

## Keyboard links to microcomputer through one-wire interface

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▶ A previous Design Idea described a single-wire-plus-ground-return, keypad-to-microcontroller interface in which a single pulse represents each keystroke (Reference 1). The pulse's width is proportional to the key's numerical value, and the microcontroller identifies the pressed key by measuring the pulse width. Component tolerances and the accuracy of the microcontroller's internal oscillator limit the original design to keypads with 16 or fewer keys—that is, four rows by four columns or smaller cross-point-key matrix. This Design Idea illustrates how a relatively simple mod-

ification applies the method to much larger keypads. (The following description omits a few details from the original Design Idea, which you can find online at [www.edn.com/article/CA512131](http://www.edn.com/article/CA512131).)

You can divide a large keypad or keyboard into sections of 12 keys each (Figure 1). Each section connects to a separate comparator circuit, which detects a keystroke and generates a trigger pulse using the monostable circuit of IC<sub>4</sub>, and you can add more sections in the same manner. Diodes D<sub>1</sub>, D<sub>2</sub>, and D<sub>3</sub> couple and isolate the comparators' outputs to Pin 2 of IC<sub>4</sub>. Each keypad

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section connects the same selection of timing-resistor values to the monostable. Therefore, each keypad section's output pulse widths occupy the same range: 110 to 1320  $\mu$ sec.

For the microcontroller to identify an active keypad section, the circuit generates a single, double, or triple pulse, depending on whether the pressed key

