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Edited by Bill Travis

Circuit provides 4- to 20-mA loop for microcontrollers

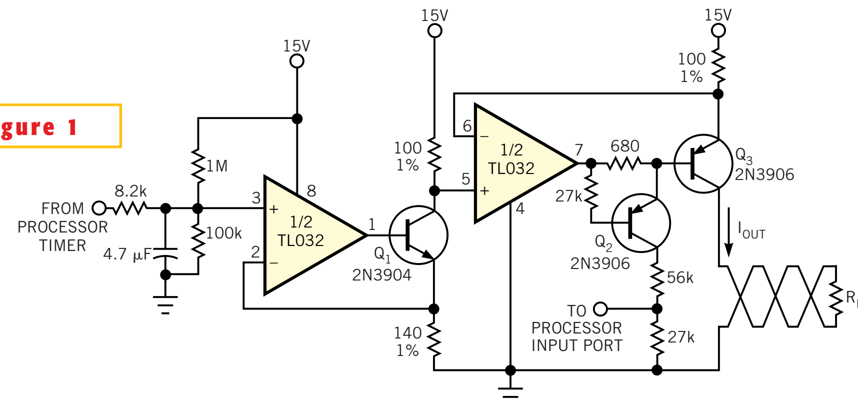
Robert Most, Dow Corning Corp, Auburn, MI

THE 4- TO 20-MA current loop is ubiquitous in the world of controls in manufacturing plants. Discrete logic, microprocessors, and microcontrollers easily cover the digital portions of control schemes, such as limit switches, pushbuttons, and signal lights. Interfacing a 4- to 20-mA output to a rudimentary microcontroller can be problematic. A built-in A/D converter would be nice, but such a device is sometimes unavailable in the "economy" line of these processors. Serial 4- to 20-mA chips exist but are relatively expensive and require serial programming and involve microcontroller overhead. Most lower end chips lack dedicated serial ports and require pin-programming.

This circuit is a low-cost alternative that provides not only a 4- to 20-mA output, but also a digital feedback signal that indicates an open wire in the current loop (**Figure 1**). One output-port pin sets the current, and one input-port pin monitors an open circuit in the loop wire. The circuit does not require the open-loop feedback portion of the circuit for the current loop to operate; you can omit it for further cost savings.

The circuit derives its drive from a simple timer output in the microcontroller. The duty cycle of the timer determines the output current of the circuit. The input RC network in front of the first operational-amplifier signal conditions the pulse train from the processor, so that the op amp interprets it as a dc voltage. In addition, the network ensures that the minimum input voltage is close to 100 mV, even if the input is at ground potential. This minimum voltage ensures that the feedback loop of the first op amp does not fold back to the positive rail when you cut off npn transistor Q_1. If you use a dual supply, the transistor has the additional voltage swing below ground potential to keep it in its active region and does not cut off. The emitter resistor of npn transistor Q_1 sets the current span of the circuit. With a 5V drive from the microcontroller, the output current is 20 mA. A grounded input results in less than 1 mA. A duty cycle of 12.5% drives the loop at 4 mA and exhibits linear control to full scale. Although it may not be mandatory, most current loops prefer a grounded return path. The purpose of the second operational amplifier is to provide a current source, rather than the current sink of the first stage, and the grounded return path. Hence, pnp transistor Q_3 provides this high-side drive. Bipolar-junction transistors Q_1 and Q_3 meet cost considerations, but you could also use MOSFETs for slightly better performance.

Figure 1



This configuration provides both a 4- to 20-mA loop and an open-circuit indication.

The open-loop feedback portion of this circuit lets the microcontroller know that a fault condition exists on the line. The processor can then execute alarm, shutdown, or other control functions to mitigate possible safety concerns. When an open-loop condition occurs, Q_3 shunts the entire loop current back through its emitter-base junction and through the 680Ω resistor to the op amp. The voltage developed across the 680Ω resistor turns on Q_2, resulting in a logic-one feedback to the microcontroller. Note that the open loop requires at least 1 mA of current for the open indication to function, which is below the normal 4 mA—a "zero" output condi-

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tion for this type of control system.

Response time for a step change is approximately 500 msec, which is acceptable for most current-loop control devices, such as control valves. If the microcontroller you select has a built-in A/D converter, response time can decrease by a couple of orders of magnitude with the elimination of the input-filtering network. Op-amp selection is important if you use a single-supply topology. An operational amplifier that can maintain stability close to its negative, or ground, rail is an important asset.□

