



# Analog Dialogue

A forum for the exchange of circuits, systems, and software for real-world signal processing

Volume 38, Number 2, 2004



**Bridge-Type Sensor Measurements ... see page 6**

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## Editors' Notes

These words are being written in the second half of the thirty-eighth year of *Analog Dialogue's* existence in print (Volume 38). That number, 38, is a relatively undistinguished number: it's not a prime number—and even expressed as the product of two primes,  $2 \times 19$ , it is still not particularly interesting. One could perhaps stir up a little interest by noting that it's also  $2 \times (20-1)$  or  $2 \times (2 \times 2 \times 5 - 1)$ —or  $2 \times (2^2(2^2+1)-1)$ . But somehow, round numbers seem more interesting historically. Take, for example, 35 ( $=5^2+(5 \times 2)$ ), which is the number of years of service the undersigned will soon complete as the editor of this publication. For such occasions, in the same way as for alumni reunions, we celebrate only multiples of five years, while the in-between years, though equally long, go unrecognized (yet they inexorably accumulate with the passage of time).



Much more interesting numerically, our editorial colleague, Scott Wayne, also celebrating a quintennium, has been at Analog Devices for a quarter century, 25 ( $=10^2/2^2$ ) years. His personal reflections about that passage appear below.

In 1969 (Volume 3), the fourth year of existence for Analog Devices, our product catalog overwhelmingly comprised high-performance modular discretely assembled operational amplifiers; later augmented by our first data-converter family—inherited in our acquisition of Pastoriza Electronics, Inc. during that year. Five years later (Volume 8, 1974), solidly entrenched in the semiconductor op amp and analog-circuits business—and having become the world's recognized leader in data converters—we introduced the AD7520, the first monolithic CMOS 10-bit DAC.

In 1979 (Volume 13), we acquired Computer Labs, in Greensboro, NC—extending our converter leadership into the high-speed market. (That year we also introduced a “box”—the ill-fated MACSYM integrated measurement-and-control system—which even had its own language—but within a few years, it had the misfortune to find itself in competition with products using the PC!). By 1984 (Volume 18), we had already introduced our first DSP ICs and were rapidly consolidating our lead in conversion ICs.

1989 (Volume 23) was the year in which we attained the across-the-board high ground in high speed, featuring the fast-slewing-and-settling AD843 and AD844 op amps, the 120-MHz AD640 log amp, 300-MSPS AD9028 8-bit flash ADC, and the 500-MHz AD834 analog multiplier. In 1993 (Volume 28), our highly successful SHARC®, entered the floating-point DSP fray, and ADI pushed the door open wider in high-speed RF analog with devices such as the 500-MHz AD831 mixer.

## Analog Dialogue

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*Analog Dialogue* is the free technical magazine of Analog Devices, Inc., published continuously for 38 years—starting in 1967. It discusses products, applications, technology, and techniques for analog, digital, and mixed-signal processing. It is currently published in two editions—*online*, monthly at the above URL, and quarterly *in print*, as periodic retrospective collections of articles that have appeared online. In addition to technical articles, the online edition has timely announcements, linking to data sheets of newly released and pre-release products, and “Potpourri”—a universe of links to important and rapidly proliferating sources of relevant information and activity on the Analog Devices website and elsewhere. The *Analog Dialogue* site is, in effect, a “high-pass-filtered” point of entry to the [www.analog.com](http://www.analog.com) site—the virtual world of *Analog Devices*. In addition to all its current information, the *Analog Dialogue* site has archives with all recent editions, starting from Volume 29, Number 2 (1995), plus three special anniversary issues, containing useful articles extracted from earlier editions, going all the way back to Volume 1, Number 1.

If you wish to subscribe to—or receive copies of—the print edition, please go to [www.analog.com/analogdialogue](http://www.analog.com/analogdialogue) and click on <subscribe>. Your comments are always welcome; please send messages to [dialogue.editor@analog.com](mailto:dialogue.editor@analog.com) or to these individuals: Dan Sheingold, Editor [[dan.sheingold@analog.com](mailto:dan.sheingold@analog.com)] or Scott Wayne, Managing Editor and Publisher [[scott.wayne@analog.com](mailto:scott.wayne@analog.com)].

And just five years ago (1999, Volume 33), the communications world was revolutionized by the Othello direct-conversion radio chip set. Also in that year, we initiated joint publication of *Analog Dialogue*, online and in print, achieving at last the benefits of timely publication, and direct contact with our readers, while retaining the relative permanence of print.

Dan Sheingold [[dan.sheingold@analog.com](mailto:dan.sheingold@analog.com)]

### 25 YEARS AT ADI

Dan Sheingold and I recently celebrated our respective 35<sup>th</sup> and 25<sup>th</sup> anniversaries at Analog Devices. At our service award dinner we learned that there were approximately 200 Massachusetts-based ADI employees who were also celebrating 15-, 20-, 25-, 30-, or 35-year anniversaries in 2004. This got me to thinking: If this were a typical year, at least 1,000 employees, or about 1/3 of ADI's Massachusetts work force, would have more than 10 years' tenure. How did this compare to other companies?



Consulting the US Department of Labor website, I was surprised to learn that this was not unusual. While the median length of time that employees have been with their current employer is only 3.7 years, fully 31% of workers have 10 or more years under their belt.

So, what makes an employee and a company want continue their association for such as long time? For me, it was a combination of such things as great co-workers, opportunities to do new things, constant challenges, continuous learning, and solid corporate management.

Trained as an electrical engineer, my first 20 career years were spent in design, starting with high-resolution converters in 2" × 2" modules. As customer demands changed and our technology evolved, I worked on our division's earliest hybrid and compound-monolithic converter designs, later branching out into signal-conditioning and isolation products. I then transferred to the semiconductor division, working on bipolar and CMOS IC designs.

Five years ago I joined corporate marketing, where I am now doing teaching, editing, and publishing. Rather than feeling trapped in a single job for 25 years, I have progressed through a series of jobs, continually learning new things, encountering new challenges, and enjoying the benefits of closer contact with customers and a view of our products and fellow engineers with an extended horizon.

The biggest factor that many people cite as contributing to their longevity with a company is the *people*. (It may also be why people leave some companies.) Throughout my career at Analog, I have been privileged to work with some of the best, brightest, and nicest people that you could hope to meet—people who seem more like friends or extended family than simply co-workers.

An additional way that companies motivate employees to stick around (and do their best) is by empowering them to make the decisions that they need to get their jobs done on a day-to-day basis with a minimum of oversight or bureaucracy.

There are, of course, more tangible considerations, such as salary, vacations, and other concrete benefits. They must be competitive with those of other employers. Although long-timers feel that they can get more spiritual satisfaction by staying with a company, yet others feel that they can get more material benefit (or eliminate sources of dissatisfaction) by leaving to join a new company.

We'll welcome your feedback on what motivates people to stay with companies—or jump around—and to what degree they have that choice. In the newspapers, we seem to read only the horror stories—but perhaps that's what sells papers!

Scott Wayne [[scott.wayne@analog.com](mailto:scott.wayne@analog.com)]

# Current Measurement in Solenoids for Automotive Control Systems

By Scott Beversdorf [scott.beversdorf@analog.com]  
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## Solenoids in Automotive Control Applications

A solenoid is a linear motor with a fixed range of travel. Solenoids may be designed for simple on-off applications, acting much like relays. For example, they are used this way in *starters* and *door locks*.

On the other hand, a *linear*, or *proportional*, solenoid is one whose position can be controlled in a precise manner. They are used to operate pistons and valves for accurate control of fluid pressure or flow in applications such as *transmissions* and *fuel injection*.

Transmissions require accurate and smooth control of pressure on clutches to change gears, and for controlling the locking torque converter. Electronically controlled transmissions may contain more than eight linear solenoids, all of which require smooth, accurate control. Common-rail diesel fuel-injection applications, with pressures in excess of 2000 psi, may require one linear solenoid per cylinder—and one at the fuel pump—to adjust pressure accurately to maintain predictable injector fuel flow.

## Example: Electronic Transmission Control

The automatic transmission is one system in which electronic control is largely supplanting mechanical control because of improvements in drive quality and fuel efficiency. Previous improvements in fuel efficiency and acceleration came with the introduction of the locking torque converter. More recently, a combination of software and hardware using electronically controlled solenoids allowed easier adjustment of the shift algorithms, and provided additional benefits in transmission-shift smoothness and quality.

Overall, electronic control of the transmission allows for a simpler, more reliable, and less costly electromechanical system. Electronic transmission control systems improve the control of transmission shift points, with less abrupt gear shifting and improved shift smoothness. In addition, the flexibility of the electronic control allows for better adaptability to changing conditions. Electronic control of shift points with finer resolution allows better acceleration, improved economy, better load control, and reduced emissions, with minimal effort by the driver. In addition, the electronic control allows the transmission to shift more smoothly with varying load and acceleration.

With an electronic control system it is possible to affect the shift-control algorithm by a variety of inputs in addition to shaft speed, vacuum, and driver input. Some of these parameters include spark advance, injector parameters, input speed sensors, shift selection by wire, engine speed, throttle position, torque-converter speed/lock, ATF temperature, engine temperature, wheel-slip sensors, and inertial sensors. Combining these kinds of inputs allows a wide variety of shift optimization points, adapted to the overall operating conditions. To use these inputs most effectively, it is necessary to have a system benefiting by precise and infinitely adjustable electronic control of the shift points and shift speed.

Hydraulic control is still used to change gears in the electronically controlled automatic transmission. In contrast to the mechanical system (see Appendix), electronic control of the hydraulics in the electromechanical system is executed by linear solenoids that vary the hydraulic pressure applied to the actuators attached to the clutch packs. In order for this to work, it is extremely important to have accurate and repeatable control of the solenoid opening—which in turn allows for accurate, repeatable control of the shift points through the application of precise amounts of hydraulic fluid.

## Determining Solenoid Position

The *linear solenoid's* position is controlled in a feedback loop. For example, a valve's downstream pressure can be monitored and used as a feedback signal to compare with the setpoint, adjusting the pulse-width modulation (PWM) duty cycle to control the solenoid. However, it may be difficult, impractical, or very costly to measure the downstream pressure.