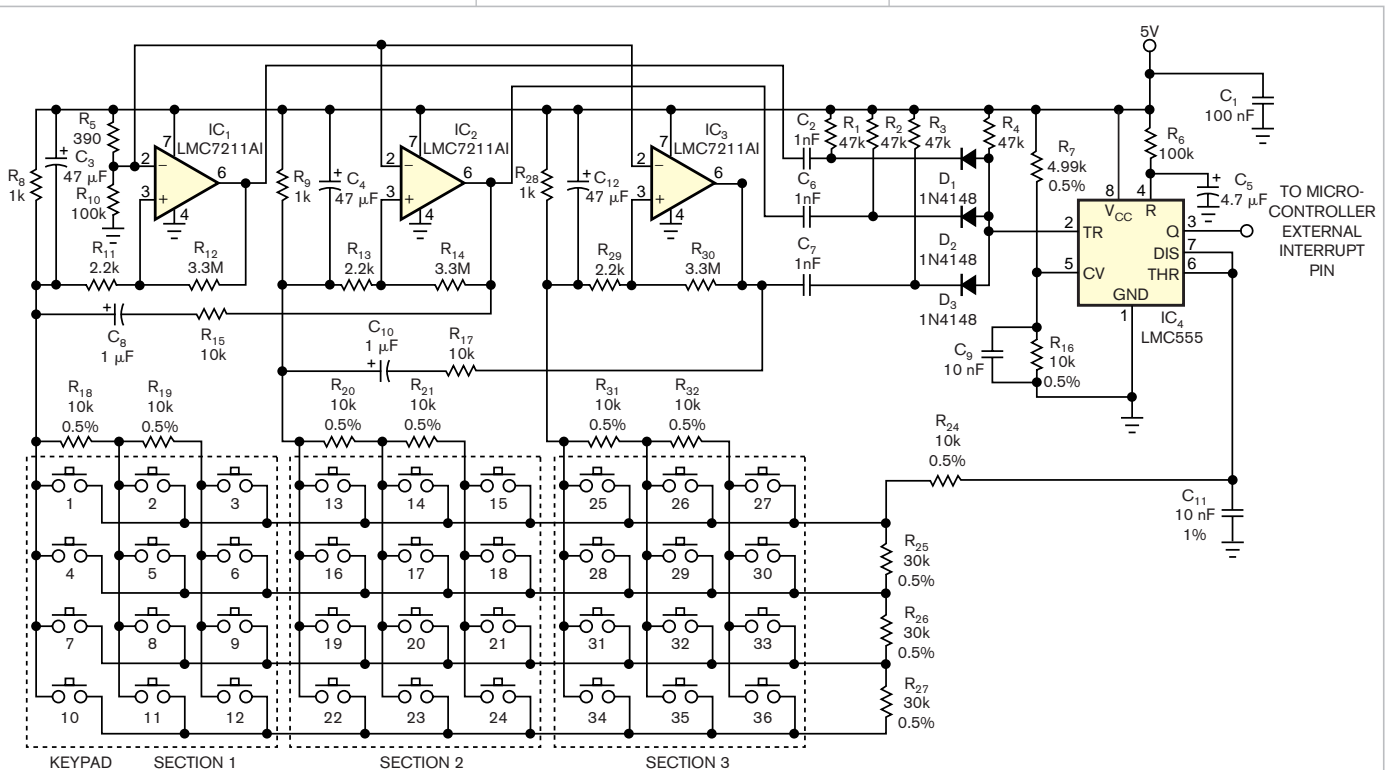


Figure 1 A keyboard encoder uses multiple variable-width pulses to communicate keys' addresses to a microcontroller using a single-wire interface.

resides in section 1, 2, or 3, respectively. The microcontroller identifies the pressed key by measuring the width of the first pulse and identifies the keypad section by counting the number of pulses. Implementing the multiple-pulse-encoding scheme requires no additional active components. Pressing a key in Section 1 generates a single pulse. Pressing a key in Section 2 causes comparator IC<sub>2</sub>'s output to go low and triggers IC<sub>4</sub> to generate the first pulse. At the same time, R<sub>15</sub> and C<sub>8</sub> apply IC<sub>2</sub>'s output transition to comparator IC<sub>1</sub>'s input. As IC<sub>2</sub>'s output, Pin 6, goes low, C<sub>3</sub> starts to charge, and, after a delay of about 2 msec, IC<sub>1</sub> triggers IC<sub>4</sub> to generate the required second pulse. Because the time-constant prod-

uct of R<sub>15</sub> × C<sub>8</sub> is much longer than 2 msec, C<sub>8</sub> does not charge significantly during the 2-msec interval. However, C<sub>8</sub> does charge to a full 5V during the interval in which the key is pressed, which allows comparator IC<sub>1</sub> to recover to its steady state.

In a three-section keypad, pressing a key in the third section activates comparator IC<sub>3</sub>, which in turn triggers IC<sub>4</sub> to produce the first pulse. After feedback through R<sub>17</sub> and C<sub>10</sub> produces a 2-msec delay, IC<sub>2</sub> triggers IC<sub>4</sub> to produce a second pulse. After yet another 2-msec delay, IC<sub>1</sub> triggers IC<sub>4</sub> to produce the third and final pulse. Although somewhat arbitrary, a 2-msec delay provides sufficient margin over the maximum key-pressed pulse width of

1.32 msec. The interrupt routine provides additional timing margin by allowing as much as 3 msec between pulses.

**Listing 1**, available in the online version of this Design Idea at [www.edn.com/050929di1](http://www.edn.com/050929di1), represents a modified version of the interrupt routine in **Reference 1** and supports the circuit in **Figure 1** for any number of keypad sections. For a three-section-keyboard implementation, the routine returns a key number from 1 to 36. **EDN**

**REFERENCE**

■ Schleicher, Israel, "Single-wire keypad interface frees microcontroller-I/O pins," *EDN*, March 31, 2005, pg 75, [www.edn.com/article/CA512131](http://www.edn.com/article/CA512131).

