Volume 41, Number 3, 2007

Analog Dialogue

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

In This Issue

Analog Dialogue Since 1967

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ANALOG DEVICES

Editors' Notes

IN THIS ISSUE

Modern op amps and in-amps provide great benefits to the designer, and a great many clever, useful, and tempting circuit applications have been published. But all too often, in one's haste to assemble a circuit, some very basic issues are overlooked, leading to the circuit not functioning as expected—or perhaps at all. The article on page 3, destined to be a classic, will discuss a few of the most common application problems and suggest practical solutions.

Medical ultrasound systems, among the more complex signal processing devices in use today, are used for real-time detection of health problems and general diagnostic procedures. Slower than radar and faster than sonar, ultrasound systems have become increasingly portable, evolving from cart-based systems into palm-sized devices. The article on page 10 discusses some of the necessary ingredients of compactness.

As electret microphones become smaller, their element capacitance decreases. JFET preamps no longer suffice, as their relatively large input capacitance attenuates the signal from the microphone. Much can be gained by replacing JFET amplifiers with CMOS analog and digital circuitry, including easier gain setting, multiple functional modes, direct digital outputs, enhanced sound quality, and higher noise immunity (see page 13).

40TH ANNIVERSARY CD

The centerfold of this issue depicts the cover of *40 Years of Analog Dialogue*, our 40th anniversary CD. Sure to become a valuable collector's item, the CD contains a complete set of PDF reproductions of (every page of) every issue published in print during its first 40 years, from Volume 1, Number 1, in 1967, through Volume 40, Number 4, in 2006. Articles can be accessed by issue or by author. In addition, Editors' Notes, articles in the "Ask The Applications Engineer" series, and Barrie Gilbert's "The Wit and Wisdom of Dr. Leif" can be accessed directly.

If you're on our North American print distribution list, you should have received a copy with this magazine. If you're on our European print distribution list, you should be receiving a copy soon. For all other readers in North America who want a copy, please contact your local sales office, or send your name and mailing address, along with a check or money order for $50.00, made out to Analog Devices, Inc., to this address:

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A READER RESPONDS

In the *Analog Dialogue* Volume 41, Number 2 Editors' Notes, you wrote:

> "The second problem is that the GPS loses its signal when I travel through Boston's Big Dig tunnels, making it unable to maintain its bearings or to provide instructions for lane splits and exits. Too bad it doesn't include an ADI accelerometer, whose inertial navigation capabilities could provide positional information during these temporary signal losses."

Now, extend that problem to coal mines, where GPS is useless "ALL" of the time. How do you tell a miner's location "after" an anomalous event like an earthquake, when the mine is literally a different place? Way points might be damaged, so they can't be relied on. Coal absorbs most radio frequencies, so like GPS, RF is nearly useless.

The physical environment also presents several challenges: dust so thick that you cannot see through it; irreplaceable batteries that must be capable of lasting through a 12-hour work shift; intrinsic safety, which requires that the total available energy must not be capable of igniting a methane gas atmosphere; and the need for a small, lightweight device that can be carried for 12 hours.

Some possible areas for research include underground and underwater radio antennas, RFID-radar, black hole antennas, and NIST WWVB radio broadcasts. More information can be found at the Location Challenge (http://www.wearablesmartsensors.com/location_challenge.html).

Do any of your readers have an idea that no one has yet thought of?

Bob Paddock [bpaddock@designer-iii.com]

Dan Chimes In

A productive place to start may be with locating tunnels used by escaping convicts, terrorists, and smugglers of dope, other contraband, and people. As a medium for R&D, there are many more of them to be found, and if we can solve that problem, it may be a big step toward locating the (less frequently) lost miners. A solution to this problem could perhaps even allow us to follow miners (and terrorists) around in real time.

The answer is (im?)possibly through gravitational waves, but I don't know of any likely technologies for exploiting them.

WHAT WAS THE LIGHTNING EMPIRICIST?

In June 1952, George A. Philbrick Researches, Inc., a company soon to become the originator of a new product category, the (octal-socket plug-in) differential operational amplifier, published Volume 1, Number 1, of *The Lightning Empiricist*, "A Journal for devotees of high-speed analog computation, those enthusiasts for the new doctrine of Lightning Empiricism, publishable aperiodically and distributed without charge … and offering items of interest and value on such computational topics as applications, techniques, and new or improved components."

Its first editor was George A. Philbrick himself; later, from 1957 through 1966 (the year that Philbrick was acquired by Teledyne, Inc.) it was edited by Dan Sheingold—who also wrote much of the foreword to the first Teledyne/Philbrick issue, that would be published in March 1969, after his departure to become editor of *Analog Dialogue*. You can find quite a few issues of *The Lightning Empiricist* at www.philbrickarchive.org.

Dan's favorite issue was one published in 1964, featuring an article containing many good ideas with the lengthy title: "IMPEDANCE AND ADMITTANCE TRANSFORMS Using Operational Amplifiers—Transadmittance, Transimpedance, Positive and Negative Self-Impedance through Active Circuits, including references to Photomultiplier and ion-current amplifiers; Current sources and generators; and Negative resistors and capacitors for dynamic compensation with Single-Ended, Differential, and Inverted Amplifiers." Even after more than 40 years, one may find a few fresh ideas in it at www.philbrickarchive.org/1964-1_v12_no1_the_lightning_empiricist.htm.

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Analog Dialogue

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Analog Dialogue is the free technical magazine of Analog Devices, Inc., published continuously for 41 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** site—the virtual world of *Analog Devices*. For history buffs, the *Analog Dialogue* archives include all regular editions, starting with Volume 1, Number 1 (1967), plus three special anniversary issues.

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

If All Else Fails, Read This Article

Avoid Common Problems When Designing Amplifier Circuits

By Charles Kitchin [charles.kitchin@analog.com]

INTRODUCTION

Modern operational amplifiers (op amps) and instrumentation amplifiers (in-amps) provide great benefits to the designer, compared with assemblies of discrete semiconductors. A great many clever, useful, and tempting circuit applications have been published. But all too often, in one's haste to assemble a circuit, some very basic issue is overlooked that leads to the circuit not functioning as expected—or perhaps at all.

This article will discuss a few of the most common application problems and suggest practical solutions.

Missing DC Bias Current Return Path When AC-Coupled

One of the most common application problems encountered is the failure to provide a dc return path for bias current in ac-coupled operational- or instrumentation-amplifier circuits. In Figure 1, a capacitor is connected in series with the noninverting (+) input of an op amp to ac couple it, an easy way to block dc voltages that are associated with the input voltage (V_{IN}). This should be especially useful in high-gain applications, where even a small dc voltage at an amplifier's input can limit the dynamic range, or even result in output saturation. However, capacitively coupling into a high-impedance input, without providing a dc path for current flowing in the + input, will lead to trouble!

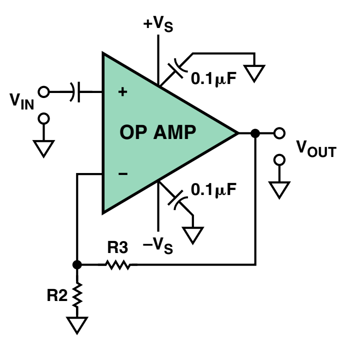

Figure 1. A malfunctional ac-coupled op-amp circuit.

What actually happens is that the input bias currents will flow through the coupling capacitor, charging it, until the common-mode voltage rating of the amplifier's input circuit is exceeded or the output is driven into limits. Depending on the polarity of the input bias current, the capacitor will charge up toward the positive supply voltage or down toward the negative supply. The bias voltage is amplified by the closed-loop dc gain of the amplifier.

This process can take a long time. For example, an amplifier with a field-effect-transistor (FET) input, having a 1-pA bias current, coupled via a 0.1-µF capacitor, will have a charging rate, I/C, of $10^{-12}/10^{-7}$ = 10 µV/s, or 600 µV per minute. If the gain is 100, the output will drift at 0.06 V per minute. Thus, a casual lab test (using an ac-coupled scope) might not detect this problem, and the circuit will not fail until hours later. Obviously, it is very important to avoid this problem altogether.

