Edited by Brad Thompson

Pushbuttons and digital potentiometer control boost converter

Simon Bramble, Maxim Integrated Products Inc, Wokingham, UK

Digitally controlled potentiometers are useful for generating analog control voltages under the control of a microcontroller. In some applications, manual pushbutton switches could replace a microcontroller and simplify product design. Mechanical switches exhibit contact bounce, and, when a user actuates them, they may open and close many times before reaching a stable state.

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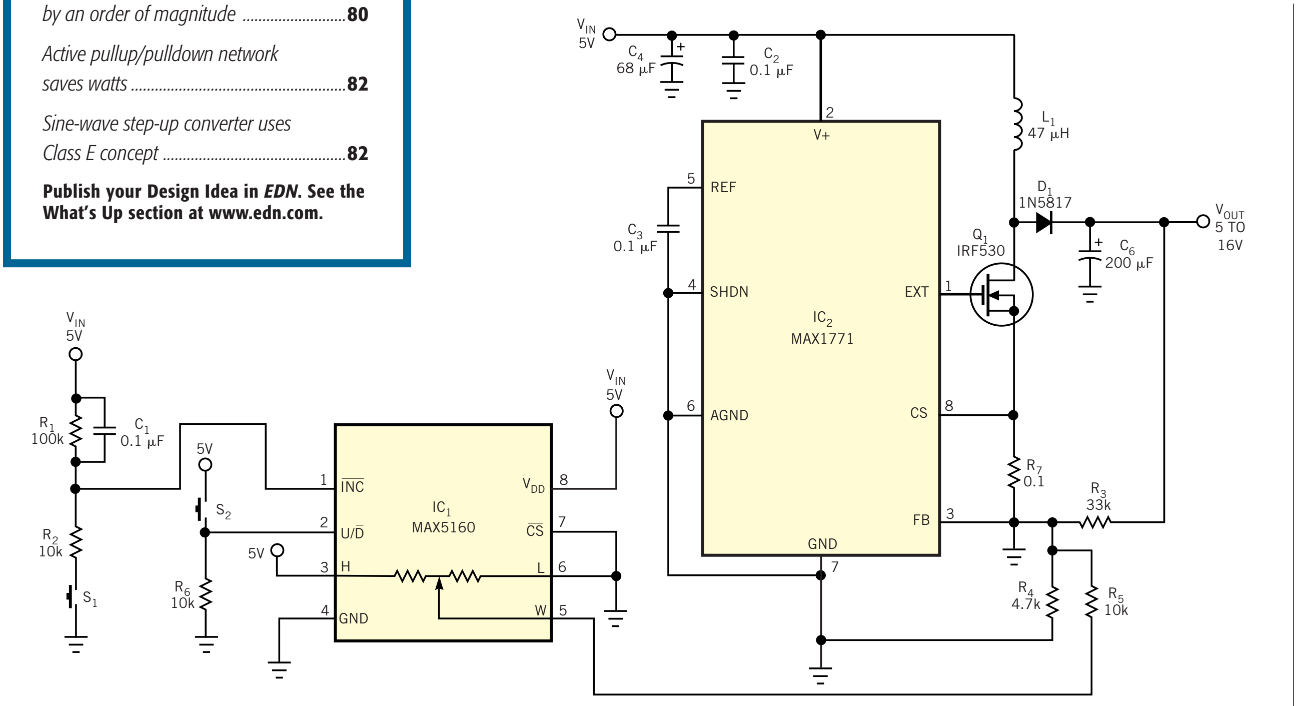
A digital potentiometer's control inputs lack switch-debouncing capabilities, and its up/down control is not suited for pushbutton operation. **Figure 1** illustrates solutions to these issues and shows how to use a digital potentiometer to control a boost converter.

The potentiometer, IC_1, a MAX5160M, presents an end-to-end resistance of 100 kΩ. To increment the wiper's position, W, you press and hold the $U/\overline{D}$ pushbutton, S_2, to pull the $U/\overline{D}$ pin high and then press and release pushbutton S_1 to pulse the $\overline{INC}$ input. Similarly, you decrement the wiper position by releasing S_2 and pulsing S_1.

