Measuring nanoamperes

Paul Rako - April 26, 2007

Thousands of applications require a circuit to measure a small current. One of the most common is the measurement of photodiode current to infer the light impinging on the diode. Scientific applications, such as CT (computer-tomography) scanners, gas chromatographs, and photomultiplier and particle and beam monitoring, all require low-level current measurements. In addition to these direct applications, the manufacturers of semiconductors, sensors, and even wires must measure extraordinarily low currents to characterize their devices. Leakage current, insulation-resistance measurements, and other parameters require consistent, accurate measurements to establish data-sheet specifications.

Few engineers realize, however, that the data sheet of a part is a contractual document. It specifies the behavior of the device, and any disputes over the operation of the part always come down to the specs on the data sheet. Recently, a customer of a large analog-IC company threatened legal action against the manufacturer, claiming that the parts he had purchased exhibited far higher operating currents than the submicroampere levels that the company specified. It turned out that the PCB (printed-circuit-board)-assembly house was properly washing the board but that assemblers were picking up the PCB and leaving fingerprints on a critical node. Because it could measure these tiny currents, the semiconductor company proved that its parts were working correctly; the leakage current was due to dirty PCBs.

The difficulty with measuring small currents is that all kinds of other effects interfere with the measurement (see sidebar "History of current measurements"). This article looks at two breadboard circuits that must handle surface leakage, amplifier-bias-current-induced errors, and even cosmic rays. As in almost all circuits, EMI (electromagnetic interference) or RFI (radio-frequency interference) can induce errors, but, at these low levels, even electrostatic coupling can cause a problem. As the currents you measure drop into the femtoampere range, the circuits are subject to even more interfering effects. Humidity changes the value of capacitors and causes higher surface leakage. Vibrations induce piezoelectric effects in the circuit. Minor temperature variations, even from a room fan, cause temperature gradients in the PCB that give false readings. Even room light can degrade the accuracy of measurements; light from fluorescent fixtures can enter the glass ends of a detector diode and cause interference (Reference 1).

Small currents require accurate measurement if you want to characterize the performance of quartz-crystal oscillators. Jim Williams, a staff scientist at Linear Technology and longtime EDN contributor, shares a circuit he designed for a customer who needed to measure the rms current in a 32-kHz watch crystal (Figure 1). One difficulty with this measurement is that even a FET probe’s 1-pF loading can affect the crystal oscillation. Indeed, one of the goals of current measurement is to establish the sizing of the low-value capacitor you use with every crystal oscillator. A further difficulty of this measurement is that it must measure accurately and in real time at 32 kHz, which rules out the use of an integrating capacitor. The signal is a complex ac signal that the system
designer must convert to an rms value for evaluation.

“Quartz-crystal rms operating current is critical to long-term stability, temperature coefficient, and reliability,” says Williams. The necessity of minimizing introduced parasitics, especially capacitance, complicates accurate determination of rms-crystal current, especially in micropower-crystal types, he says. Figure 2’s high-gain low-noise amplifier, he explains, combines with a commercially available closed-core current probe to permit the measurement, and an rms-to-dc converter supplies the rms value. The dashed lines indicate a quartz-crystal test circuit that exemplifies a typical measurement situation. Williams uses the Tektronix CT-1 current probe to monitor crystal current and introduce minimal parasitic loading. A coaxial cable feeds the probe’s 50Ω output to A₁; A₁ and A₂ take a closed-loop gain of 1120, and the excess gain over a nominal gain of 1000 corrects for the CT-1’s 12% low-frequency gain error at 32.768 kHz.