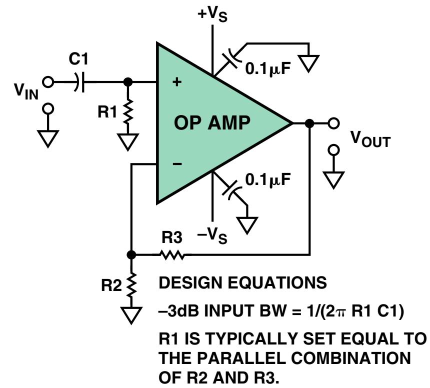

Figure 2. Correct method for ac coupling an op-amp input for dual-supply operation.

Figure 2 shows a simple solution to this very common problem. Here, a resistor is connected between the op-amp input and ground to provide a path for the input bias current. To minimize offset voltages caused by input bias currents, which track one another when using bipolar op amps, R1 is usually set equal to the parallel combination of R2 and R3.

Note, however, that this resistor will always introduce some noise into the circuit, so there will be a trade-off between circuit input impedance, the size of the input coupling capacitor needed, and the Johnson noise added by the resistor. Typical resistor values are generally in the range from about 100,000 Ω to 1 MΩ.

A similar problem can affect an instrumentation amplifier circuit. Figure 3 shows in-amp circuits that are ac-coupled using two capacitors, without providing an input-bias-current return path. This problem is common with instrumentation amplifier circuits using both dual- (Figure 3a) and single (Figure 3b) power supplies.

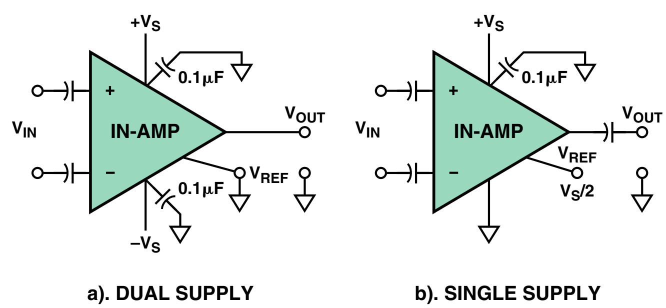

Figure 3. Examples of nonfunctional ac-coupled in-amp circuits.

The problem can also occur with transformer coupling, as in Figure 4, if no dc return path to ground is provided in the transformer's secondary circuit.

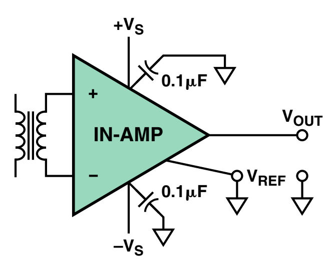

Figure 4. A nonfunctional transformer-coupled in-amp circuit.

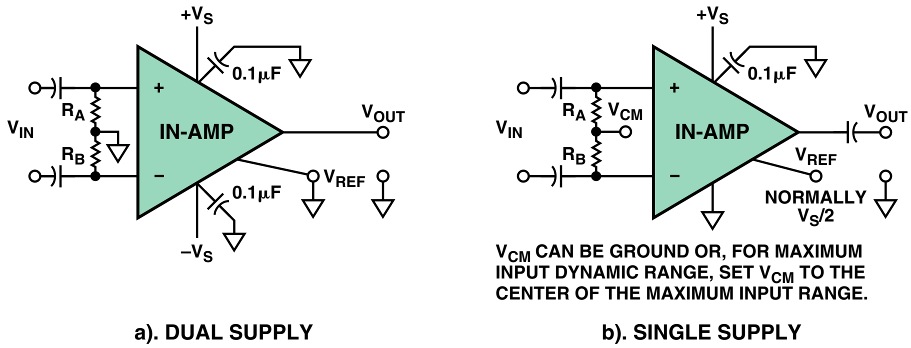

Figure 5. A high-value resistor between each input and common supplies the necessary bias-current return path. a. Dual supply. b. Single supply.

Simple solutions for these circuits are shown in Figure 5 and Figure 6. Here, a high-value resistance (R_A, R_B) is added between each input and ground. This is a simple and practical solution for dual-supply in-amp circuits.

The resistors provide a discharge path for input bias currents. In the dual-supply example of Figure 5a, both inputs are now referenced to ground. In the single-supply example of 5b, the inputs may be referenced either to ground (V_{CM} tied to ground) or to a bias voltage, usually one-half the maximum input voltage range.

The same principle can be used for transformer-coupled inputs (Figure 6), unless the transformer secondary has a center tap, which can be grounded or connected to V_{CM}.

In these circuits, there will be a small offset-voltage error due to mismatches between the resistors and/or the input bias currents. To minimize such errors, a third resistor, about $^1/_{10}{}^{th}$ their value (but still large compared to the differential source resistance), can be connected between the two in-amp inputs (thus bridging both resistors).

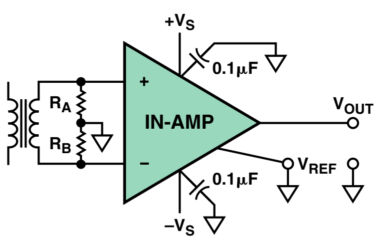

Figure 6. Correct method for transformer input coupling to an in-amp.

Supplying Reference Voltages for In-Amps, Op Amps, and ADCs

Figure 7 shows a single-supply circuit where an in-amp is driving a single-ended analog-to-digital converter (ADC). The amplifier's reference provides a bias voltage corresponding to zero differential input, and the ADC reference provides the scale factor. A simple RC low-pass antialiasing filter is often used between in-amp output and ADC input to reduce out-of-band noise. Often designers are tempted to use simple approaches such as resistance dividers to supply the in-amp and ADC reference voltages. This can lead to errors in the case of some in-amps.

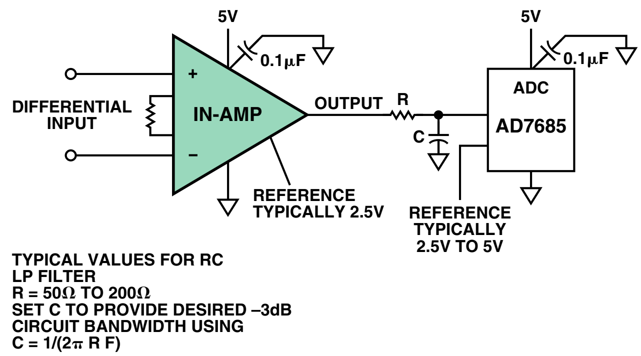

Figure 7. An in-amp drives an ADC in a typical single-supply circuit.

Correctly Providing In-Amp Reference Voltage

A common assumption is that the in-amp's reference-input terminal is at high impedance, since it's an input. So a designer may be tempted to connect a high-impedance source, such as a resistive divider, to the reference pin of an in-amp. This can introduce serious errors with some types of instrumentation amplifiers (Figure 8).

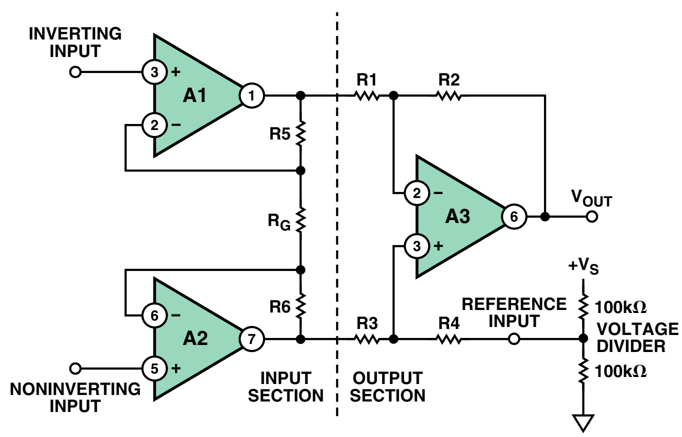

Figure 8. Incorrect use of a simple voltage divider to directly drive the reference pin of a 3-op-amp instrumentation amplifier.

For example, a popular in-amp design configuration uses three op amps connected as above. The overall signal gain is

$$G = \left(1 + \frac{R_5}{R_G} + \frac{R_6}{R_G}\right)\left(\frac{R_2}{R_1}\right) \text{ where } \frac{R_2}{R_1} = \frac{R_4}{R_3}$$

The gain for the reference input (if driven from low impedance) is unity. However, in the case shown, the in-amp has its reference pin tied directly to a simple voltage divider. This unbalances the symmetry of the subtractor circuit and the division ratio of the voltage divider. This would reduce the in-amp's common-mode rejection and its gain accuracy. However, if R4 is accessible, so that its resistance value can be reduced by an amount equal to the resistance looking back into the paralleled legs of the voltage divider (50 kΩ here), the circuit will behave as though a low-impedance voltage source equal to (in this example) one-half the supply voltage were applied to the original value of R4, and the subtractor's accuracy would be maintained.

