Edited by Bill Travis

Buffer adapts single-ended signals for differential inputs

Randall Carver, Analog Devices Inc, Greensboro, NC

DC COUPLING of single-ended signals into differential-input, single-supply ADCs can be challenging. The input signal requires level shifting from ground to $V_S/2$ as well as single-ended-to-differential conversion. In addition, you must balance the differential inputs of the ADC to cancel even-order harmonics and common-mode noise. Systems often require this signal translation to take place without injecting dc bias currents back into the signal source. Processing wideband signals with large dynamic range (12- to 14-bit ADCs) can also add to the circuit complexity. Wideband amplifiers address nearly all these issues, but their standard implementation requires the use of ac coupling.

This Design Idea describes a new circuit that eliminates this requirement through the use of an external dc feedback loop. It also allows the lower end of the passband to extend to dc. The basis of the circuit is a simple level-shifting circuit (**Figure 1**). Tying two series resistors between V_S and a signal source attenuates the signal by a factor of two and biases it to $V_S/2$. The center tap is buffered; single-sided supply circuits can then process the signal. Two additional series resistors connected between the source and a negative supply of equal value remove dc bias currents from the source.

The circuit of **Figure 2** expands upon this simple concept by replacing the supply voltages $\pm V_S$ with precise $\pm V_{DC}$ levels that track one another. In addition, this design implements differential signaling by doubling the number of level-shifting resistors. You produce the $\pm V_{DC}$ levels by subtracting the 2.4V ADC reference signal (CML pin) from the common-mode level of the amplifier, which you form by summing the two amplifier outputs through equal-value resistors. The circuit amplifies, filters, and inverts the difference to create the $\pm V_{DC}$ levels. The dc feedback-loop gain of approximately 1040 allows the amplifier to track the output common-mode level to within (2.4V/1040)=2.3 mV of the ADC's reference (CML) signal. The addition of this external dc feedback path allows you to open the VOCM pin of IC_1 and de-

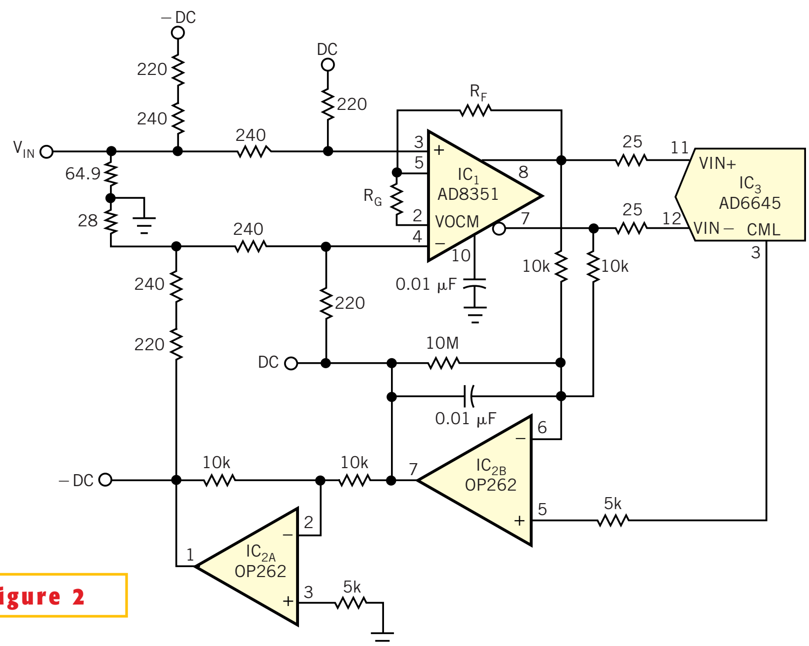

Figure 2

This circuit is a wideband, dc-coupled, single-ended-to-differential buffer.

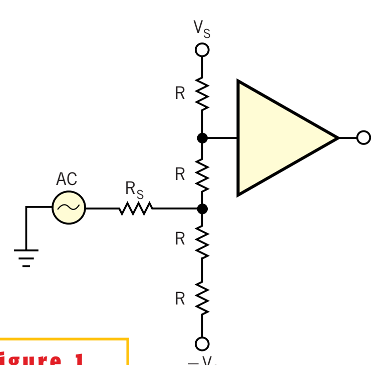

Figure 1

This simple circuit level-shifts ac signals to accommodate the $\pm V_S$ supplies.

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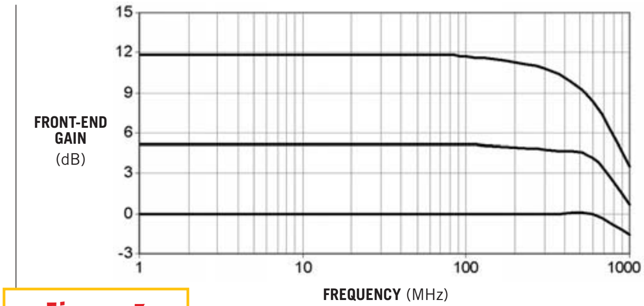


Figure 3

This graphic shows the frequency response of the circuit in Figure 2 for various values of gain.

