



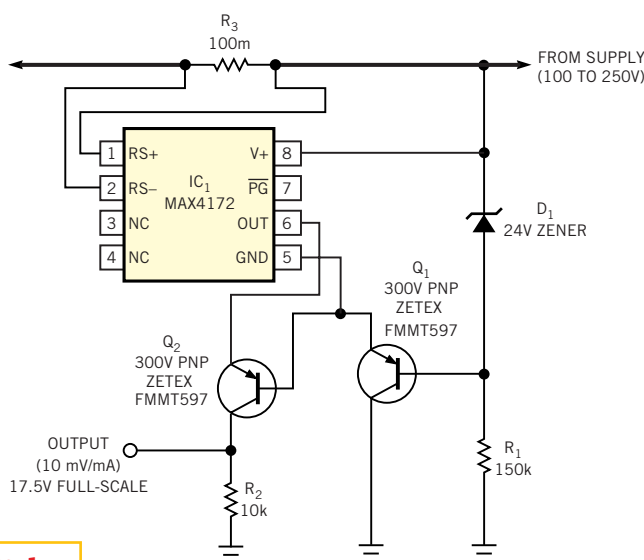
Edited by Bill Travis

## High-side current monitor operates at high voltage

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**T**HE SIMPLEST technique for measuring current in an actuator or a motor is to monitor the ground current with a resistive element between the load and the ground. Because the device and its associated electronics share a ground potential, you need to amplify only the ground-current signal. This approach, however, does not detect device short circuits to ground, which can overload the high-side drive circuitry. To avoid such potential fault conditions, you should use a high-side current monitor to detect short circuits

and similar faults that can occur following the current monitor. High-side current monitoring has advantages, but it finds limited use because of the dearth of devices able to handle the high voltage levels—24V to many hundreds of volts—prevalent in the industry. Off-the-shelf devices can operate to 32 and 76V, but even 76V is insufficient for many applications. **Figure 1** shows a simple way to adapt a standard 32V device for use at any



**Figure 1**

**This circuit enables a 36V current-monitoring IC to operate at common-mode voltages as high as 130V.**

voltage level, subject to limitations of the external components. (The components in **Figure 1** can accommodate 130V.)

The accuracy of the circuit is better than 1% for load currents greater than 30 mA. IC<sub>1</sub>'s current-output stage allows easy implementation of the current mirror needed for level-shifting the output signal to ground. Thus, you can easily monitor the ground-referenced signal by using an A/D converter or a comparator. The circuit monitors load current in the presence of a 130V-dc common-mode level. You must ensure that you do not violate IC<sub>1</sub>'s absolute maximum rating—36V with respect to the ground pin—for the RS+, RS-, and V+ pins. For that purpose, zener diode D<sub>1</sub> limits the voltages between the V+, RS+, and GND pins to 24V. Thus, the typical voltage between these pins is 24V minus the V<sub>BE</sub> of Q<sub>1</sub>, or 23.3V. The zener-diode current for this circuit is approximately 700 μA. Note that the manufacturer's suggested

bias current is 500 μA, but the zener diode's di/dt slope goes *negative* below 300 μA, a condition that can introduce noise or even oscillation. The minimum specified bias—300 to 500 μA—sets the maximum value of R<sub>1</sub>, and the maximum allowed power dissipation for R<sub>1</sub> and D<sub>1</sub> combined sets the minimum value for R<sub>1</sub>. Thus, for supply rails of 100 to 250V, a reasonable R<sub>1</sub> value is 150 to 225 kΩ—150 kΩ in this case.

Q<sub>1</sub> and R<sub>1</sub> form a shunt regulator. The design uses Q<sub>1</sub> because of its maximum V<sub>CE</sub> rating of -300V, high gain of 100V/V at 1 mA, and its