## Added components improve switching-regulator stability

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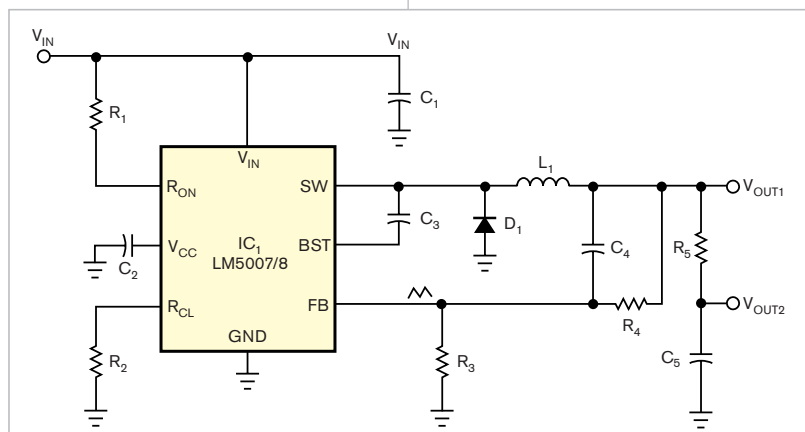
This Design Idea shows how adding one or two passive parts can reduce a hysteretic constant-on-time switched-mode voltage regulator's output voltage ripple and reduce its susceptibility to variations in external load capacitance and ESR (equivalent series resistance). The regulator operates much like a pure hysteretic switcher. The device's internal one-shot sets its pass transistor's on-time, making it less prone to frequency runaway but still susceptible to noise injected at the regulator's FB (feedback) pin. To switch cleanly with predictable frequency and duty cycle, the switcher requires application of approximately 50 to 100 mV of ripple voltage to the feedback pin. This Design Idea shows how four circuit implementations using National's LM5007 and LM5008 regulators satisfy the important feedback-ripple requirements by adjusting on-resistance and maintaining a nearly constant switching frequency as input voltage varies.

**Figure 1** shows a basic buck regulator whose output capacitor, C<sub>5</sub>, presents

a high internal ESR, R<sub>5</sub>. Note that the designer cannot access R<sub>5</sub> and V<sub>OUT2</sub>. Inductor L<sub>1</sub>'s ripple current flows through R<sub>5</sub> and C<sub>5</sub> and produces a certain amount of ripple voltage at V<sub>OUT1</sub>. Although common, this simple design presents two problems: First, feedback resistors R<sub>4</sub> and R<sub>3</sub> form a voltage divider that reduces the output ripple

presented to IC<sub>5</sub>'s FB pin. Thus, 50 mV of ripple at the pin may correspond to excessive ripple voltage at V<sub>OUT1</sub>. Adding a compensation capacitor, C<sub>4</sub>, forces the output-ripple voltage to appear at the feedback pin. Second, a typical pc board may include many low-ESR ceramic bypass capacitors that attenuate ripple voltage to a level that destabilizes the circuit.

Suppose that you replace C<sub>5</sub> in **Figure 1** with a low-ESR capacitor and add R<sub>5</sub> as a discrete resistor. You can connect the external load to V<sub>OUT2</sub> to reduce output-ripple voltage at the load



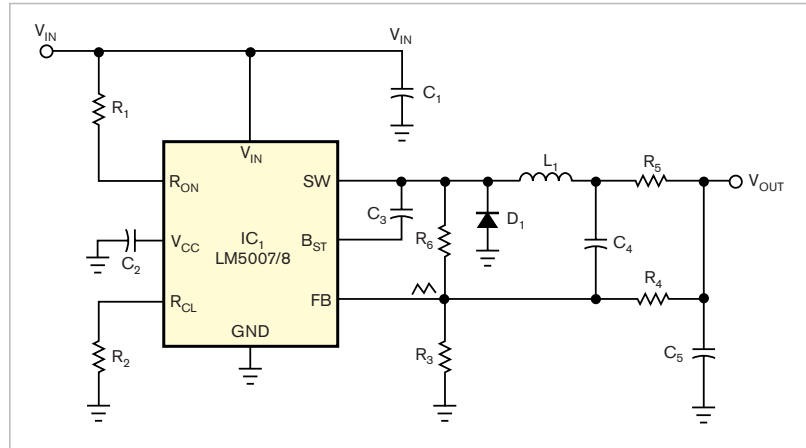
**Figure 1** This basic switched-mode voltage regulator reduces its input voltage to a lower value.

and increase the circuit's immunity to added load capacitance. The regulator's feedback voltage derives from  $V_{OUT1}$ , and  $C_4$  reduces output-ripple losses in the feedback divider,  $R_4$  and  $R_3$ . Unfortunately, this frequently used design introduces new problems: As output current increases,  $V_{OUT2}$  falls below  $V_{OUT1}$  and degrades load-voltage regulation. Second,  $R_5$  carries full load current and dissipates power, reducing overall efficiency. Using a large, high-wattage fractional-ohm resistor at  $R_5$  increases product cost and regulator-package dimensions.

