Edited by Brad Thompson

Interrupt-driven keyboard for MCS-51 uses few components

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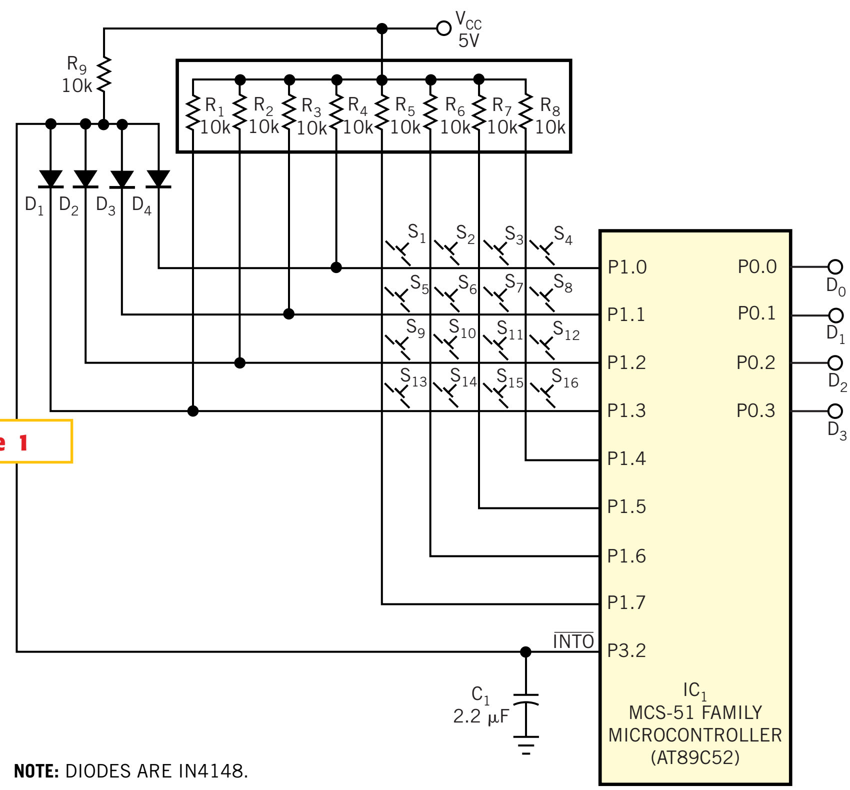
Designers of microcontroller-based products that require a keypad for user data entry can select from dedicating an input line for each key, continuously polling the keypad's x and y lines, or generating an interrupt whenever a user presses a key. Although conceptually simple, dedicating lines to a keypad can tie up most of the microcontroller's I/O resources. Continuously polled keypads can burden the microprocessor's resources and consume excessive amounts of battery power.

The third method, an interrupt-driven keypad, offers several benefits. First, using interrupts frees the microcontroller to perform other tasks or to switch into an idling or power-down mode while awaiting the next key closure. Second, using interrupts helps reduce electromagnetic interference produced by continuously scanning the keypad's lines. **Figure 1** shows an interrupt-driven keypad implementation that's based on Atmel's AT89C52 version of the popular MCS-51 family of microcontrollers. Here, the rows of a 16-key keypad, S_1 through S_{16}, implemented as a 4×4-key matrix connect to the lower nibble (P1.0 to P1.3) of IC_1's Port 1. The keypad's columns connect to IC_1's Port 1 upper nibble (P1.4 to P1.7) and a network of four diodes (D_1 through D_4) and a 10-kΩ resistor, R_9. The junction of R_9 and the diodes' anodes connects to Port Pin 3.2 and generates an interrupt whenever the user presses a key.

Initially, Port 1's lower nibble sits high at logic one, and the upper nibble is grounded at logic zero, applying reverse bias to the diodes and pulling the $\overline{\text{Into}}$ signal high. Pressing a key applies forward bias to the diode corresponding to that row and causes $\overline{\text{Into}}$ to go low, generating an external interrupt to the microcontroller. Upon receiving an interrupt and after a 20-msec software-debouncing interval, the microprocessor sequentially reads the row and column lines. Capacitor C_1 provides a hardware-based debouncing interval of approximately 25 msec.