Minimize the short-circuit current pulse in a hot-swap controller

Jim Sherwin and Thong Huynh, Maxim Integrated Products, Sunnyvale, CA

Because of internal circuit-breaker delay and limited MOS-gate pulldown current, many hot-swap controllers do not limit current during the first 10 to 50 μsec following a shorted output. The result can be a brief flow of several hundred amperes. A simple external circuit can counter this problem by minimizing the initial current spike and terminating the short circuit within 200 to 500 nsec. A typical 12V, 6A, hot-swap-controller circuit contains, as do many others, slow and fast comparators with trip thresholds of 50 and 200 mV (**Figure 1**). The 6-mΩ sense resistor, R_S, allows a nominal slow-comparator trip at 8.3A for overload conditions and a fast-comparator trip at 33.3A for short circuits. Only circuit resistances limit the initial short-circuit current spike during a period that includes the fast-comparator delay and the 30 μsec it takes to complete interruption of the short circuit by discharging M_1's gate capacitance. Various elements, such as R_S and the on-resistance of M_1, contribute to the circuit resistances. The waveform recorded during a short circuit indicates a peak current of 400 from the 2.4V peak across R_S, decreasing to 100A in 28 μsec (**Figure 2**).

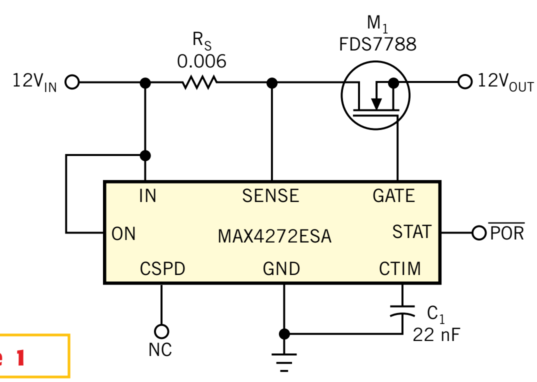


Figure 1

A typical hot-swap controller circuit exhibits a 30-μsec short-circuit current pulse of 400A peak.

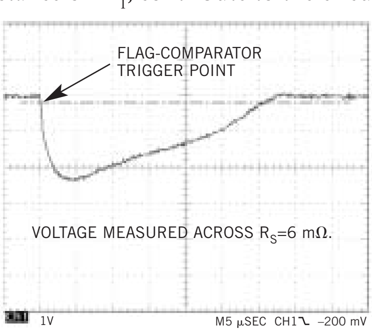


Figure 2

The short-circuit current in Figure 1 is 400A, decreasing to 100A in 28 μsec.

You can limit the short-circuit current duration to less than 0.5 μsec by adding a Darlington pnp transistor, Q_1, to speed the gate discharge (**Figure 3**). D_1 allows the gate to charge normally at turn-on, but, at turn-off, the controller's 3-mA gate-discharge current is directed to the base of Q_1. Q_1 then acts quickly to discharge the gate, in less than 100 nsec. Thus, the high-current portion of the short circuit is limited to slightly more than the fast comparator's delay time of 350 nsec. The apparent reverse overshoot current and the steep rise in the waveform of **Figure 4** arise from parasitic series inductance in the sense-resistor chip. The circuit of **Figure 5** can limit short-circuit current to approximately 100A for less than 200 nsec. The pnp transistor, Q_{1A}, which triggers when the voltage across R_S reaches approximately 600 mV, drives the npn transistor, Q_{1B}, to quickly discharge M_1's gate capacitance. The steep voltage waveform aids quick triggering of the pnp transistor.

The oscilloscope's ground lead introduces an artifact, which appears as the leading-edge oscillation in **Figure 6**. Again, as in **Figure 4**, the apparent reverse-overshoot current and the steep rise in the waveform of **Figure 6** arise from parasitic series inductance in the sense-resistor chip. C_2 connects between the gate and source of M_1 to reduce the positive-transient step voltage applied to the gate during a short circuit. Zener diode D_1 reduces $I_{D(ON)}$ by limiting V_{GS} to less than the 7V available from the MAX4272. Although D_1

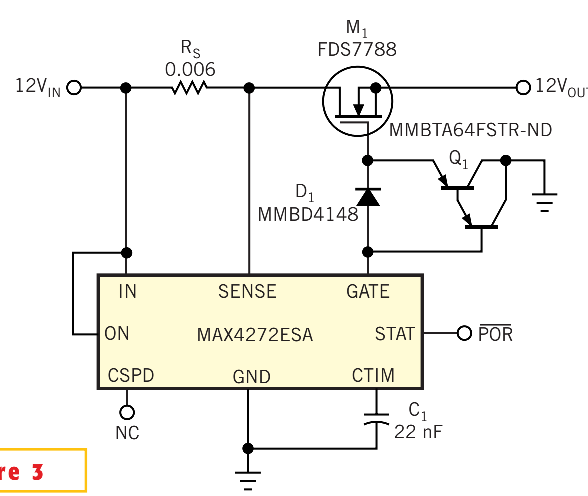


Figure 3

The addition of Q_1 increases the gate-pulldown current, limiting the short-circuit-current duration to less than 0.5 μsec.

