High-Side Current Sensing with Wide Dynamic Range: Three Solutions

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Introduction

Current sensing is a critical function, necessary for precision closed-loop control in applications such as motor control, solenoid control, communications infrastructure, and power management. End uses range from safety-critical automotive and industrial applications to handheld devices, where power and efficiency are essential. Precision current monitoring allows designers to obtain critical instantaneous information, such as motor torque—based on motor current—efficiency of a dc-to-dc converter, or bias current in a base station’s LDMOS (laterally diffused MOS) power transistor, or diagnostic information, such as shorts to ground.

To understand the key trade-offs, options, and challenges faced by system designers when choosing the most accurate, cost-effective current sensor for a circuit board, we take a close look at current sensing in LDMOS bias current monitoring in cellular base station power amplifiers and other relevant applications.

Current monitoring is necessary in base station power amplifiers, especially with the more complex modulation methods used in 3G and LTE, where the peak-to-average power ratio varies from 3.5 dB (about 2.2 to 1) for 3G W-CDMA to 8.5 dB (about 7.1 to 1) for LTE OFDM—compared with 3 dB (about 2 to 1) for the most popular 2G single-carrier GSM. One of the control-loop functions is to monitor the LDMOS bias current, which allows the bias of the LDMOS to be properly modulated for a given power output. Typically, this dc bias current has a wide dynamic range based on operating, maximum, or off-peak operation. To the designer, this means that an accurate current sensor is needed to monitor a current that can range from 50 mA (or as little as 15 mA)1 to 20 A, while the drain of the LDMOS is biased at a high voltage ranging from 28 V to 60 V. Using a shunt resistor to monitor this current means that the designer is limited to a very small shunt that will not dissipate too much power when the LDMOS current is 20 A. As an example, even a 10 mΩ shunt will dissipate 4 W at maximum current.

While shunt resistors are available to handle this power, lower power dissipation might be a requirement of the board. But the choice of such low resistance values means that at low currents, say 50 mA, the voltage across the 10 mΩ shunt is extremely small (500 μV), making it a challenge to monitor accurately—with a circuit that must also withstand high common-mode voltage.

This article will focus on providing current-sensing solutions that can help designers accurately monitor wide-ranging dc currents in the presence of high common-mode voltages. Special attention will also be devoted to temperature performance, a critical parameter that is often not easy to calibrate but must be faced in the case of outdoor power amplifiers. Three optional solution approaches are described here—in order of decreasing design complexity—that provide viable high-accuracy, high-resolution current sensing for a variety of applications.

1. Use discrete components, such as op amps, resistors, and Zener diodes, to build a current sensor. This solution employs the AD8628 zero-drift amplifier as its key component.

2. Attain an increased level of integration by using a high voltage bidirectional current shunt monitor, such as the AD8210, with additional external components to extend dynamic range and accuracy.

3. Employ an application-optimized device such as the newly available AD8217, an easy-to-use, highly integrated zero-drift current sensor with an input common-mode voltage range of 4.5 V to 80 V.

Configuring a Standard Op Amp for High-Side Current Sensing

Figure 1 shows an op-amp-based discrete solution employing the AD8628. The same setup is valid with other op amps as well, but specific features that are required include low input offset voltage, low offset voltage drift, low input bias current, and rail-to-rail input- and output swing capability, if possible. Other recommended amplifiers include the AD8538, AD8571, and AD8551.

This circuit monitors the high-side current, I. The amplifier is biased on via the Zener diode, which, in this case, is rated at 5.1 V. Its use ensures that the amplifier operates safely at the high common-mode level and that its supply voltage remains steady and within the allowed supply limits, while the output is converted to a current by the MOSFET—and to a ground-referenced voltage by resistor RL. The output voltage can thus feed converters, analog processors, and other ground-referenced components—such as op amps or comparators—for further signal conditioning.