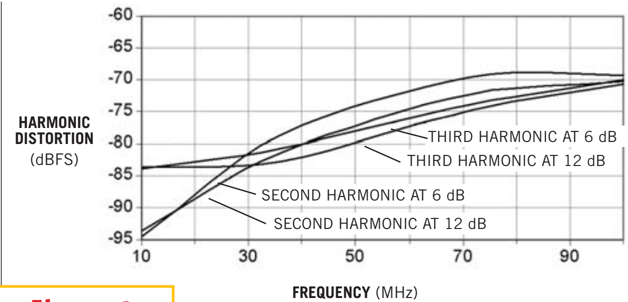


Figure 4

This plot shows harmonic distortion versus frequency for a −1-dB (referred to full-scale) input to the ADC at 80 MHz.

couple it to ground, disabling the AD8351's internal dc feedback path.

The level-shifting resistors have a ratio of 1.09-to-1 to reduce the required swing of the $\pm V_{DC}$ levels to ± 2.4 $[(1.09+1)/1.09] = \pm 4.6$V. The design uses accurate networks with excellent tracking to ensure good CMRR (common-mode-rejection ratio) and minimize the injection of dc bias currents into the source. IC_2 uses a rail-to-rail feedback amplifier to allow the use of ±5V supplies. The remaining circuits are powered from 5V. Resistor R_G varies the overall gain of the front end. For a front-end gain of 0 dB, the bandwidth extends beyond 1 GHz (**Figure 3**). After you determine the required gain, you adjust resistor R_F to balance the two differential signals into the ADC. **Table 1** shows typical values of R_G and R_F for various gain levels. The 64.9Ω resistor provides for a 50Ω source impedance. The 28Ω resistor provides for a balanced input that the amplifier sees. You can accommodate a differential-input signal structure by replacing the 28Ω resistor with a 64.9Ω resistor and tying the additional negative input signal to the junction of the new 64.9Ω resistor and the two 240Ω level-shifting resistors. This differential-input structure allows you to remove R_F. The circuit maintains the excellent distortion performance of the AD8351 amplifier, allowing the circuit to drive 12- and 14-bit ADCs with minimal degradation of the ADC's dynamic range (**Figure 4**).□

TABLE 1—RESISTOR VALUES FOR VARIOUS FRONT-END GAINS

R_G (Ω)	R_F (Ω)	Front-end gain (dB)
56.2	1540	12
154	698	6
1000	316	0

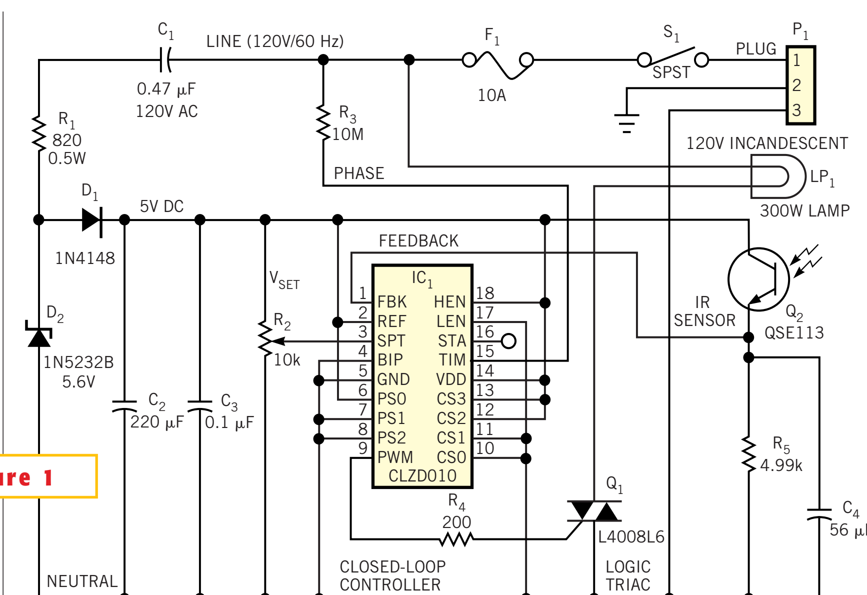
Triac lighting and heating controller uses few parts

David Caldwell, Flextek Electronics, Carlsbad, CA

The triac lighting-control circuit in **Figure 1** is small and inexpensive because load and housekeeping power come directly from the line voltage, thereby eliminating bulky, expensive supplies. The CLZD010 closed-loop controller maintains constant light intensity by automatically adjusting the timing of the triac's firing until the feedback signal and setpoint command are equal. The 5V supply is a charge pump that energizes C_1 on the negative swing of the line voltage and then transfers charge to C_2 on the positive swing. Zener diode D_2, minus the forward drop of rectifier D_1, sets the 5V. Triac Q_1 is a latching switch that conducts in either direction until you remove gate drive and load current drops below its holding threshold, which occurs at the zero-crossing point of the line voltage.