Williams investigates the validity of this gain-error correction at one sinusoidal frequency—32.768 kHz—with a seven-sample group of Tektronix CT-1s. He reports that device outputs are collectively within 0.5% of 12% down for a 1-µA, 32.768-kHz sinusoidal input current. Although these results tend to support the measurement scheme, Williams contends that it is worth noting that Tektronix measured the results. “Tektronix does not guarantee performance below the specified –3 dB, 25-kHz low-frequency roll-off. A₃ and A₄ contribute a gain of 200, resulting in total amplifier gain of 224,000. This figure results in a 1V/µA scale factor at A₄ referred to the CT-1’s output. A₅’s LTC1563-2 32.7-kHz bandpass-filtered output feeds A₆ through an LTC1968-based rms-to-dc converter that provides the circuit’s outputs,” he says. The signal-processing path, Williams explains, constitutes an extremely narrowband amplifier tuned to the crystal’s frequency. Figure 3 depicts typical circuit waveforms. According to Williams, the crystal drive at C₁’s output (upper trace), causes a 530-nA rms crystal current that the A₄’s output (middle trace) and the rms-to-dc-converter input (lower trace) represent. “Peaking visible in the middle trace’s unfiltered presentation derives from parasitic paths shunting the crystal,” he says.

Williams' circuit provides several lessons. Measuring nanoamperes is difficult even when using integrating techniques. This problem was far more difficult, because he had to complete the measurement in real time. Further complicating matters was the fact that this ac measurement required a bandwidth of 32 kHz to capture the bulk of energy in the oscillator current waveform. Williams addressed these problems by using a sensor. The Tektronix CT-1 sensor (Reference 2) can cost as much as $500, but, without a good sensor, Williams would not have been able able to recover the signal from all the noise. In addition to good sensitivity, the CT-1 has a 50Ω output impedance that allows for lower noise-signal paths than would a high-impedance output. Another important principle that this example demonstrates is that it is essential to limit the bandwidth of the signal path. By making a narrowband amplifier chain, Williams discarded all the noise contributions from frequencies that were not in his area of interest. Finally, Williams used good low-noise design principles in the circuit. Wiring critical nodes in air minimizes leakage paths, and the LT1028 is perhaps the lowest noise amplifier available from any manufacturer when working from 50Ω source impedance.

**Femtoampere bias current**

Paul Grohe, an application engineer at National Semiconductor, provides another remarkable example of measuring tiny currents. Years ago, National decided to sell the LMC6001, an amplifier that had a guaranteed bias current of 25 fA, implying that National needed to measure the bias current of each part to verify the specification. The test department could not accommodate test equipment in the setup; all the circuitry had to fit onto a standard probe card. Grohe and engineering colleague Bob Pease built a proof-of-concept fixture to demonstrate the feasibility of a
small test circuit that could resolve to 1 fA (Figure 4). Many books and resources discuss using an integrating capacitor to measure small currents (Reference 3). The principle is that a small current can charge a small capacitor and that you can read that voltage to infer the current. In some cases, the current is an external current from a sensor. In this case, the current is leaving the amplifier-input pin. Figure 5 shows a simple theoretical circuit in which the amplifier is measuring its own bias current.

The reality of measuring small currents is far more involved than the figure would suggest. First, Grohe could not use the part itself to measure its own bias current. If he had tried to use the part itself as the integrator, there would have been no way to calibrate the effects of a socket and other leakages associated with the test fixture. Doing so required a separate low-bias-current part as the integrator (Figure 6). Using a CMOS LMC660 amplifier ensured that the bias-current contribution would be less than 2 fA. By employing this technique, Grohe could simply remove any DUT (device under test), and the integrator would then have measured its own bias current as well as all the leakages from the test socket and the PCB on which the integrator was mounted.

Figure 7 shows that Grohe did not insert the DUT into a socket and that none of the pins are in contact with a PCB. To minimize leakage, Grohe brought up just two power pins as long, separate individual sockets that he did not mount to a PCB. Likewise, he hooked the pin to be tested to a socket and a 2-in. flying lead and connected that pin-and-socket combination to the integrating-amplifier input. To keep the DUT from running as an open loop, Grohe soldered together two sockets to bridge the output pins, which are suspended in air. Air currents can carry charged ions that can give false readings, so Grohe enclosed the entire DUT in a shielded copper-clad box.