A time-delay network comprising R_1, R_2, and C_1 masks S_1's switch bounce, which would otherwise toggle the wiper's position between V_{DD} and approximately 0V. When you press S_1, capacitor C_1 charges via R_2 and causes the $\overline{INC}$ pin to ramp slowly toward 0V, thereby removing the effects of S_1's contact bounce. The R_1C_1 time constant requires that you depress S_1 for several milliseconds before the $\overline{INC}$ input takes effect.

In this application, switching converter IC_2, a MAX1771, operates as a standard boost converter and increases its 5V input to a higher voltage positive output. You can use **Equation 1** to set IC_2's out-



A digital potentiometer, IC_1, and two pushbutton switches, S_1 and S_2, let you adjust the regulated output voltage of boost converter IC_2 over a 10V range.

Figure 1

put to a nominal 12V output without the digital potentiometer:

$$V_{OUT}=\frac{1.5\times(R_3+R_4)}{R_4}. \quad (1)$$

Connecting IC_1's wiper via 10-kΩ resistor R_5 to IC_2's FB (feedback) node sets IC_2's voltage-feedback level. Although inclusion of feedback resistors R_3, R_4, and R_5 and digital potentiometer IC_1 complicates the precise calculation of IC_2's output voltage, you can simplify the math by calculating the output voltages at the potentiometer's extreme settings. Thus, with IC_1's wiper set to 0V, R′ equals the paralleled resistance of R_4 and R_5, and IC_2's maximum output voltage becomes **Equation 2**:

$$V_{MAX}=\frac{1.5\times(R_3+R')}{R'}, \quad (2)$$

or $V_{MAX}=16.84V$.

With IC_1's wiper set to 5V, you can attempt to calculate the minimum output voltage by summing voltages into the feedback node:

$$V_{MIN}=\left[\left\{\frac{V_{FB}}{R_4}-\left(\frac{5-V_{FB}}{R_5}\right)\right\}\times R_3\right]+V_{FB}. \quad (3)$$

which simplifies to $V_{MIN}=0.48V$. However, **Equation 3** provides an incorrect value for V_{MIN} because a boost converter's output voltage cannot go below its input voltage. You can approximate V_{MIN} by substituting a value of 10 kΩ for R_5 in **Equation 3** and solving for V_{MIN}: $V_{MIN}=4.93V$. Refer to the manufacturer's application notes for additional component information.□

Microcontroller protects dc motor

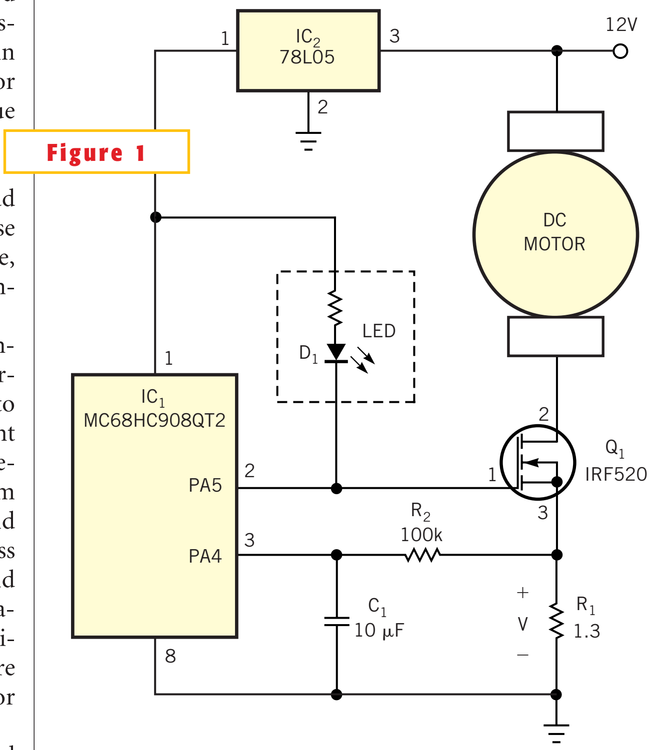
Abel Raynus, Armatron International Inc, Malden, MA

Although any overloaded dc motor can draw excessive current and sustain damage, a cooling fan's motor is particularly vulnerable due to fouling by dust, insects, or misplaced objects. A few fans include built-in overload protection, and others can use an external warning device, such as Microchip's TC670 fan-failure detector.

