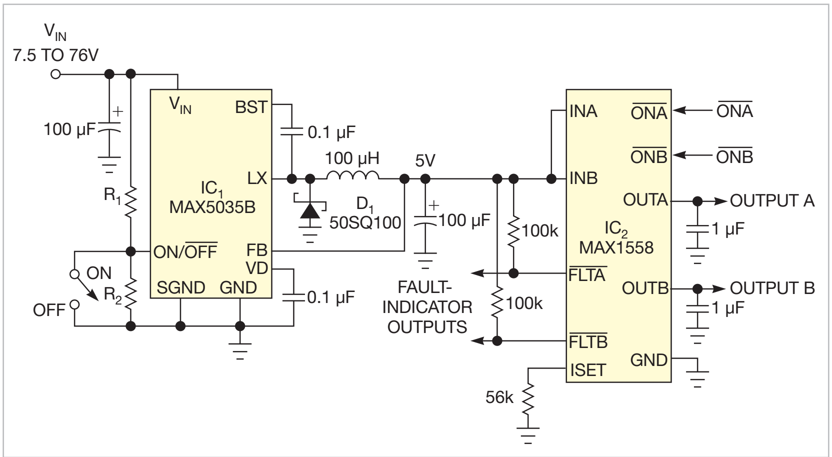

Figure 1 An automotive USB power supply generates two regulated, supply-voltage outputs from an unregulated input.

IC_2 distributes to two outputs the 5V that IC_1 generates. It not only distributes power but also protects against overload conditions. Most portable equipment receives its power or charge through a USB (Universal Serial Bus) interface, whose current limit is 500 mA. Because IC_2 targets use in USB applications, it latches off any port that tries to deliver more than 500 mA but does not affect the other port. Automatic-restart capability ensures that the port recovers automatically after the removal of the overload condition.

Figure 2 shows the protection feature in action. Output B has a constant load of 300 mA, and Output A switches between a 100-mA load and a 600-mA overload. IC_2 switches off Output A after an overload but allows a 20-msec delay to avoid responding to brief transients (**Figure 2a**). The circuit removes the overload 80 msec later; after another 20-msec delay, the automatic-restart feature brings Output A back online. Output B is unaware of the problem in Output A (**Figure 2b**). The fault-indicator output, however, goes low to indicate a problem in Channel A.

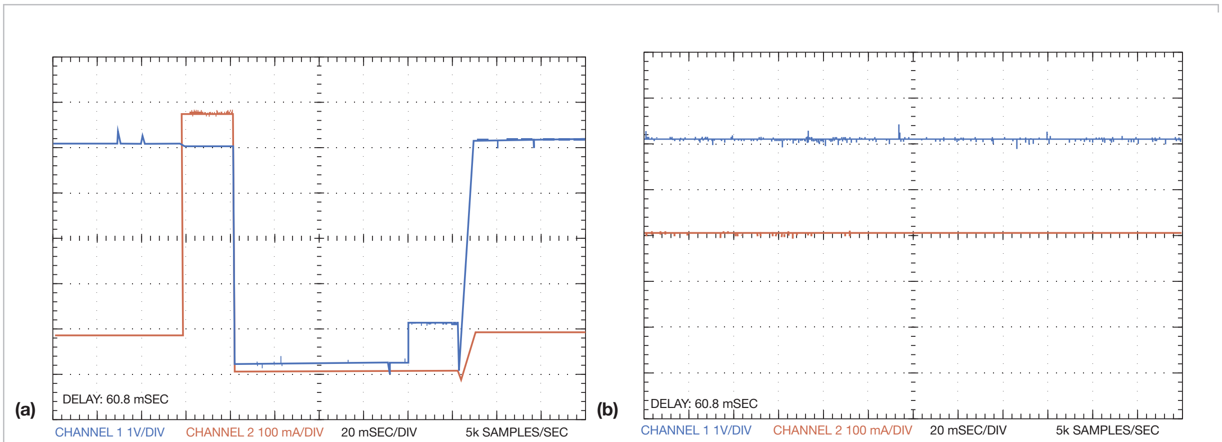

Figure 2 These current/voltage waveforms from Figure 1 show that an overload on Output A (a) has no effect on Output B (b).

This circuit is small because it requires few external components. You can build it into a cigarette-lighter plug or place it in a small space behind the dashboard. **EDN**

Use LEDs as photodiodes

Raju R Baddi, Raman Research Institute, Bangalore, India

The simple circuit in **Figure 1**, which can be powered with a 3.6V nickel-cadmium rechargeable battery, lets you use an LED to detect light. The circuit consumes practically no quiescent power. Two LEDs act as photodiodes to detect and respond to ambient light. When ambient light is present, the upper LED, a small, red, transparent device covered with a black pipe, has a higher effective resistance than the lower, large, green LED. The voltage drop across the input of the NAND gate is less than its threshold voltage for logic 1, making the output of the NAND gate low. When the ambient light goes off, the voltage drop across the reverse-biased green LED increases, forcing the NAND gate's output high.