A practical alternative is to establish the position of the solenoid by measuring the current through the solenoid. This is possible because the force imposed by the mechanical load on a solenoid is directly proportional to the magnetic field, which, in turn, is directly proportional to the current through the coil. Proportional control of the solenoid is achieved by a balance of the forces between the spring-type load and the solenoid's magnetic field, which can be determined by measuring the current through the solenoid.

## PWM Solenoid Control

The solenoid is powered by using a microcontroller-generated pulse-width modulated input signal to rapidly open and close a FET switch in series with the solenoid and a voltage source (the car's battery). The average voltage is determined by the ratio of the waveform's *on* time to the pulse period. Changes in the pulse width and the solenoid's mechanical load cause the average current flowing through the solenoid to change. The average current is indicative of the amount of solenoid movement, and thus, fluid pressure and flow.

The relationship between solenoid movement and average current for a particular PWM waveform is established through *characterization*. While it is true that the magnetic force directly relates to the current through the solenoid, the actual mechanical force and movement are not so closely correlated, since they depend on the construction of the solenoid and the nature of the load. So, characterization is required to correlate the average current to the solenoid opening.

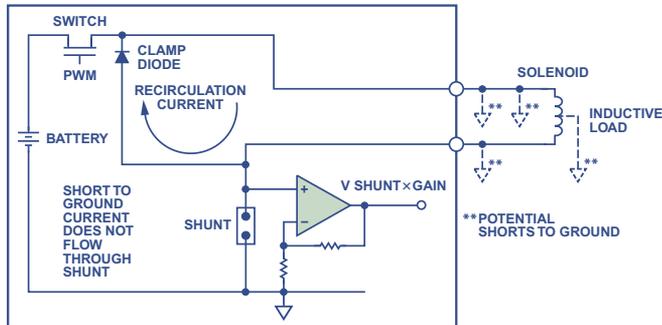
For example, the PWM ratio must be increased when the solenoid is first energized to overcome static friction. Once static friction is overcome, a different PWM relationship is used to move it in and out.

## Measuring the Current Through the Coil

The current is thus an important indication of the solenoid's state. The most effective method of measuring the solenoid current is to measure the voltage across a resistive shunt connected in series with the solenoid, the battery, and the switch. There are several different ways to configure this series circuit for switching and voltage measurement.

### Low-Side Current Sense with High-Side Drive

The circuit in Figure 1 shows a switch, connected to the high (ungrounded) side of the battery, in series with the solenoid coil and the grounded resistive shunt. A reversed diode is connected across the coil to clamp (i.e., short-circuit) the inductive voltage generated by the coil when the current is turned *off*. Using a ground reference for the shunt allows an inexpensive op amp—with indifferent common-mode specifications—to be used in the electronic control unit (ECU) to measure the voltage across the shunt.



**HIGH-SIDE DRIVE/LOW-SIDE SENSE**  
 • LOWER COST AMPLIFIER  
 • HIGHER COST SWITCH  
 • HIGHER COST DRIVE FOR SWITCH  
 • SHORT-CIRCUIT CURRENT DOES NOT FLOW THROUGH SHUNT WHICH DOES NOT ALLOW FOR DIAGNOSTICS OF SHORT TO GROUND (SHOWN OUTSIDE THE ECU), OR SOLENOID CONDITION POTENTIALLY DAMAGING WIRING AND THE SWITCH.

Figure 1. Electronic control unit with high-side switching and low-side sensing.

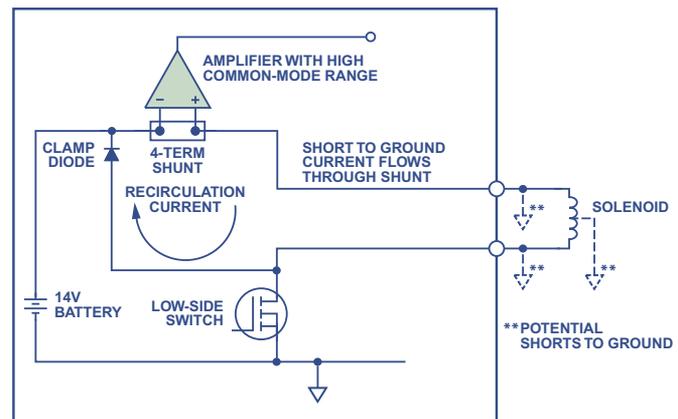
When considering this approach, the designer must take into account these disadvantages:

1. The solenoid recirculating current is not included in the measurement, so this circuit provides an inaccurate measure of the average current through the solenoid coil. The solenoid recirculation current can be an aid to detecting solenoid health; if some of the windings have begun to short, the condition can be seen by measuring the passively controlled recirculation current.
2. Because the switch is on the high side, it is more expensive to buy and to drive. The PWM drive requires careful level translation between the logic output of the microcontroller and the gate.
3. Additional circuitry is needed to detect shorts to ground because the short-circuit currents do not flow through the shunt. Damage to the wiring and the FET can occur if shorts to ground (Figure 1) are not detected.
4. The measurement may be unstable because, in practice, *ground* is not the *ideal* universal connection, graphically rendered as a little inverted triangle. In the actual application, “ground” may not in fact be *ground*. Voltage drops caused by currents between the op-amp’s ground and the shunt ground can cause significant errors.

### High-Side Current Sensing with a Low-Side Switch

A better method of driving the solenoid is to use a switch that is referenced to ground (Figure 2), allowing the use of a less expensive, low-side switch.

Significantly better diagnostics are possible because the solenoid recirculation current is included in the measurement. In addition, the drive is less expensive, because it is not necessary to perform level translation to the gate.



**LOW-SIDE DRIVE/HIGH-SIDE SENSE**  
 • LOWER COST SWITCH  
 • LOWER COST DRIVE FOR SWITCH BECAUSE THE GATE IS GROUND REFERENCED  
 • NEED TO USE A DIFFERENTIAL AMPLIFIER WITH COMMON-MODE RANGE THAT EXTENDS ABOVE THE BATTERY  
 • ALLOWS DIAGNOSTICS AND PROTECTION FOR SHORTS TO GROUND

Figure 2. Electronic control unit with low-side switching and high-side sensing.

However, the amplifier must have high, common-mode rejection, and it must be able to stand off significant common-mode voltage (CMV). The voltage level at the shunt in this example varies from battery voltage to battery-voltage-plus-a-diode-drop. Here’s the explanation: When the switch is closed, the common-mode voltage level at the shunt remains at the low impedance battery voltage. When the switch is opened, the voltage across the solenoid reverses because of the solenoid’s inductance, and causes the common-mode voltage level to include the clamping diode drop—while the transient current flows—before settling to the battery voltage.

An important benefit of this drive method is that it allows detection of shorts to ground, because the high-side current flows through the shunt—as seen in Figure 2.

A key concern when using this current sensing method is that the high side of the battery is always connected to the solenoid. This could allow the solenoid to switch unexpectedly if there is an intermittent short to ground. Also, the continuous presence of the voltage at the solenoid could cause excessive corrosion over time.

### High-Side Current Sensing with a High-Side Switch

A configuration that minimizes the possibility of unexpected solenoid activation and excessive corrosion is shown in Figure 3, where both the switch and the shunt are connected on the high side. This removes the battery voltage from the solenoid when the switch is *off*, preventing damage from potential shorts to ground, and allows the recirculating current to be included in the measurement. The battery voltage is removed from the load when the switch is open, so corrosive effects caused by the voltage differential are eliminated.

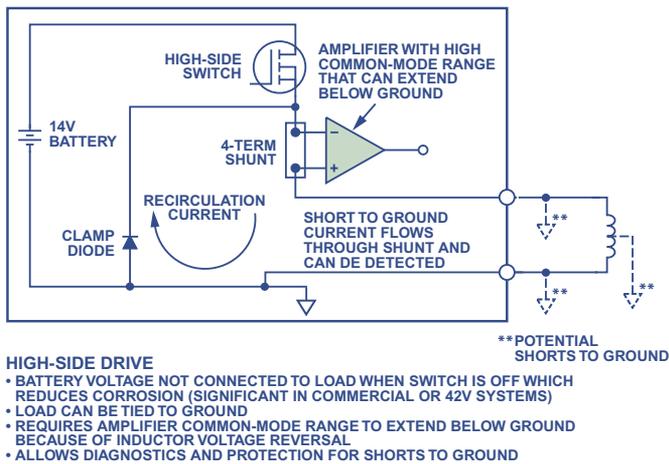


Figure 3. Electronic control unit with low-side solenoid and high-side switching and sensing.

In this case, however, the voltage reversal across the solenoid when the switch is opened will cause a much wider swing of common-mode voltage, from battery high-side voltage to one diode drop below ground (the reverse voltage is limited by the clamping diode). The amplifier used in this application must thus be able to provide accurate measurement of the shunt voltage (current), ignoring the large, rapid swing of common-mode voltage when the switch goes off.

As with the low-side switching, high-side sensing configuration (Figure 2), it is possible to measure shorts to ground, because all of the solenoid current from the high side flows through the shunt, as seen in Figure 3.

### A Simple High-Side Current Measurement Circuit

Fortunately, the AD8200 single-supply difference amplifier, with all the desirable properties for this application—*complete in a single IC package*—is available from Analog Devices. Figure 4 shows an example of the AD8200 applied in an ECU to measure current on the high side in this type of application. Here, the AD8200 is used to amplify and filter the small, differential voltage from the shunt, while rejecting the large common-mode swings mentioned above. The AD8200 can be used in any of the configurations shown earlier.

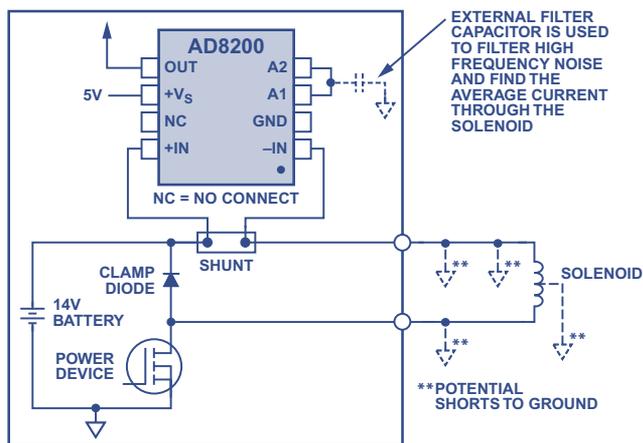


Figure 4. Electronic control unit, using the AD8200, with low-side solenoid and high-side switching and sensing.