The next issue was selecting an integrating capacitor. Initially, Grohe felt that the best capacitor would be an air-dielectric capacitor, so he fashioned two large plates, measuring about 4×5 in., for the integrator capacitor. The size of this capacitor accounts for the size of the second copper-clad box on which the DUT box is mounted. Using a large capacitor proved to be a bad idea. The large area provides an ample target for cosmic radiation, creating ionized charges that interfere with the measurement (Figure 8). Grohe then minimized the capacitor’s size while still using a good dielectric. It occurred to him that RG188 coax cable uses Teflon insulation. A 2-in. section of this cable provided the 10 pF for the integration capacitor (Figure 9). As a further benefit, the outside braiding would serve as shield. Grohe therefore hooked it to the low-impedance-output side of the amplifier. With the switch to this capacitor, the cosmic rays struck only once every 30 seconds or so. Grohe took the integrated measurement for 15 seconds and, by taking five measurements, negated their effect. Any ionizing radiation sources, even an old watch with a radium dial, can cause cosmic-ray problems. Note that Grohe pried up the input pin of the amplifier to prevent leakage from the PCB.

Before taking a measurement, you need to reset the integrating capacitor to zero. Using a semiconductor switch is impractical, because of leakage currents and the 5- to 20-pF capacitance most analog switches offer. That capacitance exhibits the varactor effect, as well; it changes with applied voltage, further complicating measurement. To minimize these problems, Grohe used a Coto-reed relay. Knowing that the coil might couple to the internal reed when the relay was open, he specified a relay with an electrostatic shield. Much to his dismay, there still was a large jump in the measurement when the relay opened due to charge injection. It turns out that you can also look at a reed relay as a transformer, with the reed assembly representing a single turn. This phenomenon explains the failure of the electrostatic shield to prevent the interference. Magnetic fields inducing voltages in the high-impedance side of the circuit caused the charge injection. The relay does not open instantaneously, and the pulse needed to energize the coil makes a significant current injection just before the relay opens. Grohe minimized this problem by characterizing the absolute minimum
voltage swing needed to operate the relay he had installed. It turned out that the relay would pull in with 3.2V and drop out with 2.7V. He used a set of resistor taps on an LM317 adjustable regulator to control the output between these two values. By choosing not to energize the relay with a full 5V, he minimized the jump in the integrator output and made it repeatable. He then nulled out the jump by injecting a small current into the second gain-stage amplifier.

The gain stages are two low-noise amplifiers—the LMV751 or perhaps a chopper amplifier, such as the LM2011, would be suitable. Grohe sent this gained-up signal to a digital scope, which could record data and subtract the slope of the calibration run from the test runs to give a valid measurement. Grohe used two LS123-style one-shot circuits—one to trigger the relay and another to provide a suitable and repeatable time delay that triggered the digital scope.

Grohe also understood that good low-noise-design principles also include the power rails to the parts, so he chose not to power the relay or digital circuits from the same power he used for the integrator and DUT. He used a handful of fixed and variable regulators to provide ±5V for the DUT and integrator, 8V for the relay-drive circuit, and a separate 5V for the digital circuits.

Using this circuit, Grohe was easily able to resolve 1 fA of current and found that most of the LMC6001 parts he tested had less than 5 fA of bias current, far exceeding the spec. He used this breadboard as the basis for a production-test circuit mounted on a standard probe card. (See reference 4, reference 5, and reference 6 for more about his design, including a video of the system.)

Grohe would not use this circuit to measure femtoampere currents in his lab. “I would wheel out the Keithley 2400 electrometer,” he says (Editor’s note). “We would have used that instrument to test the LMC6001 in manufacturing had the fab allowed us to use external test equipment.” His faith in Keithley is well-placed. The company offers free on its Web site an excellent article on measuring attoamperes (Reference 7), as well as a book on delicate measurements (Reference 8).

**DDC112**

Grohe and Pease’s integration approach is not limited to laboratory setups. Texas Instruments has created a line of parts that can measure in the femtoampere range and provide a digital output to boot. The line includes a single-channel DDC101 as well as the improved-sensitivity, dual-channel DDC112, which provides for external integrating capacitors. The four- and eight-channel DDC114 and DDC118 have a charge sensitivity of 12 pC (Reference 9). The sample rate for these 20-bit parts reaches 3 kHz.

