A popular category of aiming/pointing aids is the reflex, or “red-dot,” sight. This system finds use in such diverse applications as astronomy, archery, and shooting. In the reflex sight, light from an internal source—typically a high-intensity red LED—reflects from a curved, transparent optical (reflex) element through which you view the target. The result of this geometry is that the image of the LED (the red dot) appears superimposed on the target image, thus indicating the point of aim. When you correctly adjust the aiming point of the telescope, bow, or gun, the target and LED images coincide. The reflex sight offers several advantages over competing pointing technologies, such as telescopic and open sights. These benefits include rapid and intuitive target acquisition, noncritical eye positioning, and a wide field of view.

For best sight performance, the intensity of the red-dot light source must at least roughly match the illumination level of the target. Otherwise, if the source is too dim, the aim-point dot loses itself in the brightness of the target. If too bright, the dot flares, and its apparent size increases, obscuring the point of aim and making precise pointing difficult or impossible. For this reason, most reflex sights require manual adjustment of the source intensity. Although this adjustment is effective enough, the time and attention needed to optimize intensity with a manual control detracts from the fast and intuitive target-acquisition capabilities of the red dot. The circuit in Figure 1 uses phototransistor $Q_1$ to sense target brightness and automatically adjust the LED output. The circuit maintains near-constant dot size over a wide range of ambient-light levels.

Potentiometer $R_1$ divides $Q_1$’s photocurrent, $I_p$, between the LED driver, $Q_2$, and the bias transistor, $Q_3$ (connected as a diode). The adjustment of $R_1$ therefore determines the ratio between drive current, $I_1$, and ambient (target) intensity over the range of 1 to B, where B is the beta of $Q_2$ (greater than 100). The prototype of the intensity-control circuit was packaged in a small plastic enclosure attached to the side of a Compasceco Inc (www.com pasceco.com) Tech Force model 90 30-mm objective reflex sight. The light shield mimics the field of view of the sight, so the light that $Q_1$ samples represents the target intensity visible through the sight. Proper adjustment of $R_1$ results in good compensation of dot intensity for a wide range of both incandescent and natural light. The circuit effectively maintains a constant angular dot diameter of 4 minutes of arc under outdoor ambient lighting ranging from dark overcast to full sunlight. The circuit also delivers similar performance under indoor incandescent-lighting conditions. Compensation with fluorescent lamps, however, is less satisfactory because of the absence of an adequate near-infrared component in the spectrum of these light sources. You could probably fix this shortcoming by using a suitable visible-light filter in front of $Q_1$.

This simple circuit automatically adjusts red-dot intensity in reflex optical sights.

Figure 1

Circuit controls intensity of reflex optical sights

indicator features expanded scale
16-bit ADC provides
19-bit resolution
Microcontroller emulates numerically controlled oscillator
Method simplifies testing high-Q devices

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Test equipment for a production line should be user-friendly (“idiot-proof”) and should offer minimal test time. In many cases, the test fixture must give an operator only one answer: pass or fail. Usually, two indicators assume this role: green for pass and red for fail. In most applications, a sensor transforms the tested parameter to a voltage; the test fixture must measure this voltage and display the result. Sometimes, an operator needs to observe the dynamics of the tested parameter to verify that the results are inside the permitted “green zone.” For example, when evaluating a regulated system’s behavior, the operator is often interested in measuring the parameter’s deviation and estimating its average value after the process reaches steady state (Figure 1). In this situation, using an analog meter with marked red and green zones is preferable to using a digital or bar-graph LED display.

Assume that the range of the tested voltage is 4.75 to 5V. To make a voltmeter with maximum 5V reading by using a 100-μA dc meter, you would use a series resistor of 50 kΩ. The meter scale is linear, and the tested voltage zone represents only 5% of full-scale (Figure 2). It is difficult for an operator to observe the meter reading inside such a narrow zone. It would be desirable to expand the test zone to, say, 90% of full-scale (Figure 3). The circuit in Figure 4 does just that. When the tested voltage, $V_{\text{TEST}}$, is lower than the threshold voltage, $V_2$, the diode, $D_1$, does not conduct, and the voltmeter comprises the microammeter and resistor, $R_1$. When the tested voltage surpasses the threshold $V_2$, the diode conducts, and resistor $R_2$ connects in parallel with $R_1$. The voltmeter’s impedance decreases, thereby expanding the measurement scale. You can calculate the values of resistors $R_1$, $R_2$, and $R_3$ as follows:

1. The voltage at the beginning of the tested zone, 4.75V, should consume 10% of the scale, with a corresponding current of 10 μA. Hence, neglecting the internal meter impedance, $R_1$ equates to $4.75V/10\,\mu A$, or 475 kΩ.
2. After the measured voltage exceeds the threshold level, 4.75V, the voltmeter impedance equals to $R_{1,2} = (5-4.75)V/(100-10)\,\mu A$, or 2.8 kΩ.
3. Hence, $(R_{1,2})/(R_1-R_2) = (475\times2.8)/(475-2.8) = 2.8 \, k\Omega$, and $R_3 = R_2(V_{CC}/V_2-1) = 2.8 (5/4.75-1) = 147\Omega$.