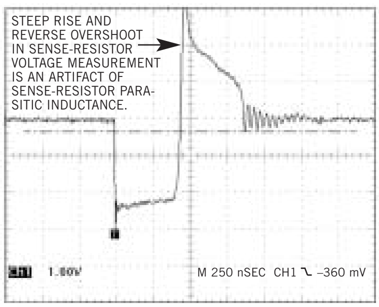


Figure 4 The steep rise and reverse overshoot in Figure 3's circuit are artifacts of sense-resistor parasitic inductance.

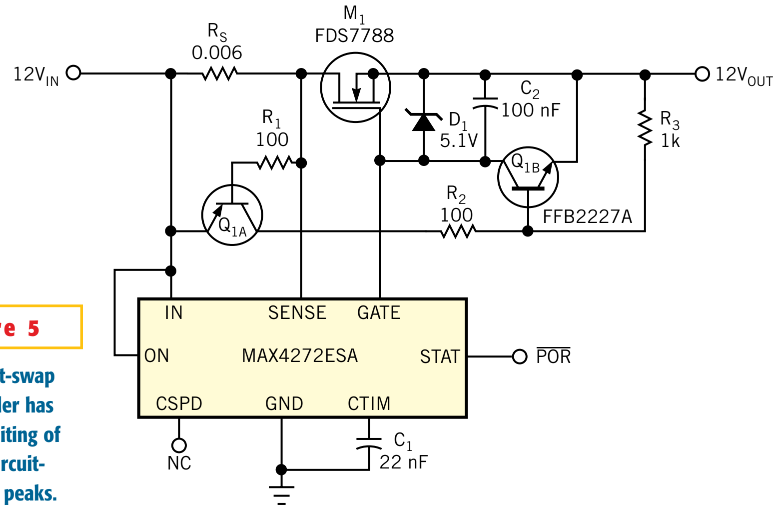


Figure 5 This hot-swap controller has fast limiting of short-circuit-current peaks.

is rated at 5.1V when biased at 5 mA, it limits V_{GS} to approximately 3.4V in this circuit because only 100 µA of gate-charging current (zener-bias current) is available from the IC. The limited V_{GS} lowers $I_{D(ON)}$—at some expense to on-resistance—and allows a quicker turn-off of M_1. You could also use D_1 and C_2 to some advantage in **figures 1** and **3**, to reduce $I_{D(ON)}$ during short circuits.

Either of the two circuits can protect a backplane power source by minimizing the energy dissipated when a hot-swap-controller circuit incurs a short circuit. The simpler circuit (**Figure 3**) dramatically shortens the short-circuit-current interval to somewhat less than 500 nsec, and the slightly more complex circuit (**Figure 5**) reduces the peak short-circuit current to 100A, as well as truncating the pulse width to less than 200 nsec. You can apply either technique to most hot-swap-controller circuits. Individual results vary according to the impedance of the power source, the impedance of the short circuit, and the quality and attack time of the short circuit itself. Note that it is inordinately difficult to achieve a repeatable low-resistance short circuit by manual manipulation of a shorting bar. You require careful layout and low-ESR capacitors to create a power source with very low ESR.□

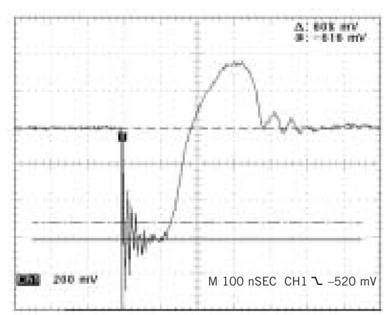


Figure 6 This waveform depicts the short-circuit-current peaks for the circuit in Figure 4.

Reduce EMI by sweeping a power supply's frequency

John Betten, Texas Instruments, Dallas, TX

Switching power supplies can be notorious noise generators. You should prevent this noise, which is conducted, radiated, or both, from returning to the input source, where it can potentially wreak havoc on other devices operating from the same input power. The goal of an EMI (electromagnetic-interference) filter is to block this noise and provide a low-impedance path back to the noise source. The larger the noise, the greater the size, expense, and difficulty of the filter design. Power supplies that operate at a fixed frequency have their largest EMI emission at this fundamental, fixed frequency. Emissions also occur at multiples of the switching frequency but at diminished amplitudes. The simple circuit in **Figure 1** makes the switching converter operate over multiple frequencies rather than one, thereby reducing the time average at any one frequency. This scheme effectively lowers the peak emissions.

