ALTHOUGH RAIL-TO-RAIL OPERATIONAL AMPLIFIERS ARE NOW AN ACCEPTED WAY TO SENSE HIGH-SIDE DC CURRENT, A DEPENDABLE CIRCUIT REQUIRES METICULOUS ANALYSIS AND DESIGN.

Sense dc current from the high side

It is often desirable to perform current sensing on the high side of a load. However, engineers must overcome design hurdles to successfully realize such an implementation. Now, with the availability of rail-to-rail op amps, the circuit in Figure 1 has become a popular means of providing high-side dc-current sensing. This circuit is popular for several reasons. Its single-supply operation and wide supply-voltage range lets you adapt it to high- or low-side current sensing. The input CMRR (common-mode rejection ratio) is also as good as the basic rejection of the op amp itself and does not depend on resistor matching. With easily set gain and range, the circuit produces measurement accuracies of ±1% or better. It also operates over a wide temperature range with proper component selection and requires no “special-function” or single-source ICs.

The virtue of the rail-to-rail op-amp that this case exploits is that the input common-mode range can “reach” all the way to the positive supply rail. Most regular op amps have an input-voltage range that works up to only within about a volt or two of the positive supply rail. Be careful when selecting a rail-to-rail op amp. Manufacturers can use this term to refer to the input-voltage range, the output-voltage range, or both. In this application, the main feature of the rail-to-rail op amp is that its input range includes the positive supply rail. Some op amps that have been around for a while have this feature, which is necessary for high-side sensing, but these devices aren’t classified as rail-to-rail; one example is the LF355 FET input-type op amp.

The circuit in Figure 1 works by sensing the drop across the current-sense resistor and then adjusting the operating point of the output transistor to produce the same current through $R_{OUT}$ and the feedback path that is produced through $R_{IN}$. Due to the inverting response of the transistor in this case, you must direct the feedback path back to the noninverting input for an overall negative feedback response. The transfer function for this circuit is:

$$V_{OUT} = I \cdot R_S \cdot \left( \frac{R_{OUT}}{R_{IN}} \right)$$

This circuit presents a couple of potential sources of error. Obviously, you want to make sure the current-sensing resistor is as accurate as necessary for the application. Additionally, you want to make sure that the gain resistors ($R_{OUT}/R_{IN}$) are accurate. $R'_{IN}$ is optional and is there to further reduce any offset error due to the input bias current of the op amp. Beyond that, you should give some consideration to the output transistor. If you use a bipolar junction transistor, the additional base current contributes to an output error. If the transistor has a beta of, say, 100.
(β=100), you can expect the output to be high by 1% (that is, 1/β, expressed in percent). You could use a Darlington transistor to significantly reduce this error. Additionally, you could use a MOS-FET, which would present no such equivalent base current error, because the source and drain current in such a device are exactly identical. However, because there would likely be some error due to the leakage current through the MOS-FET, a low-leakage device would be the best choice. Such leakage would result in errors only when measuring zero current or very small current values but would not produce a “gain” error in the sense that a bipolar junction transistor would.

The last problem to address is input and output filtering. In most cases, adding a capacitor in parallel with \( R_{\text{OUT}} \) is more than adequate. Doing so gives the familiar filtering response of:

\[
\frac{f_{3\text{dB}}}{\text{Hz}} = \frac{1}{2\pi R_{\text{OUT}} C_{\text{OUT}}}.
\]

If the output must respond rapidly to quick changes in load current, you want to make sure the output filter supports the necessary rise/fall-time requirement. You can quickly estimate this value with the equation:

\[
\tau = (R_{\text{OUT}} C_{\text{OUT}}).
\]

If your circuit must operate in a nasty environment in which it is subject to high levels of radiated and conducted noise, ensure that the sense leads are as short as possible and reduce the loop area between these leads. Resist the temptation to place a capacitor directly between the inverting and the noninverting leads of the op amp; doing so could lead to serious stability problems. If you insist on additional filtering at the input, use the arrangement shown in Figure 2. \( C_{\text{DIFF}} \) helps to limit the bandwidth of the differential noise that would otherwise be amplified and treated as a legitimate signal. In Figure 2, for the values given, the 3-dB bandwidth is about 800 Hz. Tailoring the output stage (\( R_{\text{OUT}} \) and \( C_{\text{OUT}} \)) further assists filtering. The op amp does an adequate job of rejecting common-mode noise, such as 60-Hz noise, but its ability to reject common-mode noise diminishes with increasing frequency. You add the common-mode capacitors (\( C_{\text{CM}} \)) to pick up where the op amp leaves off, in this respect. In the example given, \( f_{3\text{dB-CM}} \) is set at around 160 kHz. Resist the temptation to use larger values of \( C_{\text{CM}} \) in the hope of further improving common-mode rejection. If the \( C_{\text{CM}} \) capacitors are too large and poorly matched in value, the common-mode noise transforms into differential-mode signals (due to the capacitor-value mismatch), and the amplifier then treats it as a legitimate signal.