In many products, it's essential not only to detect an overloaded motor, but also to switch off the motor to prevent failure. Although you can design a protection system around the TC670, a low-end microcontroller can offer a less expensive, more flexible, and easier to implement alternative. If a product includes a microcontroller, only two spare pins are necessary for motor protection.

Figure 1 shows a dedicated protection circuit based on a small microcontroller and a power FET. This project uses an eight-pin flash-memory MC68HC908QT2 from Freescale and an IRF520A FET from Fairchild to control a dc brushless-fan motor rated for 0.72A at 12V dc. A high or low output voltage on output PA5 of IC_1 controls Q_1, an N-channel FET that in turn controls the motor. Current through the motor develops a voltage, V, that's proportional to the motor current across sense resistor R_1. A lowpass filter comprising R_2 and C_1 reduces noise on the sense voltage you apply to input PA4 and IC_1's built-in A/D converter. Voltage regulator IC_2 provides stable 5V power for IC_1.

Figure 1



NOTE: D_1, A LITE-ON LTL-4223-R2, INCLUDES A BUILT-IN RESISTOR.

Controller for dc brushless fan motor features minimal parts count.

Under normal operation, the voltage across R_1 measures approximately 0.52V. When the motor undergoes an overload, voltage increases until it reaches a preset upper limit of 0.85V. The output on PA5 then drops to a low level, switching off transistor Q_1 to stop the motor, and lighting D_1 to indicate the overload. When the motor stops, it draws no current, and the sense voltage falls below the minimum threshold value of 0.3V. The microcontroller's output on PA5 remains low and holds off Q_1, a state that it maintains indefinitely until you cycle the circuit's power off and then on.

The control program, in assembly language for the MC68HC908, features a straightforward algorithm adaptable to other microcontrollers that include an A/D converter. A 2.5-sec delay routine prevents the motor from starting until the system's power-supply voltage stabilizes. You can download the listing from the online version of this Design Idea at www.edn.com.□

Improved Kelvin contacts boost current-sensing accuracy by an order of magnitude

Craig Varga, National Semiconductor Corp, Phoenix, AZ

Many power-supply designs rely on accurately sensing the voltage across a current-sense element. Multiphase regulators use the sense voltage to force current sharing among phases, and single-phase regulators to control the current-limit setpoint. As internal complexity and clock speeds increase, processors impose narrower operating margins for power-supply voltages and currents, which in turn make accurate current sensing critically important. The most accurate of several available methods involves inserting a low-value current-sensing resistor in the power supply's output path. Another popular technique uses the parasitic resistance of a switching regulator's output inductor as the sense element. For either method, currents of 20A or more per power-supply phase impose a sense-resistance limit of approximately 1 mΩ. Precision resistors of 1% accuracy are available at reasonable cost, but an error of 1% of 1 mΩ amounts to only 10 μΩ.

The resistance of solder joints that attach a sense resistor or inductor can easily exceed 10 μΩ and, worse yet, can vary significantly during a production run. In the past, discrete four-wire resistors provided separate high-current and sense-voltage connections, allowing accurate Kelvin sensing and excluding voltage drops that the high-current connections introduce. Unfortunately, four-wire sense resistors or inductors are unavailable in low cost SMD packages. Thus, most power-supply designers use two-wire sense components and apply a Kelvin-connection pc-board-layout technique (**Figure 1**). However, test results reveal that applying conventional Kelvin sensing techniques to low-value resistors introduces transduction errors as high as 25%—an unacceptable error margin for designs that require high accuracy.

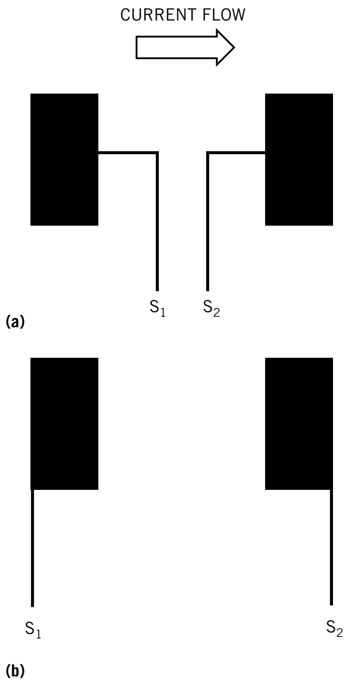


Figure 1 Conventional solder pads offer two choices for Kelvin connections: on the pads' inner edges (a) or outer corners (b).