ability to handle 500 mW of power. Output current is proportional to the voltage difference, V<sub>SENSE</sub>, between RS+ and RS-:  $I_{OUT} = g_m \times V_{SENSE}$ , where  $V_{SENSE} = R_{SENSE} \times I_{LOAD}$ . Transconductance for IC<sub>1</sub> is 10 mA/V. If the maximum monitored load current, I<sub>LOAD</sub>, is 4A, and R<sub>SENSE</sub> is 10 mΩ, then the maximum I<sub>OUT</sub> is 10 mA/V × 10 mΩ × 4A = 400 μA. Thus, I<sub>OUT</sub> is proportional to I<sub>LOAD</sub>, and the maximum expected output is 400 μA. For applications of wide dynamic range in which V<sub>SENSE</sub> can approach the absolute maximum rating of the differential pair, 700 mV, you should protect the sense pins by adding series resistors between R<sub>SENSE</sub> and RS+ and between R<sub>SENSE</sub> and RS-. You should select the resistor values to limit input currents to within 10 mA when the RS+ to RS- difference is 700 mV.

I<sub>OUT</sub> is now proportional to I<sub>LOAD</sub>, but, for easy monitoring, you must level-shift it to ground by using the Q<sub>1</sub>-Q<sub>2</sub>

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current mirror.  $Q_2$ 's high gain forces the collector current to closely approximate the emitter current which, when you apply it to  $R_2$ , produces a measurable voltage at  $V_{OUT}$ . As with  $Q_1$ ,  $Q_2$  needs a

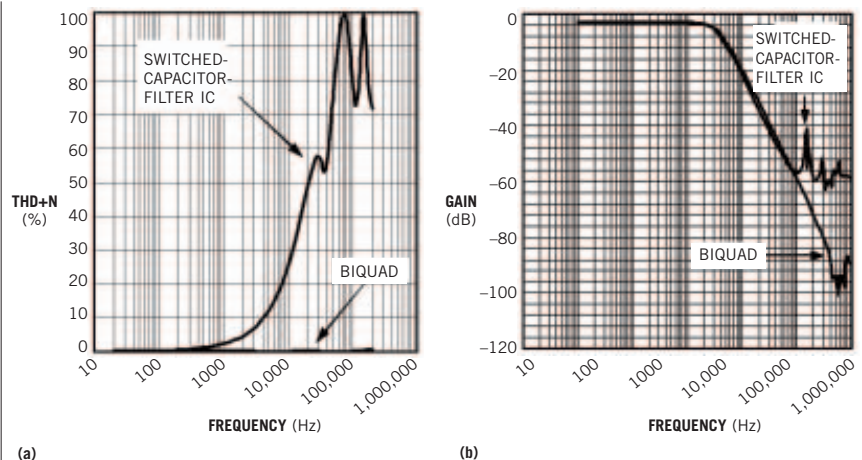
maximum  $V_{CE}$  rating of  $-240V$ . The device in **Figure 1** is rated at  $-300V$ .  $V_{OUT}$  now equals  $I_{OUT} \times R_2$ . (The actual output current at  $Q_2$ 's collector is slightly less, because of  $Q_2$ 's base current.) At

$I_{LOAD} = 4A$ ,  $V_{OUT} = 400 \mu A \times 10 k\Omega = 4V$ . You can accommodate designs with lower or higher operating voltages by properly selecting  $Q_1$ ,  $Q_2$ , and the base resistor,  $R_1$ . □

