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Use an off-the-shelf signal source as a jitter/wander generator

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E NSURING THAT new networked products, such as routers, gateways or DSLAMs (digital-subscriber-line-access multiplexers) meet stringent timing specifications usually requires a specialized jitter/wander generator. As a substitute, you can use a standard function generator equipped with PM (phase modulation) or FM (frequency modulation) to measure jitter and wander tolerance. This Design Idea describes how to convert PM and FM parameters (phase deviation, frequency deviation, and modulating frequency) into jitter/wander parameters—amplitude in UIs (unit intervals) and frequency.

Network-communications engineers use the terms "jitter" and "wander" to describe phase noise in digital signals. "Wander" refers to phase noise at frequencies below 10 Hz, and "jitter" refers to phase noise at frequencies at or above 10 Hz. Defining phase noise requires specifying both its amplitude and its frequency.

As **Figure 1** shows, if you observe a clock with phase noise on an oscilloscope triggered by a clock of the same frequency but without phase noise, the rising and falling edges of the noisy clock appear blurred—that is, not clearly defined in time. If the clock has low frequency-phase noise (wander), the rising and
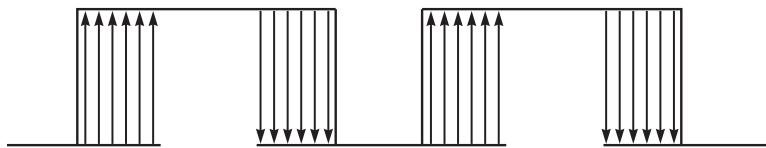

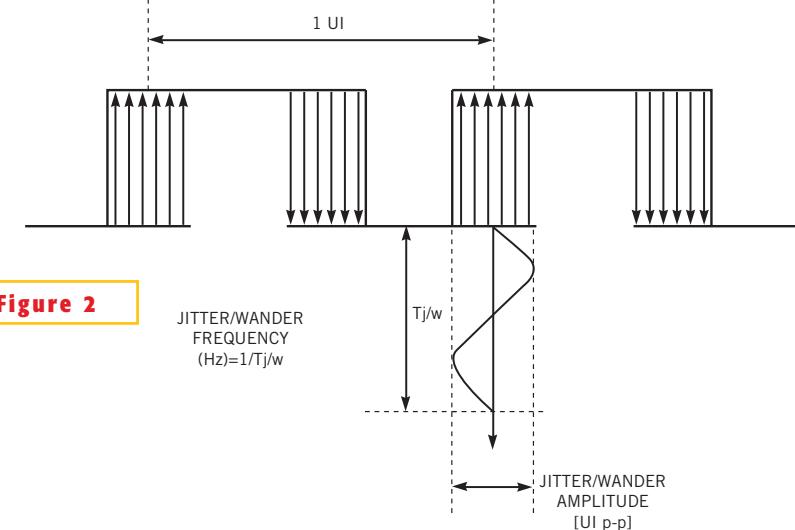**Figure 1**   The effects of jitter or wander appear as position modulation of a pulse's leading and trailing edges.

Figure 2

JITTER/WANDER FREQUENCY (Hz)=1/Tj/w

1 UI

Tj/w

JITTER/WANDER AMPLITUDE [UI p-p]

Applying an external signal source to a signal generator's phase- or frequency-modulation input produces a jittery output waveform.

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falling edges move back and forth at a rate equal to the wander frequency. The range of this movement defines the jitter/wander amplitude.

Figure 2 illustrates an instance of sine-wave-shaped FM of jitter or wander. You can express jitter or wander amplitude in UIs; one UI is equal to the clock period. For example, the amplitude of the jitter/wander in **Figure 2** is 0.25 UI p-p.

You can use a signal generator to generate waveform jitter and wander by connecting a low-frequency signal source to the signal generator's PM or FM input. **Equation 1** applies to both FM and PM and describes the general form of an an-

gle-modulating signal:

$$s(t) = A\cos[2\pi f_C t + \theta(t)]. \qquad (1)$$

Although in digital communications, s(t) usually approximates a square-wave function, using a square wave instead of a sine wave complicates the math but doesn't affect the process of angle modulation. For simplicity, this Design Idea uses a sine-wave function for s(t).