The circuit pulses the triac gate for 100

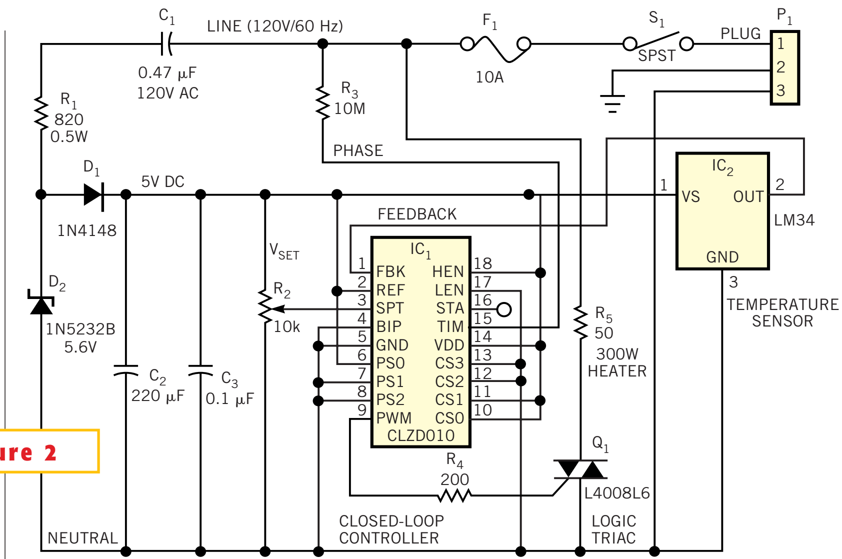
Figure 1



This closed-loop lighting-control system uses a few inexpensive parts.

µsec to turn on the load for the remainder of each 60-Hz half-cycle, so higher power accrues by turning on earlier in the half-cycle. R_3, at the timing pin of the controller chip, detects line phase. Controller pins CS3 to CS0 set the closed-loop configuration for an application. You can easily modify the lighting-control circuit for thermal control (**Figure 2**). Closed-loop timing is 134 sec for optimized temperature response, using controller pins CS3 to CS0. The circuit initially drives the heater at high power levels until the temperature nears its final value and then reduces the power to avoid overshoot. The CLZD010 controller is available from Flextek Electronics (www.flex-tek.com).□

Figure 2



A slight modification adapts Figure 1's circuit to for heating control.

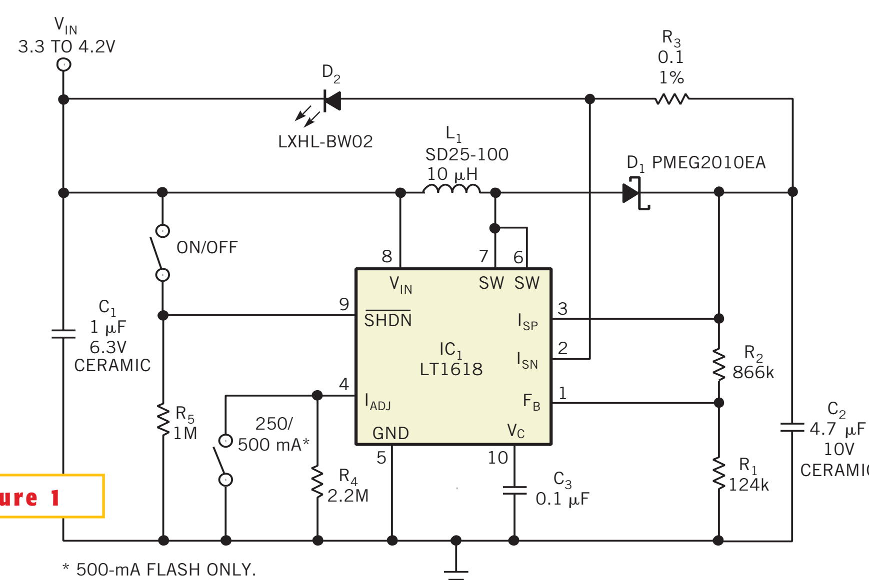
Constant-current, constant-voltage converter drives white LEDs

Keith Szolusha, Linear Technology Corp, Milpitas, CA

LEDs usually take their drive from a constant dc-current source to maintain constant luminescence. Most dc/dc converters, however, deliver a constant voltage by comparing a feedback voltage to an internal reference via an internal error amplifier. The easiest way to turn a simple dc/dc converter into a constant-current source is to use a sense resistor to convert the output current to a voltage and use that voltage as the feedback. The problem is that 500 mA of output current with a 1.2V drop—the typical reference voltage—in the sense resistor incurs relatively high power losses and, thus, a drop in efficiency.