In this design, the microcontroller's software returns a binary-formatted input corresponding to the pressed key's number as sensed at Port P1 (P1.0 to P1.3). As the commented assembly-language routine, available with the online version of this Design Idea at www.

Figure 1



Adding an interrupt-driven keypad to an MCS-51 family microcontroller requires only a few components.

edn.com, explains, the software ignores invalid key combinations. Idle and power-down modes available in CHMOS (complementary high-density MOS) versions of the MCS-51 family save power and thus make these microcontrollers ideal choices for battery-operated devices. For example, a 5V, 12-MHz Atmel AT89C52 consumes approximately 25 mA in active mode, 6.5 mA in idle mode, and only 100 μA in power-down mode. Any enabled interrupt can switch the microprocessor from idle to active modes.

However, recovery from power-down mode to active mode normally requires a hardware reset—an apparent limitation of the MCS-51 microcontroller. However, an earlier Design Idea overcomes the problem and allows use of an interrupt-driven keypad even in hardware-reset-based systems (**Reference 1**).□

Reference

1. Chrzaszcz, Jerzy, "Use 8051's power-down mode to the fullest," *EDN*, July 6, 2000, pg 138, www.edn.com/article/CA47001.html.

Simplified white-LED flasher operates from one cell

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The white-LED flasher in **Figure 1** offers an alternative approach to a previously published idea (**Reference 1**). Targeting use in portable appliances and products powered by a single cell, this circuit flashes an LED to provide a highly visible warning signal—for example, to indicate power on, battery low, or another eye-catching visual signal. However, white LEDs typically present a forward-voltage drop of 3 to 5V. Given that a single cell's end-of-life output voltage decreases to 1V, flashing a white LED from such a low voltage demands special circuit techniques. The circuit in **Figure 1** exploits the low-voltage capabilities of Fairchild's universally configurable two-input NC7SV57 logic gate. The NC7SV-57 operates from a supply voltage as low as 0.9V, a key requirement for single-cell applications.

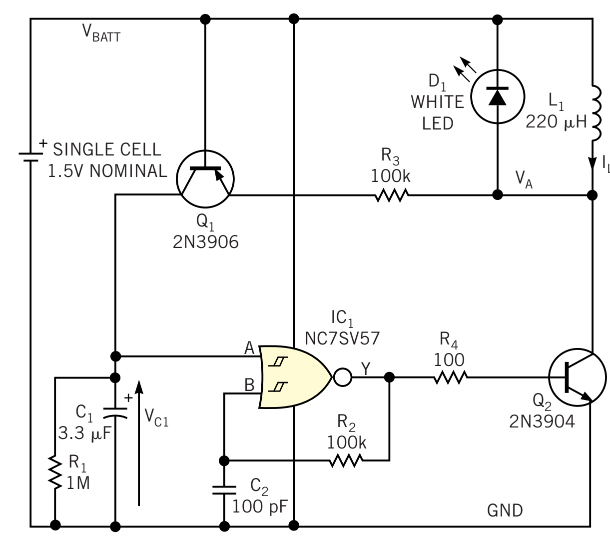
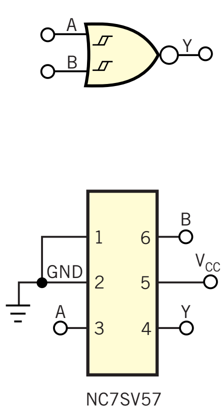


Figure 1 A few inexpensive components can light a white LED from a single primary cell.

Figure 2



Connecting pins 1 and 2 to ground configures the NC7SV57 as a two-input NOR gate.