The AD8200 uses a single +5 V supply and has an input common-mode voltage range extending from  $-2\text{ V}$  to  $+24\text{ V}$ , with load dump to  $+44\text{ V}$ . If higher common-mode range is required, use of another member of the AD8200 family is recommended—for

example, the AD8205, with a CMV range of  $-2\text{ V}$  to  $+65\text{ V}$  and a gain of 50; or the AD8206 with a CMV range of  $-2\text{ V}$  to  $+65\text{ V}$  and a gain of 20 (the same gain as the AD8200).

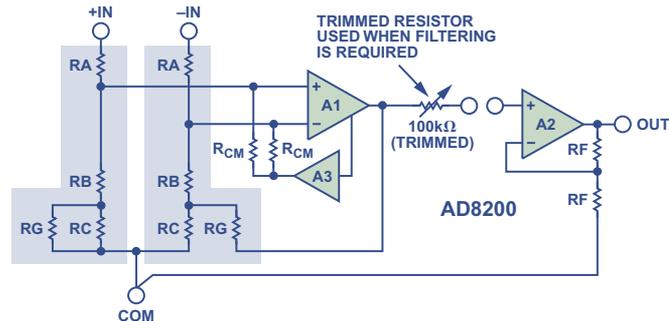


Figure 5. Functional block diagram of the AD8200.

Figure 5 is a functional block diagram of the internal wiring of the AD8200. Before designing a differential amplifier using an inexpensive op-amp and some external resistors, consider that achieving the performance required to measure solenoid current accurately enough for a control application requires a circuit built with resistors that are precisely matched to within 0.01%. The AD8200 has internal laser-trimmed resistors that allow for this level of precise operation when handling both ac and dc voltages. Typical offset- and gain drift in the SOIC package are  $6\text{ }\mu\text{V}/^\circ\text{C}$  and  $10\text{ ppm}/^\circ\text{C}$ , respectively. The device also provides a minimum, common-mode rejection of 80 dB from dc to 10 kHz.

Besides its availability in the SOIC package, the AD8200 is also available in die form. Both package options are specified over wide temperature ranges, making the AD8200 well suited for use in many automotive and industrial platforms. The SOIC package is specified for  $-40^\circ\text{C}$  to  $+125^\circ\text{C}$  and the die is specified from  $-40^\circ\text{C}$  to  $+150^\circ\text{C}$ .

The AD8200 also features an externally accessible 100-k $\Omega$  resistor at the output of the preamplifier, which can be used with external capacitance for low-pass filter applications, and with external resistors for establishing gains other than the preset gain of 20.

## APPENDIX

### Mechanical Transmission Control

The older method of controlling transmission shift points involved complex, speed-dependent hydraulic circuits. Shifting was done through the use of changes in hydraulic pressure in a complex valve body. The hydraulic pressure was modulated by a governor attached to the output shaft. Centrifugal force moved the governor, releasing transmission oil and increasing the pressure in the valve body. The method of adapting to changing driving conditions typically consisted of forcing the transmission to downshift under heavy acceleration or loading with mechanical actuators.

When the driver needed more acceleration, the requirement was typically transmitted via a down-shift mechanism, consisting of a rod running from the throttle control in the engine compartment to the side of the transmission. The rod moved a lever, closing off a series of channels in the throttle body. This forced the transmission to downshift under heavy acceleration until, at speed, the governor overrode the down-shift mechanism.

Adapting to changes in load was accomplished using a vacuum modulator. With increasing load on the engine, the vacuum change caused a rod to slide in or out of the valve body, changing the transmission shift point and shift speed. While these methods of controlling shift points and shift smoothness worked, there was little that could be done to adjust these parameters to more variable conditions because of the characteristics of the mechanical system used to control them.

# Bridge-Type Sensor Measurements Are Enhanced by Autozeroed Instrumentation Amplifiers with Digitally Programmable Gain and Output Offset

By Reza Moghimi [reza.moghimi@analog.com]

Resistance-based bridge circuits are widely used to provide electrical outputs from sensors measuring physical variables, such as pressure and force. These outputs are usually small—and thus require amplification to bring them up to levels required by A/D converters in measurement-and-control systems. This article describes a versatile new IC instrumentation amplifier that has very low drift and many useful features, and discusses its role in solving problems of bridge instrumentation, using pressure measurement as an example.

## Pressure Sensing

Figure 1 illustrates the functionality of a typical electrical output pressure instrument. Pressure is sensed by the motion of mechanical elements, such as diaphragms, Bourdon tubes, bellows, and capsules, all of which deflect when pressure is applied. This deflection causes a change in the resistances of a strain gage.

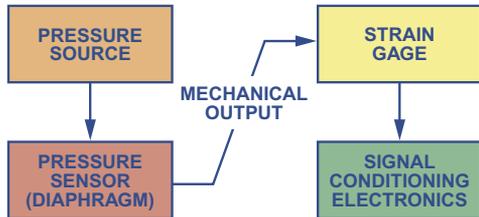


Figure 1. Converting pressure to an electrical signal.

The most popular pressure sensors use strain gages in a Wheatstone resistance-bridge configuration in which all four elements are variable—as shown in Figure 2D—thus providing optimal linearity and sensitivity. When pressure is applied to the diaphragm, two gage elements of the bridge are subjected to tension; the other two elements are subjected to compression. The corresponding changes in resistance are a measure of the incident pressure. The bridge is excited by a constant voltage or current, producing an electrical signal.

In one form of *gauge-pressure sensor* technology, the strain-gage elements are bonded to a metal diaphragm and produce a full-scale resistance variation, typically of the order of 0.1% of the base resistance. When a constant voltage or current is applied to the bridge, the change in the resistance over that range produces a linear unbalance, which is measured as a differential voltage (or current). In another technology, *semiconductor strain gages* are bonded into a silicon diaphragm, and can generate a much larger response—in many cases about 1% of the base resistance.

A good detailed explanation of strain gages and various bridge configurations can be found in the Analog Devices *Sensor Signal Conditioning* manual.<sup>1</sup>

## Signal Conditioning

The signal produced by the bridge is generally small and is subject to noise, offset, and gain errors. Before the bridge output can be digitized, it must be amplified, with offset adjusted to match the span of the A/D converter, and filtered, to remove noise. Although the signal conditioning block shown in Figure 3 could be built with operational amplifiers and discrete circuit elements, instrumentation amplifiers have proved to save on parts cost, circuit-board area, and engineering design time.

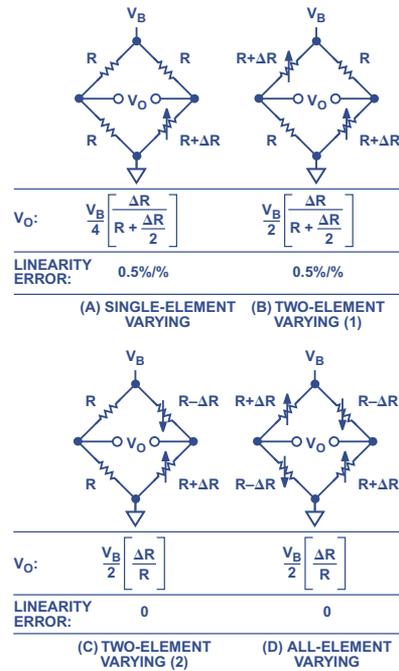


Figure 2. Bridge with four resistive elements.

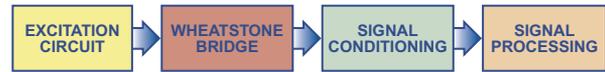


Figure 3. Pressure measuring instrumentation.

In typical pressure sensor applications, a resistive bridge outputs a differential signal—with a span of tens or hundreds of millivolts—that is proportional to the applied pressure and the excitation voltage applied to the bridge. The Honeywell 26PC01SMT series microstructure pressure sensor, for example, has a  $\pm 1.0$ -psi full-scale span. With 5 V applied, it would have a zero-pressure null offset of  $\pm 2$  mV, a full-scale output span anywhere in the range of  $\pm 14.7$  mV to  $\pm 18.7$  mV, and a 2.5-V common-mode level. In order to accurately resolve this small differential output voltage in the presence of the high common-mode voltage, an instrumentation amplifier's ability to reject common-mode signal is essential. For example, 12-bit readout resolution calls for an LSB of less than  $10 \mu\text{V}$  ( $35 \text{ mV}/4096$ ), or about 101 dB below the common-mode level.

## Manual Bridge Compensation

Wheatstone bridges in pressure sensors are typically manually compensated to remove their offset and span errors. This requires the instrumentation manufacturing process to include steps to trim the offset, offset temperature drift, span, and span temperature drift. Figure 4 illustrates a bridge with resistors added to compensate for these errors. Trimming out these errors is time consuming and expensive. Alternatively, the offset can be adjusted by applying a programmable dc voltage from a D/A converter to the reference pin of an instrumentation amplifier. Offset correction is required since the offset would otherwise reduce the available dynamic range of the ADC.

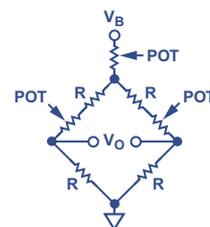


Figure 4. Manual bridge compensation.

Gain uncertainties in pressure sensors make gain adjustment a requirement in most instrumentation amplifier-based systems. This was traditionally done by adding a trim potentiometer in series with the external gain resistor of the instrumentation amplifier. To achieve higher levels of performance over a wider temperature range, system designers turned to software-controlled gain compensation.