This approach cannot be used if the in-amp is provided as a closed single package (an IC). Another consideration is that the temperature coefficients of the resistors in the voltage divider should track those of R4 and the other resistors in the subtractor. Finally, the approach locks out the possibility of having the reference be adjustable. If, on the other hand, one attempts to use small resistor values in the voltage divider in an effort to make the added resistance negligible, this will increase power supply current consumption and increase the dissipation of the circuit. In any case, such "brute force" is not a good design approach.

Figure 9 shows a better solution, using a low-power op-amp buffer between the voltage divider and the in-amp's reference input. This eliminates the impedance-matching and temperature-tracking problem and allows the reference to be easily adjustable.

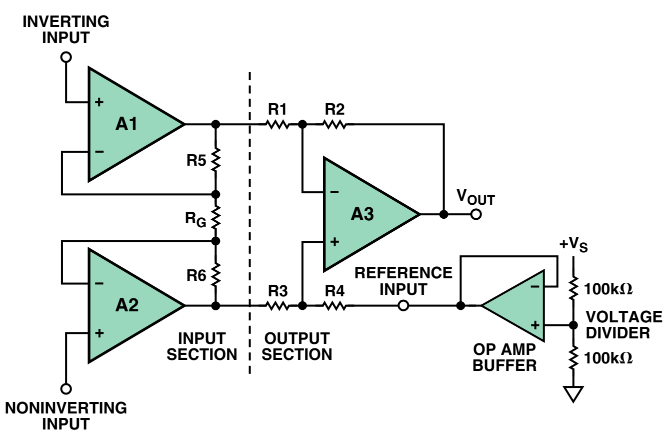

Figure 9. Driving the reference pin of an in-amp from the low-impedance output of an op amp.

Preserving Power-Supply Rejection (PSR) When Amplifiers Are Referenced from the Supply Rail Using Voltage Dividers

An often overlooked consideration is that any noise, transients, or drift of power-supply voltage, V_S, fed in through the reference input will add directly to the output, attenuated only by the divider ratio. Practical solutions include bypassing and filtering, and perhaps even generating the reference voltage with a precision reference IC, such as the ADR121, instead of tapping off V_S.

This consideration is important when designing circuits with both in-amps and op amps. Power-supply rejection techniques are used to isolate an amplifier from power supply hum, noise, and any transient voltage variations present on the power rails. This is important because many real-world circuits contain, connect to, or exist in environments that offer less-than-ideal supply voltage. Also, ac signals present on the supply lines can be fed back into the circuit, amplified, and under the right conditions, stimulate a parasitic oscillation.

Modern op amps and in-amps all provide substantial low-frequency power-supply rejection as part of their design. This is something that most engineers take for granted. Many modern op amps and in-amps have PSR specs of 80 dB to over 100 dB, reducing the effects of power-supply variations by a factor of 10,000 to 100,000. Even a fairly modest PSR spec of 40 dB isolates supply variations from the amplifier by a factor of 100. Nevertheless, *high-frequency* bypass capacitors (such as those in Figure 1 through Figure 7) are always desirable and often essential.

In addition, when designers use a simple resistance divider on the supply rail and an op-amp buffer to supply a reference voltage for an in-amp, any variations in power-supply voltage are passed through this circuitry with little attenuation and add directly to the in-amp's output level. So, unless low-pass filtering is provided, the normally excellent PSR of the IC is lost.

In Figure 10, a large capacitor has been added to the voltage divider to filter its output from power-supply variations and preserve PSR. The –3-dB pole of this filter is set by the parallel combination of R1/R2 and capacitor C1. The pole should be set approximately 10 times lower than the lowest frequency of concern.

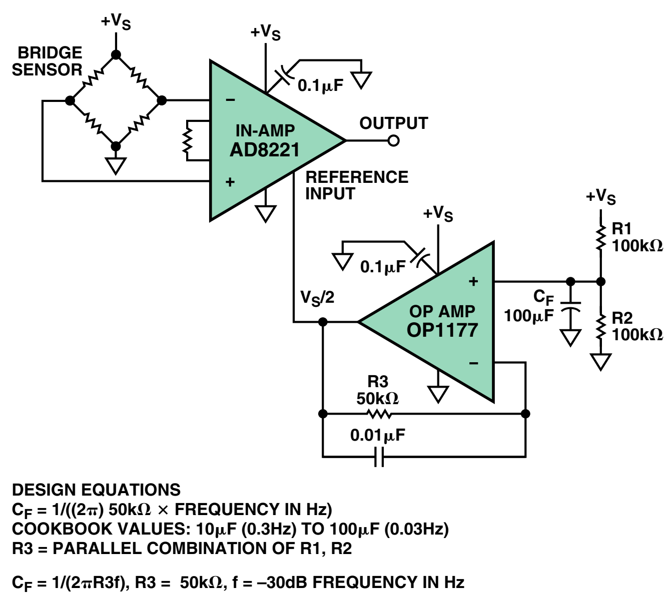

Figure 10. Decoupling the reference circuit to preserve PSR.

The "cookbook" values shown provide a –3-dB pole frequency of approximately 0.03 Hz. The small (0.01-μF) capacitor across R3 minimizes resistor noise.

The filter will take time to charge up. Using the cookbook values, the rise time at the reference input is several time constants (where $T = R_3C_f = 5$ s), or about 10 to 15 seconds.

The circuit of Figure 11 offers a further refinement. Here, the op-amp buffer is operated as an active filter, which allows the use of much smaller capacitors for the same amount of power-supply decoupling. In addition, the active filter can be designed to provide a higher Q and thus give a quicker turn-on time.

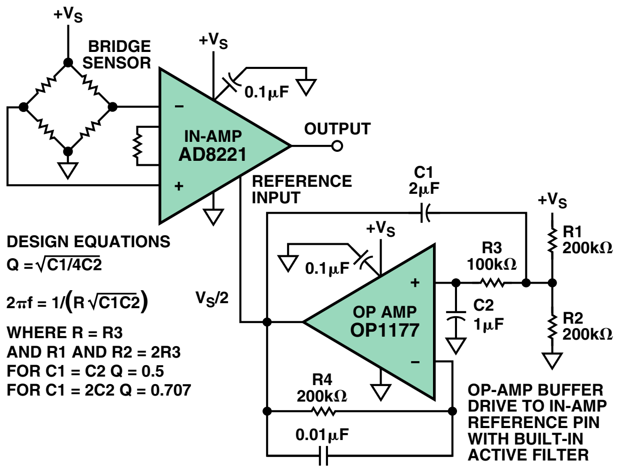

Figure 11. An op-amp buffer connected as an active filter drives the reference pin of an in-amp.

Test results: With the component values shown, and 12 V applied, a 6-V filtered reference voltage was provided to the in-amp. A 1-V p-p sine wave of varying frequency was used to modulate the 12-V supply, with the in-amp gain set to unity. Under these conditions, as frequency was decreased, no ac signal was visible on an oscilloscope, at VREF or at the in-amp output, until approximately 8 Hz. Measured supply range for this circuit was 4 V to greater than 25 V, with a low-level input signal applied to the in-amp. Circuit turn-on time was approximately 2 seconds.

Decoupling Single-Supply Op-Amp Circuits

Finally, single-supply op-amp circuits require biasing of the input common-mode level to handle the positive and negative swings of ac signals. When this bias is provided from the power-supply rail, using voltage dividers, adequate decoupling is required to preserve PSR.

A common and incorrect practice is to use a 100-kΩ/100-kΩ resistive voltage divider with a 0.1 µF bypass capacitor to supply $V_S/2$ to the noninverting pin of the op amp. Using these values, power-supply decoupling is often inadequate, as the pole frequency is only 32 Hz. Circuit instability ("motor-boating") often occurs, especially when driving inductive loads.