For PM, the phase $\theta(t)$ in **Equation 1** is proportional to the modulation signal:

$$\theta(t) = D_{PM}\cos(2\pi f_m t), \qquad (2)$$

where D_{PM} is the phase deviation (peak variation of the phase), and f_m represents the modulating frequency, which is also the jitter/wander frequency. The rela-

tionship between phase deviation and jitter/wander amplitude is straightforward, and you can obtain it from:

$$\text{JITTER/WANDER}[\text{UI p-p}] = D_{PM}/180°, \quad (3)$$

where D_{PM} has units of radians in communications theory, but, for convenience, most signal generators specify units of degrees instead.

For FM, the phase $\theta(t)$ in **Equation 1** is proportional to the integral of modulating signal.

$$\theta(t) = 2\pi D_{FM} \int \cos(2\pi f_m t) dt, \quad (4)$$

where D_{FM} is the frequency deviation (peak variation of the frequency) and f_m is the modulating frequency. The FM modulating frequency is the same as the jitter/wander frequency. **Equation 5** yields the jitter/wander amplitude:

$$\text{JITTER/WANDER}[\text{UI p - p}] = \frac{D_{FM}}{\pi f_m}, \quad (5)$$

which derives from **Equation 6**:

$$\theta(t) = 2\pi D_{FM} \int \cos(2\pi f_m t) dt =$$
$$2\pi D_{FM} \frac{1}{2\pi f_m} \cos\left(2\pi f_m t - \frac{\pi}{2}\right). \quad (6)$$

Therefore, the peak-to-peak deviation of $\theta(t)$ is:

$$\theta_{PP} = 2\frac{2\pi D_{FM}}{2\pi f_m} = \frac{2D_{FM}}{f_m}, \quad (7)$$

in which the factor of 2 originates from the peak-to-peak amplitude of a sine-wave function. To get jitter/wander in peak-to-peak unit intervals, divide **Equation 7** by the period of the sine-wave function, 2π. Thus, you get **Equation 6**, which is valid only when the modulating signal comprises a sine wave because the integral of a sine wave is also a sine-wave function shifted in phase. Fortunately, most jitter/wander-tolerance tests almost exclusively use sine-wave modulation.

Some signal generators specify modulation in phase and frequency span instead of phase and frequency deviation. For PM, the span is the peak-to-peak variation of the phase, and, for FM, the span is the peak-to-peak variation of the frequency. That is, the span equals twice the deviation for both PM and FM. In this case, the jitter/wander amplitude for PM is:

$$\text{JITTER / WANDER}[\text{UI p - p}] = \frac{\text{SPAN}_{PM}}{360°}, \quad (8)$$

and it is:

$$\text{JITTER / WANDER}[\text{UI p - p}] = \frac{\text{SPAN}_{FM}}{2\pi f_m}. \quad (9)$$

for FM.□

Temperature controller saves energy

Tito Smailagich, ENIC, Belgrade, Yugoslavia

GIVEN THE HIGH COST of electrical power, replacing a conventional on/off temperature control with a proportional controller can often save energy and money. **Figure 1** shows a low-cost, high-efficiency, time-proportional temperature controller for a residential water heater. An Analog Devices ADT14,

IC_1, serves multiple functions as a temperature sensor, quad-setpoint, programmable analog temperature monitor and controller. Resistors R_1, R_2, R_3, R_4, and R_5 adjust desired temperature at setpoints SETP1, SETP2, SETP3, and SETP4, which IC_1 compares with the actual temperature from its internal sensor. The

ADT14's active-low open-collector outputs drive Input Port A of IC_2, an 8-bit Motorola/Freescale 68HC908QT4 microcontroller that provides 4 kbytes of flash memory, 128 bytes of RAM, and an on-chip clock oscillator.