Available in a six-lead SC70 package, this IC features Schmitt-trigger inputs. You can configure the device to implement AND, NAND, NOR, Exclusive-NOR, and invert logic functions. Connecting the pins as in **Figure 2** produces a two-input NOR gate. Upon initial power application, capacitor C_1 is uncharged, and its voltage, V_{C1}, is 0V. With C_1 holding Input A low, IC_1 temporarily becomes an inverter and forms a simple astable relaxation oscillator at a frequency that C_2 and R_2 determine. The square-wave signal at IC_1's output drives switching transistor Q_2.

When IC_1's output pulses high, Q_2 turns on and saturates, sinking current, I_L, through inductor L_1. The inductor current ramps up at a rate determined mainly by V_{BATT}; L_1's inductance; and Q_2's on-time, t_{ON}. During this interval, LED D_1 and Q_1's base-emitter junction are reverse-biased. Provided that the inductor does not saturate, current I_L ramps up linearly and reaches a peak value, $I_{L(PEAK)}$ at the end of the on-time.

When IC_1's output goes low and Q_2 turns off, L_1 generates back EMF to forward-bias D_1 and raises its anode voltage, V_A, to a value greater than V_{BATT}. Current circulates through L_1 and D_1 and ramps down to zero as L_1's stored energy decays. Values of 100 pF for C_2 and 220 kΩ for R_2 set the astable oscillator's frequency at approximately 20 to 30 kHz. On each cycle, a current pulse with a peak value, $I_{L(PEAK)}$, flows through the LED. Due to the high repetition frequency and persistence of vision, the LED appears to be continually on.

Without Q_1 and R_3, the astable circuit would oscillate without interruption and continuously illuminate the LED. However, Q_1 and R_3 provide a charging path for C_1. On any cycle of the astable square wave during which V_A rises above V_{BATT}, Q_1's base-emitter junction becomes forward-biased via R_3, allowing a pulse of current to flow through Q_1 into C_1. The magnitude of this current pulse depends largely on D_1's forward-voltage drop and R_3's resistance. Each current pulse applied to C_1 slightly increases the capacitor's charged voltage. When the charged voltage eventually exceeds IC_1's upper input-threshold voltage, V_{TU}, at Input A, the astable oscillator shuts off, and IC_1's output goes low. In response, Q_2 and the LED turn off, and current pulses cease to flow through Q_1.

Next, C_1 discharges through R_1, and

V_{C1} decays at a rate that only the R_1-C_1 time constant determines. D_1 remains off until V_{C1} falls below the Schmitt trigger's lower threshold voltage, V_{TL}; astable oscillations recommence; and the LED again turns on and repeats the illumination cycle.

For the values of C_1 and R_1 in **Figure 1**, the LED's on-time is roughly proportional to R_3's resistance. A relatively small value produces a short "blink," whereas a higher resistance leads to an on-time of several seconds. The LED's off-time depends only on the values of C_1 and R_1. To ensure that V_{C1} can exceed IC_1's upper input-threshold voltage, R_1's resistance must exceed R_3's resistance.

You can use a small-signal pnp transistor with good current gain for Q_1, because its specifications aren't critical. However, to ensure that most of the battery voltage appears across L_1 when Q_2 conducts, use a switching transistor with low collector-emitter saturation voltage for Q_2. If the saturation voltage is low enough to neglect, you can calculate the peak current in L_1 using $V_{BATT} \times t_{ON}/L_1$. The LED's intensity is proportional to its average forward current and thus determined in part by $I_{L(PEAK)}$. For optimum LED brightness, select the on-time and L_1 to maximize $I_{L(PEAK)}$ but not to exceed D_1's and L_1's maximum current ratings. The actual value of L_1 isn't critical, but values of 100 to 330 μH provide good performance and reasonable efficiency.