The circuit in **Figure 1** is a self-starting oscillator with an oscillation frequency of approximately 500 Hz. When you apply power, C_3 begins to charge up from 0V, and the output of the TL331 comparator is in a high-impedance state because its noninverting input sees a higher voltage than that of the inverting input. As C_3 charges, its voltage crosses the voltage reference of the R_1-R_6 divider, and the comparator output trips to a low state. The voltage on R_6 instantly drops to a lower reference level because R_5 is now in parallel with R_6. C_3 begins to discharge toward this new reference level because R_3 is simultaneously in parallel with C_3. The cycle repeats after C_3 discharges to the voltage on R_6 when the comparator output reopens. You must carefully select the components to ensure that the two voltage-reference states of R_6 are lower than the upper and lower possible charge states of C_3. The circuit uses C_3 to adjust the oscillator frequency; you should select C_3 to have a lower value than C_2. The oscillator's frequency is approximately equal to

$$f = \frac{1}{-2R_2C_3 \ln\left[1 - \frac{\Delta V_{REF}}{V_{BIAS}}\right]}.$$

Capacitor C_2 ac-couples the ramp voltage of C_3 into the UCC3813's oscillator pin. The injected signal adds to the charging current of C_T during its positive portion (ac signal), thus increasing the controller's operating frequency. During the injected signal's negative portion, some of C_T's charging current disappears, slowing the controller's operating frequency. **Figure 2** shows the effects of the injected signal on the charging of C_T. R_4 controls the magnitude of the current that is injected. Reducing R_4's value increases the range, or spread, of the operating frequency around its nominal fixed frequency. The injected signal's oscillation frequency, which C_3 sets, controls the frequency-sweep rate.

The differential EMI-current measurement of **Figure 3** (1 dBμV=1 dBμA) shows the before-and-after effects of adding the frequency-shifting oscillator. This design easily achieves a 10-dBμA reduction with a 12-kHz sweep window. A wider window further reduces EMI, but the modulator frequency may be noticeable in the converter's output ripple voltage. It is also desirable to make the injected ramp voltage as linear in shape as possible to prevent the switching converter from spending excess time at its switching-frequency limits. The nonlinearity can result in an EMI response with two distinct frequencies. You must take care not to operate the circuit below the power converter's low-frequency limits, or saturation of magnetics may occur. This circuit demonstrates a low-cost, small-area approach to reducing conducted-EMI emissions.□

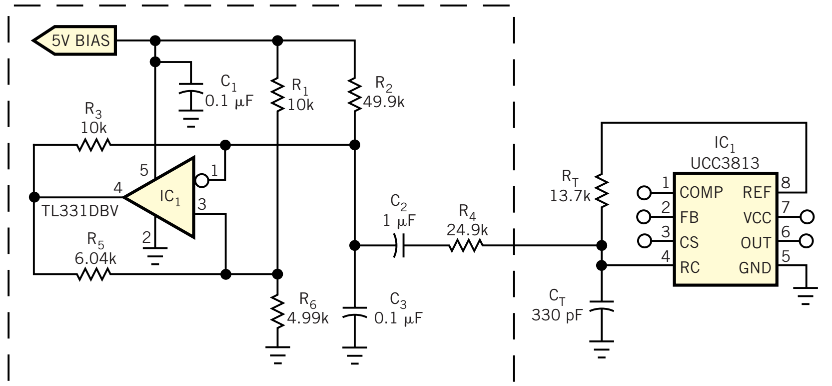


Figure 1 A low-frequency oscillator ramp, injected into the RC pin, modulates the supply's switching frequency.

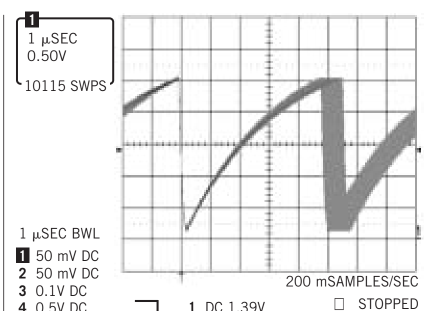


Figure 2 The external oscillator varies the charging of the timing capacitor.

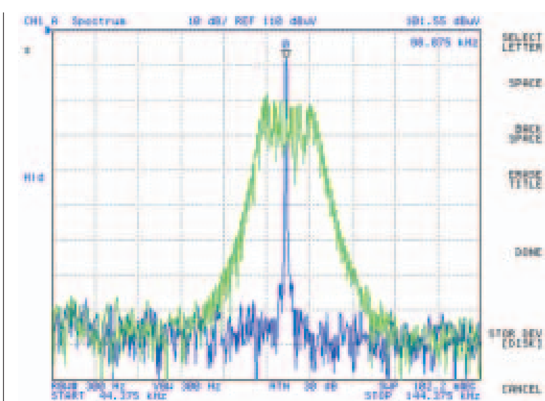


Figure 3 The EMI of the flyback converter differs with and without external modulation.