### Instrumentation Amplifier Errors

Figure 5 shows some typical error sources that occur when a sensor is interfaced to a signal conditioning circuit. The bridge amplifier bias current flows through the output resistance of the bridge. Any unbalance in the resistance or bias current produces an offset error. This error, multiplied by the gain, appears at the output. In addition, the offset voltage and bias current are functions of temperature. Other important errors to consider are the amplifier's gain accuracy, nonlinearity, and noise. A high-performance amplifier with low input offset voltage, bias current, offset-voltage TC, and bias-current TC is required for bridge-measurement applications.<sup>2</sup>

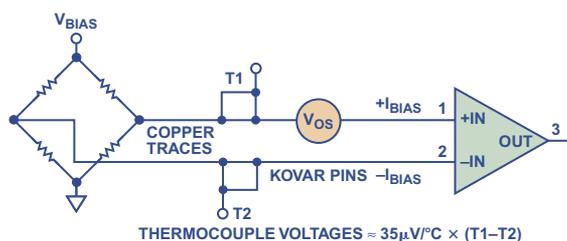


Figure 5. Amplifier error sources.

### The AD8555 to the Rescue

Measurement system designs that use bridge circuits, such as pressure sensors, will benefit by the AD8555, a new zero-drift digitally programmable amplifier from Analog Devices.<sup>3</sup> Its instrumentation amplifier configuration employs three autozeroed amplifiers (A1, A2, A3), as shown in Figure 6.

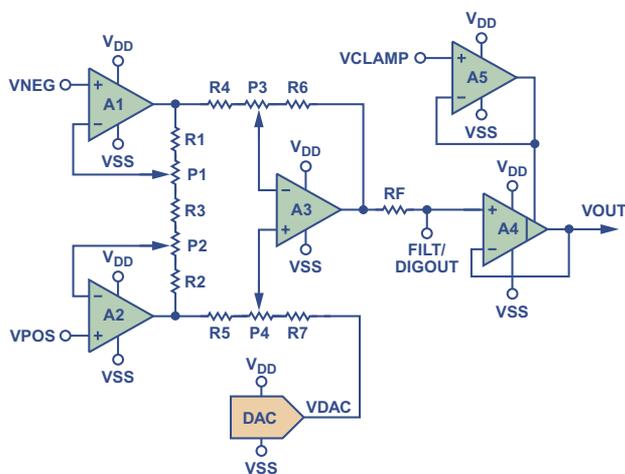


Figure 6. Functional block diagram of the AD8555 instrumentation amplifier.

To avoid loading the sensor's bridge, the differential inputs feature high impedance and low bias current at both terminals ( $V_{POS}$  and  $V_{NEG}$ ). Autozero techniques minimize offset and offset drift by continuously correcting for amplifier-generated dc errors. This results in a 10- $\mu$ V maximum input offset voltage over a temperature range of  $-40^{\circ}\text{C}$  to  $+125^{\circ}\text{C}$ , with maximum drift of just 65 nV/ $^{\circ}\text{C}$ .

Gain, ranging from 70 to 1280, is programmed in steps of less than 1 (with better than 0.4% resolution)—via a single-wire serial interface—by adjusting the gains of the two stages individually. In the DigiTrim<sup>®</sup> process, the gain setting is locked in place by blowing polysilicon fuses.<sup>4</sup> The first-stage gain is trimmed from 4.00 to 6.40 in 128 steps by a 7-bit-code that adjusts both P1 and P2; and the second-stage gain is set from 17.5 to 200 by an 8-step, 3-bit-code that adjusts P3 and P4. The adjustment values can be temporarily programmed, evaluated, and readjusted for optimum calibration accuracy—before the settings are permanently fixed.

The AD8555 also has an 8-bit DAC-programmable offset, which can be used to compensate for offset errors in the input signal and/or add a fixed bias to the output signal. This bias is used, for example, to handle bipolar differential signals in a single-supply environment. The output offset voltage may be set with a resolution of 0.39% of the voltage difference between the supply rails, ( $V_{DD}-V_{SS}$ ).

$$V_{out} = (Gain) \times (V_{diff}) + V_{offset}$$

and

$$V_{offset} \approx \left( \frac{N_{DAC-code} + 0.5}{256} \right) \times (V_{DD} - V_{SS}) + V_{SS}$$

where

$$Gain = (1^{st}\text{-stage gain}) \times (2^{nd}\text{-stage gain})$$

$V_{diff}$  is the differential input voltage to be measured

$N_{DAC-code}$  is the numerical value of the DAC's input code.

Like the gain, the output offset can be temporarily programmed, evaluated, and readjusted; and then can be permanently set by blowing the fuse links.

Single-supply operation has become an increasingly desirable characteristic of modern sensor amplifiers. Many of today's data acquisition systems are powered from a single low voltage supply. The AD8555 operates from single-supply voltages from 2.7 V to 5.5 V. The output of amplifier A4 swings to within 7 mV of either supply rail.

Fault detection in the AD8555 amplifier protects against open, shorted, and floating inputs. Any of these conditions triggers a circuit that causes the output voltage to be clamped to the negative supply rail ( $V_{SS}$ ). Shorts and floating conditions are also detected on the  $V_{CLAMP}$  input. Using an external capacitor, the AD8555 can implement a low-pass filter to limit its dc-to-400-kHz output frequency range.

### The AD8555's Role in Sensor Bridge Signal Conditioning

Systems engineers would prefer that all pressure sensors with the same part number exhibit nearly identical performance. Typically, off-the-shelf sensors do not meet such requirements with adequate accuracy. One way to achieve consistency from sensor to sensor is extensive trimming during the manufacturing process. If the behavior of these sensors is repeatable over temperature, a better way may be to use this new generation of programmable amplifiers to provide compensation.

The AD8555, with its zero-drift instrumentation amplifier, can provide amplification, gain setting and trim, offset setting and trim, and clamping—all established digitally. It can be used to compensate for offset and gain errors in bridge-type sensors, as well as providing an indication of sensor malfunction. It enables

adjustments using software, making compensation using trim potentiometers an outdated art in a manufacturing environment. In the many cases where sensors are used in harsh and crowded environments, measurements benefit from the wide temperature range and space-saving package size of the AD8555, which is housed in a 4-mm-by-4-mm lead-frame chip-scale package (LFCSP). Because the AD8555 is capable of driving very large capacitive loads, it can be placed close to the sensor and at a distance from the signal processing circuitry. Its high levels of programming flexibility and dc accuracy distinguish it from all other solutions.

### Application Example

In a given batch of pressure sensors, sensor-to-sensor variations cause offset voltages of 20% of the sensor's full-scale output. Sensor-to-sensor gain can vary by almost 2 to 1. The following example shows how we can use the AD8555 to compensate for the offset and gain variations, while maximizing the dynamic range of the signal provided to the A/D converter. Figure 7 shows the application circuit, including the bridge, AD8555, and AD7476 A/D converter.

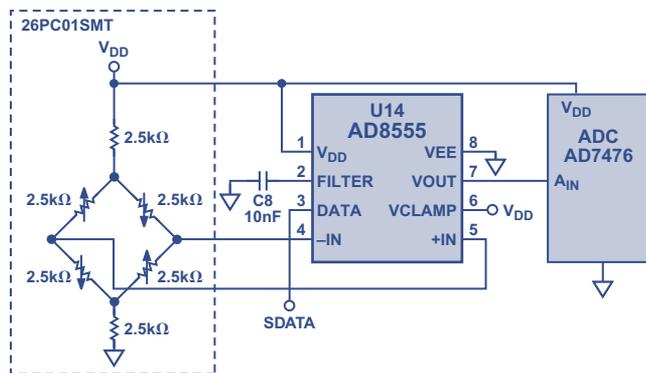


Figure 7. Pressure-sensing application example.

### Sensor characteristics

The 26PC01SMT (surface-mount technology) pressure sensor is a Wheatstone bridge-based, printed circuit board-mountable device, which together with the AD8555, would appear to provide complete programmable pressure measurement and conditioning in a small footprint. Based on data sheet specifications at 10 V and 25°C, these are the expected sensor characteristics with a 5-V source:

| Measurement Type                          | Gage, Vacuum Gage, Differential |
|---|---------------------------------|
| Pressure Range, psi                       | ±1.0                            |
| Input Resistance Range, kilohms           | 5.5 to 11.5                     |
| Output Resistance Range, kilohms          | 1.5 to 3.0                      |
| Output Voltage Span, mV, ±1 psi           | ±14.7 min, ±16.7 typ, ±18.7 max |
| Span Shift Over Temperature, %            | ±1.5 typ, ±4.5 max              |
| Null Offset, mV                           | -2 to +2                        |
| Null Shift, mV, 25°C to 0°C; 25°C to 50°C | ±1.0 max                        |
| Linearity Error, % of Span                | ±0.50 typ, ±1.75 max            |
| Repeatability and Hysteresis Errors, %    | 0.20 typ                        |
| Overpressure, lb/in <sup>2</sup>          | 20 max                          |

### Amplifier characteristics

The signal conditioning circuitry also introduces errors which appear as an offset. The error contributions of the AD8555 are listed in the following table.

| Parameter                  | Error   | PPM  |
|----------------------------|---|------|
| Input Offset Voltage       | $(2 \mu\text{V} + 2 \text{ mV}/150) \times 150$     | 458  |
| Input Offset Current       | $2500 \text{ ohm} \times 200 \text{ pA} \times 150$ | 15   |
| Gain Error                 | 0.5%  | 5000 |
| CMRR                       | 100 dB  | 750  |
| Gain Nonlinearity (1)      | 50 ppm  | 50   |
| 0.1–10 Hz 1/f Noise (2)    | $0.7 \mu\text{V p-p} \times 150$                    | 21   |
| Total Unadjusted Error     | ~7 bits   | 6294 |
| Total Adjusted Error (1+2) | ~14 bits  | 71   |

Conditions:  $R_{\text{bridge}} = 2500 \text{ ohm}$ , Full scale =  $\pm 16.7 \text{ mV}$ ,  $A_V = 150$ ,  $V_{\text{OFF}} = 2.5 \text{ V}$ ,  $V_{\text{OUT}} = 0 \text{ V to } 5 \text{ V}$

The table shows that the dominant error sources are static errors appearing at the input of the AD8555. These will be trimmed out along with the corresponding variations in the sensor. Errors caused by current noise, gain drift, and offset drift are minimal, and can be ignored. The remaining errors, which cannot be trimmed out, are noise and gain nonlinearity.