Available at *EDN*'s online version of this Design Idea at www.edn.com, **List-**
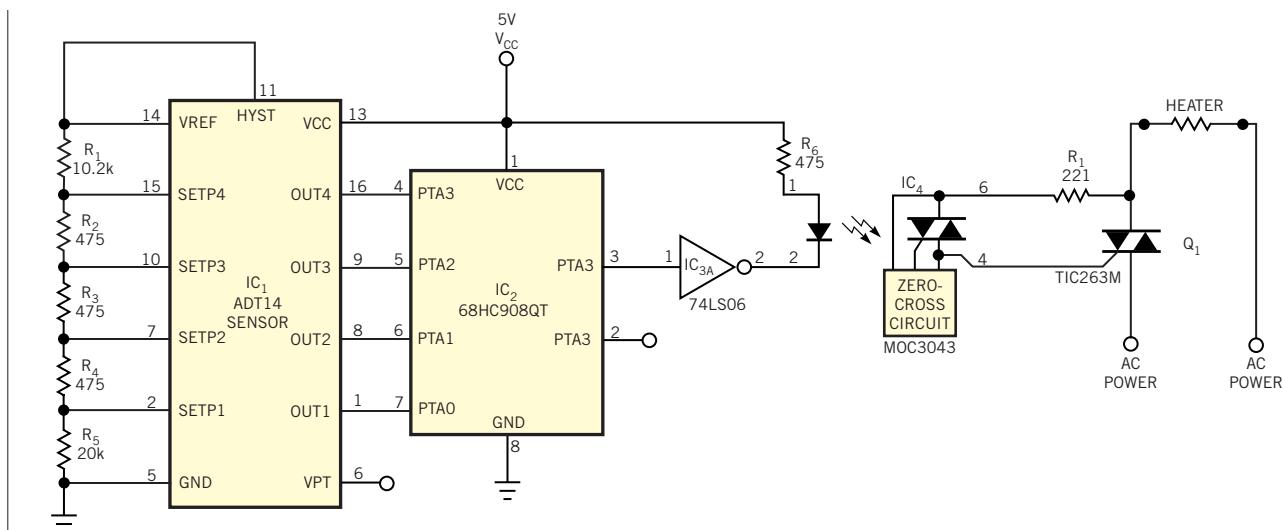

Figure 1 This proportional temperature controller features a minimal parts count.

ing 1 contains commented assembly-language software. When you load it into the microcontroller's flash memory, the software provides the time-proportional control algorithm. When IC_1's OUT1, OUT2, OUT3, and OUT4 outputs are inactive, IC_2 switches its output PTA4 to a totally on-state, 100% duty cycle for maximum heating. **Listing 2** at the Web version of this Design Idea contains an assembled version of the software, and **Listing 3** presents the hex code for programming IC_2.

When IC_1's OUT1 output is active, IC_2 produces a 75%-duty-cycle output on PTA_4. In similar fashion, when IC_1's OUT2 output goes active, IC_2 produces a 50%-duty-cycle output on $PTA4$, and when IC_1's OUT3 output goes active, IC_2 produces a 25%-duty-cycle output on

PTA_4. When IC_1's OUT4 output goes active, IC_1 disables the output on PTA4 to produce a totally off state (0% duty cycle). **Table 1** summarizes the relationship of IC_2's inputs and output duty cycle.

To minimize component count, IC_2's internal oscillator generates a 12.8-MHz clock that divides to produce a sample pulse whose basic width is 0.1 sec for each 1% of output on-time. One cycle of output comprises 100 samples for a total duration of 10 sec. Thus, for a 25% duty cy-

cle, IC_2's output PTA4 generates a 2.5-sec on interval followed by a 7.5-sec off interval. One section of an open-collector hex inverter, IC_{3A}, a 74LS06, drives optocoupler IC_4, an MOC3043, which features an internal zero-crossing circuit and pilot triac. Power triac Q_1, a TIC263M rated for 600V and 25A, controls application of power to the water heater's 2-kW resistive heating element. For best results, place IC_1 in close thermal contact with the water heater's inner tank.□

TABLE 1—IC₂'S LOGIC STATES VERSUS OUTPUT DURATIONS

PTA3	PTA2	PTA1	On (%)	Off (%)	On (sec)	Off (sec)
1	1	1	100	0	10	0
1	1	1	75	25	7.5	2.5
1	1	0	50	50	5	5
1	0	0	25	75	2.5	7.5
0	0	0	0	100	100	10