Figure 12 (noninverting) and Figure 13 (inverting) show circuits to accomplish $V_S/2$ decoupled biasing for best results. In both cases, bias is provided at the noninverting input, feedback causes the inverting input to assume the same bias, and unity dc gain also biases the output to the same voltage. Coupling capacitor C1 rolls the low-frequency gain down toward unity from BW3.

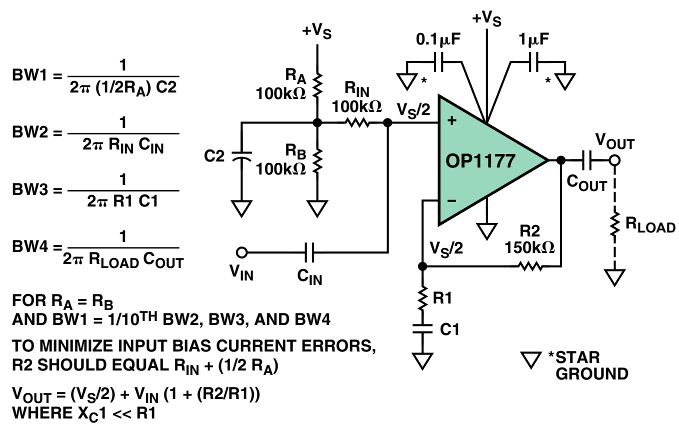

Figure 12. A single-supply noninverting amplifier circuit, showing correct power-supply decoupling. Midband gain = 1 + R2/R1.

A good rule of thumb when using a 100 kΩ/100 kΩ voltage divider, as shown, is to use a C2 value of at least 10 µF for a 0.3-Hz, –3-dB roll-off. A value of 100 µF (0.03-Hz pole) should be sufficient for practically all circuits.

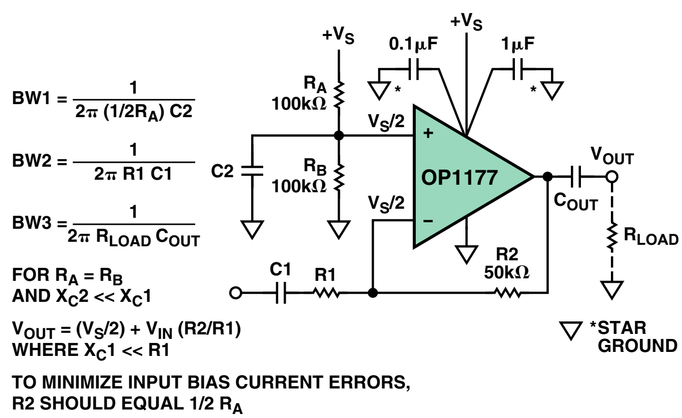

Figure 13. Proper decoupling for a single-supply inverting-amplifier circuit. Midband gain = –R2/R1.

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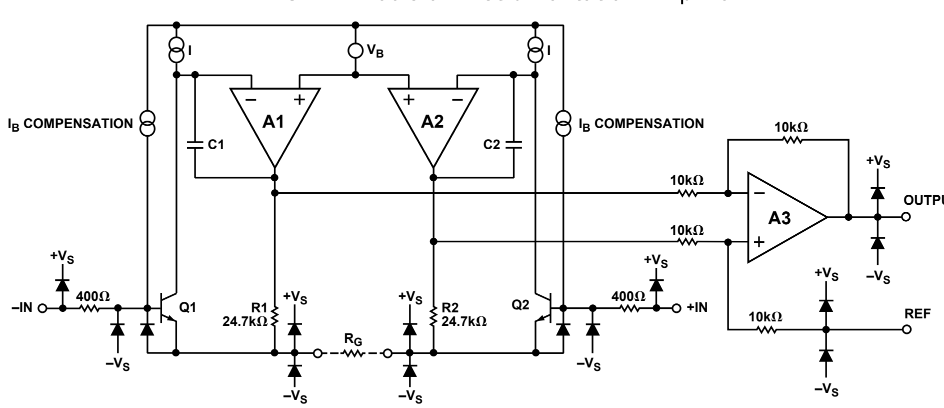
AD8221 Precision Instrumentation Amplifier



PRODUCT INTRODUCTIONS: VOLUME 41, NUMBER 3

Data sheets for all ADI products can be found by entering the model number in the Search box at www.analog.com

July

ADCs, Sigma-Delta, 24-bit, 312-/156-kSPS, on-chip buffers **AD7764/AD7765**
ADCs, Successive-Approximation, dual, 2-channel, 12-/14-bit, 1-MSPS **AD7366/AD7367**
Amplifiers, Power, Class-D audio **ADAU1590/ADAU1592**
Controller, Synchronous Buck, 2-/3-phase, 6-bit VID code **ADP3197**
Controller, Synchronous Buck, 2-/3-/4-phase, 8-bit VID code. **ADP3192A**
Converter, DC-to-DC, synchronous, step-down, 600-mA output **ADP2102**
DACs, Current-Output, dual, 8-/10-/12-/14-/16-bit, 250-MSPS ... **AD974x**
Front-End, Mixed-Signal, broadband modem **AD9869**
Modulator, Quadrature, 2300-MHz to 3000-MHz.......... **ADL5373**
Synthesizer, Direct Digital, 14-bit, 1-GSPS **AD9910**
Upconverter, Quadrature, 14-bit, 1-GSPS.......... **AD9957**

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ADC, Pipelined, dual, 14-bit, 80-/105-/125-/150-MSPS, 1.8-V operation **AD9640**
Amplifier, Power, 2.3-GHz to 2.4-GHz, WiMAX.......... **ADL5570**
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Generator, Clock, 12-output, on-chip VCO **AD9517-x**
Modulator Quadrature, 3000-MHz to 4000-MHz **ADL5374**
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Sensor, Impact, High-*g*, programmable, digital output **ADIS16204**
Supervisor, Microprocessor, high-accuracy, monitors up to four voltages.......... **ADM6710**
Transceiver, RS-232, 460-kbps, ESD-protected, 3.3-V operation **ADM3101E**

September

ADCs, Successive-Approximation, 24-bit, 128-/64-/32-kSPS, 16-/18-bit INL.......... **AD7766/AD7767**
ADCs, Successive-Approximation, dual, 2-channel, 12-/14-bit, 1000-/500-kSPS **AD7366/AD7367**
ADC, Successive-Approximation, 16-bit, 1-MSPS, 7-mW power consumption.......... **AD7980**
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Amplifier, Operational, low-power, high-precision, CMOS, 16-V operation **AD8663**
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Converter, Synchronous Buck, 2-/3-/4-phase, 8-bit VID code.......... **ADP3198A**
DACs, Voltage-Output, 32-channel, denseDAC™, 16-/14-bit, serial-input **AD5372/AD5373**
DACs, Current-Output, dual, 12-/14-/16-bit, 1-GSPS.......... **AD9776A/AD9778A/AD9779A**
DAC, Voltage-Output, 40-channel, denseDAC, 14-bit, serial-input.......... **AD5371**
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Display Interface, Analog, 10-bit, high-performance **AD9984A**
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Processors, Blackfin, industrial and peripheral-intensive applications.......... **ADSP-BF538/ADSP-BF538F**
Processor, SigmaDSP® audio, 28-/56-bit, two ADCs, four DACs.......... **ADAU1401**
Receiver, Differential, triple, adjustable line equalization **AD8123**
Regulator, Low-Dropout, adjustable-output, 1-A loads.......... **ADP1708**
Regulators, Low-Dropout, fixed-output, 1-A loads.... **ADP1706/ADP1707**
Signal Processor, CCD, 12-bit, *Precision Timing™* core, on-chip V-driver.......... **AD9920A**
Signal Processor, CCD, 2-channel, 14-bit, *Precision Timing* core.......... **AD9978**
Signal Processor, CCD, 2-channel, 14-bit, *Precision Timing* core, on-chip V-driver.......... **AD9990**
Switch, HDMI/DVI, 3:1, equalization, pre-emphasis.......... **ADV3000**
Switches, Crosspoint, 32 × 16, buffered, video, 600 MHz **AD8104/AD8105**
Switches, Crosspoint, triple, 16 × 5, buffered, video, 450-/500-MHz **AD8177/ AD8178**
Switches, Crosspoint, triple, 16 × 9, buffered, video, 450-/500-MHz **AD8175/ AD8176**
Supervisor, dBCool®, remote thermal monitor and fan controller.......... **ADT7473-1**