**Figure 2** shows a rearranged buck-mode-switching regulator with two additional components. Assume that  $C_5$ 's ESR is negligible and that  $R_6$  is open. Now,  $R_5$  remains inside the feedback loop, and  $C_4$  couples voltage ripple from the inductor side of  $R_5$  to feedback. This action stabilizes the regulator's operation and requires almost no output-ripple voltage, and  $R_5$  introduces only a small reduction in efficiency. Taking dc feedback at the load preserves the circuit's excellent load regulation.

In another scenario, suppose that you replace  $R_5$  with a short circuit and select values for  $R_6$  and  $C_4$  that provide the desired amount of ripple voltage at the feedback pin. This configuration produces almost no output-voltage ripple and eliminates  $R_5$ 's power losses. Load regulation suffers because, as load current increases, inductor  $L_1$ 's ESR introduces voltage droop at  $V_{OUT1}$  and forces the voltage at switching node SW slightly higher. However, designers can select  $L_1$  for a low ESR that minimizes its effects on load regulation and can make  $R_6$ 's resistance larger than that of  $R_4$ .

The following examples compare output ripple, circuit losses, and component count for the design scenarios. Assume that input voltage is 50V, out-



**Figure 2** Adding two components,  $R_6$  and  $C_4$ , helps reduce output-ripple voltage and improve output-voltage regulation.

put voltage is 5V, output current is 400 mA, switching frequency is 480 kHz, and desired minimum feedback ripple is 50 mV p-p. Select  $L_1$  to operate at a ripple current of 200 mA. Solving for  $L_1$ , you obtain:

$$L_1 = \frac{(V_{IN} - V_{OUT}) \times t_{ON}}{I_{RIPPLE}}$$

$$\frac{(50 - 5) \times t_{ON}}{0.2}$$

$$t_{ON} = \frac{V_{OUT}}{V_{IN} \times f_S} = \frac{5}{50 \times 480 \text{ k}\Omega} = 0.21 \mu\text{SEC.}$$

Substituting on-time into the above equation yields

$$L_1 = \frac{45 \times 0.21}{0.2} = 47 \mu\text{H.}$$

Select a Coilcraft DO1813P-473HC with ESR of  $0.47\Omega$  based on its small pc-board footprint. For  $C_5$ , choose a ceramic capacitor that's large enough to limit the ripple voltage on  $V_{OUT}$  to less than 10 mV p-p. Given the maximum ripple voltage and a known triangle-wave current drive, calculate a value for  $C_5$ :

$$C_5 = \frac{I_{RIPPLE}}{(8 \times V_{RIPPLE} \times f_S)} = \frac{0.2}{(8 \times 0.010 \times 480 \text{ kHz})} = 5.2 \mu\text{F.}$$

For  $C_5$ , you can use TDK's 10- $\mu\text{F}$  C3216X7R0J106 ceramic capacitor that presents an ESR of 3 m $\Omega$  or less. Because the internal reference voltage for the LM5007 or LM5008 is 2.5V, set feedback resistors  $R_3$  and  $R_4$  to 1 k $\Omega$  to divide the regulator's 5V output to 2.5V. Next, select values for  $R_5$ ,  $C_4$ , and  $R_6$  to compare results for each design. In the first scenario, to provide 100 mV of ripple at  $V_{OUT1}$  and 50-mV ripple at the feedback pin in the circuit of **Figure 1**, the design requires a value of  $0.25\Omega$  for  $R_5$ . Adding  $C_4$  changes the value of  $R_5$  to  $0.125\Omega$  to provide 50-mV ripple at  $V_{OUT1}$  and the feedback pin. You calculate a value for  $C_4$  that passes the ripple current:

$$C_4 = \frac{10}{(2\pi(R_4 \parallel R_3) \times f_S)} = \frac{10}{(2 \times 3.14159 \times 500 \times 480 \text{ k}\Omega)} = 6.6 \text{ nF,}$$