## Digital potentiometers enable programmable biquadratic filter

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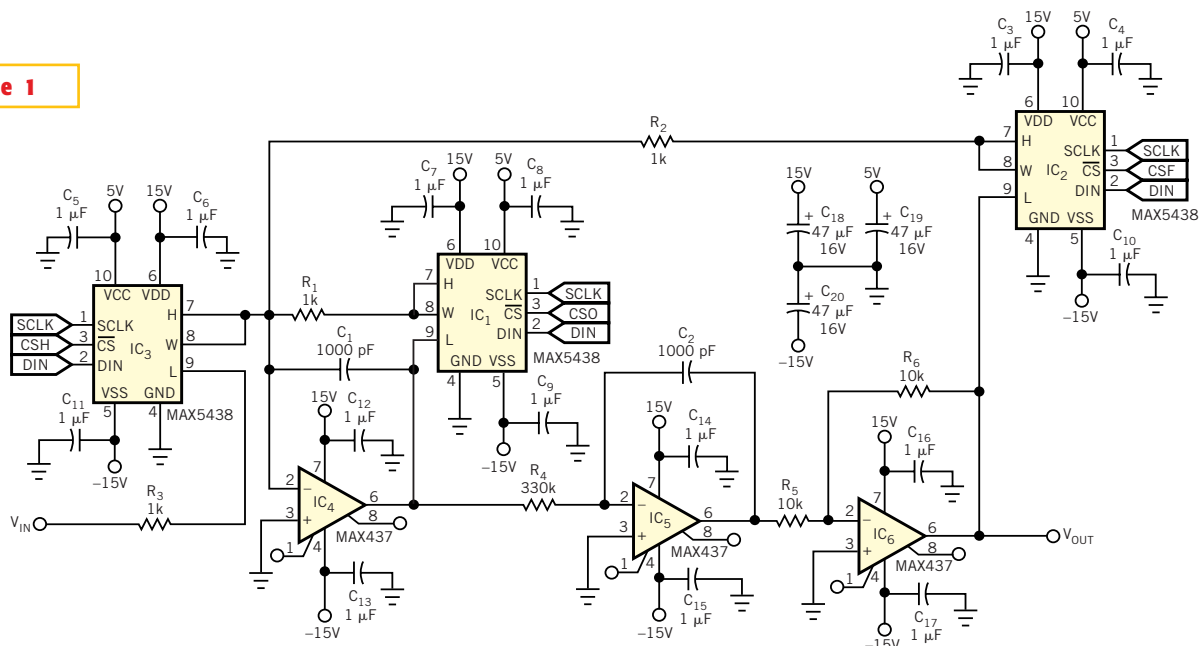
**O**F THE MANY TYPES of analog filters available to designers, few allow easy adjustments of the filter parameters. The biquadratic, or biquad, filter is an exception, however. You can change that filter's corner frequency ( $\omega_0$ ),  $Q$ , and gain ( $H$ ) by adjusting the values of three resistors. For that purpose, the lowpass biquad circuit of **Figure 1** includes three digital potentiometers configured as variable resistors in the feedback loops. Altering the settings of these potentiometers changes the filter characteristics. The circuit produces corner frequencies of 5.5 to 55 kHz;  $Q$  values of 0.055 to 5.5, depending on the selected corner frequency; and gain of 1 to 100, also depending on the selected corner frequency. To tune the biquad filter, you set a corner fre-



**Figure 2**

Noise (a) and low bandwidth (b) plague switched-capacitor filters. The biquad filter of **Figure 1** maintains less than 1% THD+N over the range 20 Hz to 200 kHz.

**Figure 1**



Digital potentiometers adjust the corner frequency,  $Q$ , and gain for this biquad analog filter.

quency  $\omega_0$  in radians per second by adjusting digital potentiometer IC<sub>2</sub> via the SPI interface. In the same fashion, set Q by adjusting IC<sub>1</sub> and gain by adjusting IC<sub>3</sub>.

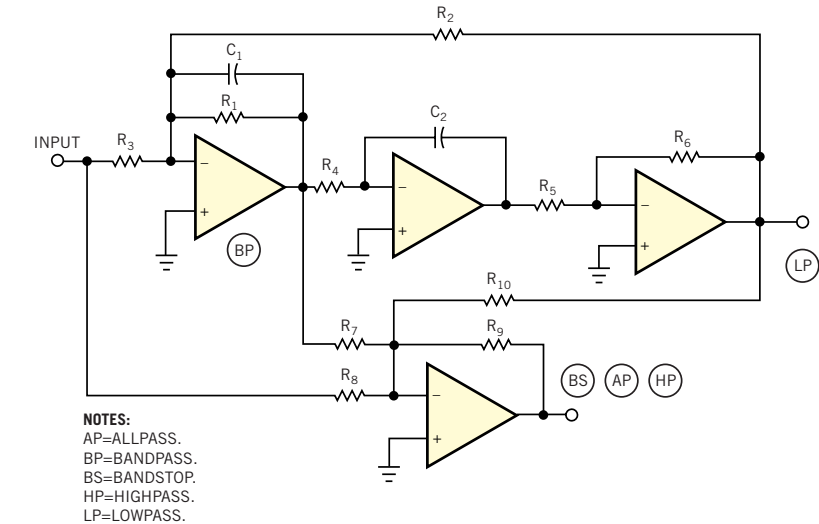
Note that adjusting Q does not affect the corner-frequency setting, and a gain adjustment does not affect the settings of Q or the corner frequency. The three **equations** below demonstrate this orthogonal tuning for the biquad filter:

$$\omega_0^2 = \frac{1}{(R_2 + R_{IC2})R_4C_1C_2}$$

$$Q = \sqrt{\frac{(R_1 + R_{IC1})^2 C_1}{(R_2 + R_{IC2})R_4C_2}}$$

$$H = \frac{R_2 + R_{IC2}}{R_3 + R_{IC3}}$$

where  $R_{IC1}$ ,  $R_{IC2}$ , and  $R_{IC3}$  are the input resistances of IC<sub>1</sub>, IC<sub>2</sub>, and IC<sub>3</sub>, respectively. The circuit in **Figure 1** is substantially more complex than the switched-capacitor approach usually integrated into an IC, but the switching noise and low bandwidth of a switched-capacitor filter are unacceptable in many applications (**Figure 2**). A biquad filter offers better frequency and noise performance in ex-



**Figure 3**

The standard biquad filter circuit produces lowpass and bandpass responses, and the addition of a fourth op amp produces a highpass response. Removing R<sub>10</sub> and adjusting various component values produce a notch or bandstop response or an allpass response.

change for more pc-board real estate. And, because monolithic switched-capacitor filters are usually expensive, the biquad circuit of **Figure 1** may be a cost-competitive solution. Many filter applications require higher supply voltages, bipolar operation, or both, so the single 5V supply associated with most switched-capacitor filters may be inadequate for a giv-

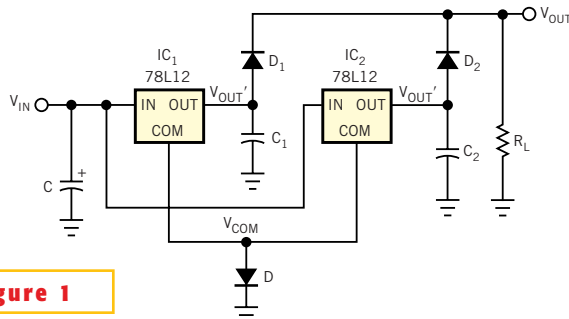
en application. You can implement  $\pm 15V$  voltage rails using digital potentiometers and high-voltage op amps, such as the MAX5438 and MAX437. The biquad filter is not limited to the lowpass response. You can implement highpass, bandpass, bandstop, and allpass filters by adding a fourth op amp to selected terminals of the original lowpass design (**Figure 3**). □

## High-current supply uses standard three-terminal regulator

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**V**OLTAGE-REGULATOR design for high output currents can be a critical and difficult task. Although voltage regulators with 1A maximum output current are simple to design, thanks to 78xx three-terminal voltage regulators, at output currents higher than 1A output, problems arise. For high output currents, voltage

regulators usually use parallel-connected power transistors. The design of the operating points of these power transistors is a difficult task, because the transistors need critical power resistances at the collector and the emitter to set the dc operating point. Both the power transistors and the



**Figure 1**

Two 7812s connect in parallel to double the available output current to 2A.

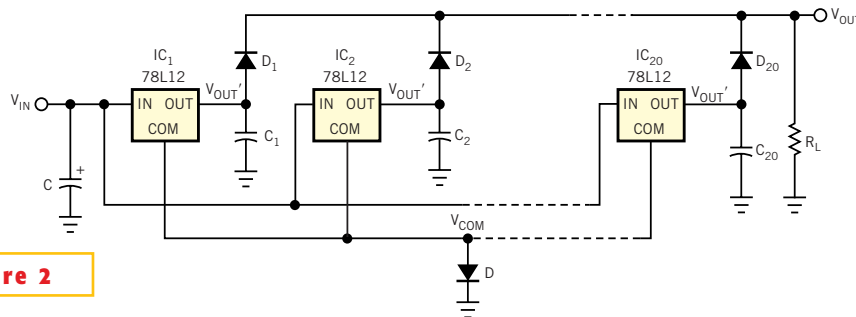
power resistors dissipate high power, so the design needs heat-sinking. This Design Idea implements a simple voltage regulator that delivers high output current. The basic idea is to parallel-connect multiple

three-terminal regulators. These 78xx regulators each handle 1A and are available in 5, 6, 8, 9, 10, 12, 15, 18, and 24V versions. In **Figure 1**, two 7812 regulators are connected in parallel.