One approach is to use an external op amp to amplify the voltage drop across a low-value resistor to the given reference voltage. This method saves converter efficiency but significantly increases the cost and complexity of a simple converter by using additional components and board space. A better approach is to use the LT1618 constant-current, constant-voltage converter, which combines a traditional voltage-feedback loop and a unique current-feedback loop to operate as a constant-voltage, constant-current dc/dc converter.

Figure 1



The LT1618 white-LED driver supplies 250-mA constant current and 500-mA flash from a lithium-ion battery.

Figure 1 shows the LT1618 driving a 1W, white Lumileds (www.lumileds.com) LXHL-BW02 Luxeon LED.

You need no external op amps for this compact approach. The LXHL-BW02 has a forward voltage of 3.1 to 3.5V for 250 mA of current. Although the maximum dc rating of the LED is 350 mA, you can pulse it up to 500 mA for a camera flash. R_4 is set for a 250-mA torch or dimming operation. The I_{ADJ} (current-adjustment) pin provides the ability to dim the LED during normal operation by varying the resistor setting or injecting a PWM signal. Access to both the positive and the negative inputs of the special internal constant-current amplifier allows you to place the sense resistor anywhere in the converter's output or input path and provide constant output or input current. Without access to both inputs, you would need a ground-referenced sense resistor, some additional level-shifting transistors, or an op amp. In this case, the floating sense resistor's value is only 100 mΩ; at 500 mA, it consumes an average of 50 mW of power. The sense resistor connects directly to the positive and the negative input pins of the LT1618.

Although the LT1618 conventionally serves as a high-frequency boost converter with the load connected between V_{OUT} and ground, this method of tying the load from V_{OUT} back to V_{IN} allows the IC to drive the LXHL-BW02 from a lithium-ion battery input. Tying the load back to V_{IN} allows the forward voltage of the LED (the load voltage) to be either above or below the input voltage as the battery voltage changes. This topology avoids the need for an additional inductor. This design uses one small, low-cost inductor, matching the all-ceramic capacitors and low-profile IC. Tying the load back to V_{IN} increases the inductor current by summing both the input and the output currents. The internal switch losses double, and the overall efficiency of the approach is approximately 70% over the input-voltage range. Even at this efficiency, it is difficult to match the compactness and low cost of this approach.□

Sequential state machine aids in automatic control

Abel Raynus, Armatron International, Malden, MA

Creating a sequential state machine is a common approach for automatic-control design. The method finds wide use for controlling sequential processes in industry, robotics, and measuring. The concept of a state machine is simple: A number of states describe a process under control; each state produces some output signals and advances to the next state according to received input signal.

Two kinds of state machines exist. In a Moore machine, only the current state determines an output. A Mealy machine's output depends on both the current state and the inputs. A state diagram uniquely and completely specifies a Moore machine (**Figure 1**). Based on this diagram, you can create a microcontroller program that assigns five assembler directives to each state: an address offset, the outputs, a time to remain in this state (delay value), the next state to go to for an input of zero, and the next state to go to for an input of one.

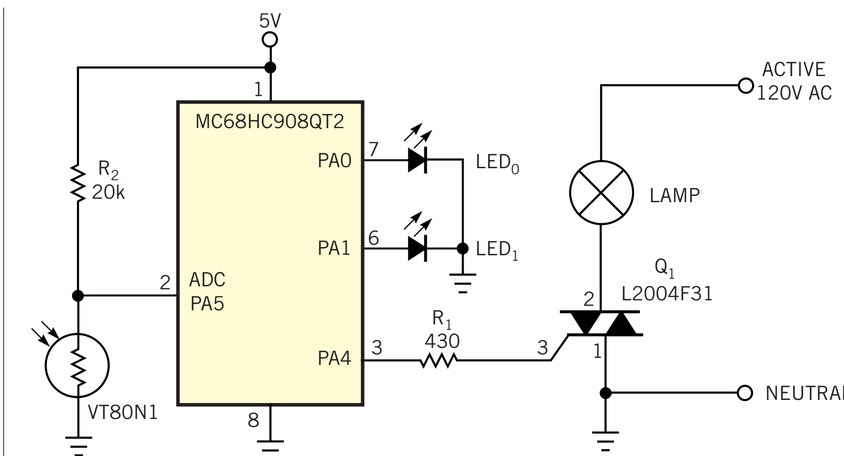


Figure 2 This simple circuit uses a Moore state machine to control a phototimer.