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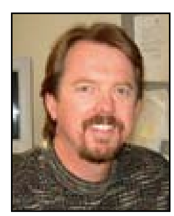

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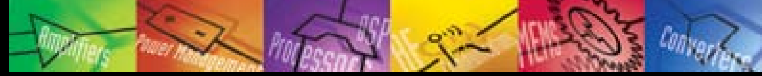
40 years of Analog Dialogue 1967 2006



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8-Channel, 12-Bit, 10-MSPS to 50-MSPS Front End:

The AD9271—A Revolutionary Solution for Portable Ultrasound

By Rob Reeder [rob.reeder@analog.com]
Corey Petersen [corey.petersen@analog.com]

Medical ultrasound systems[1] are among the more sophisticated signal processing machines in widespread use today. Though analogous to radar or sonar, they operate at RF speeds that are orders of magnitude slower than radar and faster than sonar. Since the development of early cart-based ultrasound systems, the medical industry has used this real-time technology for both early detection of health problems and general diagnostic procedures. Over time, ultrasound systems have become increasingly portable, with some even evolving into ultracompact palm-sized devices. In the not-so-distant future, an ultrasound system could become a specialized personal digital assistant (PDA)—though not quite as common as the doctor's stethoscope. We will discuss here some of the necessary ingredients of compactness.

Ultrasound System Architectures

A commonly used approach to image acquisition in ultrasound systems is *digital beamforming* (DBF). Beamforming, as applied to medical ultrasound, is defined as the phase alignment and summation of signals that are generated from a common source, but received at different times by a multi-element ultrasound transducer. With arrays of 16 to 32 (or more) receiver channels phase-shifted and summed together to extract coherent information, beamforming has two functions: it imparts directivity to the transducer—enhancing its gain—and defines a focal point within the body, from which the location of the returning echo is derived. In its simplest state, a DBF system block diagram can look something like Figure 1. The output of each sensor element is amplified, converted to digital, and arranged in sequence. Multiple channels are spatially summed to develop an image.

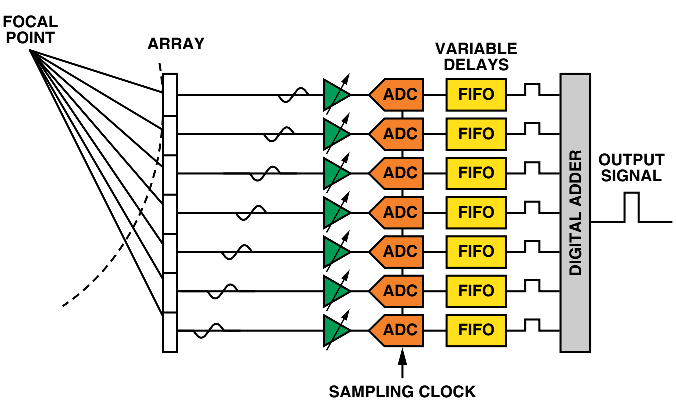

Figure 1. Simplified block diagram of a typical DBF system.

DBF architectures are preferred over earlier analog beamforming systems (ABF)—which use variable delay lines and analog summation before conversion—because they tend to have better channel-to-channel matching characteristics and are more flexible. Once the signal is acquired, its quality can be enhanced by performing digital operations such as beam steering and coherent signal summation. Bringing the digital engine closer to the ultrasonic sensors enables significantly finer adjustments to be made than could be achieved in the analog system. DBF is today's most commonly used architecture, even though significant challenges include high power consumption (due to the large number of channels) and size—due to the sheer number of components usually needed to acquire and produce an accurate signal.

Until recently, most DBF systems were assembled from many components, using discrete solutions and multiple ICs. The *receive* (Rx) signal chain consists primarily of a low-noise amplifier (LNA), which functions as a preamplifier; a variable-gain amplifier (VGA), which functions as a *time-gain* amplifier—to compensate for attenuation of the return signal by body tissues as a function of time (as a proxy for depth); an antialiasing filter (AAF); and an analog-to-digital converter (ADC). Multiple copies of these components are required in common digital-beamforming architectures. Increasing the number of channels improves the dynamic range, as long as the channel noise is random or uncorrelated. A range of 64 to 256 channels is common for high-end systems, while a range of 16 to 64 channels is more common for portable, mid- to low-end ultrasound systems.

Why the Push for Portable?

Many demanding applications can realize the benefits of a lightweight portable compact device that delivers real-time scanning. Obviously, field emergency medical service (EMS) teams will have quicker access to a patient, and will be able to send in results before arriving at the emergency room. If the ride is long, a doctor can diagnose remotely while awaiting the patient in the ER. During routine office visits, general practitioners can perform scans of the patient as part of an examination, without requiring a specialist.

Increased portability offers opportunities to use these devices to provide a better grade of medical service in remote areas and villages that may not have reliable electrical power.

Veterinarians find portable ultrasound to be useful for onsite diagnosis of larger animals and pets. It is also useful at swine and cattle ranches for onsite diagnosis.

Ultrasound in nondestructive testing and preventive maintenance is also a growing market. Examples include systems deployed to scan bridge beams, bearings in industrial machinery, and oil pipelines. Inspection costs can be reduced and critical downtime for expensive equipment can be avoided. Portable scanning equipment in industrial plants can be vital for catching potentially catastrophic problems before they arise.

Adoption of portable ultrasound of course carries a cost, both for acquisition of these devices that diagnose, scan, and analyze, and also for training of users of these new devices. But, in a great many such cases, the benefits vastly outweigh the costs.

Saving Space, Power, and Money with the AD9271

An essential subsystem from Analog Devices designed to satisfy the compactness requirement, the tiny, 14-mm × 14-mm × 1.2-mm, AD9271[2] (Figure 2) brings together all of the required signal chain blocks for acquiring *eight* channels of data, with a dramatic reduction in board space and power. In comparison to a solution employing discrete elements, the AD9271 reduces the total area per channel by more than 1/3, and power dissipation by more than 25%—consuming only 150 mW per channel at 40 MSPS. The AD9271 also offers a host of customizing options—available through a serial port interface—allowing further optimization of power and configurability, depending on the application.

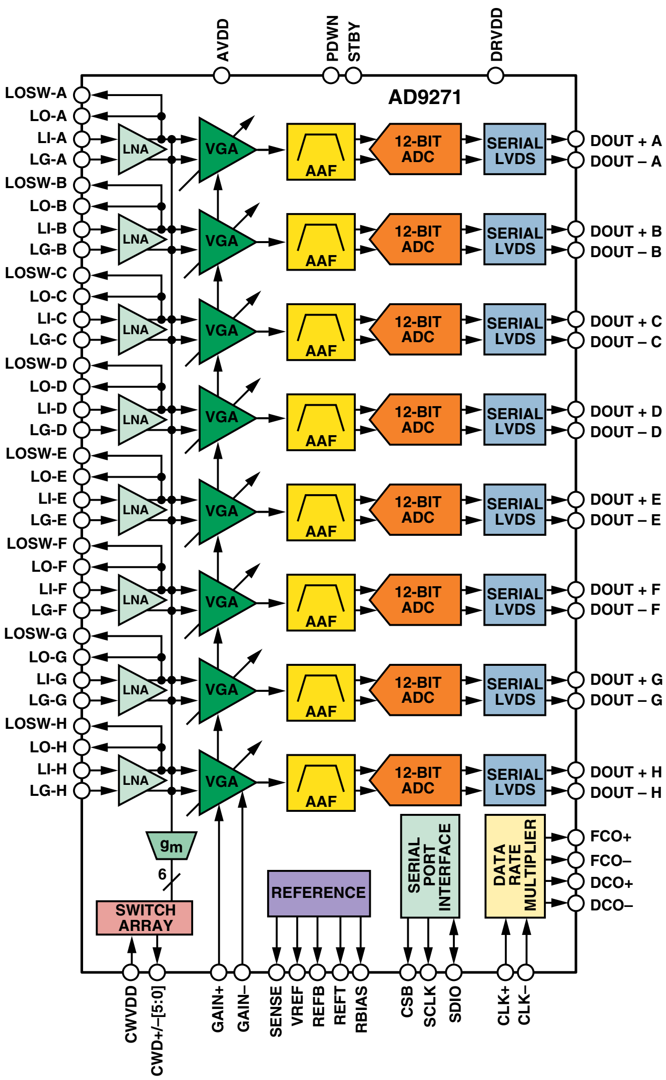

Figure 2. AD9271 block diagram.