In this configuration, the voltage across RG is equal to the voltage across RSHUNT because feedback via the MOSFET maintains both high-impedance op-amp inputs at the same voltage. The current through RG flows through the FET and RL to develop VOUTPUT. The relationship between the current, I, flowing through the shunt resistor and VOUTPUT is expressed by Equation 1:

\[
V_{\text{OUTPUT}} = \frac{I \times R_{\text{SHUNT}}}{R_G} \times R_L
\]

RSHUNT Selection: The maximum value of RSHUNT is limited by the allowable power consumption at maximum current. The minimum value of RSHUNT is limited by the input range and error budget of the op amp. Normally, the value of RSHUNT is from 1 mΩ to 10 mΩ for monitoring currents greater than 10 A. If a single resistor cannot meet the power consumption requirement or is too large for the PCB, RSHUNT may have to be made up of multiple resistors in parallel.

Figure 1. Discrete high-current sensing solution using an operational amplifier.
Consider the LDMOS drain current monitoring example, with 28-V common mode and $I_{DSS}$ of 100 nA. The mirror of the minimum current through $R_L$ should be at least 20 times $I_{DSS}$. This results in

$$R_{G_{-MAX}} = \frac{I_{MIN} \times R_{SHUNT}}{20 \times 100 \text{nA}}$$

The minimum $R_G$ is limited by allowable mirror current power consumption at maximum load current

$$R_{G_{-MIN}} = \frac{I_{MAX} \times R_{SHUNT}}{I_{MIRROR_{-MAX}}}$$

$R_{BLAS}$ Selection: The current through $R_{BLAS}$ divides to produce the op amp’s quiescent current and the essentially constant Zener-diode voltage, $V_D$, which determines the op amp’s supply voltage. Make sure the current flowing through the Zener diode does not exceed its maximum regulated current, $I_{Z_{-MAX}}$, when the amplifier current, $I_{SUPPLY}$, is essentially zero and $V_{IN}$ is the maximum:

$$R_{BLAS_{-MIN}} = \frac{V_{IN_{-MAX}} - V_Z}{I_{Z_{-MAX}}}$$

To ensure a stable diode voltage, the current flowing through it should be higher than its minimum operating current, $I_{Z_{-MIN}}$, when $I_{SUPPLY}$ is the maximum and $V_{IN}$ is the minimum:

$$R_{BLAS_{-MAX}} = \frac{V_{IN_{-MIN}} - V_Z}{I_{Z_{-MIN}} + I_{SUPPLY_{-MAX}}}$$

The Zener diode and $R_{BLAS}$ are the key components in this solution because they take away the high common-mode voltage from the following circuits and allow the use of a low-voltage precision op amp. For best voltage stability, the Zener diode should have low dynamic resistance and low temperature drift.

$R_1$ Selection: $R_1$ is used to limit the amplifier input current if internal transients exceed the op amp’s power supply voltage. A 10-kΩ resistor is recommended.

The offset voltage, $V_{OS}$, and offset current, $I_{OS}$, of the chosen op amp are critical, especially with low values of shunt resistance and at low load currents. $V_{OS} + I_{OS} \times R_1$ must be smaller than $I_{MIN} \times R_{SHUNTS}$ or the amplifier may be saturated. Therefore, a rail-to-rail input amplifier with zero crossover distortion is preferred for optimal performance.

Another issue to consider for this discrete solution is temperature drift. Even if an amplifier with zero drift is used, it is very difficult or costly to optimize drifts caused by discrete components: the Zener diode, MOSFET, and resistors. From Table 1, the maximum $I_{DSS}$ of the MOSFET changes from ~10 μA to ~60 μA as the operating temperature changes from 25°C to 125°C with $V_{GS} = 0$ V and $V_{DS} = −50$ V. This drift will degrade the system accuracy over temperature, especially when the monitored current is low. The drift of the Zener diode will impact the stability of the amplifier’s power supply, so the amplifier used should have high power-supply rejection (PSR).