**TABLE 1** CIRCUIT OPTIONS

Scenario	Figure	$R_5$ ( $\Omega$ )	$R_6$ ( $\Omega$ )	$C_4$ (nF)	$V_{OUT}$ , $R_6$ (mV)	Voltage droop (mV)	Power dissipation (mW)	Issues
1	1	0.25	0	0	100	0	0	Load capacitance and stability
1	1	0.125	0	6.6	50	0	0	Load capacitance and stability
2	1	0.25	0	0	6	100	40	Voltage droop and power dissipation
2	1	0.125	0	6.6	6	50	20	Power dissipation
3	2	0.125	0	6.6	6	0	20	Power dissipation
4	2	0	30k	6.6	6	7	0	Voltage droop

where  $R_4 \parallel R_3$  represents the value of the parallel combination of  $R_4$  and  $R_3$ .

In the second scenario,  $V_{OUT2}$  droops to 100 mV at full load but exhibits only 6 mV of ripple without  $C_4$  in the circuit. Losses in  $R_5$  amount to 40 mW. Adding  $C_4$  delivers 50 mV of ripple to the feedback pin and  $V_{OUT1}$ ; setting  $R_5$  to  $0.125\Omega$  reduces  $R_5$ 's power loss to 20 mW. In the third scenario,  $R_6$  is an open circuit, and the design in **Figure 2** requires an  $R_5$  value of  $0.125\Omega$  to provide 50-mV ripple at FB.  $V_{OUT1}$  exhibits no voltage droop at full load current and has only 6 mV of ripple voltage.  $R_5$ 's power loss is 20 mW.

In the fourth case, a short circuit replaces  $R_5$ , and the design in **Figure 2** requires a value for  $R_6$  that increases the voltage across  $C_4$  to provide 50 mV of ripple voltage at FB. Use the following equations to calculate the value:

$$R_6 = \frac{(V_{IN}-2.5)}{C_4 \times 50 \text{ mV} \times V_{IN} \times ((f_S / V_{OUT}))} = 30 \text{ k}\Omega.$$

$$R_{\text{RIPPLE}} = (V_{IN}-2.5)/(C_{\text{RIPPLE}} \times 50 \text{ mV} \times V_{IN} \times f_S / V_{OUT}) = 30 \text{ k}\Omega.$$

With  $R_6$  in the circuit,  $V_{OUT}$  drops slightly because  $R_6$  and  $R_4$  effectively connect in parallel. To compensate, you can slightly increase  $R_4$  so that the new value of  $R_4$  in parallel with  $R_6$  equals  $R_4$ 's original value. Thus,

$$R_{4(\text{NEW})} = \frac{R_{4(\text{OLD})} \times R_6}{R_6 - R_{4(\text{OLD})}} = \frac{1000 \times 30}{29} = 1.034 \text{ k}\Omega.$$

In this instance, you may decide not to use the new value of  $R_4$  because adding  $R_6$  raises  $V_{OUT}$  by only 85 mV.

Adding  $R_6$  produces 6 mV of ripple at  $V_{OUT}$  and little or no loss in  $R_5$ , but load regulation will not be perfect.

Inductor  $L_1$ 's ESR of  $0.47\Omega$  introduces a voltage drop of about 200 mV at full load, which increases the voltage at the junction of  $L_1$  and  $R_6$  and also reduces  $V_{OUT}$  to maintain 2.5V at FB. You can calculate the magnitude of the change by multiplying  $L_1$ 's voltage drop of 200 mV by the ratio of  $R_4$  to  $R_6$ :

$$V_{\text{OUT(DROOP)}} = \frac{200 \text{ mV} \times R_4}{R_6} = \frac{0.2 \times 1}{30} = 0.0067\text{V}.$$

Note that output-voltage droop and power dissipation become more significant in designs that deliver higher output current or lower output voltage (**Table 1**). **EDN**

## Electromechanical damping stabilizes analog-meter readings

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Before shipping moving-coil meters, manufacturers may short-circuit the meters' terminals with a length of wire, which provides effective electromagnetic damping and results in better immunity to external mechanical vibration and shocks that can occur during transportation. This Design Idea applies essentially the same principle to analog meters under normal operating conditions. Connecting a meter to a voltage source with low internal resistance applies electromagnetic damping and makes the meter's readings more stable. Increased immunity to external vibration and shock takes on importance in mobile- or portable-system applications and especially in automotive devices.