As an example of a Moore machine, consider a phototimer, which automatically lights a lamp for a given time when it becomes dark. After that time expires, it switches off the lamp and waits for the next night to repeat the process (**Figure 2**). The design uses an eight-pin Motorola (www.motorola.com) MC68HC908QT2 microcontroller; **Listing 1** at www.edn.com shows the program code. The photocell determines the Moore machine's input signal. The threshold of darkness is equal to 1.6V, which corresponds to a value of 52H after built-in analog-to-digital conversion. The sensitive triac, Q_1, model L2004F31 from Teccor (www.teccor.com) activates the lamp. The triac needs a gate current of 3 mA from the microcontroller can drive load current as high as 4A rms at a maximum voltage of 200V. The two LEDs with built-in resistor, LTL-4231T-R1s from LiteOn (www.liteon.com), indicate a current state in binary code. Thus, the Moore machine has two outputs—the lamp and the state indicator—both of which the Port A setting determines. The delays occur at 2 sec for states 1 and 3, and 4 minutes for state 2. In a real project, you can set any delay, even several hours or days. The delay affects only the complexity of the timer-interrupt routine. You can download the program code in **Listing 1** from the Web version of this Design Idea at www.edn.com.□

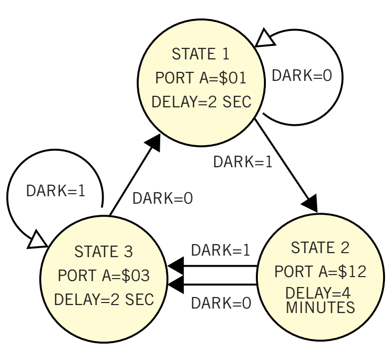


Figure 1 This state diagram uniquely and completely specifies the workings of a Moore state machine.

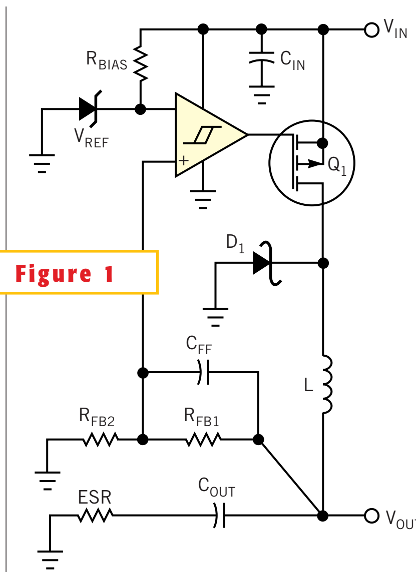
Hysteretic regulators provide high performance at low cost

Wayne Rewinkel, National Semiconductor, Phoenix, AZ

Hysteretic voltage regulators offer the potential advantages of simplicity, fast response, 100%-duty-cycle operation, high efficiency at light loading, and low cost. They need no loop-compensation components to add delays; thus, response time to a load change is less than one switching cycle. What's the catch? You must be able to accept a switching frequency that is not precisely controlled and a sensitivity to noise that requires layout skill.

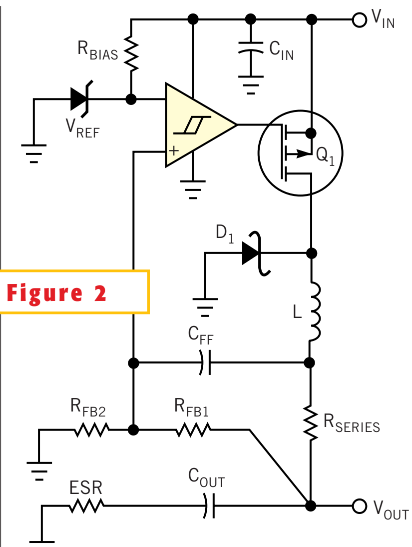
Figure 1 shows a simple hysteretic switching regulator made from a comparator with a fixed hysteresis and a PFET. The comparator switches on the PFET whenever V_{OUT} falls to its low threshold and off again when V_{OUT} rises to its high threshold. The time V_{OUT} lingers between the thresholds determines the on-time and, hence, the switching frequency. The inductor's ripple current flowing through the ESR of C_{OUT} provides a triangular voltage-ripple waveform, which produces predictable operation.

Figure 1



This hysteretic regulator suffers from unpredictability of the switching frequency because of C_{OUT} ESR variance.

Figure 2



An added series resistor makes this circuit's switching frequency more predictable.