For large values of C_1, R_1, and R_3, the LED flashes at a fairly slow rate. With values of 3.3 μF for C_1, 1 MΩ for R_1, and 100 kΩ for R_3, the circuit produces a flash rate of approximately 0.4 Hz with V_{BATT} of 1.6V. Reducing V_{BATT} to 0.8V results in little flash-rate variation. At V_{BATT} of 1.6V, the LED delivers brightness that remains even when battery voltage decreases to 0.8V. The circuit continues to operate when V_{BATT} decreases to 0.65V, although the LED dims considerably at that level.

The NC7SV57's guaranteed operating voltage ranges from 0.9 to 3.6V, which allows operation from one or two alkaline or rechargeable nickel-cadmium cells or from a single 3V lithium cell. Texas Instruments' SN74LVC1G57 offers the same logic functions but operates over a slightly higher supply voltage range of 1.65 to 5.5V. To eliminate flashing operation, simply omit C_1, R_1, R_3, and Q_1. To turn the LED on or off, you can apply a gating signal to IC_1's Input A.□

Reference

1. Smith, Anthony, "Single cell flashes white LED," *EDN*, Dec 11, 2003, pg 84, www.edn.com/article/CA339716.html.

Low power voltage-to-frequency converter makes a wireless probe for testing an inductive power supply

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Powering portable telemetry systems for long-term monitoring presents interesting design challenges. Batteries are unsuitable for certain critical applications, and, in these circumstances, designers typically use wireless inductive links to transmit both power and data. An inductive link comprises an RF transmitter that drives a fixed primary coil, and a loosely coupled secondary coil that supplies power to the portable circuitry. For design engineers, measuring transmitted power takes on considerable importance because it imposes limits on the amount of circuitry that designers can include in the portable circuitry. Unfortunately, classical test equipment is poorly suited to the task. Standard voltage probes pick up noise that the primary coil induces, and, in some applications, the portable circuits are hermetically encapsulated in small enclosures that prevent entry of a cable or a probe.

The circuit in **Figure 1** reduces noise effects because its VFC (voltage-to-frequency converter) produces a PPM (pulse-position-modulated) output signal, V_{OUT}, that integrates, or averages, noise. In addition, the design uses "load modulation" to eliminate wired connections. When the PPM signal drives on MOSFET switch Q_1, the switch connects an additional loading network comprising D_2 and the series combination of R_{SF} and R_{SV} across the secondary coil, L_S. A load-modulation receiver connects to the primary coil and recovers the PPM signal. When you build it with surface-mounted components, the VFC circuit occupies a board area of only 238 mm^2.

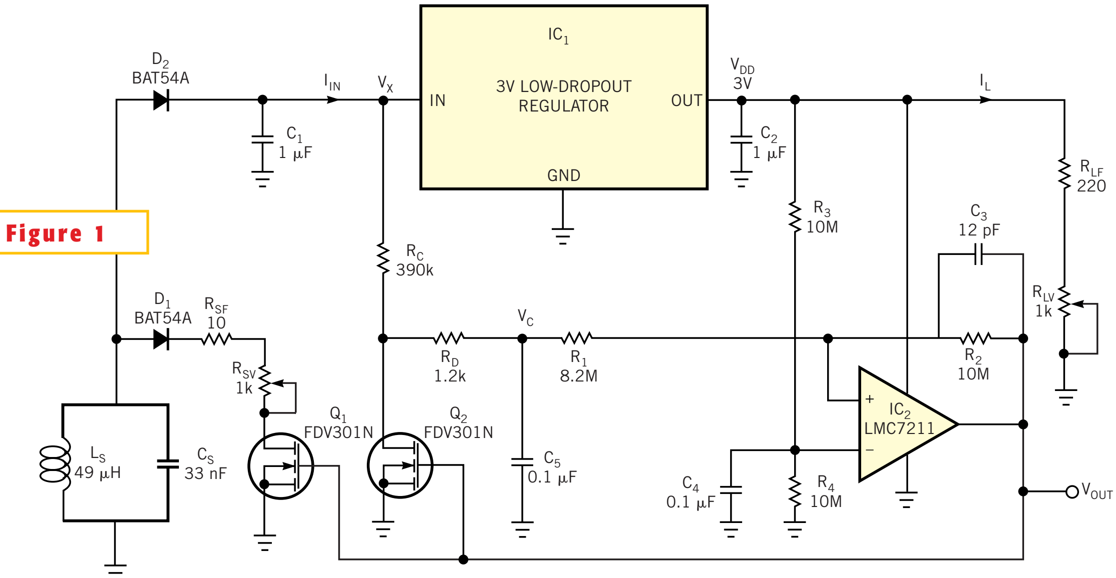
To understand the circuit's operation, assume that a 125-kHz sinusoidal magnetic field induces approximately 4 to 16V in secondary coil L_S. To improve power-transfer efficiency, L_S and C_S form a tuned, 125-kHz tank circuit having a loaded Q factor, Q_L, of approximately 8. Schottky diode D_1 rectifies the voltage induced in L_S, and C_1 provides lowpass filtering. The resultant dc voltage, V_X, powers low-dropout regulator IC_1, which supplies a constant 3V to VFC IC_2 and the load resistors, R_{LF} and R_{LV}. Trimmer potentiometer R_{LV} sets the output current at 2.5 to 13.5 mA.