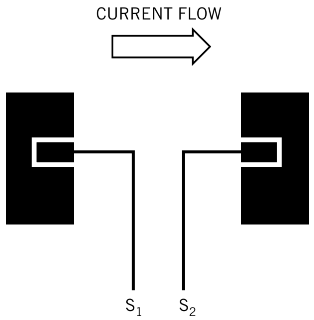


Figure 2 Modified pads provide isolated Kelvin connections that eliminate errors introduced by voltage drops across soldered connections carrying high current.

So, what's a power-supply designer to do? The answer involves a slight variation on an old idea that requires only a minor change in a sense resistor's mounting footprint. To compare performance of conventional Kelvin connections versus the proposed method, a test board includes three pc-layout footprints for installation of 1-mΩ, 1%-accurate, surface-mount resistors. In all three patterns, current enters and exits the resistor via traces (not shown) on the pads' left and right sides, respectively.

In **Figure 1a**, applying a current of 4.004A produces a sense voltage at the Kelvin terminals of 4.058 mV, a 1.35% error. At 8.002A, the sense voltage at the Kelvin terminals measures 8.090 mV, a 1.1% error. In **Figure 1b**, a current of 4.004A produces a sense voltage at the Kelvin terminals of 5.01 mV, a 25% error. At 8.002A, the sense voltage at the Kelvin terminals measures 9.462 mV for an error of 18.2%. **Figure 2** shows an improved component footprint. Each large solder pad includes a central cutout area that partially surrounds a narrow pad that solders directly to the sense element and thus carries no current. This approach removes from the sense path the large-area solder joints that mount the part and carry high load current.

When you apply a current of 4.002A to the pads in **Figure 2**, voltage at the Kelvin terminals measures 4.004 mV, a 0.05% error. At 8.003A, the sense voltage measures 8.012 mV, an error of only 0.11% and an order of magnitude improvement over **Figure 1a**. Sense-voltage variation over temperature should greatly improve, and solder-thickness variation no longer affects the sense voltage. Best of all, the technique costs nothing to implement.

Obviously, the technique in **Figure 2** works only with terminations sufficiently wide to allow dividing the solder pad into three sections and still retain adequate soldering area to handle the high-current connections. However, for many designs, this simple technique can significantly improve the accuracy of current sharing, V-I load-line characterization, and current-limit setpoints.□

Active pullup/pulldown network saves watts

JB Guiot, DCS Dental AG, Allschwil, Switzerland

The control circuit in **Figure 1** presents a relatively low input resistance and thus imposes a low value on external pulldown resistor R_1 to provide the desired low-level input voltage. In turn, R_1 wastes power by drawing a relatively high current through switch S_1. For example, suppose that the control circuit presents an input resistance of 2.2 kΩ and requires a logic-low input of 5V or less. At a V_{CC} of 24V, R_1 must not exceed 500Ω for a current drain of 24/500=48 mA. The power dissipated in R_1 is thus $24^2/0.5=1152$ mW, which requires a 2W resistor for reliable operation.

In this application, the system includes three controls with three input circuits, which present a total current of 432 mA and adds approximately 10W to the power budget. To reduce wasted power, it uses an active pulldown circuit (inside the dashed line in **Figure 2**). As long as switch S_1 remains closed, PNP transistor Q_1's base voltage exceeds its emitter voltage due to diode D_1's forward voltage drop. Thus, Q_1 doesn't conduct, and the control circuit's input voltage rests at $V_{CC}-0.7V$ (D_1's forward voltage drop).