Calculator program finds closest standard-resistor values

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ALTHOUGH IT MAY not appear obvious to newcomers to the electronics-design profession, components' values follow one of several progressions that divide a decadewide span into equally spaced increments on a logarithmic scale. For example, when you plot the values of 1, 2.2, and 4.7 on a logarithmic scale, they divide the range 1 to 10 into three roughly equal increments (1...2...5). To meet requirements for greater precision, resistor manufacturers offer parts in several additional series. The most precise series divide a decade into 24, 48, or 96 increments by computing $10^{n/m}$, where $n=1...(m-1)$,

and $m=24$, 48, or 96, and then rounding the values to two or three digits. The results are the R_{24}, R_{48}, and R_{96} series and respectively contain 24, 48, or 96 values per decade.

You can use a Hewlett-Packard HP-48 or HP-49 calculator and one of the following programs written in RPN (Reverse-Polish Notation) to compute the nearest standard value that's closest to a required value. You enter a required resistor value, and the program returns the closest higher or lower value in the selected series. **Table 1** lists a few examples.

Each program acts as an operator by processing the first line of the calculator's stack and returning the new value in the same line of the stack. The R_{48} and R_{96} series are mathematically exact, and their programs consist of only a single line of code. The **List-**

ings at the Web version of this Design Idea at www.edn.com show the code. The values of the older R_{24} series are not as strictly rounded, and the program is thus somewhat more complex.

Note that the values of other components, such as capacitors, inductors, and zener diodes, also follow preferred-value series, making these programs universally applicable. You can view an earlier version of a standard-value calculator for IBM-compatible PCs at *EDN*'s online version of Design Ideas. David Kirkby of the Department of Medical Physics, University College London, UK, wrote the program in C. *EDN* first presented it, "Resistance calculator yields precise values," in the Aug 3, 1995, issue. You can read the instructions at www.edn.com/archives/1995/080395/16di5.htm. Note that certain portions of the software may require rewriting for better operation on today's PCs.□

TABLE 1–INPUT VALUE IN SELECTED SERIES RETURNS A CLOSEST VALUE OF:

47.8	R_{24}	47
490	R_{24}	510
12.2	R_{96}	12.1
12.3	R_{96}	12.4

Reduce voltage-reference output noise by half

Alfredo H Saab and Steve Logan, Maxim Integrated Products Inc, Sunnyvale, CA

Reducing low-frequency (1/f) noise generated by an IC voltage reference can prove difficult. In theory, adding a lowpass filter to a reference's output reduces noise. In practice, a lowpass RC filter for suppression of noise frequencies below 10 Hz requires large values of series resistance and shunt capacitance. Unfortunately, a high-value series resistor introduces resistance errors and thermal noise, and a shunt capacitor's leakage resistance forms an unpredictable and unstable shunt path. Together, the two components form a noisy and temperature-dependent voltage divider that directly affects the reference's accuracy and long-term stability. In addition, pc-board surface contaminants can add yet another possible leakage path and error source.

You can stack multiple voltage references in series to reduce their 1/f noise. The references' dc outputs add linearly, and their uncorrelated internal noise sources add geometrically. For example, consider a stack of four voltage references, each comprising a dc reference source, V_{REF}, in series with a random-noise generator, V_{NOISE}. Adding four reference sources produces the following outputs: $V_{REFTOTAL} = 4 \times V_{REF}$, and $V_{NOISETOTAL} = \sqrt{4 \times (V_{NOISE})^2} = 2 \times V_{NOISE}$. The original ratio of noise voltage to dc reference voltage thus divides in half.

Figure 1 illustrates a method of adding multiple references to produce a single, less noisy reference voltage. The resistors are parts of a highly stable metal-film network, and buffer amplifier IC_5 offers low noise, low input-offset voltage, and low offset-temperature coefficients.