The two 7812s operate independently, and each delivers a maximum of 1A. D<sub>1</sub> and D<sub>2</sub> separate and isolate the two regulators. The output voltage is the regulator's nominal output minus a diode drop:  $V_{OUT} = V_{REG} - V_D$ . The output voltage of the regulators is

the specified value if the COM pin connects to ground (0V). To increase the output voltage of the circuit in **Figure 1** to the desired value, the COM pin must connect to a voltage point one diode drop

above ground.  $C_1$ ,  $C_2$ , and  $C_3$  are filter capacitors. **Figure 2** shows a voltage regulator that uses 20 7812s and delivers 20A. All the diodes are 1N4007 types.  $C = 47,000 \mu\text{F}$ , and all the numbered capacitors have a value of  $4700 \mu\text{F}$ . The 7812s all connect to a heat sink that a small fan ventilates. You can expand the concept of this Design Idea to hundreds of amperes. □



## Figure 2

**A total of 20 7812 regulators extend the capability of Figure 1's circuit to 20A.**

## Synchronous flyback circuit provides high-efficiency conversion

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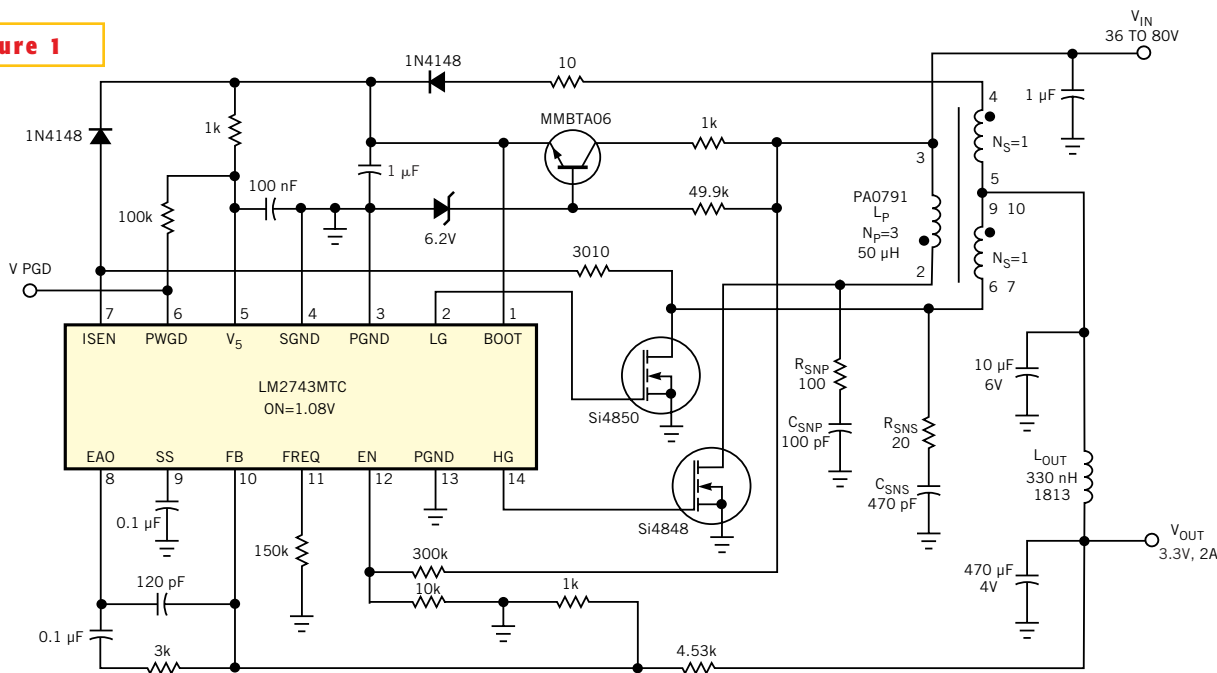
**B**UCK REGULATORS are usually the first choice when you design non-isolated step-down regulators unless the ratio of  $V_{IN}$  to  $V_{OUT}$  is greater than 10, the input voltage is high, or both. Low duty cycle can be problematic for FET drivers and cause current-mode control loops to lose control. Efficiency can fall dramatically to 60 to 70% at low  $V_{OUT}$  and current of only a few amps. The efficiency loss arises from switching losses, because the upper switch always sees full load current. **Figure 1** presents a cir-