This type of light detector is highly power-efficient and is ideal for battery applications. You can use the NAND gate's logic output to drive an LED driver or a relay driver, or you can connect it to a microcontroller.

Place the circuit so that sufficient light falls on the green sensor LED. Doing so avoids any voltage build-up near the junction that could be close to the NAND gate's threshold voltage. The NAND gate's power consumption rises sharply at the threshold voltage. When the gate's input voltage is within the defined limits for the logic state, its power consumption is extremely small. EDN

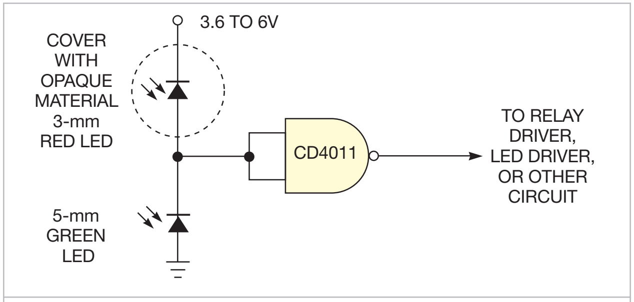

Figure 1 An LED's resistance changes with ambient light, which changes a voltage that drives a logic gate.

Circuit achieves constant current over wide range of terminal voltages

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Analog-circuit design often requires a constant-current sink. An example would be for a TRIAC (triode-for-alternating current) dimmer holding current in fluorescent or solid-state lighting. Other examples include a precise current sink at the end of a long line, such as a cable or an ADSL (asymmetric digital-subscriber-line) modem, which produces a "signature" current value that alerts the device at the source end, such as an exchange office or a cable center, that the remote equipment is attached. The trick is to make a circuit that gives a constant current over a variety of terminal voltages.

A common circuit for achieving this task uses a sense resistor, a transistor, and a power device. **Figure 1** shows the circuit using a power transistor, Q_1. The circuit provides an approximate constant current at high voltages, but it doesn't enter regulation until it reaches nearly 60V due to the base current the transistor requires. **Figure 2** shows the circuit using a MOSFET, Q_2, for the power device. With a MOSFET, you can use smaller biasing resistors, and the circuit comes into regulation at a much lower terminal voltage.

Unfortunately, the current-sense resistor, R_1, in **figures 1** and **2** doesn't

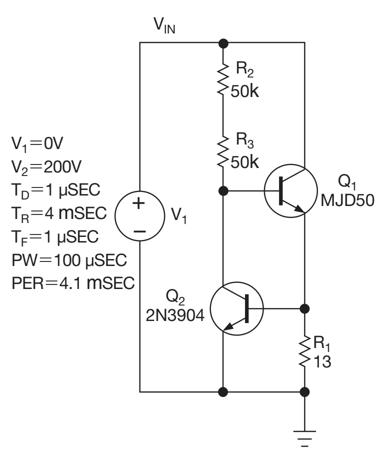

Figure 1 Resistor R_1 sets a constant current through Q_1.

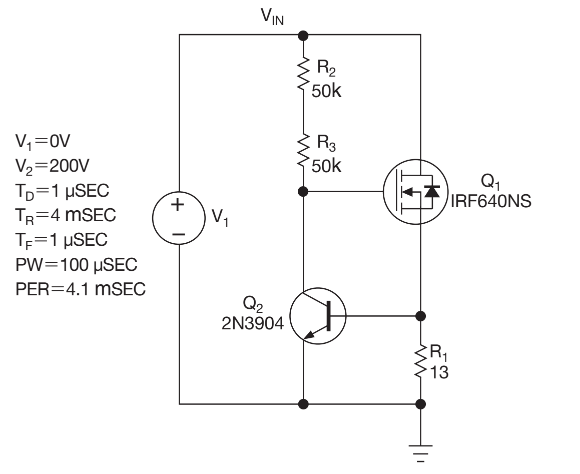

Figure 2 This circuit substitutes a MOSFET for Q_1 in Figure 1 and uses smaller resistors.