Get just enough boost voltage

Kieran O'Malley, On Semiconductor, East Greenwich, RI

Adding a current-mirror circuit to a typical boost circuit allows you to select the amount of boost voltage and to ensure a constant difference between the input and the output voltages (**Figure 1**). This circuit is useful for high-side-drive applications, in which a simple voltage doubler is unacceptable because of the voltage range of the components involved or where the input voltage can vary widely. You can also use the circuit at the front end of a power supply to ensure that the PWM controller has enough voltage to start correctly in low-input-voltage con-

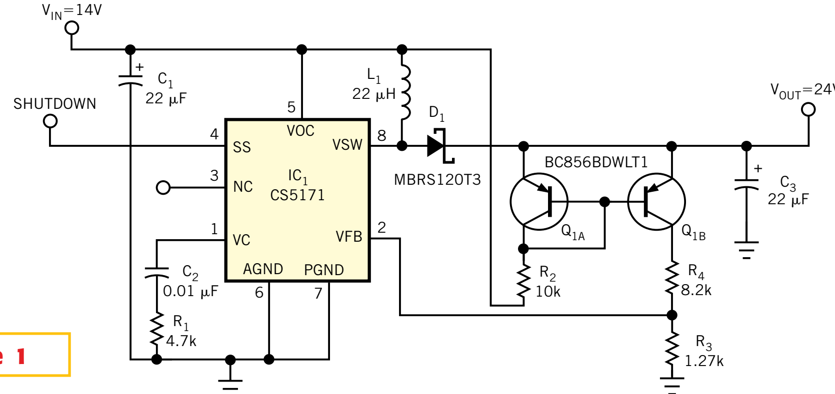


Figure 1 Adding a current-mirror circuit to a boost circuit allows you to get just enough boost voltage.

ditions. The circuit maintains a 10V difference between V_{IN} and V_{OUT}, but you can easily change it to provide other voltages. The PWM circuit in **Figure 1** is the CS5171 from On Semiconductor (www.onsemi.com), but you can use the idea with any boost circuit. The current-mirror circuit, comprising the dual-pnp transistor, Q_1, and the associated resistors, establishes a current that depends on the voltage difference between V_{IN} and V_{OUT}. The dual-pnp transistor has a V_{CEO} of 65V. In this case, V_{IN}=14V (nominal), so you need V_{OUT} to be 24V (nominal). First, calculate a value for R_2, thus establishing the reference current. If you select a reference current of 1 mA, you obtain

$$\frac{(V_{IN}-V_{OUT}-V_{BE}(Q_{1A}))}{1\ \text{mA}}=R_2$$

$$\frac{(14\text{V}-24\text{V}-0.6\text{V})}{1\ \text{mA}}=9.4\ \text{k}\Omega.$$

Because the output voltage is not critical, you use a 10-kΩ resistor.

Q_{1B} mirrors the current and sets up the feedback voltage to the PWM circuit. The CS5171 has an internal voltage of 1.28V (typical), so R_3 yields the correct feedback voltage when the current flowing through it is 1 mA. In this case, by selecting 1.27 kΩ for R_3, you obtain an output voltage of 24V. As V_{IN} varies, V_{OUT} tracks it and maintains a 10V difference between the input and the output. R_4 helps reduce the power dissipation in Q_{1B}.□

Processor's PWM output controls LCD/LED driver

Joe Neubauer, Maxim Integrated Products, Sunnyvale, CA

The PWM (pulse-width-modulation) output available from many microprocessors is based on an internal 8- or 16-bit counter and features a programmable duty cycle. It is suitable for adjusting the output of an LCD driver (**Figure 1**), a negative-voltage LCD driver (**Figure 2**), or a current-controlled LED driver (**Figure 3**). The circuit comprises simply the PWM source, capacitor C, and resistors R_D and R_W. For CMOS circuits, you calculate the open-circuit output voltage as $V_{CONT}=D\times V_{DD}$, where V_{CONT} is the control circuit's output voltage, D is the PWM duty cycle, and V_{DD} is the logic-supply voltage. The control circuit's output impedance is the sum of the resistor values R_D and R_W: $R_{CONT}=R_D+R_W$. For the circuit of **Figure 1**, the output voltage, V_{OUT}, is a function of the PWM average voltage, V_{CONT}:

$$V_{OUT}=V_{REF}\left[1+\frac{R_1}{R_2}\right]+\frac{(V_{REF}-V_{CONT})\times R_1}{R_{CONT}},$$

where V_{REF} is the reference voltage at the feedback input.