Further, designers must take into account the low power efficiency of this solution: A lot of power is consumed by $R_{BLAS}$. For example, if the bus common-mode voltage is 28 V, the Zener-diode voltage output is 5.1 V, and $R_{BLAS}$ is a 1000-Ω resistor, the circuit will dissipate more than 0.52 W of undesirable power. This adds to the power-consumption budget and must be taken into account.

High-Side Current Sensing with the AD8210 and External Components

Figure 2a shows a simplified block diagram of the AD8210 integrated high-voltage bidirectional current shunt monitor; Figure 2b shows a unidirectional application with an external voltage reference.
The differential currents through Q1 and Q2 are converted into a differential voltage by R1 and R2. A2 is configured as an instrumentation amplifier. The differential voltage is converted into a single-ended output voltage by A2. The gain is internally set to 20 V/V with precision-trimmed, thin film resistors.

The output reference voltage is easily adjusted using the VREF1 and VREF2 pins. In a typical configuration for handling bidirectional current flow, VREF1 is connected to VCC while VREF2 is connected to GND. In this case, the output is centered at VCC/2 when the input signal is 0 V, so with a 5-V supply the output is centered at 2.5 V. The output will be greater or less than 2.5 V, depending on the direction of the current through the shunt resistor.

This configuration works well for charge/discharge applications, but if the user needs to utilize the entire output range to measure a unidirectional current flow, then the circuit of Figure 2b shows a typical way that an external source may be used to set the range. Here a resistive divider is buffered by an op amp to drive the VREF1 and VREF2 pins, connected together, to offset the output.

It is difficult for the amplifier alone to monitor the load current as it comes close to zero. With a 5-V supply, the AD8210 specifies a linear output range with a minimum output of 50 mV and a maximum output of 4.9 V. Consider an application where the maximum output of 4.9 V. For example—the AD8603, AD8601, AD8603, AD8605, AD8613, AD8691, and AD8655.

Recently, Analog Devices introduced the AD8217, a high-voltage current sensor featuring zero drift and 500-kHz bandwidth, designed specifically to enhance resolution and accuracy over wide temperature-, input common-mode-, and differential voltage ranges. Figure 3a shows a simplified block diagram of the device; Figure 3b shows it in a typical application.

For measuring very small currents through a small shunt resistor, the AD8217 features a minimum 20-mV output range over temperature, an improvement over the AD8210’s 50-mV range. Thus, if the minimum load current being monitored across the shunt produces 20-mV minimum output, which is 1-mV minimum input, from the current sensor, then the user can choose the AD8217 configured as shown in Figure 3b. The relationship between the output voltage of the AD8217 and the input current can be calculated as Equation 3:

$$V_{OUT} = I \times R_{SHUNT} \times 20$$

The AD8217 features an internal low-dropout regulator (LDO), which provides a constant-voltage supply to the amplifier. The LDO withstands the high common-mode voltage that can vary from 4.5 V to 80 V, essentially performing a similar function to that of the Zener diode in Figure 1.

The AD8217 features a factory-set gain of 20 V/V, with a maximum ±0.35% gain error over the entire temperature range. The initial offset, specified ±300 μV over temperature, and the miniscule temperature drift, ±100 nV/°C, will enhance any error budget. The buffered output voltage directly interfaces with any typical analog-to-digital converter. Regardless of the common mode,
the AD8217 provides the correct output voltage when the input differential is at least 1 mV. Using a 10-mΩ shunt resistor, as above, the minimum current can be as low as 100 mA.

The single-chip solution avoids the temperature drift and power consumption issues of the discrete solution.

**Performance Results Compared**

The following section will display test results obtained comparing the three different methods. The input current through the shunt was adjusted by changing both the input voltage and load resistance. In the data, initial calibration has been performed in order to remove the initial gain and offset errors associated with all the parts used in our board.