The combined total current drain of the low-dropout regulator and the VFC measures a few tens of microamperes and is negligible compared with the output current. Hence, I_{IN} approximately equals I_L. **Equation 1** expresses the dc output power that the inductive power supply produces:

$$P_X \approx V_X \times I_L = V_X \frac{3V}{R_{LF} + R_{LV}}. \qquad (1)$$

This **equation** shows that the output current is constant and therefore the dc output power, P_X, is proportional to the dc voltage, V_X. After setting a known initial output current adjustment via R_{LV}, you can test the inductive power supply's

Figure 1



A low-power VFC and load modulator measure power generated by a wireless-telemetry power source.

output ability by measuring the transmitted dc voltage that the VFC digitizes. To minimize power consumption, component count, and pc-board area, a simple passive integrating network comprising R_C, R_D, and C_5 replaces the classical op-amp integrator that constitutes a typical VFC's input stage.

The VFC generates a constant-amplitude sawtooth voltage whose rising slope is proportional to V_X across integrating capacitor C_5. When the capacitor's voltage reaches a high reference voltage, switch Q_2 rapidly discharges the capacitor to a low reference voltage. This action produces a free-running waveform whose frequency is proportional to the input voltage, V_X. A noninverting Schmitt trigger comprising comparator IC_2; its positive-feedback network, R_1, R_2, and C_3; and supply-voltage splitter R_3, R_4, and C_4 defines the high- and low-level reference voltages, as **equations 2** and **3** calculate.

$$V_{TH} = \frac{V_{DD}}{2}\left(1+\frac{R_1}{R_2}\right), \quad (2)$$

$$V_{TL} = \frac{V_{DD}}{2}\left(1-\frac{R_1}{R_2}\right). \quad (3)$$

Equation 3 shows that, to reset the integrated voltage almost to 0V, the value of R_1 must be slightly lower than that of R_2. Using standard values of E12-series resistors and taking into account power-consumption constraints, select a value of 8.2 MΩ for R_1 and 10 MΩ for R_2. Replacing these values in **equations 2** and **3** yields, respectively:

$$V_{TH} = 2.73V \approx V_{DD};\ V_{TL} = 0.27V \approx 0V. \quad (4)$$

To understand the VFC's operation, assume that, at start-up, capacitor C_5 is fully discharged. Consequently, comparator IC_2's output, V_{OUT}, is low and MOSFET switches Q_1 and Q_2 are off. Under these conditions, current through R_C and R_D begins to charge C_5 toward V_X with a time constant of $\tau_C=(R_C+R_D)\times C_5$. When capacitor C_5's voltage reaches the Schmitt trigger's upper threshold voltage at time t_X, the comparator's output, V_{OUT}, rises to V_{DD} and turns on MOSFET switches Q_1 and Q_2. Switch Q_2