Bear in mind that the initial charge on filter capacitor C produces a turn-on transient. The capacitor forms a time constant with R_{CONT}, which causes the output to initialize at a voltage higher than that intended. You can minimize this overshoot by scaling the value of R_D as high as possible with respect to R_1 and R_2. As an alternative, the microprocessor can disable the LCD until the PWM volt-

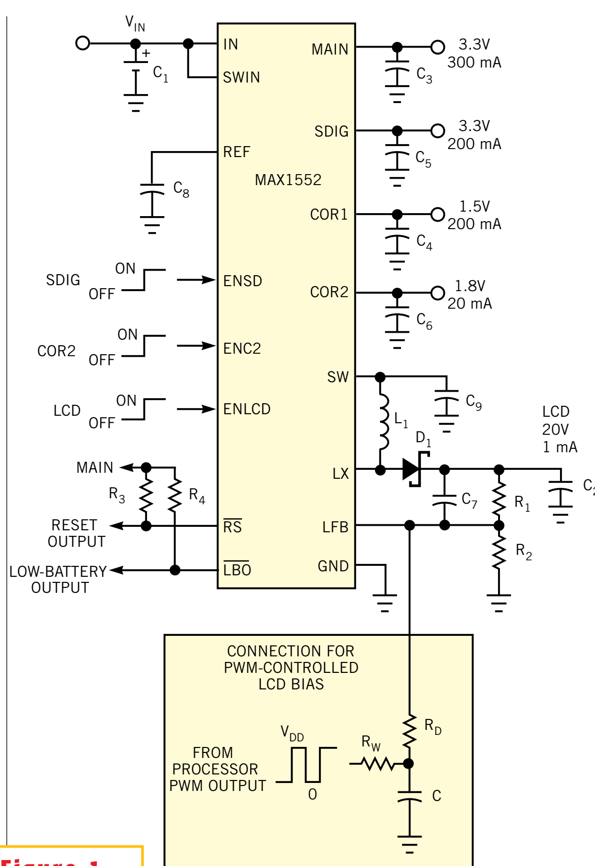


Figure 1

This simple circuit provides positive-output voltage LCD drive.

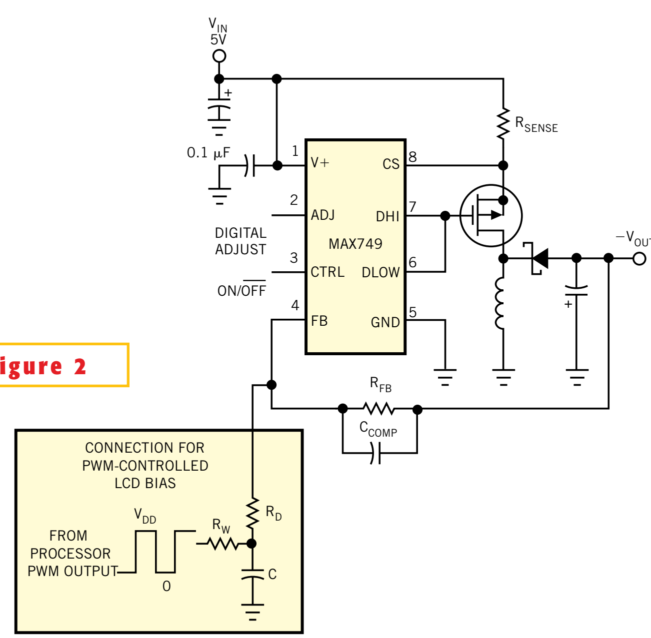


Figure 2

This configuration provides negative-output-voltage LCD drive.

age stabilizes. For **Figure 2**, the output voltage, V_{OUT}, is a function of the PWM average voltage, V_{CONT}:

$$V_{OUT} = V_{REF} + \frac{(V_{REF} - V_{CONT}) \times R_{FB}}{R_{CONT}},$$

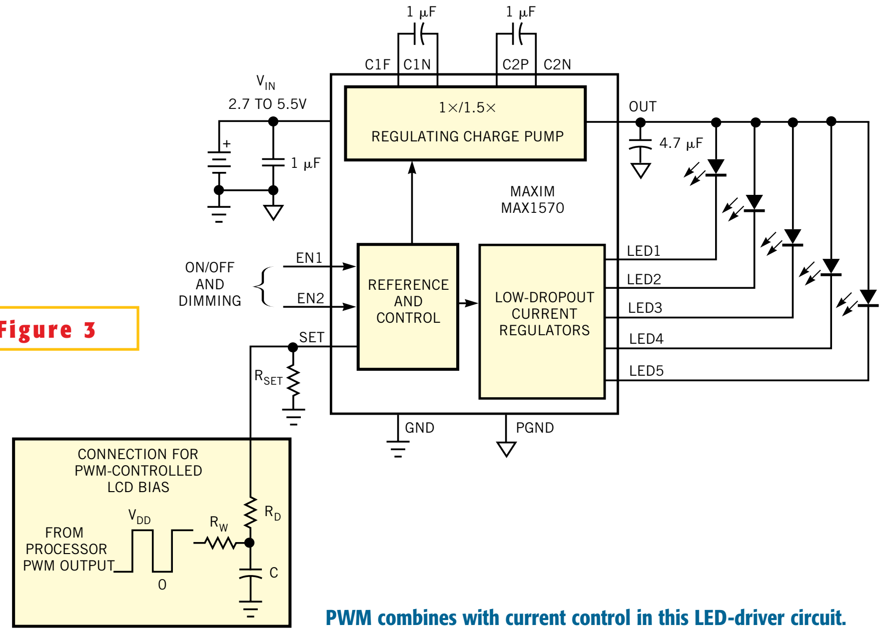
where V_{REF} is the reference voltage at the feedback input. For **Figure 3**, the output current is a function of the PWM average voltage, V_{CONT}:

$$I_{OUT} = \left[\frac{V_{REF} + \frac{(V_{REF} - V_{CONT}) \times R_{SET}}{R_{CONT}}}{R_{SET}} \right] \times K,$$

where V_{REF} is the reference voltage at the Set output and K is the current-scaling factor.

R_D isolates the capacitor from the feedback loop in the PWM-control methods. Assuming a stable voltage at the feedback point, the following equation defines the lowpass filter's cutoff frequency: $f_C = 1/(2\pi\ RC)$, where $R = R_D || R_W$. To minimize ripple voltage at the output, you should set the cutoff frequency at least two decades below the PWM frequency.□

Figure 3



PWM combines with current control in this LED-driver circuit.

Method provides automatic machine shutdown

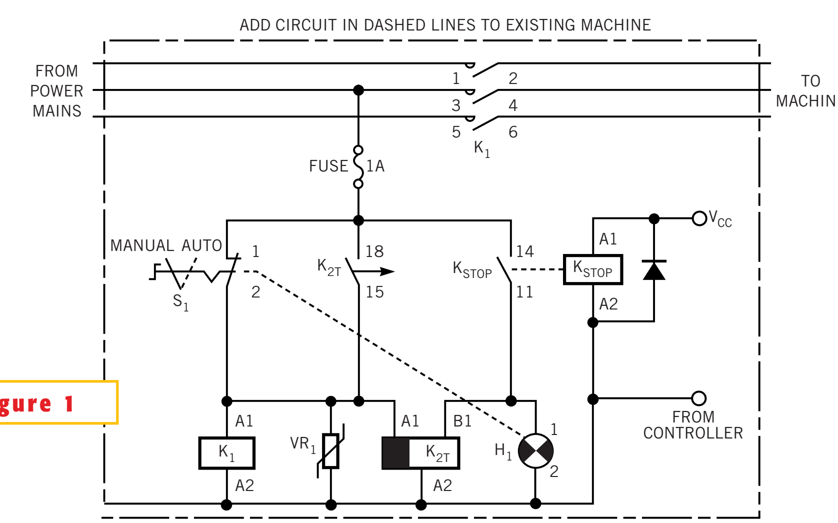
Jean-Bernard Guiot, Mulhouse, France

Some machines need to run for long periods and therefore may finish their work in the middle of the night or during the weekend. For the time remaining, until the operator returns, the machines stay idle, uselessly consuming power. This Design Idea allows a machine to completely shut itself down after finishing its work. In addition, the method allows for informing the machine operator by phone. You insert the circuit into the area that **Figure 1** indicates as a dashed line into the main supply line of the machine. The relay, K_{STOP}, connects to a free output of the programmable controller of the machine. You must program the controller in such a way that relay K_{STOP} is energized as long as the process is running. In normal operation, switch S_1 stays in manual position; thus, the power contactor, K_1, is on, and the machine receives power. When an operator starts the process, relay K_{STOP} energizes, and the indicator, H_1, lights, signaling the operator that switch S_1 is ready for operation. The timer relay, K_{2T}, is also on, closing its contact 18-15. Switching S_1 to automatic now has no effect.

At the end of the process, relay K_{STOP} and indicator H_1 turn off. Because contact 18-15 of relay K_{2T} incurs a delay before opening, because K_{2T} is a time-delay relay, the machine stays on during the delay time. This delay allows a second contact of K_{STOP} to control an automatic telephone dialer (not shown) to inform the remotely located operator and allows the process to finish supplementary tasks, such as cooling down, removing chips, allowing coolant to flow back into the tanks, for example.

Once the delay time expires, the contacts of K_{2T} open, K_1 turns off, and the machine completely turns off. The varistors, V_{R1}, suppresses voltage spikes. You must select V_{R1}, K_1, K_{2T}, and H_1 in accordance with the power-mains voltage and the power rating of the machine. You select K_{STOP} according to the controller's output (the relay coil) and the power-mains voltage (the relay contacts). The circuit has worked satisfactorily in hundreds of machines over a five-year period.□