Because noise places a limit on how accurately a sensor's signal can be amplified, high resolution measurement of low-level signals requires a low noise, low drift amplifier. The AD8555 has a voltage noise density floor of  $32 \text{ nV}/\sqrt{\text{Hz}}$  at 1 kHz. The noise from dc to 10 Hz is  $0.7 \mu\text{V}$  peak-peak.

The bridge, AD8555, and A/D converter are all excited by a +5-V supply. The full-scale output span of the bridge will be anywhere in the range of  $\pm 14.7 \text{ mV}$  to  $\pm 18.7 \text{ mV}$ . Its offset will be between  $-2 \text{ mV}$  and  $+2 \text{ mV}$ . Matching the 5-V full-scale input span of the A/D converter requires gain settings from 134 to 170. With the offset set to 2.5 V, the amplifier output will range from 0 V to 5 V as the pressure varies between  $-1 \text{ psi}$  and  $+1 \text{ psi}$ .

First set the gain to 134—the minimum required by this sensor. With a 0-psi input, adjust the offset until the amplifier output is at 2.5 V. This compensates for the null offset of the sensor and the amplifier error terms. Apply a 1-psi input, and adjust the gain so that the amplifier output voltage is  $5 \text{ V} - 1 \text{ LSB}$ . The output offset is a function of the gain, so the offset and gain adjustments must be done iteratively. As an alternative, the required gain can be calculated after the output span is measured. The offset is adjusted after the gain has been set, and thus only needs to be adjusted once.

### CONCLUSION

The AD8555 simplifies the design of bridge measurement systems by integrating a zero-drift instrumentation amplifier with programmable gain, programmable output offset, fault detection, output clamping, and low-pass filtering, thus providing the complete signal conditioning path between a sensor and an A/D converter. An evaluation kit containing sample pieces and evaluation software and hardware, is available from Analog Devices. ▶

### REFERENCES

<sup>1</sup>“Bridge Circuits”: from *Practical Design Techniques for Sensor Signal Conditioning*, edited by Walt Kester, Norwood, MA: Analog Devices, Inc., 1999, Section 2. [http://www.analog.com/Analog\\_Root/static/pdf/amplifiersLinear/training/Sensor\\_sect2.pdf](http://www.analog.com/Analog_Root/static/pdf/amplifiersLinear/training/Sensor_sect2.pdf)

<sup>2</sup>Analog Bridge Wizard™: *A Tool for Design and Product Selection* <http://www.analog.com/Wizard/bridge/bridgeUserEntry/>

<sup>3</sup>AD8555 Description and Data sheet: <http://www.analog.com>. Enter <AD8555> in SEARCH box.

<sup>4</sup>Polysilicon fuses. One advantage of using polysilicon fuses is their reliable performance over temperature—a requirement in automotive applications, for example. Once the user is satisfied with the offset and gain settings, the trim circuit is locked out by blowing the *master fuse* to prevent any accidental re-trimming.

## Practical Techniques to Avoid Instability Due to Capacitive Loading

By Soufiane Bendaoud [soufiane.bendaoud@analog.com]  
Giampaolo Marino [giampaolo.marino@analog.com]

**Q:** ADI has published a lot of information on dealing with capacitive loading and other stability issues in books, such as the amplifier seminar series, in earlier issues of *Analog Dialogue*, and in some design tools. But, I need a refresher—NOW.

**A:** OK. Here goes!

Capacitive loads often give rise to problems, in part because they can reduce the output bandwidth and slew rate, but mainly because the phase lag they produce in the op amp’s feedback loop can cause instability. Although some capacitive loading is inevitable, amplifiers are often subjected to sufficient capacitive loading to cause overshoots, ringing, and even oscillation. The problem is especially severe when large capacitive loads, such as LCD panels or poorly terminated coaxial cables, must be driven—but unpleasant surprises in precision low-frequency and dc applications can result as well.

As will be seen, the op amp is most prone to instability when it is configured as a unity-gain follower, either because (a) there is no attenuation in the loop, or (b) large common-mode swings, though not substantially affecting accuracy of the signal gain, can modulate the loop gain into unstable regions.

The ability of an op amp to drive capacitive loads is affected by several factors:

1. the amplifier’s internal architecture (for example, output impedance, gain and phase margin, internal compensation circuitry)
2. the nature of the load impedance
3. attenuation and phase shift of the feedback circuit, including the effects of output loads, input impedances, and stray capacitances.

Among the parameters cited above, the amplifier output impedance, represented by the output resistance,  $R_O$ , is the one factor that most affects performance with capacitive loads. Ideally, an otherwise stable op amp with  $R_O = 0$  will drive any capacitive load without phase degradation.

To avoid sacrificing performance with light loads, most amplifiers are not heavily compensated internally for substantial capacitive loads, so external compensation techniques must be used to optimize those applications in which a large capacitive load at the output of the op amp must be handled. Typical applications include sample-and-hold amplifiers, peak detectors, and driving unterminated coaxial cables.

Capacitive loading, as shown in Figures 1 and 2, affects the *open-loop* gain in the same way, regardless of whether the active input is at the noninverting or the inverting terminal: the load capacitance,  $C_L$ , forms a pole with the open-loop output resistance,  $R_O$ . The loaded gain can be expressed as follows:

$$A_{loaded} = A \left( \frac{1}{1 + j \frac{f}{f_p}} \right), \text{ where } f_p = \frac{1}{2\pi R_O C_L}$$

and  $A$  is the unloaded open-loop gain of the amplifier.

The  $-20$  dB/decade slope and  $90^\circ$  lag contributed by the pole, added to the  $-20$  dB slope and  $90^\circ$  contributed by the amplifier (plus any other existing lags), results in an increase in the rate of closure (ROC) to a value of at least  $40$  dB per decade, which, in turn, causes instability.

This note discusses typical questions about the effects of capacitive loads on the performance of some amplifier circuits, and suggests techniques to solve the instability problems they raise.

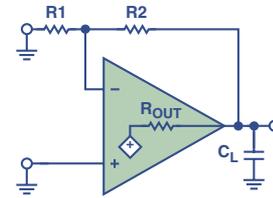


Figure 1. A simple op amp circuit with capacitive load.

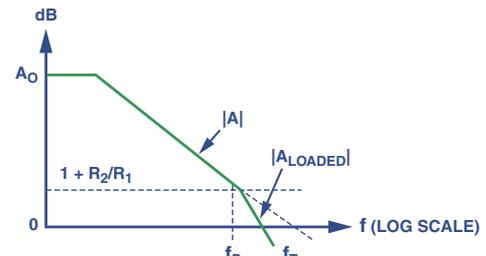


Figure 2. Bode plot for the circuit of Figure 1.

**Q:** So, different circuits call for different techniques?

**A:** Yes, absolutely! You’ll choose the compensation technique that best suits your design. Some examples are detailed below. For example, here’s a compensation technique that has the added benefit of filtering the op amp’s noise via an RC feedback circuit.

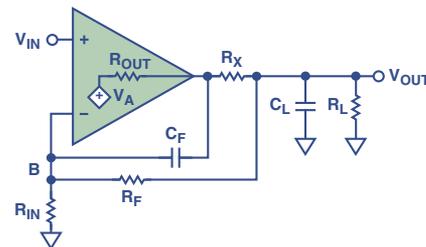


Figure 3. In-the-loop compensation circuit.

Figure 3 shows a commonly used compensation technique, often dubbed *in-the-loop* compensation. A small series resistor,  $R_x$ , is used to decouple the amplifier output from  $C_L$ ; and a small capacitor,  $C_f$ , inserted in the feedback loop, provides a high frequency bypass around  $C_L$ .

To better understand this technique, consider the redrawn feedback portion of the circuit shown in Figure 4.  $V_B$  is connected to the amplifier’s minus input.

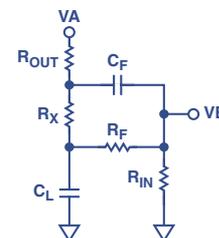


Figure 4. Feedback portion of the circuit.

Think of the capacitors,  $C_f$  and  $C_L$ , as open circuits at dc, and shorts at high frequencies. With this in mind, and referring to the circuit in Figure 4, let's apply this principle to one capacitor at a time.

**Case 1 (Figure 5a):**

With  $C_f$  shorted,  $R_x \ll R_f$ , and  $R_o \ll R_{in}$ , the pole and zero are functions of  $C_L$ ,  $R_o$ , and  $R_x$ .

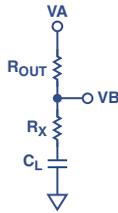


Figure 5a.  $C_f$  short-circuited.

Thus,

$$\text{Pole Frequency} = \frac{1}{2\pi(R_o + R_x)C_L}$$

and

$$\text{Zero Frequency} = \frac{1}{2\pi R_x C_L}$$

**Case 2. (Figure 5b)**

With  $C_L$  open, the pole and zero are a function of  $C_f$ .

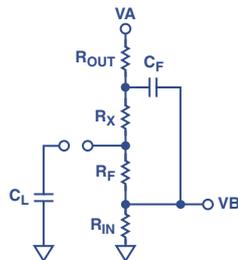


Figure 5b.  $C_L$  open-circuited.

Thus,

$$\text{Pole Frequency} = \frac{1}{2\pi \left[ (R_x + R_f) \parallel (R_o + R_{in}) C_f \right]}$$

$$\text{Zero Frequency} = \frac{1}{2\pi (R_x + R_f) C_f}$$

By equating the pole in Case 1 to the zero in Case 2, and the pole in Case 2 to the zero in Case 1, we derive the following two equations:

$$R_x = \frac{R_o R_{in}}{R_f} \text{ and}$$

$$C_f = \left( 1 + \frac{1}{|A_{cl}|} \right) \left( \frac{R_f + R_{in}}{R_f^2} \right) C_L R_o$$

The formula for  $C_f$  includes the term,  $A_{cl}$  (amplifier closed-loop gain,  $1 + R_f/R_{in}$ ). By experimenting, it was found that the  $1/A_{cl}$  term needed to be included in the formula for  $C_f$ . For the above circuit, these two equations alone will allow compensation for any op amp with any applied capacitive load.

Although this method helps prevent oscillation when heavy capacitive loads are used, it reduces the closed-loop circuit bandwidth drastically. The bandwidth is no longer determined by the op amp, but rather by the external components,  $C_f$  and  $R_f$ , producing a closed-loop bandwidth of:  $f_{-3\text{ dB}} = 1/(2\pi C_f R_f)$ .