The AD9271 embodies an 8-channel signal chain, each channel comprising a low-noise amplifier (LNA), variable-gain amplifier (VGA), antialiasing filter (AAF), and analog-to-digital converter. This is the *receive* chain commonly used to process return pulses in *pulsed-wave* mode: *B-mode* scanning for gray scale imaging, and *F-mode*, which is a color overlay on the B-mode display, to show blood flow. In pulsed-wave mode, the transducer alternates between transmission and reception in order to develop a periodically updated two-dimensional image.

Another common form of imaging is *continuous-wave* (CW) *Doppler*, or D-mode, for showing blood flow velocities and their frequencies. As the name suggests, the image is produced using continuously generated signals, where one-half the transducer channels are transmitting and the other half are receiving. CW has the advantage of measuring high velocities of blood flow accurately, but it lacks the depth and penetration found in traditional pulsed-wave systems. Since each method has its own benefits and limitations, depending on the application, modern ultrasound systems commonly use both modalities—and the AD9271 is applicable to both. In particular, it allows the user to operate in continuous-wave Doppler mode by employing an integrated crosspoint switch. This crosspoint switch allows channels of similar phase to be coherently summed into groups for phase alignment and summation. The AD9271 supports delay lines for low-end systems, and the AD8339[3] quad demodulator with programmable phase adjustment for the best performance. The AD8339 allows finer adjustments to phase alignment and summation in order to increase image accuracy. This device easily connects externally, allowing the user to compact more of the signal chain required for signals that need very large dynamic range.

Dynamic Range and Noise Requirements

As the high-frequency acoustic signals penetrate through the body, they are attenuated by about 1 dB/cm/MHz. For example, with an 8-MHz probe and 4-cm depth penetration—and accounting for both outgoing and return attenuation—the signal amplitude variation from the internal tissues will differ by 64 dB (or 4 × 8 × 2) from reflections near the surface (Further Reading 2). Adding 50 dB of imaging resolution, and accounting for losses due to bone, cables, and other mismatches, the desired dynamic range approaches 119 dB. To put this into perspective, a 0.333-V p-p full-scale signal with a 1.4-nV/$\sqrt{\text{Hz}}$ noise floor in a 10-MHz bandwidth implies an 88-dB input dynamic range. Additional dynamic range is achieved by using multiple channels [10 × log(N channels)], e.g., 128 channels increases the dynamic range by 21 dB. This establishes a practical limit for dynamic range between 100 dB and 120 dB.

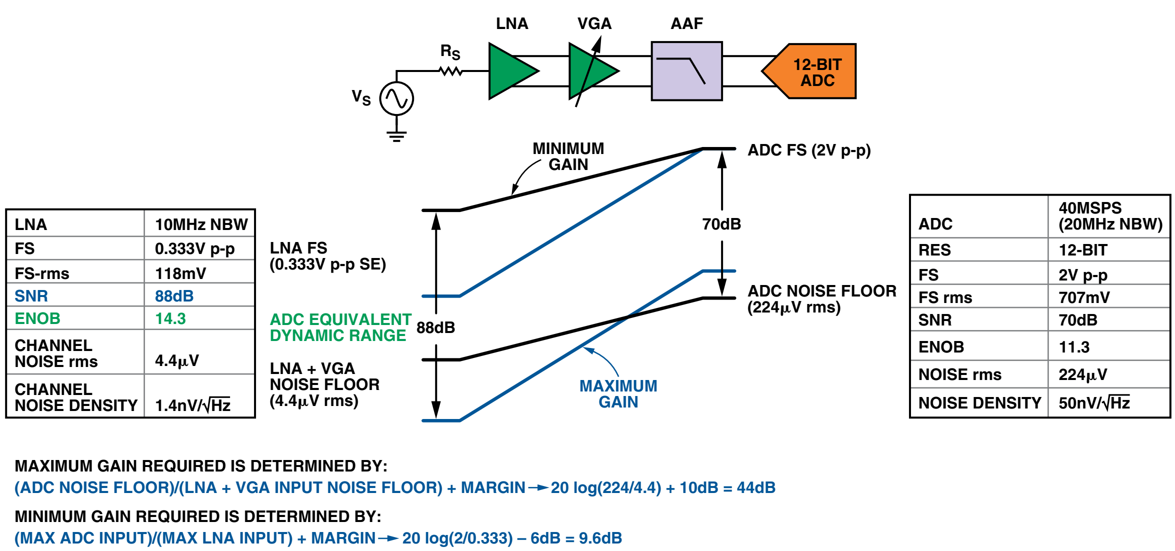

Figure 3. TGC gain requirements for 12-bit ADC.

Table 1. Comparison of Solutions Using Analog Devices Components

Product	LNA Input Range	LNA Input Noise	Total Channel Input Noise (Without ADC)	Input Dynamic Range for Channel (@10-MHz BW)
AD8332	550 mV p-p	0.74 nV/√Hz	0.82 nV/√Hz	97 dB
AD8335	625 mV p-p	1.2 nV/√Hz	1.3 nV/√Hz	95 dB
AD9271	400 mV p-p	1.4 nV/√Hz	1.65 nV/√Hz	89 dB
	333 mV p-p	1.2 nV/√Hz	1.44 nV/√Hz	88 dB
	250 mV p-p	1.1 nV/√Hz	1.31 nV/√Hz	87 dB

The achievable dynamic range is limited by the front-end components. Since the entire dynamic range is not needed at all instants of time, an ADC with less than full dynamic range can be used by sweeping the gain of the VGA to match the attenuation of the received reflection over time (proportional to depth of penetration). This is called time-gain compensation (TGC). The LNA sets up the equivalent dynamic range that can be mapped into the ADC. The AD9271 has an equivalent dynamic range of 88 dB in a 10-MHz bandwidth (158 dB/√Hz), allowing it to handle both very small and large signals (echoes) from the tissue being scanned, as shown in Figure 3. The full scale of the LNA should be large enough not to saturate from the near-field signal; and the lower the noise floor, the higher the dynamic range.

As power requirements must increase to handle lower noise levels, some compromise must be made in portable applications because of power constraints. While the 88-dB dynamic range of the AD9271 is better than competitive solutions, it is still less than higher-power VGA products, such as the AD8332,[4] with its 0.72-nV/√Hz input-referred noise, as shown in Table 1. Notice that the AD8332 has the lowest input-referred noise and highest input dynamic range of the solutions shown. No one approach is ideal. Although digital processing is an essential feature of all solutions today, the specific implementation and choice of components are proprietary to each ultrasound manufacturer.

CONCLUSION

For both medical and industrial applications, there is a growing trend towards portable ultrasound. All such systems have similar requirements for compactness and portability in remote locations. The AD9271 makes portability increasingly realizable by combining, in a tiny IC package, eight channels of the *receive* signal chain applicable to *both* pulsed- and continuous-wave Doppler systems. The AD9271 is destined to spawn a family of products providing options with lower power requirements or lower noise, pushing the boundaries even further in future generations.