Figure 1



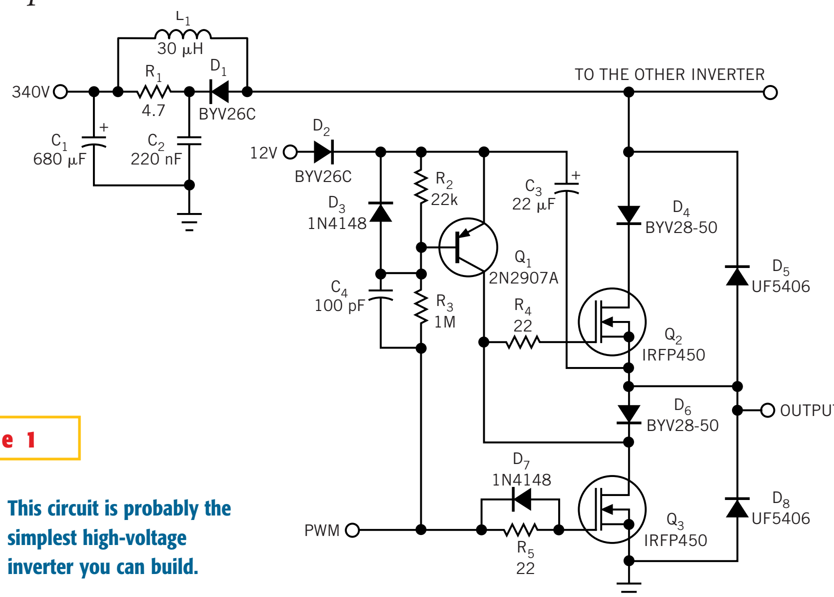
This circuit allows a machine to completely shut itself off after doing its assigned task.

Circuit makes simple high-voltage inverter

Francesc Casanellas, Aiguafreda, Spain

A SIMPLE HIGH-VOLTAGE MOSFET inverter solves the problem of driving a high-side MOSFET, using a low-voltage transistor, Q_1, and a special arrangement involving D_6 (**Figure 1**). This inverter is much faster than those that optocouplers drive, so dead-time problems are minimal. The inverter has the usual blocking diodes D_4 and D_6, and the parallel diodes D_5 and D_8. Q_3 provides the turn-off signal to Q_2. When Q_3 turns on, Q_2's gate short-circuits to ground through R_4. R_4 limits current and dampens oscillations. Q_2's gate discharges quickly; only the value of R_4 limits discharge time. Q_1 stays off, thanks to R_2, and C_3 charges to 12V through D_2. The gate pulse creates a current through C_4, and D_3 protects the

Figure 1



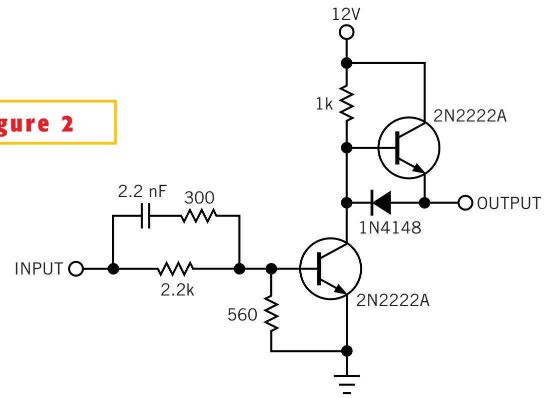
This circuit is probably the simplest high-voltage inverter you can build.

base-emitter junction of Q_1.

In the turn-on of Q_2, the following scenario occurs: When the control input, PWM, goes low, Q_3 quickly turns off, thanks to D_7. A displacement current, $C_4 \times dV/dt$, flows through C_4 to the base of Q_1. Q_1 charges the output capacitance of Q_3 and the gate capacitance of Q_2, and Q_2 turns on. C_3 supplies the collector current. If the period is long, Q_1 keeps conducting and compensating the leakage of Q_3. If D_6 were a Schottky diode, which is leaky, you would have to reduce the value of R_1. A short cross-conduction period exists between the two MOSFETs, a phenomenon that is more apparent when Q_3 turns off and Q_2 turns on. A small inductor, L_1, in series with the main supply limits the current spikes. The inductor needs a snubber comprising D_1, R_1, and C_2. Note that the inductor value is conservative and can be smaller.

The values are for a 370W, three-phase inverter with 150% overload capacity. If you change the MOSFET, the value of C_4 has to change according to the total gate charge plus the output capacitance of Q_3, which is much lower and, in fact, negligible. Q_1 amplifies the capacitor current, so C_4 is proportional to $Q_{G2} \times h_{FE1}$. Make C_4's value no higher than necessary, because the base current in Q_1 would be too high. To obtain all the speed advantages of the circuit, the PWM signal should be able to quickly drive Q_3. If necessary, you can use a buffer circuit (**Figure 2**). You can drive the circuit with a single CMOS gate. The circuit in **Figure 1** is probably the simplest high-voltage inverter you can design. It has served in thousands of three-phase motor drives from 0.37 to 0.75 kW.□

Figure 2



This buffer enhances speed at the PWM input of Figure 1's circuit.