For example, suppose that your application requires measurement of a 0 to 10V power supply (**Figure 1**). You have available a typical electromechanical

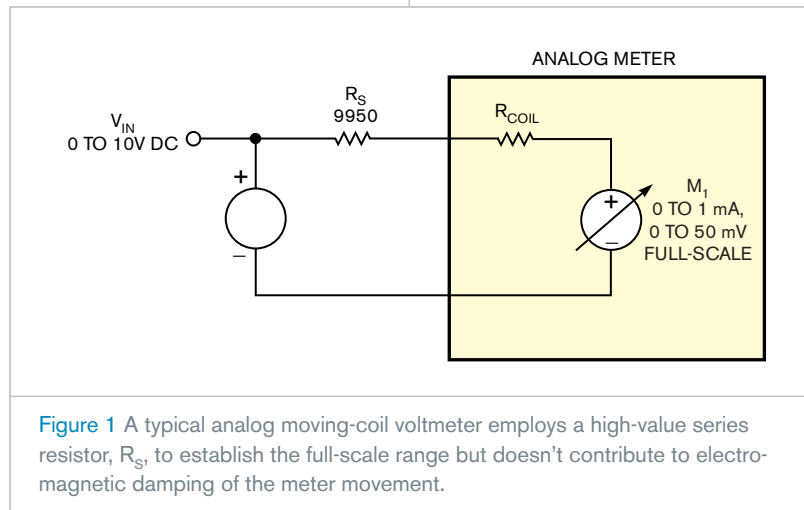
meter that presents a full-scale voltage rating,  $V_{FS}$ , of 50 mV and a full-scale current rating of 1 mA. To obtain the 10V full-scale voltage range, you add a series resistance,  $R_S$ . First, calculate the meter's internal resistance,  $R_{COIL}$ :

$$R_{\text{COIL}} = \frac{V_{\text{FS}}}{I_{\text{FS}}} = \frac{50 \text{ mV}}{1 \text{ mA}} = 50\Omega.$$

Next, calculate the multiplier resistor,  $R_S$ , as follows:

$$R_S = \frac{(V_{\text{IN}} - V_{\text{FS}})}{I_{\text{FS}}} = \frac{(10\text{V} - 0.05\text{V})}{1 \text{ mA}} = 9950\Omega.$$

The resistance of  $R_S$  typically great-

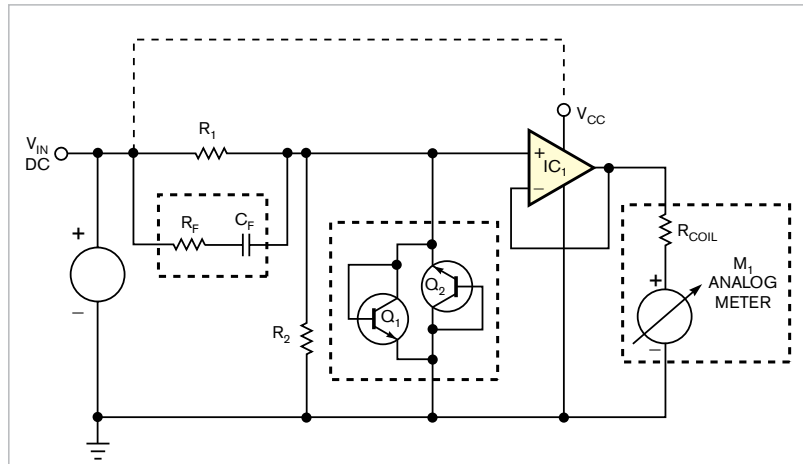


**Figure 1** A typical analog moving-coil voltmeter employs a high-value series resistor,  $R_S$ , to establish the full-scale range but doesn't contribute to electromagnetic damping of the meter movement.