A rule of thumb is to choose the value of \( C_{\text{DIFF}} \) that makes the most sense for setting the bandwidth of the input. Then, choose \( C_{\text{CM}} \) to be a factor of 10 times and preferably 100 times smaller. Make sure that those capacitors have reasonable tolerance and are well-matched. Ceramic capacitors with a low temperature coefficient are inexpensive and well-suited to the role of \( C_{\text{CM}} \). Also, keep in mind that proper decoupling across the op-amp supply pins is important, because conducted noise can also make its way through the supply—especially if the op amp is powered directly from the high-side supply line, as the figure shows.

The output imped-

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**Figure 2**

Input filtering stabilizes dc-current-sense circuits operating in noisy environments.
Design Feature  
**DC-current-sense approaches**

The design of current-sense circuits is ideal. The op amp sees only the output from a higher supply, the circuit in Figure 3 is ideal. The op amp sees only the voltage across the zener regulator, and the output transistor provides the necessary ground-referenced level shifting. The only limitation on the maximum value of the high-side voltage is that it not exceed the breakdown voltage rating of the output transistor.

Note that the op amp’s point of reference is relative to its 

V^+ supply terminal, as opposed to the V^- supply terminal in the previous circuits. It requires that you trade the npn (or n-channel FET) transistors for pnp (or p-channel FET) transistors and observe the proper orientation of the inverting and noninverting inputs. Otherwise, the same equations, discussion, filtering techniques, and principals of operation as those for the previous circuits govern this circuit. About the only downside to this circuit is that it must use a physically separate op amp if the output requires buffering.

These circuits are flexible and generally well-behaved. However, depending on your final selection of op amp, transistor, and filter arrangement, it’s always prudent to do testing and make sure there is sufficient compliance over the entire expected current measurement range to ensure stable operation. You can verify the stability by monitoring the base or gate voltage with a scope while subjecting the circuit to step changes in load current.

**Table 1** details off-the-shelf integrated current-sense devices that are based on architectures similar to those that this article describes. However, although many of the devices it lists may be “functionally similar” to one another, you are unlikely to find any device that is second-sourced. When selecting such devices, ensure the voltage rating is within your design’s supply range. Some devices allow for bidirectional current sensing and may have comparators for providing a “trip” indication. Some of the devices are actually rated for full operation over -40 to +125°C for extended industrial and automotive applications.

The IR21XX series is unique in that its floating-channel architecture allows high/low-side sensing for potentials as high as 600V. The output is not an analog voltage but rather an output pulse width proportional to sensed current. It allows for microprocessor interfacing without an ADC. Additionally, you may transfer the pulse width through an optocoupler for true isolation.

The **table** also includes some Hall-effect-type current-sense options. Hall-effect sensing allows for fully isolated current sensing just as a current transformer does; the difference is that the Hall effect suits both dc and ac current sensing.

The list in the **table** is by no means complete, and many manufacturers regularly add current-sense products to their portfolio and discontinue products at an even faster pace.

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**FOR MORE INFORMATION...**

For more information on products such as those discussed in this article, contact any of the following manufacturers directly, and please let them know you read about their products in EDN.

**Allegro MicroSystems**
1-508-853-5000
www.allegromicro.com

**Analog Devices**
1-800-447-4418
www.analog.com

**Cirrus Logic/Crystal**
1-512-851-4000
www.cirrus.com

**International Rectifier**
1-310-322-3331
www.irf.com

**Linear Technology**
1-800-350-9697
www.linear.com

**Maxim**
1-408-737-7600
www.maxim-ic.com

**Sentron AG**
+41 41 711 21 70
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**TI/Burr-Brown**
1-800-336-5236
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