A good, practical example of this compensation technique can be seen with the AD8510, an amplifier that can safely drive up to 200 pF while still preserving a 45° phase margin at unity-gain crossover. With the AD8510 in the circuit of Figure 3, configured for a gain of 10, with a 1-nF load capacitance at the output and a typical output impedance of 15 ohms, the values of  $R_x$  and  $C_f$ , computed using the above formulas, are 2 ohms and 2 pF. The square wave responses of Figures 6 and 7 show the fast response with uncompensated ringing, and the slower, but monotonic corrected response.

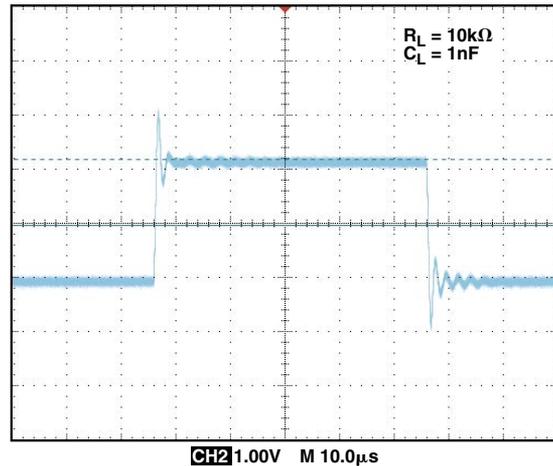


Figure 6. AD8510 output response without compensation.

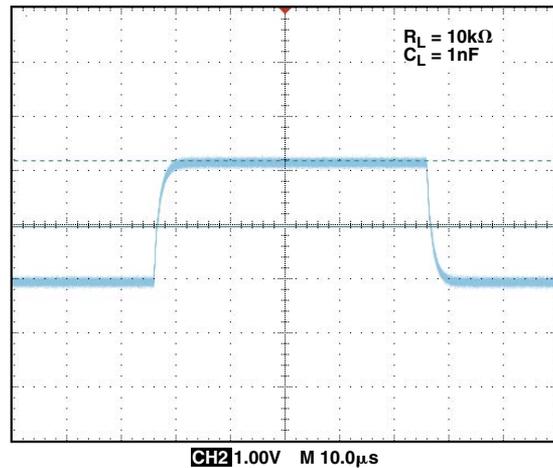


Figure 7. AD8510 output response with compensation.

In Figure 7, note that, because  $R_x$  is inside the feedback loop, its presence does not degrade the dc accuracy. However,  $R_x$  should always be kept suitably small to avoid excessive output swing reduction and slew-rate degradation.

**Caution:** The behaviors discussed here are typically experienced with the commonly used *voltage-feedback amplifiers*. Amplifiers that use *current feedback* require different treatment—beyond the scope of this discussion. If these techniques are used with a current feedback amplifier, the integration inherent in  $C_f$  will cause instability.

## Out-of-the-Loop Compensation

**Q:** Is there a simpler compensation scheme that uses fewer components?

**A:** Yes, the easiest way is to use a single external resistor in series with the output. This method is effective but costly in terms of performance (Figure 8).

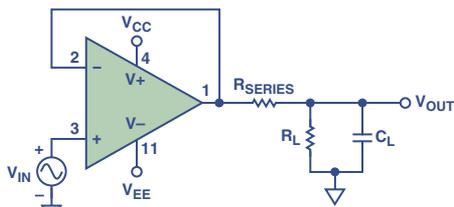


Figure 8. External  $R_{series}$  isolates the amplifier's feedback loop from the capacitive load.

Here a resistor,  $R_{series}$ , is placed between the output and the load. The primary function of this resistor is to isolate the op-amp output and feedback network from the capacitive load. Functionally, it introduces a zero in the transfer function of the feedback network, which reduces the loop phase shift at higher frequencies. To ensure a good level of stability, the value of  $R_{series}$  should be such that the zero added is at least a decade below the unity-gain crossover bandwidth of the amplifier circuit. The required amount of series resistance depends primarily on the output impedance of the amplifier used; values ranging from 5 ohms to 50 ohms are usually sufficient to prevent instability. Figure 9 shows the output response of the OP1177 with a 2-nF load and a 200-mV peak-peak signal at its positive input. Figure 10 shows the output response under the same conditions, but with a 50-ohm resistor in the signal path.

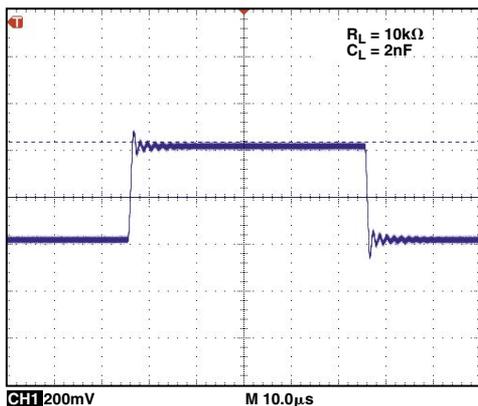


Figure 9. Output response of follower-connected OP1177 with capacitive load. Note high frequency ringing.

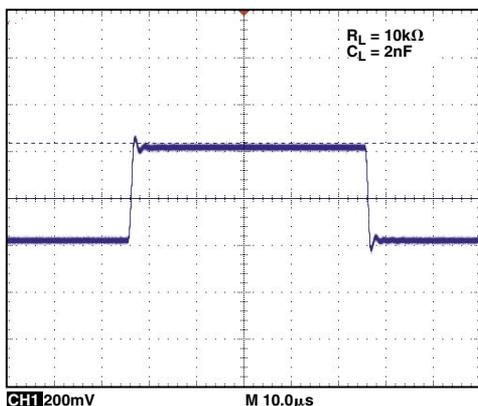


Figure 10. OP1177 output response with 50-ohm series resistance. Note reduced ringing.

The output signal will be attenuated by the ratio of the series resistance to the total resistance. This will require a wider amplifier output swing to attain full-scale load voltage. Nonlinear or variable loads will affect the shape and amplitude of the output signal.

## Snubber Network

**Q:** If I'm using a rail-to-rail amplifier, can you suggest a stabilizing method that will preserve my output swing and maintain gain accuracy?

**A:** Yes, with an R-C series circuit from output to ground, the snubber method is recommended for lower voltage applications, where the full output swing is needed (Figure 11).

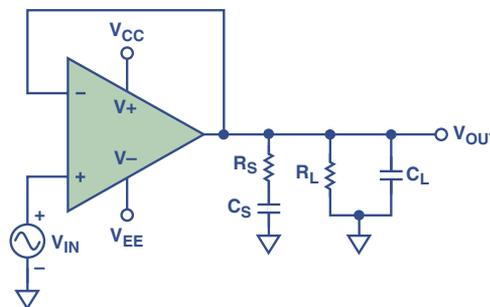


Figure 11. The  $R_S$ - $C_S$  load forms a snubber circuit to reduce the phase shift caused by  $C_L$ .

Depending on the capacitive load, application engineers usually adopt empirical methods to determine the correct values for  $R_S$  and  $C_S$ . The principle here is to resistively load down the output of the amplifier for frequencies in the vicinity at which peaking occurs—thus snubbing down the amplifier's gain, then use series capacitance to decrease the loading at lower frequencies. So, the procedure is to: check the amplifier's frequency response to determine the peaking frequency; then, experimentally apply values of resistive loading ( $R_S$ ) to reduce peaking to a satisfactory value; then, compute the value of  $C_S$  for a break frequency at about 1/3 the peak frequency. Thus,  $C_S = 3/(2\pi f_p R_S)$ , where  $f_p$  is the frequency at which peaking occurs.

These values can also be determined by trial and error while looking at the transient response (with capacitive loading) on an oscilloscope. The ideal values for  $R_S$  and  $C_S$  will yield minimum overshoot and undershoot. Figure 12 shows the output response of the AD8698 with a 68-nF load in response to a 400-mV signal at its positive input. The overshoot here is less than 25% without any external compensation. A simple snubber network reduces the overshoot to less than 10%, as seen in Figure 13. In this case,  $R_S$  and  $C_S$  are 30 ohms and 5 nF, respectively.

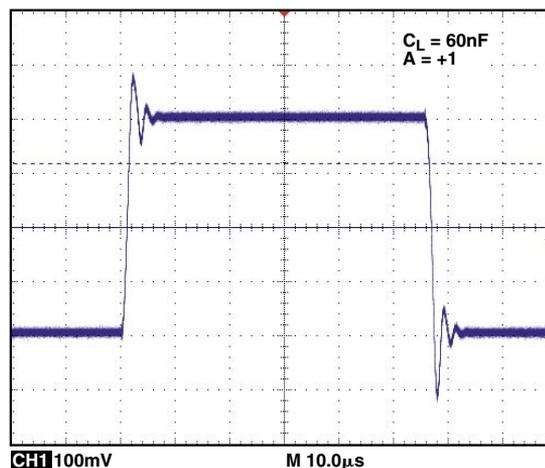


Figure 12. AD8698 output response without compensation.

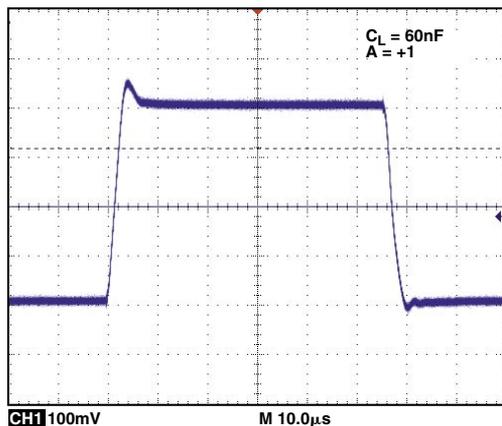


Figure 13. AD8698 output response with snubber network.

**Q:** OK. I understand these examples about dealing with capacitive loading on the amplifier output. Now, is capacitance at the input terminals also of concern?

**A:** Yes, capacitive loading at the inputs of an op amp can cause stability problems. We'll go through a few examples.

A very common and typical application is in current-to-voltage conversion when the op amp is used as a buffer/amplifier for a current-output DAC. The total capacitance at the input consists of the DAC output capacitance, the op amp input capacitance, and the stray wiring capacitance.