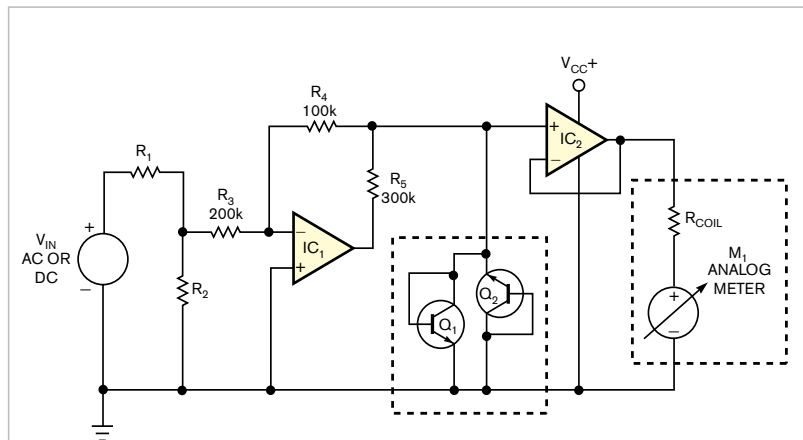
ly exceeds that of  $R_{COIL}$  and therefore significantly reduces the electromagnetic damping action on the meter movement. Although you can improve damping by shunting the meter with a capacitor, this approach also increases the meter's settling time.

**Figure 2** illustrates a better approach, in which a moving-coil meter connects to the output of an operational amplifier,  $IC_1$ , embedded in a deep negative-voltage-feedback loop. Because the op amp presents an extremely low equivalent output resistance, the meter's terminals are "virtually shorted," providing effective electromechanical damping that results in more stable readings and increased shock and vibration resistance. In **Figure 2**, the resistive voltage divider comprising  $R_1$  and  $R_2$  connected to the op amp's noninverting input determines the meter's full-scale reading. You can add  $R_F$  and  $C_F$  to form an optional highpass filter to further improve the meter's settling time. Transistors  $Q_1$  and  $Q_2$  are also optional and added as overvoltage protection. Note that, for normal operation, the transistors' forward base-emitter voltage,  $V_{BE}$ , should be several times larger than the meter's full-scale voltage,  $V_{FS}$ , which is typically 50 to 100 mV.

A rail-to-rail-capable, single-supply micropower op amp makes a good choice for this application. If the input voltage,  $V_{IN}$ , exceeds the op amp's minimum power-supply-voltage requirement, you can connect the op amp's  $V_{CC}$  pin directly to the input terminal, as the dashed line in **Figure 2** shows. In effect, the circuit combines the advantages of meter buffering and improved shock and vibration resistance with a traditional moving-coil meter's advantage of requiring no external power supply. You can choose from among many commercially available off-the-shelf, rail-to-rail output-micropower op amps that draw supply currents well below the full-scale current drain,  $I_{FS}$ , of typical moving-coil meters. For example, Maxim's MAX4289 requires as little as 1V and 9  $\mu$ A of power, and the MAX4470 requires a minimum of



**Figure 2** An operational amplifier's low output resistance provides electro-mechanical damping for a moving-coil meter for stable readings and enhanced resistance to mechanical shock and vibration. Connect  $V_{CC}$  to either an external power supply or the circuit's input terminal if  $V_{IN}$  exceeds  $IC_1$ 's minimum power-supply-voltage rating.



**Figure 3** Based on rail-to-rail operational amplifier  $IC_1$ , this circuit's input stage forms a diodeless, precision full-wave rectifier and enables the circuit of **Figure 2** to display ac- or dc-voltage measurements on a dc meter.

1.8V but only 750 nA of supply current.

Although this Design Idea has so far related only to dc-voltage measurements, you can expand the circuit to include ac- and dc-voltage measurements (**Figure 3**). In this approach, you add a precision diodeless, full-wave-rectifier stage based on a single rail-to-rail operational amplifier and resistors  $R_3$ ,  $R_4$ , and  $R_5$  (**Reference 1**). Resistors  $R_1$

and  $R_2$  determine the full-scale reading. This circuit requires an external dc power supply to drive op amps  $IC_1$  and  $IC_2$ ; voltage-limiting transistors  $Q_1$  and  $Q_2$  are optional. **EDN**

## REFERENCE

1 Bell, Alexander, "Simple Full-Wave Rectifier," *Electronic Design*, April 4, 1994, pg 78.