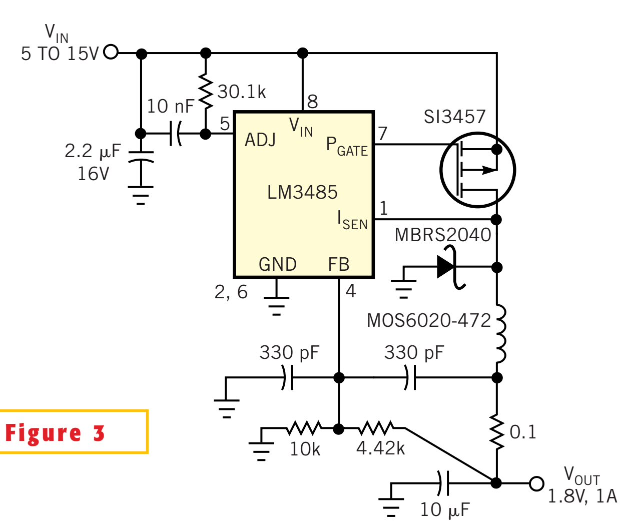
Herein lies a potential problem with simple circuits of this type. ESR is a major factor in determining switching frequency, and ESR can vary over a wide range for any given capacitor type. This variance is seldom a good thing and can lead to inductor saturation if the frequency falls too low or FET overheating arising from switching losses if the frequency rises too high. A simple solution to the ESR-variance problem is to use a ceramic C_{OUT} capacitor in series with a resistor. Although this technique works nicely in the lab, it often poses problems in the real world, in which several ceramic capacitors bypass loads.

Another approach to predictable frequency control allows the use of low-ESR capacitors (**Figure 2**). It is almost identical to **Figure 1**'s circuit except for the added resistor, R_{SERIES}, and the new connection point for C_{FF}. The inductor's ripple current induces the ac voltage present across R_{SERIES} and connects to comparator by C_{FF}. This controlled ac voltage eliminates the need for any C_{OUT} ESR. The feedback loop eliminates the dc voltage drop that R_{SERIES} creates. This new configuration produces predictable switching frequency with even zero-ESR capacitors and offers the potential of nearly zero V_{OUT} ripple at the cost of a resistor and the small added dissipation of R_{SERIES} carrying full load current.

The following equation approximates the switching frequency for either circuit, provided that C_{OUT}'s reactance at the switching frequency is lower than the ESR and C_{FF}'s reactance is much lower than R_{FB1}: $F_S=(V_{OUT}/V_{IN})(V_{IN}-V_{OUT})\times ESR/(V_{HYST}\times L+2ESR\times T_{PD}(V_{IN}-V_{OUT}))$, where ESR is the sum of C_{OUT}'s ESR and R_{SERIES}, V_{HYST} is the comparator's hysteresis voltage, and T_{PD} is the average propagation delay of the comparator plus the PFET.

You can build the circuits of **figures 1** and **2** as drawn, using a comparator, such

Figure 3



This circuit occupies an area smaller than a postage stamp.

as the LMV7219, which claims 7.5-mV built-in hysteresis, or by using a controller, such as the LM3485, which provides a current-limiting feature, wider V_{IN} range, and lower cost. You cannot overemphasize the layout sensitivity for hysteretic regulators. You cannot allow the feedback connection to pick up any stray signals. Open-core inductors are attractive for cost reasons but difficult to use, because any induced voltages from stray magnetic fields can produce unpredictable switching frequencies and ripple.

You can build the circuit in **Figure 3** in an area smaller than a postage stamp. This circuit produces output current of at least 1A, using small ceramic capacitors, a SOT-6 PFET, a 6×7-mm inductor, and an SMB-package, surface-mount Schottky diode. F_S varies from 600 to 700 kHz over a V_{IN} range of 5 to 15V for V_{OUT}=1.8V and V_{OUT} ripple less than 5 mV p-p. The 30.1-kΩ resistor and the PFET's on-resistance of 0.1Ω set the current limit to trigger at 1.5A. The no-load bias current is lower than 500 μA. Most impressive is the dynamic V_{OUT} change of only 10 mV for a load transient greater than 0.5A.□

Power-supply IC drives multiple LEDs

John Lo Giudice and Vee Shing Wong, STMicroelectronics, Schaumburg, IL

Bright LEDs are becoming prominent sources of light. They often have better efficiency and reliability than do conventional light sources. Although LEDs can operate from an energy source as simple as a battery and a resistor, driving a string of LEDs in constant-current mode can better match the luminance between the devices without needing to match LEDs for their forward-voltage drop. A switching supply also gives better efficiency than methods using a linear ballast resistor to limit the current.

The circuit in **Figure 1** uses IC_1, the integrated offline Viper22A switching regulator in a constant-current configuration to drive two to eight 1W LEDs. The circuit operates by monitoring the voltage drop across the sense resistor, R_6, and uses this voltage as feedback to regulate the current through R_6 and the LEDs. An operational amplifier in the TSM103, IC_2, monitors the voltage drop across R_{SENSE} and compares it with the 0.175V reference, which resistor divider R_5-R_7 sets, and closes the loop to maintain a 0.175V drop across the sense resistor. The output of the TSM103 drives the optocoupler, IC_3, which transfers the feedback to the Viper22A on the primary side. **Figures 2a**, **2b**, and **2c** show the completed board and the top and bottom layouts.