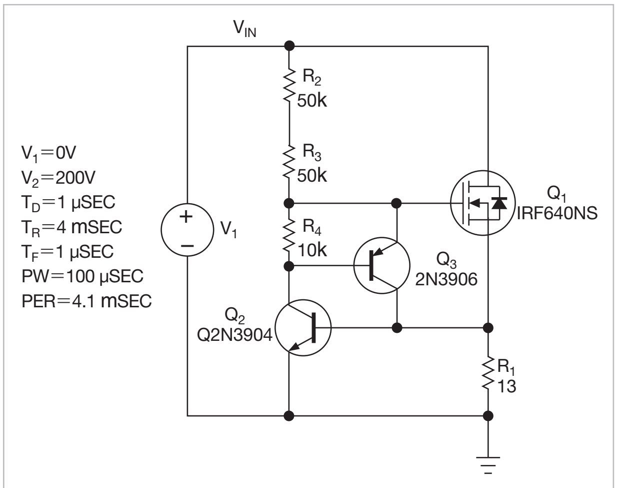

Figure 3 The addition of Q_3 and R_4 improves current regulation.

sense the bias current. As the terminal voltage increases, the terminal current also increases because of the increased bias current. A simple way to improve the regulation of both circuits is to add resistor R_4 and PNP transistor Q_3 (**Figure 3**). R_4 and Q_3 form a constant-current source to the collector of Q_2. The circuit diverts any excess bias current through the collector of Q_3 to sense resistor R_1. Thus, as the terminal voltage increases, the bias current remains relatively constant, and the current regulation appears much flatter.

The negative temperature coefficient of the base-to-emitter junction of transistor Q_2 causes another problem with this kind of circuit. The temperature coefficient is approximately −1.6 mV/°C, which causes the current value to vary widely with temperature. One way to approach this problem is to add a 6.2V zener diode, D_1, in series with the emitter of Q_2, which increases the sense voltage (**Figure 4**). A 6.2V diode has a positive temperature coefficient, which counteracts the negative temperature coefficient of the transistor. Furthermore, the total sense voltage is much larger, so 100 mV or so of voltage change with temperature does not seriously affect the regulated current. **Figure 5** shows a PSpice simulation of the circuit that uses a MOSFET for Q_1. EDN

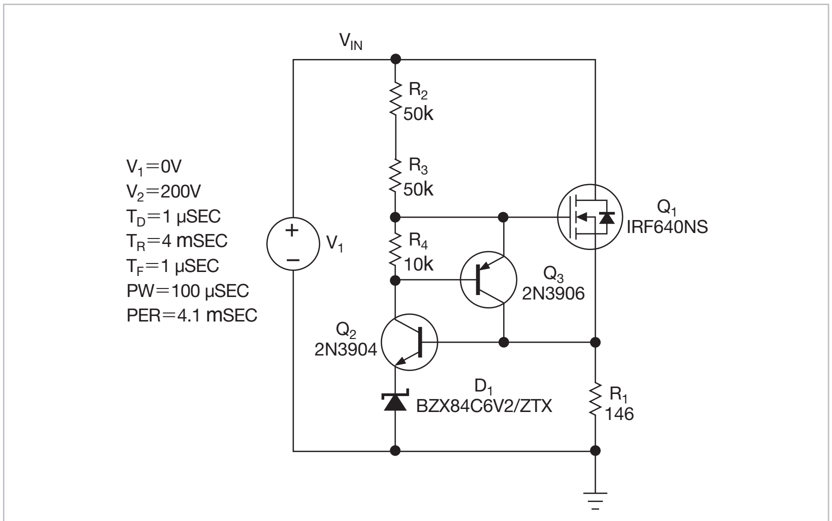

Figure 4 Adding a zener diode improves current regulation over temperature.

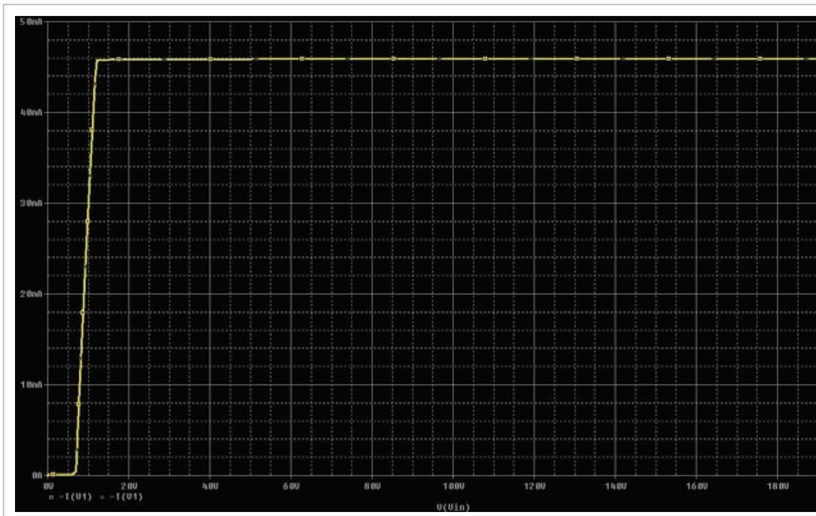

Figure 5 A constant current in Q_1 has a steep rise relative to V_{IN}.