Any data-acquisition systems require both high accuracy and a fast acquisition rate. These attributes allow the system to detect small data signals and to group more sensor channels into the same system. With more channels, the system can be smaller, less expensive, and less power-consuming. Long-distance optical communications and medical equipment, such as CT scanners, can benefit from a fast and accurate data-acquisition system. Optical power systems, such as laser pumps, need to constantly monitor their power levels. In such systems, the incoming laser-power range and the laser control-loop response time are such that the system needs a dynamic range of 90 dB or more and a sampling rate of 1 Msample/sec. In CT scanners, 16- to 22-bit resolution is necessary for the data-acquisition system to process the large dynamic range of the X-rays through various body tissues. A large number of photodetectors (more data-acquisition channels) and high data accuracy improve the image resolution.

These two examples show the need for relative accuracy, as opposed to absolute accuracy. Although it’s important to be able to detect a 10-nW change in an optical power of 1 μW, it is almost irrelevant to see the same 10-nW difference between 1 mW and 1.00001 mW. However the ADC’s accuracy appears under the integral-linearity specification as an absolute error. For the best relative accuracy, a classic solution is to use a programmable-gain amplifier in front of an ADC. The AD7677 ADC specifies ±15 ppm of full-scale nonlinearity (±1 LSB at the 16-bit level). A programmable-gain amplifier ahead of this converter must be able to settle

This system provides 19-bit accuracy by combining a programmable-gain amplifier with a 16-bit ADC.
quickly enough with the same resolution and speed as the ADC. It also must have the lowest noise possible, because the amplifier sets the SNR of the data-acquisition system. To meet these challenges, the amplifier in this design uses an AD8021, an op amp combining speed, accuracy, and fast settling time. The noise density of the AD8021 is only 2 nV/\sqrt{Hz}. Figure 1 shows how the gain settings of the programmable-gain amplifier divide the specified accuracy of the ADC. The system reaches 19-bit accuracy when the input level is low.

Relative accuracy is normally specified as parts per million of reading plus or minus the absolute minimum error. The circuit in Figure 2 can achieve a relative accuracy of 107 ppm±1.9-ppm maximum error. Analog multiplexer IC4 combines many lower bandwidth channels to take advantage of the 1M-sample/sec sampling rate of the ADC. Because the programmable-gain amplifier presents a high input impedance to the multiplexer, you can cascade the multiplexer, thus increasing the number of channels. The multiplexer also provides a simple way to calibrate the offset and gain errors at each gain setting by applying a calibration-reference voltage to one of the multiplexer’s input ports. You need to calibrate only at power-up or when operating conditions, such as temperature, change. The amplification chain comprises the multiplexer, the comparator, and the amplifier on one side and the ADC on the other side. The successive-approximation structure of the AD7677 ADC allows the individual sections in the amplification chain to work simultaneously. While the ADC converts one sample, the comparator/amplifier can settle the following channel. Therefore, the data-acquisition system can operate at the ADC’s maximum sampling rate.

Shortly after the analog multiplexer settles, the fast comparator, IC3, applies the appropriate gain setting. The comparator’s thresholds are such that the amplifier does not saturate or clip the signal after amplification by IC6 and IC7. The AD8561 comparator has a response time of 7 nsec. It integrates a latch signal that holds the gain constant during the time the amplifier settles and the ADC acquires the signal. The usual programmable-gain-amplifier configuration requires the user to predict the amplifier’s gain setting before applying the signal at the input. The programmable-gain amplifier in Figure 2 has an “autorange” feature that selects the most appropriate programmable-gain amplifier gain to maximize accuracy without incurring saturation or clipping. The comparator incorporates hysteresis to reduce gain-setting change when signals are close to the limits of an individual gain range. The circuit automatically boosts the ADC accuracy to 19 bits while maintaining a full-speed sampling rate of 1M sample/sec.

IC6 amplifies the multiplexer signal using one of two possible gain settings: 1 or 8. You can modify the feedback network to provide different gains to a maximum of 25. The analog switch, IC3, controls the gain setting. The high gain-bandwidth product of the AD8021 op amp provides more than enough bandwidth, so its compensation capacitor remains the same for all gains. Amplifier IC6 generates the differential signal for the ADC. The settling times of the comparator and the amplifier and the acquisition time of the ADC are all significantly less than the ADC’s full conversion period of 1 \mu s. The RC noise filters at the two ADC inputs, R1/C1 and R2/C2, use this extra time. These filters limit the noise bandwidth of the programmable-gain amplifier, which is the main noise source of the data-acquisition system when IC3 operates at a gain of 1.