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[2] ADI website: www.analog.com (Search) AD9271 (Go)
[3] ADI website: www.analog.com (Search) AD8339 (Go)
[4] ADI website: www.analog.com (Search) AD8332 (Go)

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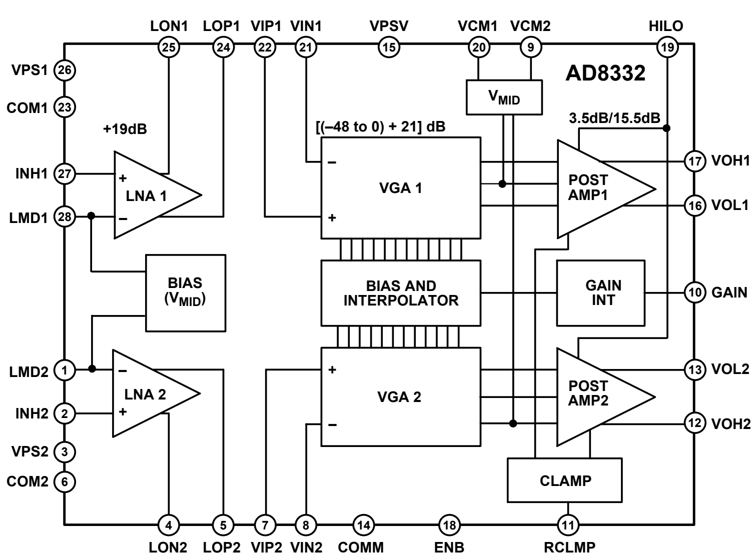
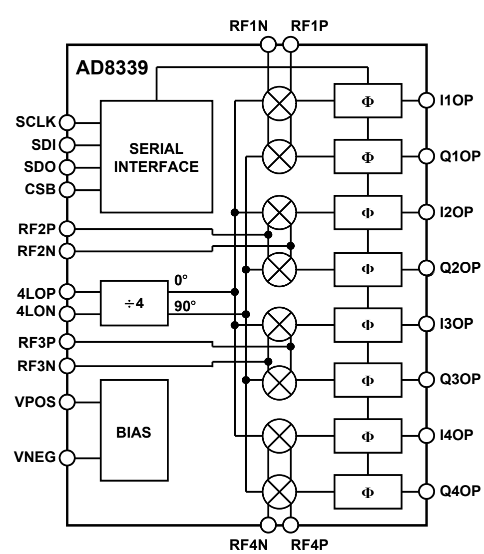
AD8332 Dual Low-Noise VGA



AD8339 Quad I/Q Demodulator



Toward More-Compact Digital Microphones

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Claus Fürst [claus.furst@analog.com]

INTRODUCTION

With more than two billion microphones sold each year, the microphone market is of interest by virtue of its volume. About half of that market is for very inexpensive, low-grade microphones for the toy market and other applications where size and performance are not critical parameters. The rest of the volume is in portable, high-end applications, such as mobile phones, headsets, digital cameras, laptops, etc. The largest players in that market are mobile-phone manufacturers, who use some 900 million devices per year. At a projected annual growth rate of 10%, mobile phones constitute the fastest-growing segment of the microphone market. Mobile phones are getting smaller while incorporating a larger number of features, calling for next-generation microphones capable of increased performance.

For many years, microphones used in telecom applications have been of the electret condenser (ECM) type. The microphone comprises a membrane, a back plate, and an electret layer. The movable membrane and fixed back plate are the plates of a variable capacitor. The electret layer stores a fixed charge corresponding to a capacitor voltage of approximately 100 V. Sound pressure will cause the membrane to move, varying the capacitance of the microphone. Since charge on the capacitor is constant, the voltage across the capacitor will vary with the changing capacitance, based on the formula for charge on a capacitor:

$$V = \frac{Q}{C}$$

Q is the charge, in coulombs, C is the capacitance, in farads, and V is the voltage, in volts. The minuscule increases and decreases in capacitance, ΔC, with sound pressure, cause proportional decreases and increases in voltage, ΔV.

$$\Delta V \cong -V\left[\frac{\Delta C}{C}\right]$$

Microphones for mobile applications are quite small, typically 3 mm to 4 mm in diameter and 1 mm to 1.5 mm in thickness. Consequently, their capacitance is also relatively small. Typical values are of the order of 3 pF to 5 pF and, in some cases, as little as 1 pF.

Having no drive strength, the signal produced by a capacitive microphone needs a buffer/amplifier prior to further processing. Conventionally, this microphone preamplifier has been implemented using a simple junction field-effect transistor (JFET). Figure 1 shows a cross-section of a packaged JFET-based ECM.

As micromachining of electret microphones has improved, microphones have become smaller, and their element capacitance has decreased. Standard JFETs no longer suffice because their relatively large input capacitance significantly attenuates the signal from the microphone cartridge element.

Fortunately, improvements in CMOS process technologies have led to improvements in amplifier circuits. Much is gained by replacing JFET-based amplifiers with CMOS analog and digital circuitry. Preamplifiers implemented in modern submicron CMOS processes have enabled, and will further enable, a wide range of improvements over traditional JFETs:

- Lower harmonic distortion
- Easier gain setting
- Multiple functional modes, including *sleep mode* for low power consumption
- Analog-to-digital conversion, enabling microphones with direct digital output
- Greatly enhanced sound quality
- Higher noise immunity

Digital-Output Microphone Preamplifier

Simple JFET-based amplifiers have inherently low power consumption, but they suffer from poor linearity and low accuracy. Thus, the main goal of improved microphone design is to combine preamplification with digital technology, increasing dynamic range through improved linearity and lower noise, while retaining very low power consumption.

Mobile phones present an inherently noisy environment. A drawback of the traditional JFET (and indeed any purely analog) solution is that analog microphone output signals can easily be corrupted by interfering signals creeping in between the amplifier and the analog-to-digital converter. Thus, incorporating analog-to-digital conversion into the microphone itself provides a digital output that is inherently less prone to corruption by interferers.

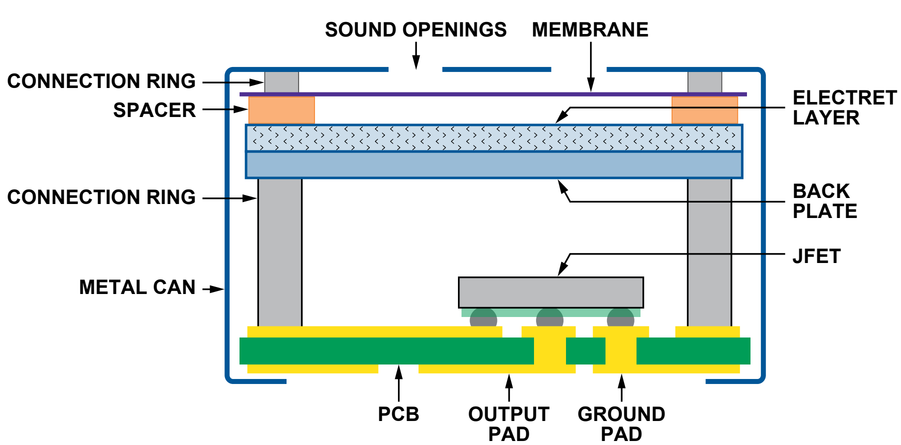

Figure 1. JFET-based microphone cross-section.

System Description

A block diagram of an integrated digital-output preamplifier and its interface is shown in Figure 2. The microphone-element signal is first amplified, and is then converted to digital by the analog-to-digital converter. These blocks receive their power from an internal regulated supply, ensuring good power supply rejection and an independent supply for the analog portion of the device.

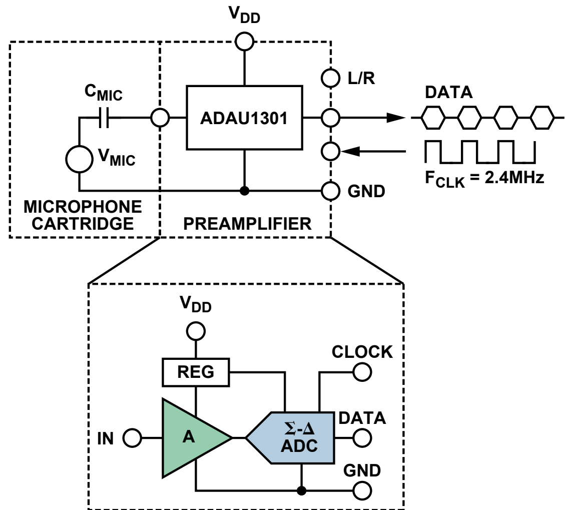

Figure 2. Digital microphone system diagram with the ADAU1301 microphone preamplifier.

The preamplifier is built in CMOS using two operational transconductance amplifiers (OTAs) in an instrumentation amplifier configuration where the gain is set using matched capacitors. This configuration, with its MOS input transistors, presents a highly desirable near-zero input admittance to the capacitive signal source. The use of capacitors for gain setting allows high gain accuracy—limited only by process lithography—and the inherently high linearity of poly-poly capacitors. The gain of the amplifier is easily set by metal-mask programming, allowing gains of up to 20 dB.