Another popular application in which significant capacitance may appear at the inputs of the op amp is in filter design. Some engineers may put a large capacitor across the inputs (often in series with a resistor) to prevent RF noise from propagating through the amplifier—overlooking the fact that this method can lead to severe ringing or even oscillation.

To better understand what is going on in a representative case, we analyze the circuit of Figure 14, unfolding the equivalent of its feedback circuit (input,  $V_{in}$ , grounded) to derive the feedback transfer function:

$$\frac{V_B}{V_A} (= \beta) = \frac{R_1}{(R_O + R_2)(1 + sR_1 C_1) + R_1}$$

which gives a pole located at

$$f_p = \frac{R_1 + R_2 + R_O}{2\pi R_1 C_1 (R_2 + R_O)}$$

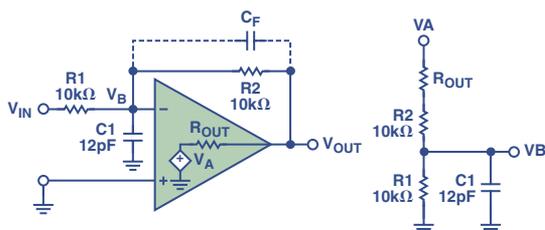


Figure 14. Capacitive loading at the input—inverting configuration.

This function indicates that the *noise gain* ( $1/\beta$ ) curve rises at 20 dB/decade above the *break frequency*,  $f_p$ . If  $f_p$  is well below the open-loop unity-gain frequency, the system becomes unstable. This corresponds to a rate of closure of about 40 dB/decade. The rate of closure is defined as the magnitude of the difference between slopes of the open-loop gain (dB) plot (–20 dB/decade at most frequencies of interest) and that of  $1/\beta$ , in the neighborhood of the frequency at which they cross (loop gain = 0 dB).

To cure the instability induced by  $C_1$ , a capacitor,  $C_f$ , can be connected in parallel with  $R_2$ , providing a zero which can be matched with the pole,  $f_p$ , to lower the rate of closure, and thus increase the phase margin. For a phase margin of 90°, pick  $C_f = (R_1/R_2) C_1$ .

Figure 15 shows the frequency response of the AD8605 in the configuration of Figure 14.

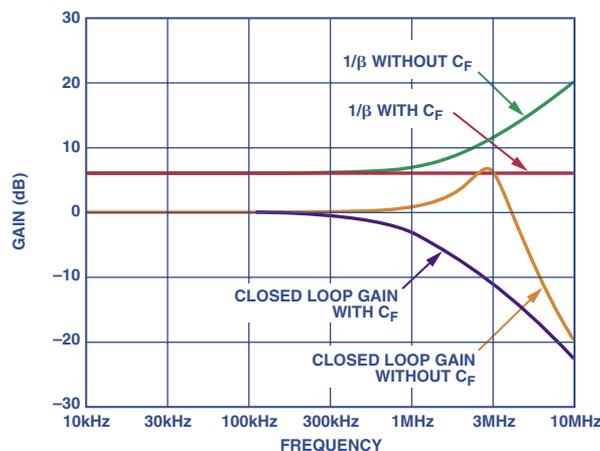


Figure 15. Frequency response of Figure 14.

**Q:** Can I predict what the phase margin would be, or how much peaking I should expect?

**A:** Yes, here's how:

You can determine the amount of uncompensated peaking using the following equation:

$$Q = \sqrt{\frac{f_u}{f_z}}, \text{ where } f_z = \frac{1}{2\pi(R_1 \parallel R_2) C_1}$$

where  $f_u$  is the unity gain bandwidth,  $f_z$  is the breakpoint of the  $1/\beta$  curve, and  $C_1$  is the total capacitance—internal and external—including any parasitic capacitance.

The phase margin ( $\Phi_m$ ) can be determined with the following equation:

$$\Phi_m = \cos^{-1} \left( \sqrt{1 + \frac{1}{4Q^4} - \frac{1}{2Q^2}} \right)$$

The AD8605 has a total input capacitance of approximately 7 pF. Assuming the parasitic capacitance is about 5 pF, the closed-loop gain will have a severe peaking of 5.5 dB, using the above equation. In the same manner, the phase margin is about 29°, a severe degradation from the op amp's natural phase response of 64°.

**Q:** How can I make sure the op amp circuit is stable if I want to use an RC filter directly at the input?

**A:** You can use a similar technique to that described above. Here's an example:

It is often desirable to use capacitance to ground from an amplifier's active input terminals to reduce high-frequency interference, RFI and EMI. This filter capacitor has a similar effect on op amp dynamics as increased stray capacitance. Since not all op amps behave in the same way, some will tolerate less capacitance at the input than others. So, it is useful in any event, to introduce a feedback capacitor,  $C_f$ , as compensation. For further RFI reduction, a small series resistor at the amplifier terminal will combine with the amplifier's input capacitance for filtering at radio frequencies. Figure 16 shows an approach (at left), that will have difficulty maintaining stability, compared with a considerably

improved circuit (at right). Figure 17 shows their superimposed square wave responses.

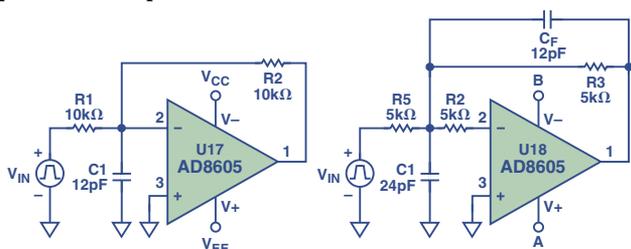


Figure 16. Input filter without (at left), and with (at right) compensation and lower impedance levels.

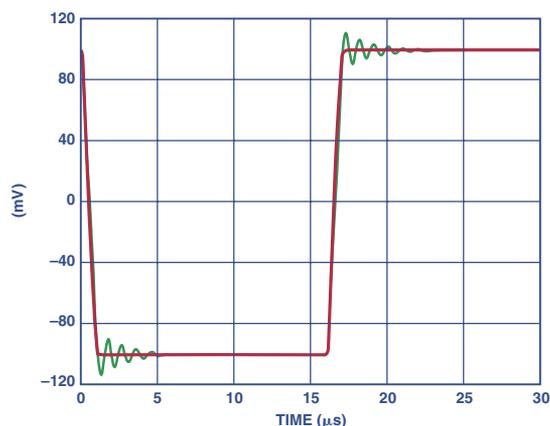


Figure 17. Comparison of output responses of the circuits in Figure 16. The circuit at left resulted in the oscillatory response.

**Q:** You mentioned earlier that stray capacitance is added to the total input capacitance. How important is stray capacitance?

**A:** Unsuspected stray capacitance can have a detrimental impact on the stability of the op amp. It is very important to anticipate and minimize it.

The board layout can be a major source of stray input capacitance. This capacitance occurs at the input traces to the summing junction of the op amp. For example, one square centimeter of a PC board, with a ground plane surrounding it, will produce about 2.8 pF of capacitance (depending on the thickness of the board).

To reduce this capacitance: Always keep the input traces as short as possible. Place the feedback resistor and the input source as close as possible to the op amp. Keep the ground plane away from the op amp, especially the inputs, except where it is needed for the circuit and the noninverting pin is grounded. When ground is really needed, use a wide trace to ensure a low resistance path to ground.

**Q:** Can op amps that aren't unity-gain stable be used at unity-gain? The OP37 is a great amplifier, but it must be used in a gain of at least 5 to be stable.

**A:** You can use such op amps for lower gains by *tricking* them. Figure 18 shows a useful approach.

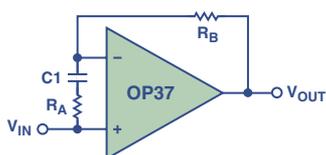


Figure 18. Unity-gain follower using an input series R-C to stabilize an amplifier that is not stable at unity-gain.

In Figure 18,  $R_B$  and  $R_A$  provide enough closed-loop gain at high frequencies to stabilize the amplifier, and  $C_1$  brings it back to unity at low frequencies and dc. Calculating the values of  $R_B$  and  $R_A$  is fairly straightforward, based on the amplifier's minimum stable gain. In the case of the OP37, the amplifier needs a closed-loop gain of at least 5 to be stable, so  $R_B = 4R_A$  for  $\beta = 1/5$ . For high frequencies, where  $C_1$  behaves like a direct connection, the op amp *thinks* it's operating at a closed-loop gain of 5, and is therefore stable. At dc and low frequencies, where  $C_1$  behaves like an open circuit, there is no attenuation of negative feedback, and the circuit behaves like a unity-gain follower.

The next step is to calculate the value of capacitance,  $C_1$ . A good value for  $C_1$  should be picked such that it will provide a break frequency at least a decade below the circuit's corner frequency ( $f_{-3\text{dB}}$ ).

$$C_1 = \frac{1}{2\pi R_A \left(\frac{f_c}{10}\right)}$$

Figure 19 shows the output of the OP37 in response to a 2-V p-p input step. The values of the compensation components are chosen using the equations above, with  $f_c = 16\text{ MHz}$

$$R_B = 10\text{ k}\Omega$$

$$R_A = R_B/4 = 2.5\text{ k}\Omega$$

$$C_1 = 1/(2\pi \times 2.5\text{E}3 \times 16\text{E}6/10) = 39\text{ pF}$$

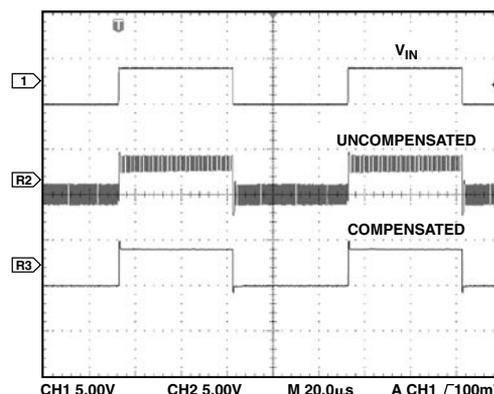


Figure 19. Unity-gain response of the OP37 with and without compensation.