Figure 3 shows the circuit’s nonlinearity. The photo shows a maximum nonlinearity of 0.44 LSB and a minimum of –0.37 LSB for the highest gain setting, which poses the most difficult challenge. This nonlinearity corresponds to a typical error of ±0.9 ppm. At a gain of 8, output noise is 85 \mu V rms. If desired, you can further reduce the noise by using software averaging. Figure 4 shows the complete data-acquisition system assembled using the AD7677 evaluation board. The printed-circuit-board area measures 15×30 mm.

Microcontroller emulates numerically controlled oscillator

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Microcontrollers commonly add intelligence or digital functions to products, but they can also provide a variety of analog signals. An 18-pin PIC 16C54 microcontroller, combined with an inexpensive, 8-bit DAC and a simple lowpass filter, can generate sine waves from dc to approximately 50 kHz with a tuning resolution of 24 bits. The accuracy and stability of the output is as good as that of the crystal driving the microcontroller. You can connect a binary data signal to one or more PIC ports to apply FSK (frequency-shift-keying), BPSK (binary-phase-shift-keying), or QPSK.

LISTING 1—16C54 NUMERICALLY CONTROLLED OSCILLATOR-EMULATION ROUTINE
Method simplifies testing high-Q devices

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The design of low-phase-noise oscillators requires careful attention to resonator unloaded Q. In the construction of a low-phase-noise, high-frequency oscillator, the goal is to achieve an unloaded-Q figure greater than 400 in a reasonable package. Also, you need to monitor the effect of the package and PCB-board arrangement. Shielding, inappropriate grounding, and some construction techniques can degrade unloaded Q, Q meters; various bridges, such as Maxwell and Hayes; and both vector and scalar impedance analyzers are useful but inconvenient-to-use test instruments. You must carefully set up test fixtures and calibration that duplicates the final environment to obtain reasonable agreement with the final measured results. A simple test set uses nothing more than the voltage-divider relation with the device under test embedded as a series trap network (Figure 1). You can measure the inductor’s value, or calculate its value from known equations based on the inductor’s form factor, such as solenoid, toroid, helical, or flat spiral. You use the inductor’s value to select $C_1$, a variable, air-dielectric high-Q capacitor. At resonance, the impedance of the inductor-capacitor combination goes to zero, so the effective load is the series resistance $R$ in parallel with the 50Ω termination resistance.

You use an RF generator and voltmeter to read the depth of the notch the trap creates. This attenuation depth is a function of the remaining finite-series resistance, a figure greater than 400 in a reasonable package. To create the other three quadrants, bit 6 determines whether to read the table forward or backward, and bit 7 specifies the output sign. An exclusive-OR operation on bit 7 with a port bit generates BPSK operation. An exclusive-OR operation on both bits 6 and 7 with port bits generates QPSK modulation. The result goes to the DAC via the PIC’s 8-bit output port. A 50-kHz lowpass filter then converts the DAC’s output into a smooth sine wave.

You can preset the output frequency or load it serially via two pins of the PIC’s 4-bit port. You obtain FSK by using an input bit to select which of two frequency-control registers to use. If the two frequencies have a large common multiple, as in minimum-shift keying, the accumulator can be shorter, leading to a higher output frequency for a given clock input. Without modulation, the firmware loop can be as short as 26 instruction times (Listing 1). You can insert non-operative instructions to make the loop-repetition rate a convenient submultiple of the crystal frequency. For example, a 31-instruction loop and a 20-MHz crystal yield a scale factor close to 104 steps/Hz.

The code takes advantage of a quirk in the 16C54’s operation: If two addresses exist on the return stack, the first copies endlessly into the second every time the routine pops the second. The initialization code puts two copies of the loop-start address into the return stack, causing all subsequent RETLW instructions to jump to the start of the loop. Indexing into the look-up table with a calculated GOTO instruction both supplies an output sample and executes a jump to restart the loop. This procedure is much faster than executing a CALL, a GOTO, a RETLW, and a further GOTO. You can download Listing 1 from the Web version of this article at www.ednmag.com.

ance of the resonator. Table 1 shows the notch attenuation for $R_S$ ranging from 0.1 to 1Ω. These values assume 50Ω source and termination impedance and the component values shown in Figure 1. Unloaded $Q$ equates to $X_L/R_S$, where $X_L$ is the reactance of the inductor, and $R_S$ is the equivalent series resistance. Figure 2 shows the notch attenuation as a function of the equivalent series resistance. In addition, a crosscheck is available: You can determine the unloaded $Q$ from the expression $Q=X_L/R_S$. For example, a solenoid inductor measuring 0.75 in. in diameter and wound with five turns of six-gauge wire has a measured inductance of 460 nH at 65 MHz. The inductor series-resonates at 65 MHz with a 13-pF capacitor. You set the signal generator at 65 MHz and use a variable, air-dielectric capacitor to fine-tune the notch at 65 MHz. The measured notch depth is 36 dB. $R_S$ is 0.4Ω, and the unloaded $Q$ is 469. You can readily notice changes in the depth of the notch with fine variations in coil position relative to conducting surfaces.