The analog-to-digital converter is a fourth-order, single-loop, single-bit Σ-Δ modulator, whose digital output is a single-bit oversampled signal. Using a Σ-Δ modulator for analog-to-digital conversion offers several advantages:

- Noise shaping shifts the quantization noise upwards, pushing much of it outside of the band of interest. Thus, high accuracy can be obtained without imposing severe matching requirements for the circuitry.
- The analog-to-digital converter uses a single-bit Σ-Δ modulator, thus making it inherently linear.
- Only one of the integrators in a single-bit, single-loop modulator requires severe design constraints. The inner-loop integrators, which have their outputs noise-shaped, have relaxed design requirements. This leads to lower power consumption.

A potential problem with higher-order Σ-Δ modulators is that they are prone to instability when the input exceeds the maximum stable amplitude (MSA). Higher-order modulators (>2) fail to return to stable operation when they become unstable due to overload, even when the input is reduced below the MSA. To counter potential instability, a digitally controlled feedback system alters the Σ-Δ noise transfer function, forcing the modulator back into stable operation.

A *power-down mode*, entered by allowing the system input clock frequency to drop below 1 kHz, lowers the current drawn by the system from 400 μA to approximately 50 μA, allowing the user to conserve power whenever the microphone is not needed. The start-up time from power-down is only 10 ms.

As a failure analysis feature, a special *test mode* enables access to various internal nodes in the circuit. A special preamble at the DATA pin during startup allows the failure analysis engineer access by switching these nodes to the DATA pin.

Noise Considerations

Three dominant noise sources in CMOS preamplifiers for capacitive microphones are flicker ($1/f$) noise, wideband white noise from the input transistors, and low-pass-filtered white noise from an input bias resistor, R_{BIAS}, needed for setting the amplifier's dc operating point. A-weighting is applied to take into account the human ear's insensitivity to low frequencies.

Flicker-noise spectral density has an inverse dependency on transistor area; its magnitude, referred to the input, is given by

$$V^2(f) = \frac{K_f}{WLC_{ox}f}$$

where K_f is a process-dependent constant, f is frequency, W is the MOS width, L its length, and C_{ox} is the gate capacitance per unit area. The $1/f$-noise amplitude can be reduced by increasing the size of the input transistors.

The input-referred white noise is inversely proportional to the transconductance, g_m, of the metal-oxide-semiconductor transistor (MOST)

$$V^2(f) = \frac{8kT}{3g_m}$$

where k is Boltzmann's constant and T is absolute temperature. For a MOST in strong inversion, $g_m \approx 2I_d/V_{eff}$, where I_d is the drain current, and the effective voltage, $V_{eff} = V_{gs} - V_{th}$, the gate-to-source voltage minus the MOST threshold voltage, V_{th}. By designing the input pair to be very wide, a bipolar-like mode of operation is imposed upon the MOST as it enters the weak inversion operating mode. Here, $g_m = I_d/(nV_T)$, where n is the slope factor (typically 1.5) and V_T is the thermal voltage. Thus, optimum white noise performance is achieved by maximizing the MOST aspect ratio.

The input bias resistor is connected to a capacitive source, so its noise will be low-pass filtered. Assuming that the noise is low-pass filtered white noise and the cutoff frequency is much smaller than the audio-band frequencies, it can be shown that the total noise power is kT/C, where C is the capacitance connected to the node.

As a consequence of the trend toward smaller microphone cartridges with lower cartridge capacitance, this noise source will increase as the microphone cartridge capacitance decreases. However, the audio-band noise power generated by the bias resistor will also depend on the cutoff frequency of the low-pass filter. The lower the cutoff frequency, the smaller the amount of the total noise power remaining in the audio frequency range. In order to keep the noise low, the value of the bias resistance will have to be increased by a factor of four for each halving of the microphone capacitance. For a 3-pF to 5-pF microphone capacitor, the resistor should have a minimum value of approximately 10 GΩ.

A good solution for implementing such large value resistors on chip is a pair of antiparallel diodes which have a very large resistance around equilibrium, typically 1 TΩ to 10 TΩ. The resistance decreases for larger signals, assuring fast settling after overload situations. Figure 3 shows the in-band noise as a function of R_{BIAS}.

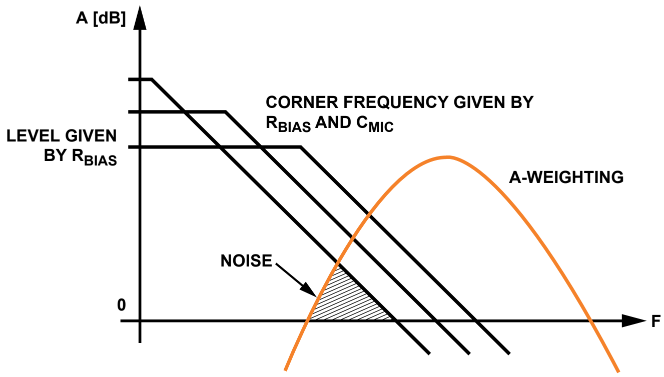

Figure 3. Noise from bias resistor.

The area of the input transistors of the preamplifier must be optimized in relation to the microphone capacitance. Although, as noted earlier, the $1/f$ noise will decrease if the input devices are made very large, the capacitive loading of the signal source will increase, attenuating the signal and reducing the wideband signal-to-noise ratio (SNR). This presents a trade-off: If the input device is made very small, the capacitive loading of the signal source becomes insignificant, but the $1/f$ noise increases dramatically, reducing low-frequency SNR. The optimum for maximizing SNR with respect to $1/f$ noise exists where the gate-source capacitance of the input device equals the microphone capacitance plus parasitic capacitance. The optimum for white noise exists where the gate-source capacitance of the input device equals one-third of the microphone capacitance plus parasitics. In practice, the best compromise is for the gate capacitance to fall between the two values.

Bootstrapping minimizes the input pad contribution to the overall chip input capacitance. As the output-referred white noise is proportional to g_m, all current-source MOSTs are biased in the strong inversion region, ensuring minimum noise contribution.

Table 1 shows the key characteristics and performance of the ADAU1301 microphone preamplifier.

Toward a Fully Integrated Digital Microphone

This digital-output amplifier fulfills the needs of ECM elements, but the combination is not fully suitable for the emerging MEMS microphone market, which will require a higher degree of integration. Since the equivalent of an electret layer does not exist among solid-state MEMS elements, the capacitive element requires an integrated high-voltage source for bias. Because the microphone element constitutes a purely capacitive load, drawing no current from the biasing reference, an extended version of this amplifier system would include a low-power on-chip charge pump, obviating the need for a stored-charge source.

CONCLUSION

A microphone preamplifier created for the mobile microphone market enables and naturally leads to the digital-output microphone. Thorough noise analysis yields an instrumentation preamplifier with low noise that attains the desired dynamic range. The low-power Σ-Δ analog-to-digital converter achieves high resolution without imposing severe design constraints. A *power-down mode* provides maximum battery life by conserving power when the microphone is not needed. A special *test mode*, designed to give the manufacturer easy access to otherwise unreachable nodes for testing, has the added benefit of making the preamplifier's analog output available for inspection.

Table 1. Typical (Unless Noted) Characteristics and Performance of the ADAU1301

Parameter	Value	Comments
Power Supply	1.64 V to 3.65 V	Functional over full range, but specified performance is at 1.8 V
Supply Current	400 μA	@ V_{DD} = 1.8 V
Max Gain Variation	$X \pm 0.4$ dBFS/V_{Peak}	X is the specified gain
Max Lower Signal Bandwidth Limit	25 Hz	
Min Upper Signal Bandwidth Limit	20 kHz	
Equivalent Input-Referred Noise Level	5 μV rms	A-weighted
Signal-to-Noise Ratio	60.6 dBFS	Calculated at –27 dBFS/Pa microphone sensitivity
Dynamic Range	>86 dB	@ THD = 10%, dependent on gain
Input Capacitance	0.1 pF	
Min Input Resistance	15 GΩ	
Start-Up Period	500 ms	Measured from time when VDD becomes 1.8 V to time when the ASIC gain settles within 1 dB of its final settled value
Max Wake-Up Period	10 ms	
Clock Frequency	1 MHz to 4 MHz	Nominal F_{clk} = 2.4 MHz
Clock Duty Cycle f_{DC}	40% to 60%	

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