Tables 1 and **2** present the noise voltages that result from stacking four each of two types of 2.5V references. Each **table** shows the 0.1- to 10-Hz noise voltage for each of the four references, IC_1 through IC_4, and for the combination. Note that the dispersion in the ratios of rms to peak-to-peak values relates to subjectivity in the method of measuring the values. In addition to lower 0.1- to 10-Hz noise, the circuit also reduces long-term drift of the reference voltage.□

TABLE 1—NOISE VOLTAGES MEASURED IN FIGURE 1 USING FOUR 2.5V MAX6037 VOLTAGE REFERENCES

Measurement points	Noise (μV rms)	Noise (μV p-p)
Reduced noise output (op amp's output to V−)	1	10
Across Reference A (IC₁) (OUT to GND)	1.9	20
Across Reference B (IC₂) (OUT to GND)	1.6	19
Across Reference C (IC₃) (OUT to GND)	1.7	20
Across Reference D (IC₄) (OUT to GND)	2.7	30

TABLE 2—NOISE VOLTAGES MEASURED IN FIGURE 1 USING FOUR 2.5V MAX6143 VOLTAGE REFERENCES

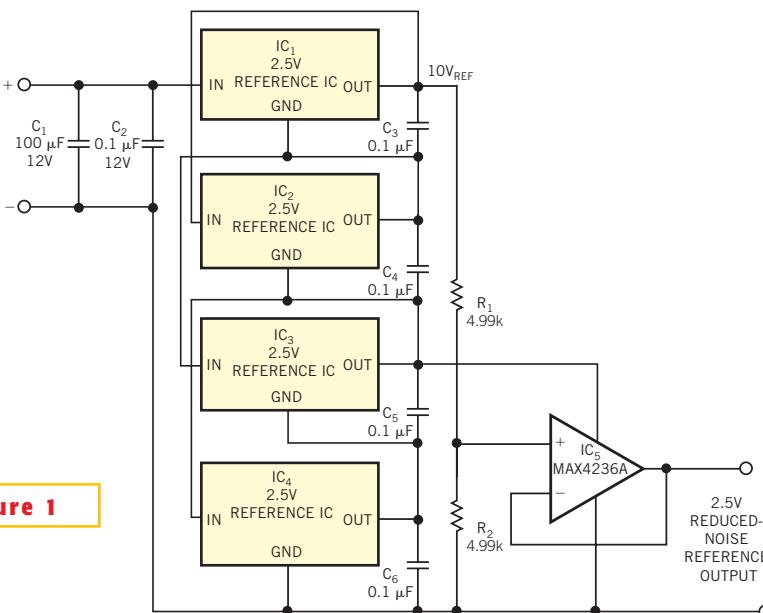| Measurement points	Noise (μV rms)	Noise (μV p-p)
Reduced noise output (op amp's output to V−)	0.27	2.2
Across Reference A (IC₁) (OUT to GND)	0.52	4.7
Across Reference B (IC₂) (OUT to GND)	0.6	4.8
Across Reference C (IC₃) (OUT to GND)	0.5	4.3
Across Reference D (IC₄) (OUT to GND)	0.55	4.7

Figure 1

Four 2.5V references, IC₁ through IC₄, produce 10V. Resistors R₁ and R₂ form a voltage divider that reduces the 10V output to 2.5V and lowers the output-noise voltage by half. Buffer amplifier IC₅ isolates the reference circuit from the load.

Touch switch needs no dc return path

Brad Albing, Philips Medical Systems Inc, Cleveland, OH

COMMON DESIGNS FOR touch switch-es detect a decrease in resistance when a user's fingertip either con-nects a contact to the circuit's common ground or supplies an injection of 60-Hz ac voltage, resulting from immersion in the electrostatic field that nearby power lines radiate. But what if no nearby pow-er lines exist and the equipment operates from a battery source, such as in an au-tomotive application, or if a galvanic con-tact to circuit common is unavailable?

The circuit shown in **Figure 1** operates by sensing an increase in capacitance that results from touching a contact. Al-though a straightforward design might require a complex circuit, the design

shown offers a low-cost approach that uses few components.

In **Figure 1**, IC_{1A} operates as a square-wave oscillator at approximately 150 kHz. The oscillator's output gets ac-coupled to potentiometer R_2 that sets the drive level and, hence, the sensitivity for the touch pad. Applying negative excursions of sev-eral volts of square-wave signal to its gate repetitively drive N-channel JFET Q_1 from conduction into cutoff. An approx-imation of the square wave swinging from 0 to 12V appears at Q_1's drain. A peak detector circuit formed by D_1, R_7 and C_4 provides sufficient dc voltage to force IC_{1B}'s output to a logic low.