You must be cognizant of physics to attempt these measurements. If the DDC112 can measure 12 pC of charge and you want to measure 12 pA of current, you need to set the integrating time to 1 second, the maximum the DDC112 allows. It is impossible to obtain a 3-kHz update rate if the part’s integration interval is a full second. However, using the part configured in this fashion yields a 20-bit value at the end of the conversion. In other words, the DDC (direct digital converter) can resolve femtoampere currents, although at reduced accuracy. The input bias of the part is 20 fA, but your system’s software can calibrate out this value, so the part should still be able to resolve to very low levels. Bear in mind that this type of sensitivity makes it difficult to calibrate the system just once in the factory and then have it work for all time. As temperature increases, the bias current increases, doubling every 10°C, and leakages as well as sensor drift can develop on your board. Providing the means for field calibration at power-up or more frequently is always a good idea when measuring currents in the femtoampere range. Texas Instruments offers evaluation boards for these parts that you can get up and running in hours, measuring currents too small for even a good handheld digital voltmeter (Figure 10).
According to Jim Todsen, product-line manager for oversampling converters at TI and patent holder on the technology that the part uses, the DDC line’s development started with the Burr-Brown ACF2101—a dual switched integrator front end that provides a single-chip option for the current-t-voltage function. The benefit of a dual integrator, Todsen explains, is that it is always collecting input current. While one integrator is sampling the input, the other side is presenting its integrated value to the ADC, and this process continues for as long as you need measurements. “After the ACF2101 converts the input current to a voltage,” he says, “a discrete high-resolution ADC digitizes it. The DDC112 brought together both the current-to-voltage function of the ACF2101 and the digitization of the high-resolution ADC in one chip.” He attributes this achievement to advances in wafer processing that allow high levels of mixed-signal integration as well as TI’s development of a high-speed delta-sigma core that can provide the required speed and resolution to measure the front-end signals. “In addition,” he notes, “we took advantage of having all the circuit elements under our control to optimize for very-low-leakage inputs and very stable performance over long integration periods.”

These applications should convince you of the difficulty of measuring small currents. They should also convince you of the value of using proven parts and equipment—whether Analog Devices’ AD549, National Semiconductor’s LMC660, TI’s DDC114 integrated circuits, Keithley’s 2400 parameter-measurement unit (Editor’s note), or Agilent’s 4156 parameter-measurement unit—in this demanding application. Remember, though, that these remarkable parts and instruments are not magic boxes. You can take advantage of them only by removing noise sources and leakage paths from your board or test setup. Understanding op-amp specifications for voltage and current noise will help you select the right part (Reference 10). In the meantime, if your boss wants to know why you need $5 or $10 for a chip or thousands of dollars for an electrometer, you can now explain that, with the challenges entailed in measuring small currents, this equipment is a bargain.

Editor’s note
After EDN published this article, some alert readers wrote in to say that the Keithley 2400 is a source meter, not an electrometer. For details, see "Keithley & TEGAM on measuring nanoamperes" from Paul Rako’s blog.

References
1. Long, James, "Sidebands be gone, or let there be (no) light," EDN, Oct 12, 2006, pg 40.
2. "AC current probes".
5. Pease, Bob, "What's All This Femtoampere Stuff, Anyhow?" Sept 2, 1993.
History of current measurements
Measuring small currents is difficult. Measuring extremely small currents is extremely difficult, and may turn you into a two problems—physics and noise. A recent monograph on measuring the physics issue (Reference 3). The measurement problem is further complicated by the need to measure 0.1 nA currents and may be required using a 1 kHz bandwidth. Several respondents pointed out the problem with noise. A femtometer is 10^-15 electrons. To be measurable, 0.001% of 1 fA is 1 pC. You'll need a needle to see the warble. The natural limit to noise is set by quantum effects, and thermal noise. If your current consists of 8 electrons per second, counting statistics predict that your SNR will drop to 0 dB in the measurement bandwidth of 1 kHz. A femtometer is 10^-15 electrons, so assuming that you want at least a 20 dB SNR (because that’s below where you haven’t yet got a measurement readout), your bandwidth is going to be limited to 0.1 Hz.