Opening S_1 reverse-biases D_1, and the base current flowing through resistor R_1 turns on Q_1, which saturates and pulls the circuit's input to $V_{CE(SAT)}$. Closed-circuit current through S_1 is thus V_{CC}/R_1. For example, with V_{CC} of 24V and R_1 having a value of 10 kΩ, I=24/10=2.4 mA, and R_1 dissipates 0.058W, or approximately 20 times less power than in **Figure 1**. In this application, current demand of the nine control-circuit inputs decreases from 432 to 22 mA and saves pc-board space by eliminating the need for using 2W. As a variant, **Figure 3** shows an active-pullup version of the circuit.□

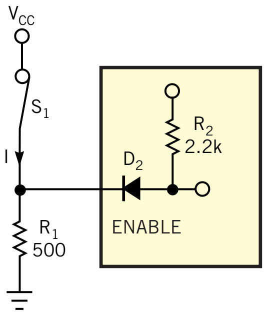


Figure 1 A conventional network requires a low-value pulldown resistor.

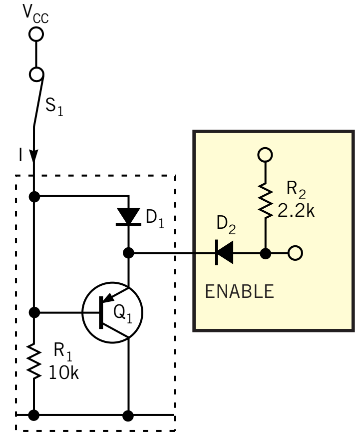


Figure 2 An active pulldown circuit substitutes a saturated transistor for a power-consuming pulldown resistor.

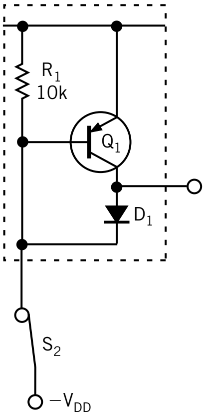


Figure 3 Rearranging the circuit and using an NPN transistor for Q_1 yields an active-pullup network.

Sine-wave step-up converter uses Class E concept

Louis Vlemincq, Belgacom, Evere, Belgium

Many power applications ranging from luminescent and fluorescent lighting to telephone-ringing voltage generators require a more or less sinusoidal-drive voltage. These applications typically require a waveform of only moderate quality, and its frequency isn't especially critical. However, avoiding waveform discontinuities that cause unwanted current peaks, excessive device dissipation, and EMC problems rules out using filtered square waves or other stepped waveforms. Sometimes, a trapezoidal drive may be acceptable but it's only a second choice at best. This Design Idea proposes a method of generating sine waves that offers a number of advantages over more complex methods:

The circuit requires only one power-switching device, and you can use an analog or a digital signal to drive the switching device. The circuit also requires only a few components: a diode, a switching transistor or a MOSFET, an inductor or a transformer, and a capacitor. Further, the design's circuit losses are low, and the switching device experiences minimal stress during operation. **Figure 1** shows the basic circuit, and **Figure 2** illustrates

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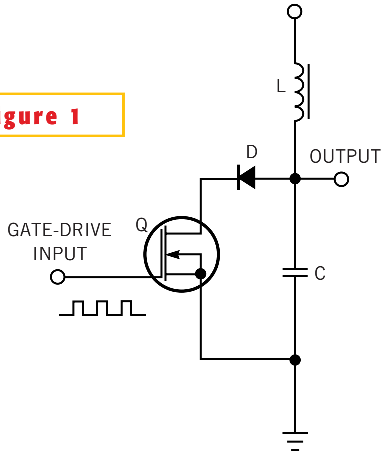
waveforms within the circuit.

In operation, the sine-wave output appears across inductor L in a series LC circuit. An external clock source produces gate drive for transistor Q at a frequency that's lower than the LC circuit's natural resonant frequency. When the transistor conducts, the inductor receives a charging current via diode D. When conduction ends, energy stored in the inductor transfers to capacitor C, and a damped oscillation begins (uppermost trace in **Figure 2**). The voltage across capacitor C approximates a sine wave (top trace). Drain current in transistor Q shows that no current flows until the capacitor voltage forward-biases diode D (middle trace). The gate-drive pulse interval is lower than the series-resonant frequency that L and C present (bottom trace).