However, if someone touches the

touch pad, any added capacitance to ground or circuit common reduces the ac drive at the FET's gate, and Q_1 continu-ously conducts. The square-wave voltage applied to D_1 decreases. The voltage on C_4 drops below the logic threshold, and IC_{1B}'s output goes high. You can adjust R_2 to set sensitivity and compensate for de-vice-to-device variations in the FET's pinch-off voltage. For novelty or nostal-gia's sake, you can use one-half of a 12AX7 dual triode as an oscillator and the remaining half in place of Q_1. Selecting plate resistors allows operation with a 12V plate power supply.□
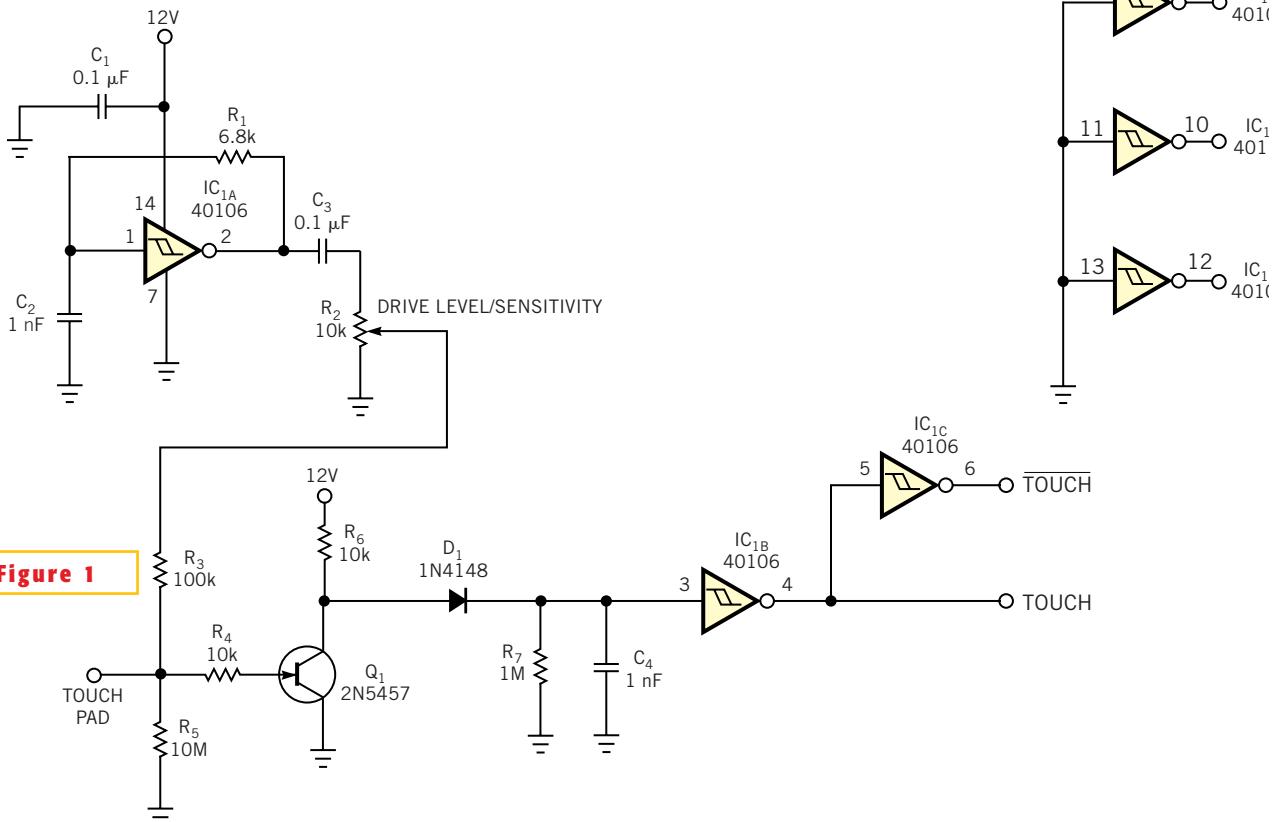

Figure 1

A low-cost touch-switch interface uses three Schmitt trigger hex inverters and a single JFET per channel.