This answer applies to the other problem with measuring extremely small currents. The noise can manifest itself as a voltage noise or a current noise. In addition, there is noise from shot effects, leakage current, and additive noise. Even the smallest noise on your board can cause thermocouple effects that interfere with your low-level measurements (Reference 4). You cannot simply slap a JFET monitor across the source and expect to turn that 1 kHz into a 1 Hz. Efforts to measure 0.001% of 1 fA is 1 pC. You’ll need a needle to see the warble. The natural limit to noise is set by quantum effects, and thermal noise. If your current consists of 8 electrons per second, counting statistics predict that your SNR will drop to 0 dB in the measurement bandwidth of 1 kHz. A femtometer is 10^-15 electrons, so assuming that you want at least a 20 dB SNR (because that’s below where you haven’t yet got a measurement readout), your bandwidth is going to be limited to 0.1 Hz.

Leakage mismatch of the ESD diodes. Like JFET leakage, the unfortunate fact is that the ESD-diode leakage also may be either into or out of the pin, depending on the part. Even parts of the same lot may have different required on every manufacturable part. These diodes have reverse leakage that does not exactly match between the parts. For example, a 100-kΩ resistor has a voltage spectral density of 4 × 10^-9 V/Hz at 100 kHz. Although JFETs have low leakage current, it is hard to beat CMOS MOSFETs for leakage. Because the gate is connected to the substrate, the reverse-biased diode exhibits leakage on the order of picoamperes. This means that the input capacitance of the op amp must be much smaller to make these voltages meaningful. Indeed, the ADM1191 power monitor can resolve down to microamperes, and nanoampere measurements are almost routine. The ability to measure picoamperes in the applications section that describes the low-noise and low-leakage design techniques necessary when measuring small currents.

Topgate technique that electrically isolated the top- and back-side gates to achieve input-bias currents, on the other hand, is desirable, because the surface of the die always contains lattice defects that contribute to noise. The downside of Topgate is that it increases the cost of the device. In the 1980s, National Semiconductor introduced the LF1546, a part that integrated the LF156 into the amplifier itself. There is a further benefit of JFETs even over MOSFETs. Due to their structure in hybrid-type packages. In the 1975, National Semiconductor introduced the LF156, a part that integrated the LF156 into the amplifier itself.

JFETs in hybrid-type packages. In the 1975s, National Semiconductor introduced the LF156, a part that integrated the LF156 into the amplifier itself. This process amplifies the ac voltage with all the benefits of rejecting the 1/f noise and then synchronously generating or amplifying noise, and the mass of the meter movement damped out all the high-frequency noise. This technique is fine for low-frequency measurements, but may not be suitable for use in the test lab. In this clever circuit, the input voltage changes the bias across two varactor capacitors. Varactor capacitors have a voltage-dependent capacitance. The bias changes the capacitance, which in turn changes the capacitance value with applied voltage. Opposite, shows the how the two capacitors create a small ac voltage when the input voltage makes one varactor capacitance bigger and the other smaller.

In the late 1960s, researchers used this same ac-amplification principle in its varactor-bridge electrometer (Reference 5). In this clever circuit, the input voltage changes the bias across two varactor capacitors. Varactor capacitors have a voltage-dependent capacitance. The bias changes the capacitance, which in turn changes the capacitance value with applied voltage. Opposite, shows the how the two capacitors create a small ac voltage when the input voltage makes one varactor capacitance bigger and the other smaller. This process amplifies the ac voltage with all the benefits of rejecting the 1/f noise and then synchronously generating or amplifying noise, and the mass of the meter movement damped out all the high-frequency noise. This technique is fine for low-frequency measurements, but may not be suitable for use in the test lab. In this clever circuit, the input voltage changes the bias across two varactor capacitors. Varactor capacitors have a voltage-dependent capacitance. The bias changes the capacitance, which in turn changes the capacitance value with applied voltage. Opposite, shows the how the two capacitors create a small ac voltage when the input voltage makes one varactor capacitance bigger and the other smaller.

Because the input of a JFET is a reversed-biased diode, JFETs exhibit leakage currents on the order of picoamperes. It is an unfortunate fact that this leakage doubles with every 10°C rise in temperature. JFETs are not used in high-temperature environments. In the 1970s, analog companies made parts that used discrete MOSFETs instead of discrete JFETs. The 1040, National Semiconductor introduced the 1040. A 100-kΩ resistance has a voltage spectral density of 4 × 10^-9 V/Hz at 100 kHz. Although JFETs have low leakage current, it is hard to beat CMOS MOSFETs for leakage.

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