During the negative-going portion of the cycle, the external clock source applies gate drive to the transistor. Diode D is still reverse-biased, and thus no current flows into Q. When the voltage across C goes positive, diode D conducts and allows current to recharge the inductor and replenish energy lost in the previous cycle. In balanced operation, energy supplied during the conduction phase replenishes energy supplied to the load and dissipated in component losses.

Figure 1



Few components are necessary to produce a pseudo sine wave.

To produce a higher peak voltage, you extend the conduction interval by raising the drive frequency or by extending the on interval. You can regulate the output voltage by applying conventional closed-loop feedback techniques to a variable-frequency clock oscillator or, in digital systems, by altering the clock's duration. For most applications in which load current is relatively fixed, such as in an electroluminescent panel lamp, an open-loop adjustment or manual control offers sufficient flexibility once you determine a clock-frequency range that corresponds to the desired degree of illumination.

For peak voltages not exceeding 10 times the power-supply voltage, you can connect the load directly to the junction of D, L, and C. You can achieve higher voltage-to-step-up ratios at the expense of applying additional voltage and current stress to L and C. Instead, you can add an isolated secondary step-up winding to the inductor. For optimum efficiency, use components designed for high-frequency power handling—for example, a polypropylene-dielectric capacitor and a low-loss inductor.

If the load consists of an electroluminescent panel that behaves as a lossy capacitor, you may be able to eliminate the use of external capacitor C. Transistor Q must obviously be able to withstand peak voltages and currents that the circuit imposes, but its specifications are otherwise relatively noncritical. No switching loss occurs at the beginning of the conduction due to the Class E mode of operation, and the output capacitor assists the device's turn-off recovery. For output voltages not exceeding approximately 50V, you can improve efficiency by selecting a Schottky or other fast-recovery diode for use in the circuit.

For powering lamps and generating telephone-ringing signals, the sine wave's "flat spots" are of little consequence because of their relatively short duration and low harmonic-energy content. You can minimize these intervals by reducing the LC ratio and thus increasing the loaded Q factor of the LC circuit. However, for a given output voltage, increasing Q factor also increases the peak current because the same amount of energy must transfer to the inductor in less time.□

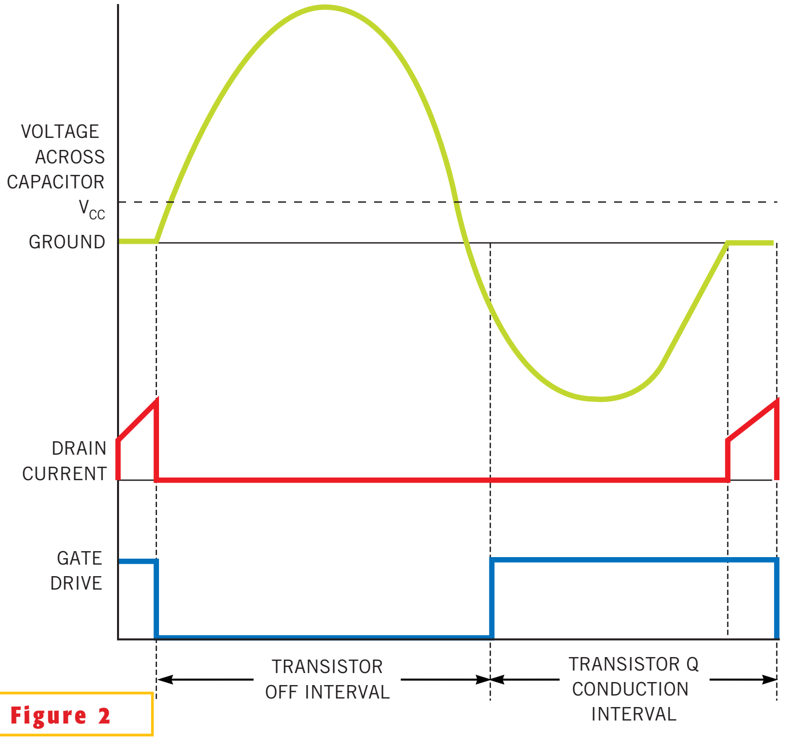


Figure 2

The circuit in Figure 1 produces these waveforms.