Figure 4 is a linearity plot of the output voltage across $R_L$ as a function of low-end values of input current flowing through $R_{SHUNT}$, measured using the circuit of Figure 1. $R_{SHUNT}$ is 10 mΩ; $R_I$ is 13 Ω; $R_{ILAS}$ is 100 Ω; $R_1$ is 10 kΩ; load resistance is 200 Ω; $R_L$ is 200 Ω; the Zener diode output is 5.1 V; the op amp is AD8628; the MOSFET is BSS84. The maximum relative error is 0.69%, and the average is 0.21% after calibration.

**Table 2. Maximum and Average Error at Different Temperatures Using the Same Correction Factor**

<table>
<thead>
<tr>
<th>Solution Circuit</th>
<th>AD8628</th>
<th>AD8210</th>
<th>AD8217</th>
</tr>
</thead>
<tbody>
<tr>
<td>–40°C</td>
<td>Max error (%) 11.982</td>
<td>2.117</td>
<td>0.271</td>
</tr>
<tr>
<td></td>
<td>Avg error (%) 4.929</td>
<td>2.059</td>
<td>0.171</td>
</tr>
<tr>
<td>+25°C</td>
<td>Max error (%) 1.806</td>
<td>0.075</td>
<td>0.103</td>
</tr>
<tr>
<td></td>
<td>Avg error (%) 0.228</td>
<td>0.039</td>
<td>0.022</td>
</tr>
<tr>
<td>+85°C</td>
<td>Max error (%) 6.632</td>
<td>3.800</td>
<td>0.918</td>
</tr>
<tr>
<td></td>
<td>Avg error (%) 5.769</td>
<td>3.498</td>
<td>0.421</td>
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</tbody>
</table>

If a temperature sensor is available for use in the system, different correction factors can be used to calibrate the data at different temperatures, but with increased component and manufacturing cost. Table 3 shows the maximum relative error and average error when using different correction factors at –40°C, +25°C, and +85°C.

**Table 3. Maximum and Average Error at Different Temperatures Using Different Correction Factors**

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<th>AD8217</th>
</tr>
</thead>
<tbody>
<tr>
<td>–40°C</td>
<td>Max error (%) 1.981</td>
<td>0.022</td>
<td>0.114</td>
</tr>
<tr>
<td></td>
<td>Avg error (%) 0.303</td>
<td>0.009</td>
<td>0.023</td>
</tr>
<tr>
<td>+25°C</td>
<td>Max error (%) 1.806</td>
<td>0.075</td>
<td>0.103</td>
</tr>
<tr>
<td></td>
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<td>0.039</td>
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</tr>
<tr>
<td>+85°C</td>
<td>Max error (%) 1.844</td>
<td>0.038</td>
<td>0.075</td>
</tr>
<tr>
<td></td>
<td>Avg error (%) 0.241</td>
<td>0.013</td>
<td>0.020</td>
</tr>
</tbody>
</table>

The temperature experiments show the high accuracy available over wide temperature ranges with devices using auto-zero technology, especially in the case of the AD8217.
Conclusion

The test results show that all three solutions can be used for wide dynamic range high-side current sensing: the outputs are linear in all three solutions, while the solution using the AD8217 leads to the best error performance without the necessity of a standalone power supply. The ±100-nV/°C offset drift characteristic also makes it ideal for the most accurate performance throughout the –40°C to +125°C temperature range. From a system design perspective, the single-chip solution can save PCB area, ease PCB layout, decrease system cost, and improve reliability. These findings apply particularly to unidirectional current sensing applications, where the load-current range is wide and dynamic range is critical.

Based on these test results, the AD8217 solution is the most suitable of the three choices for wide dynamic range unidirectional high-side current sensing and monitoring. We also note that an AD8210 solution offers operation right down to 0-V inputs—which could be beneficial for short-to-ground conditions. Note also that the AD8210 is capable of single-chip monitoring of bipolar current flow, as in charge/discharge applications.

Calibration and temperature sensing are recommended in practical system designs requiring best system performance.

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