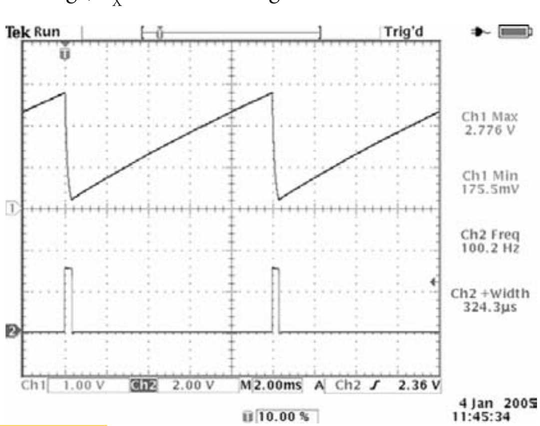
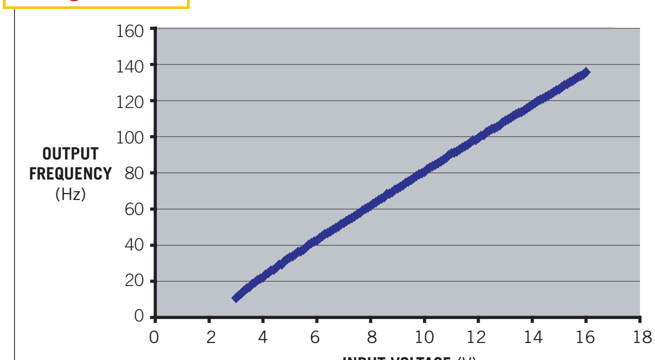


Figure 2 Oscilloscope measurements of the VFC show voltage across capacitor C_5 (upper trace) and comparator IC_2's output voltage, V_{OUT} (lower trace), for a nominal input voltage of 12V.

Figure 3



The measured transfer function of the VFC exhibits excellent linearity over a wide range of inductively coupled input voltages.

discharges C_5 through R_D at time constant of $\tau_D \simeq R_D \times C_5$. Simultaneously, Q_1 generates a load-modulation pulse.

When $V_C = V_{TL}$, the comparator's output drops to zero, restores the initial state, and repeats the sequence. As Trace 1 in **Figure 2** shows, the circuit behaves as a free-running oscillator in which the voltage across C_5 ramps up and down between the Schmitt trigger's threshold voltages. Given that the discharge-time constant, τ_D, is much less than the charging-time constant, τ_C, the discharge time, t_{ON}, is considerably shorter than the integrating time, t_X. As Trace 2 in **Figure 2** shows, the comparator's output delivers a PPM signal having a relatively short pulse of approximately 320 μsec.

Equations 5 and **6**, respectively, describe the complete expressions for calculating the pulse widths of waveforms t_X and t_{ON}:

$$t_X = (R_C + R_D) \times C_5 \times \log \left(\frac{V_X - \frac{V_{DD}}{2}\left(1 - \frac{R_1}{R_2}\right)}{V_X - \frac{V_{DD}}{2}\left(1 + \frac{R_1}{R_2}\right)} \right), \quad (5)$$

$$t_{ON} = R_D \times C_5 \times \log\left(\frac{R_2 + R_1}{R_2 - R_1}\right). \quad (6)$$

These formulas are useful for designing the VFC in **Figure 1** but yield little insight into the circuit's global-transfer function. You can apply the following approximations to simplify the calculations: Because $t_X >> t_{ON}$, the PPM output frequency is approximately $f_X \simeq 1/t_X$. In normal operation, V_X reaches relatively high values when compared with the Schmitt trigger's threshold voltages, and you can linearize the charging law of capacitor C_5 to a ramp having a constant slope (**Equation 7**).