The LED's drive current is $I_{OUT}=V_5/R_6$, where V_5 is the voltage at Pin 5 of IC_2. You could easily modify the circuit to drive 3 or 5W LEDs, as long as you respect the maximum power rating, by simply changing the sense resistor to set a higher current value. When you design a con-

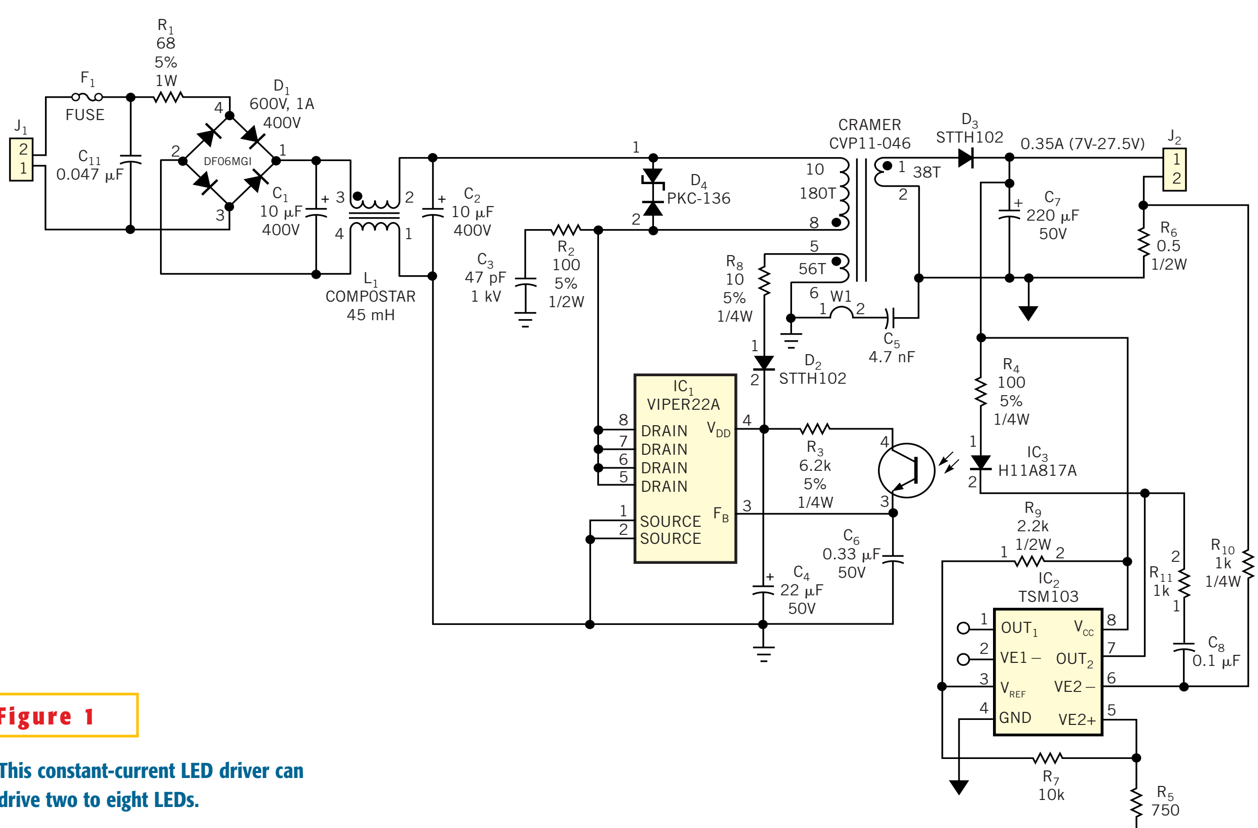


Figure 1

This constant-current LED driver can drive two to eight LEDs.

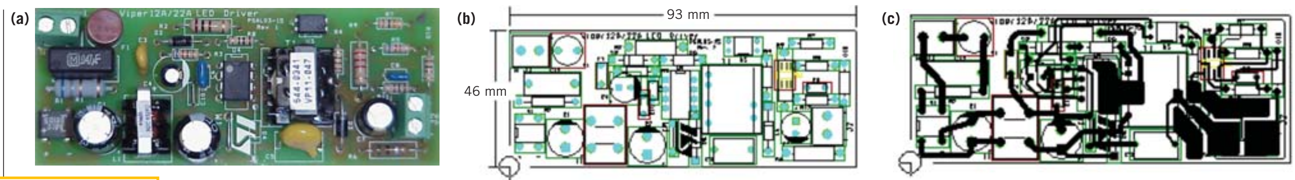


Figure 2 These photos show the completed board (a) and the top (b) and bottom (c) layouts.

stant-current power supply, the design of the transformer and the operating limitations of the circuit directly determine the output-voltage compliance of the constant-current source. For a design that can drive two to eight LEDs in series, the voltage drop across the LEDs can vary from approximately 7V for two LEDs to 28V for eight LEDs. This output voltage reflects back across the transformer and in turn changes the V_{DD} voltage to the control circuit and the peak V_{DS} across the power MOSFET.