cuit that looks a little like a buck regulator and uses a buck controller but is actually a voltage-mode, synchronous flyback circuit. The application it targets needs 3.3V at 2A with an efficiency requirement of greater than 85% and an input-voltage range of 36 to 60V. This one appeared the most promising of several evaluated technologies because of efficiency and cost advantages over buck and asynchronous-flyback approaches.

The LM2743 controller derives its power after start-up from the MMBTA06

transistor and 6.2V zener diode and from a bootstrap winding. Its EN (enable) input is a comparator for UVL (under-voltage lockout) to prevent start-up below 28V. The controller drives a synchronous switch that provides lower loss than a Schottky diode and uses the lower FET's on-resistance as the sense resistor for current limiting. The 150-k $\Omega$  resistor at Pin 11 produces a switching frequency of 250 kHz. The flyback transformer, designed by Pulse Engineering ([www.pulseengineering.com](http://www.pulseengineering.com)), is a low-

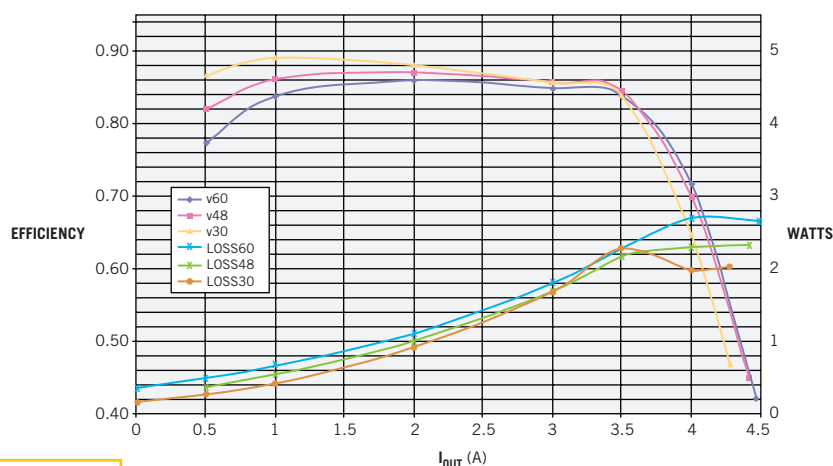
### Figure 1



**This synchronous flyback circuit provides high efficiency with wide input/output ratios.**

cost unit that provides 50  $\mu$ H of primary inductance and a 3-to-1 turns ratio in a 13L $\times$ 15W $\times$ 11H-mm footprint. Its 3-to-1 turns ratio prevents the primary switch from seeing full output current, resulting in less switching loss than that of a buck regulator. The small LC filter at the output allows a single 10- $\mu$ F ceramic capacitor to handle the high rms ripple current, and a low-cost aluminum capacitor also removes ripple and buffers load transients.

**Figure 2** plots measured data at three input voltages and several output currents for the circuit in **Figure 1**. Efficiency is displayed on the left for the three uppermost curves, and the three lower curves show total loss in watts measured by the scale on the right.  $V_{OUT}$  ripple measured 6 mV p-p at no load, rising to 20 mV p-p at 4A. The rapid fall in efficiency at 3.5A comes from current limiting. As with any switcher and especially for flyback designs, pc-board layout



**Figure 2**

The circuit in **Figure 1** provides more than 85% efficiency over a wide range of output currents.

is important. You obtain best performance with four or more layers, separate power and ground planes, and short and wide gate-drive connections. Although the circuit of **Figure 1** targets use in a 7W, single-output requirement, this synchro-

nous flyback circuit applies to a wider power range; you can easily extend it to multiple outputs by adding secondary windings. The additional outputs can use either diode rectifiers or additional FETs driven from the low-gate driver. □