**Q:** Can this approach also be used for the inverting configuration? Can I still use the same equations?

**A:** For the inverting configuration, the analysis is similar, but the equations for the closed-loop gain are slightly different. Remember that the input resistor to the inverting terminal of the op amp is now in parallel with  $R_A$  at high frequencies. This parallel combination is used to calculate the value of  $R_A$  for minimum stable gain. The capacitance value,  $C_1$ , is calculated in the same way as for the noninverting case.

**Q:** Are there drawbacks to using this technique?

**A:** Indeed, there are. Increasing the noise gain will increase the output noise level at higher frequencies, which may not be tolerable in some applications. Care should be used in wiring, especially with high source impedance, in the follower configuration. The reason is that positive feedback via capacitance to the amplifier's noninverting input at frequencies where the gain is greater than unity, can invite instability, as well as increased noise. ▶

## PRODUCT INTRODUCTION—New DigiTrim™ Amplifiers

Many of today's most popular electronics applications such as digital cameras and mobile phones are migrating to lower operating voltages, reducing error budgets and increasing component accuracy requirements. At the same time, volume for these products continues to grow, putting additional pressure on suppliers to reduce component costs. This is especially true for the amplifiers used in these applications.

The key amplifier parameter for system accuracy is input offset voltage. Various techniques including laser trim and zener-zapping have been used to adjust amplifier offset voltage and other parameters. Indeed, offset-trimming techniques have enabled the entire class of precision amplifiers to exist today. However, offset trimming and the improved accuracy it creates has not, for the most part, found its way into the high volume, low-cost CMOS amplifier segment—until now. Analog Devices has developed a novel, patented trimming process that delivers the required accuracy and performance at a very low cost. The DigiTrim technique allows production of precision CMOS amplifiers at costs up to 30% lower than competitive solutions and with higher accuracy.

**DigiTrim**—*Digital control of adjustment currents through logic circuits and weighted current sources.*

DigiTrim adjusts circuit performance by programming digitally weighted current sources. In this patented new trim method, the trim information is entered through existing analog pins using a special digital keyword sequence. The adjustment values can be temporarily programmed, evaluated, and readjusted for optimum accuracy before permanent adjustment is performed. After the trim is completed, the trim circuit is locked out to prevent the possibility of any accidental re-trimming by the end user.

The physical trimming, achieved by blowing polysilicon fuses, is very reliable. No extra pads or pins are required for this trim method, and no special test equipment is needed to perform the trimming. The trims can be done after packaging so that assembly-related shifts may be eliminated. No testing is required at the wafer level, and no special wafer fabrication process is required, so circuits can even be produced by our foundry partners. The trim circuitry scales with the process features so that as the process—and thus the amplifier circuit—shrink, the trim circuit also shrinks proportionally. The trim circuits are considerably smaller than the amplifier circuits, so their contribution to die cost is minimal. The trims are discrete, but the required accuracy is easily achieved at a very small cost increase over an untrimmed part. A simplified representation of an amplifier with DigiTrim is shown in Figure 1.

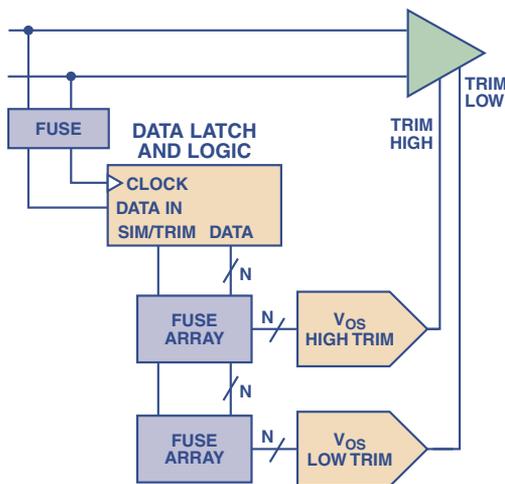


Figure 1. Simplified drawing

## DigiTrim Amplifiers

The first part used in this new scheme was the AD8602 dual, low-cost, rail-to-rail CMOS amplifier. The offset voltage, trimmed for both high and low common-mode voltage conditions, is under 500  $\mu\text{V}$  over the full common-mode input voltage range. With an 8-MHz bandwidth, 5 V/ $\mu\text{s}$  slew rate and 640  $\mu\text{A}$  per amplifier current consumption, this part supports a variety of high volume, cost-sensitive applications from bar code scanners to GSM phones. Three new DigiTrim amplifiers have recently been introduced:

The AD8615/AD8616/AD8618 single/dual/quad op amps use ADI's patented DigiTrim technology to achieve 65- $\mu\text{V}$  maximum offset voltage and 7- $\mu\text{V}/^\circ\text{C}$  maximum offset drift without laser trimming. The CMOS process provides ultralow input bias current (1 pA max), input offset current (0.5 pA max), and current noise density (0.05 pA/ $\sqrt{\text{Hz}}$ ). Other features include 24-MHz gain-bandwidth product, rail-to-rail inputs and outputs, 100-dB common-mode rejection, 90-dB power supply rejection, 6-nV/ $\sqrt{\text{Hz}}$  voltage noise, 1500-V/mV open-loop gain, and  $\pm 150\text{-mA}$  output current. This combination of features makes the AD8615/AD8616/AD8618 ideal for filters, photodiode amplifiers, buffering multiplexed inputs to ADCs, and audio applications. Specified from  $-40^\circ\text{C}$  to  $+125^\circ\text{C}$ , they operate on a single 2.7-V to 5.5-V supply, consuming only 1.7 mA per amplifier. The AD8615 is available in a 5-lead SOT-23 package; the AD8616 is available in 8-lead MSOP and SOIC packages; and the AD8618 is available in 14-lead TSSOP and SOIC packages.

Figures 2 and 3 show typical applications for the AD8616 and AD8618. Additional information on DigiTrim technology can be found at <http://www.analog.com/DigiTrim>.

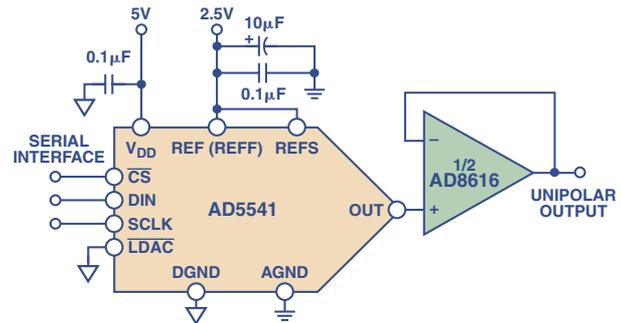


Figure 2. DAC buffer

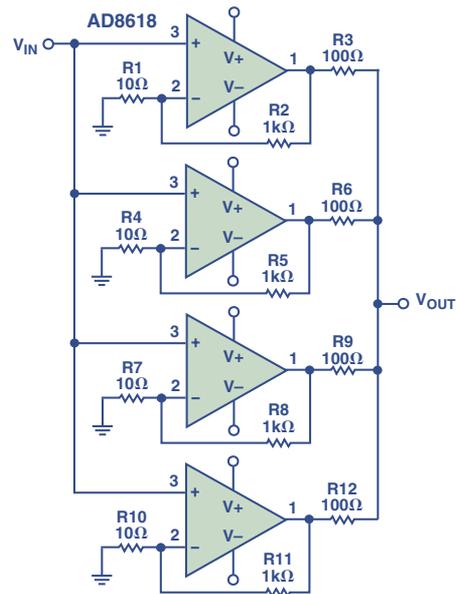


Figure 3. Noise reduction using multiple parallel paths

## PRODUCT INTRODUCTIONS: VOLUME 38, NUMBER 2

Data sheets for all ADI products can be found by entering the model number in the Search Box at [www.analog.com](http://www.analog.com)

### April

**Amplifier, Operational**, dual, CMOS,  
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**Power Supply Controller**, high-efficiency,  
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**Thermal Monitor/Controller**, dBCool™, fan and  
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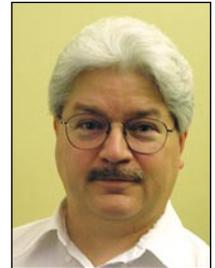
**Thermal Controller/Monitor**, dBCool, voltage  
and remote temp ..... ADT7468

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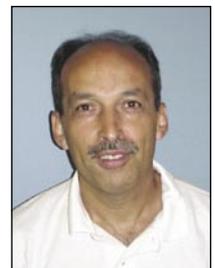
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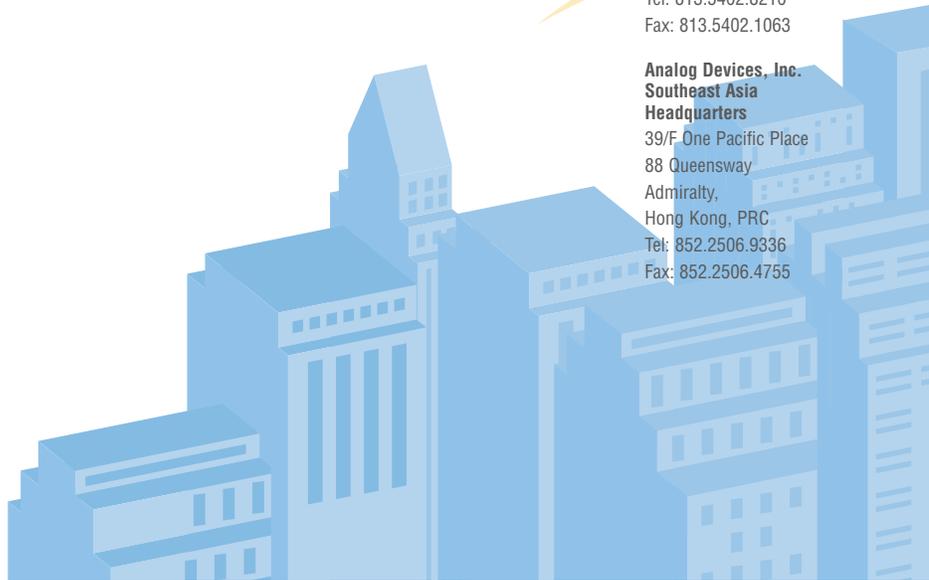
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