The designers of the transformer in this application considered three limiting factors: the allowable V_{DD} for the Viper22, which has a range of 9 to 38V for the undervoltage and overvoltage thresholds, respectively; the maximum wattage of 12W for the Viper22A; and the fact that the reflected voltage across the drain of the MOSFET, which takes account of the turns ratio $[(N_P/N_S)V_{OUT}]$, added to the input voltage, must be less than 730V.

For a design that can drive two to eight LEDs, you must design the system taking into account the fact that the reflected voltage on V_{DD} is proportional to the output voltage. To keep the reflected voltage manageable, the transformer's design uses a turns ratio of primary to secondary output voltage for the maximum number of LEDs. Using these criteria, as the number of LEDs decreases, so does the reflected voltage. If you base the transformer on two LEDs, then the reflected voltage quadruples with eight LEDs and may exceed the rating of the Viper. The turns ratio between secondary to the V_{DD} winding is set for an output voltage of two LEDs to the minimum V_{DD} voltage of 9V. As you add LEDs, V_{DD} increases proportionally until it reaches the overvoltage-shutdown point of 42V nominal.□

Supercapacitor boosts current from small battery

Yongping Xia, Navcom Technology, Torrance, CA

Some battery-powered devices require large amounts of current in a short period of time but spend most of the time in sleep (power-down) mode. The momentary large-load current demands large batteries to meet the time requirement, even though the average current consumption is low. For instance, a system operates for 1.5 sec every 10 hours and needs 500 mA at 3.3V during the operation. Although the average current is only 21 μA, small, "coin" batteries cannot drive such a heavy load.

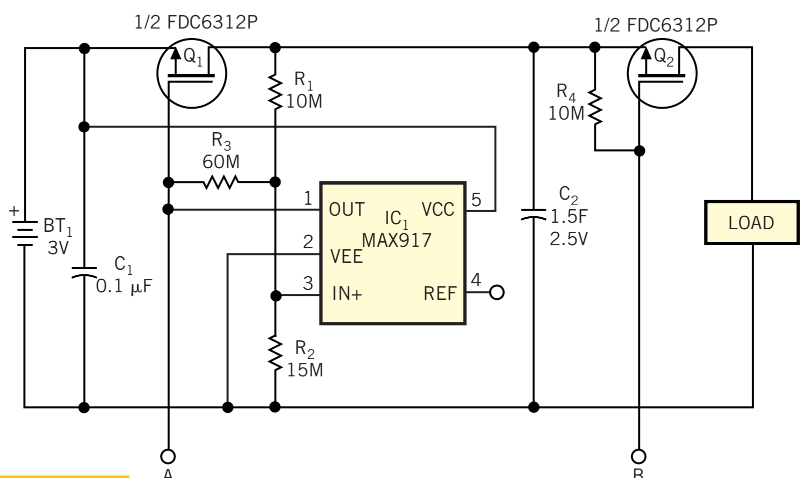


Figure 1 A supercapacitor helps a small battery to deliver large pulses of energy.

To eliminate the need for larger batteries, the circuit in **Figure 1** solves the problem by gradually building up energy in a supercapacitor. The device releases the energy when it is needed. Because the supercapacitor has low internal impedance, the momentary current can easily exceed several amperes.

Because a coin-type lithium battery delivers 3V and the supercapacitor's rated voltage is 2.5V, the circuit uses a voltage-controlled switch to cut off the battery once the voltage on the supercapacitor reaches 2.2V. This design uses a 1.5F, 2.5V supercapacitor from PowerStor (www.powerstor.com), model A10-30-2R5155. IC_1, a micropower voltage comparator with built-in 1.245V reference, senses the voltage on the supercapacitor. R_3 provides 0.5V hysteresis to the comparator. When the voltage is lower than 1.7V, the comparator's output is low and thus turns on the p-channel MOSFET, Q_1. The battery charges the supercapacitor. Once the voltage on the supercapacitor reaches 2.2V, the comparator switches high to shut off Q_1. You could use this low-to-high transition at Point A as a battery-charge-complete indicator or to trigger another device, such as a microcontroller's interrupt line.

Q_2, another p-channel MOSFET, controls the discharge of the supercapacitor. When Point B is floating, the switch is off. When an open-drain or open-collector device pulls down Point B, the switch is on. Because the voltage on the supercapacitor continuously drops when the switch is on, you can use a boost dc/dc converter to generate a constant output voltage. Select a boost converter with the lowest possible working input voltage to obtain the maximum energy from the supercapacitor. For example, you can use an LTC3402 to generate a stable 3.3V output. Once it starts, the LTC3402 can work with input voltages as low as 0.5V. The energy from the supercapacitor is $1/2V^2C$, or $1/2[(2.2V)^2\times 1.5F-(0.5V)^2\times 1.5F]=3.4J$.□