$$\frac{dV_C}{dt} \approx \frac{V_X}{(R_C + R_D)C_5}. \quad (7)$$

According to **Equation 4**, the Schmitt trigger's high and low threshold voltages are, respectively, $V_{TH} \simeq V_{DD}$ and $V_{TL} \simeq 0V$. Using these approximations, the PPM output frequency simplifies to:

$$f_X \approx \frac{1}{(R_C + R_D) \times C_5} \times \frac{V_X}{V_{DD}}. \quad (8)$$

Equation 8 shows that the circuit in **Figure 1** exhibits a voltage-to-frequency transfer function, as **Figure 3** experimentally confirms. The VFC's power consumption is low; for example, at a dc voltage of 12V, the VFC's current drain is about 36 μA.□

Simple circuit converts 5V to −10V

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A typical switched-capacitor charge pump requires no inductors, is easy to design, and can double a positive voltage or convert a positive voltage to an equivalent negative voltage. However, in some applications, only a positive supply is available, and the power-supply system must generate a negative voltage of larger magnitude than the positive supply rail's voltage. The circuit in **Figure 1** simultaneously inverts its input voltage and doubles the resulting negative output.

Normally, the MAX889T voltage inverter, IC_1, converts a positive input to a negative output voltage with an absolute magnitude lower than that of its input. But, in this circuit, Schottky diodes D_1 and D_2 and capacitors C_4 and C_5 help produce a higher output voltage. The circuit's nominal output is $V_{OUT} = -(2 \times V_{IN} - 2V_D - I_{OUT} \times R_O)$, where V_{IN} is the input voltage, V_D is a diode's forward-voltage drop, I_{OUT} is the output current, and R_O is IC_1's output resistance in free-running mode. For a 300-μA load current, the circuit's output voltage is −10V. Parasitic inductances inherent in the capacitors and pc-board traces produce a voltage overshoot that charges the output capacitors, delivering more than −11V at no load (**Figure 2**).□

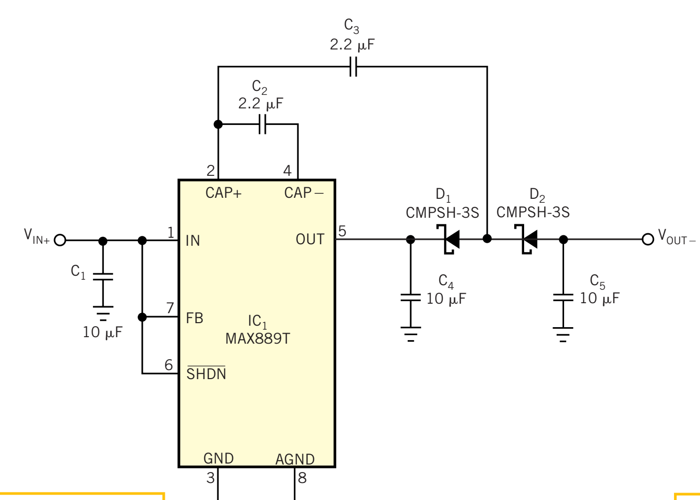

Figure 1

This switched-capacitor inverter derives −10V from 5V.

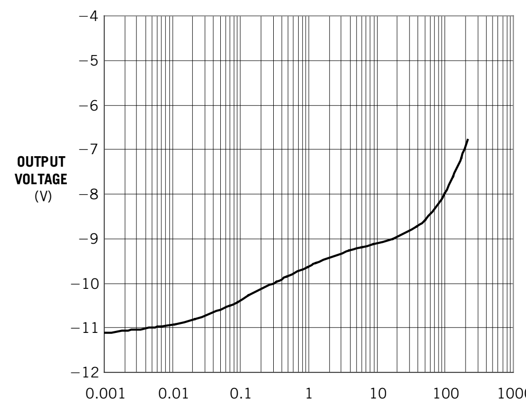

Figure 2

At light loads, the circuit in Figure 1 delivers more than −10V; at higher currents, the magnitude of the negative output voltage exceeds its positive input voltage of 5V.