design ideas

Edited by Bill Travis

Build a precision deadband circuit

V Manoharan, Kochi, India

DEADBAND CIRCUITS find applications in servo-control systems. A precision current source and a half-wave inverting rectifier form a positive deadband circuit (**Figure 1**). The REF01, IC_1, is a precision 10V voltage reference. It forms a precision current source with the addition of a unity-gain buffer (IC_{2A}) and resistor R_1. IC_{2A} forces the ground pin (Pin 4) of IC_1 to assume the potential at IC_{2A}'s noninverting input. IC_1 forces its highly accurate reference voltage (10V) across R_1, so the current I_1 through R_1 is $10V/R_1$. Because the inverting input of IC_{2B} connects

to the output of the current source, the feedback diode, D_1, becomes forward-biased with current I_1. The forward-biased diode keeps the output of IC_{2B} at approximately $-0.6V$. Because the cathode end of D_2 stays at virtual ground through R_3, D_2 remains reverse-biased. Hence, V_{OUT} remains at virtual ground (0V). Any positive voltage applied to V_{IN} further forward- and reverse-biases D_1 and D_2, respectively, and the output remains at 0V (deadband zone) for $V_{IN} > 0V$.

Negative inputs tend to forward-bias D_2 and reverse-bias D_1. This bias situation occurs only when the current through R_2 (because of the negative V_{IN}) equals or exceeds $10V/R_1$. So, the output is 0V (deadband) until V_{IN} reaches a value equal to $-10R_2/R_1$. If you choose $R_1 = 20$ kΩ, V_{OUT} remains at 0V for $V_{IN} > (-5V)$; for $V_{IN} \leq (-5V)$, $V_{OUT} = (-V_{IN} - 5V)$. **Figure 2** shows the transfer function for this scenario. C_1, C_2, and C_3 are decoupling capacitors for IC_1 and IC_2. R_4 reduces the offset voltage of IC_{2B} in the nondeadband region. You could realize an alternative
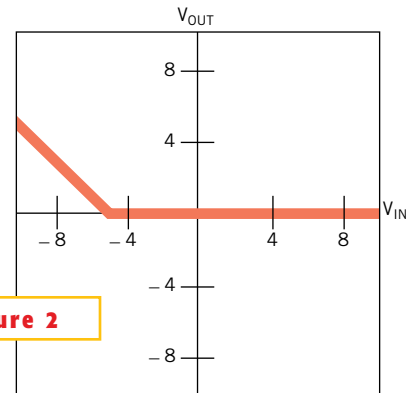

Figure 2

With $R_1 = 20$ kΩ, the circuit in Figure 1 is "dead" for input voltages higher than $-5V$.

straightforward circuit by prebiasing the inverting half-wave rectifier through a precision resistor connected to a voltage reference without using the unity-gain buffer, IC_{2A}. However, this alternative would increase the noise gain, thereby increasing the offset and noise at the output.

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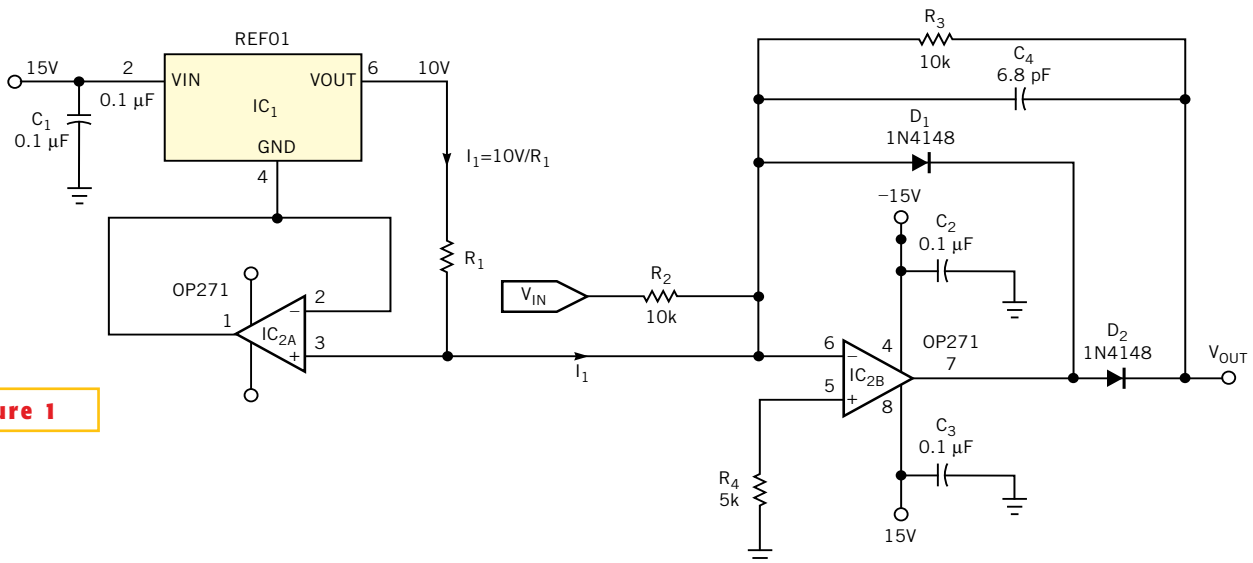**Publish your Design Idea in *EDN*. See the What's Up section at www.edn.com.**

Figure 1

This circuit exhibits a deadband for signals more positive than an arbitrary voltage, which R_1 determines.

DDS and converter form signal generator

Colm Slattery, Analog Devices, Limerick, Ireland

MANY APPLICATIONS require low-frequency signal generators that can deliver high-performance, high-resolution signals. This Design Idea presents a circuit that generates frequencies of 0 to 1 MHz. Sinusoidal, triangular, and square-wave outputs are available. You can achieve frequency resolution of better than 0.1 Hz and phase resolution of better than 0.1°; thus, you can program exact coherent frequencies. This feature is useful in digital modulation and frequency-tuning applications. The circuit uses the ADμC831 and AD9834 to generate the required frequencies (**Figure 1**). You can program the microcontroller from either a PC or a Unix-based workstation. You then program the AD9834 using a three-wire serial interface via the microcontroller. The interface word is 16 bits long.

You can program the AD9834 to provide sinusoidal, triangular, and square-wave outputs using the DDS (direct-digital-synthesis) architecture. The chip operates as an NCO (numerically controlled oscillator) using an on-chip, 28-bit phase accumulator, sine-coefficient ROM, and a 10-bit D/A converter. You typically consider sine waves in terms of their magnitude form, $A(t)=\sin(\omega t)$. The amplitude is nonlinear and is, therefore, difficult to generate. The angular information, on the other hand, is perfectly linear. That is, the phase angle rotates through a fixed angle for each unit in time. Knowing that the phase of a sine wave is linear, and, given a reference interval (clock period), you can determine the phase rotation for that period: $\Delta Phase=\omega$ dt; $\omega=\Delta Phase/dt=2\pi f$, and $f=(\Delta Phase \times f_{MCLK})/(2\pi)$, where $dt=1/f_{MCLK}$, and f_{MCLK} is the master clock.

Using this formula, you can generate output frequencies, knowing the phase and master-clock frequency. The phase accumulator provides the 28-bit linear phase. The amplitude coefficients of the output sine wave are stored in digital format in the sine-coefficient ROM. The DAC converts the sine wave to the analog domain. If you bypass the ROM, the AD9834 delivers triangular waveforms instead of sinusoidal waveforms. A

Figure 1
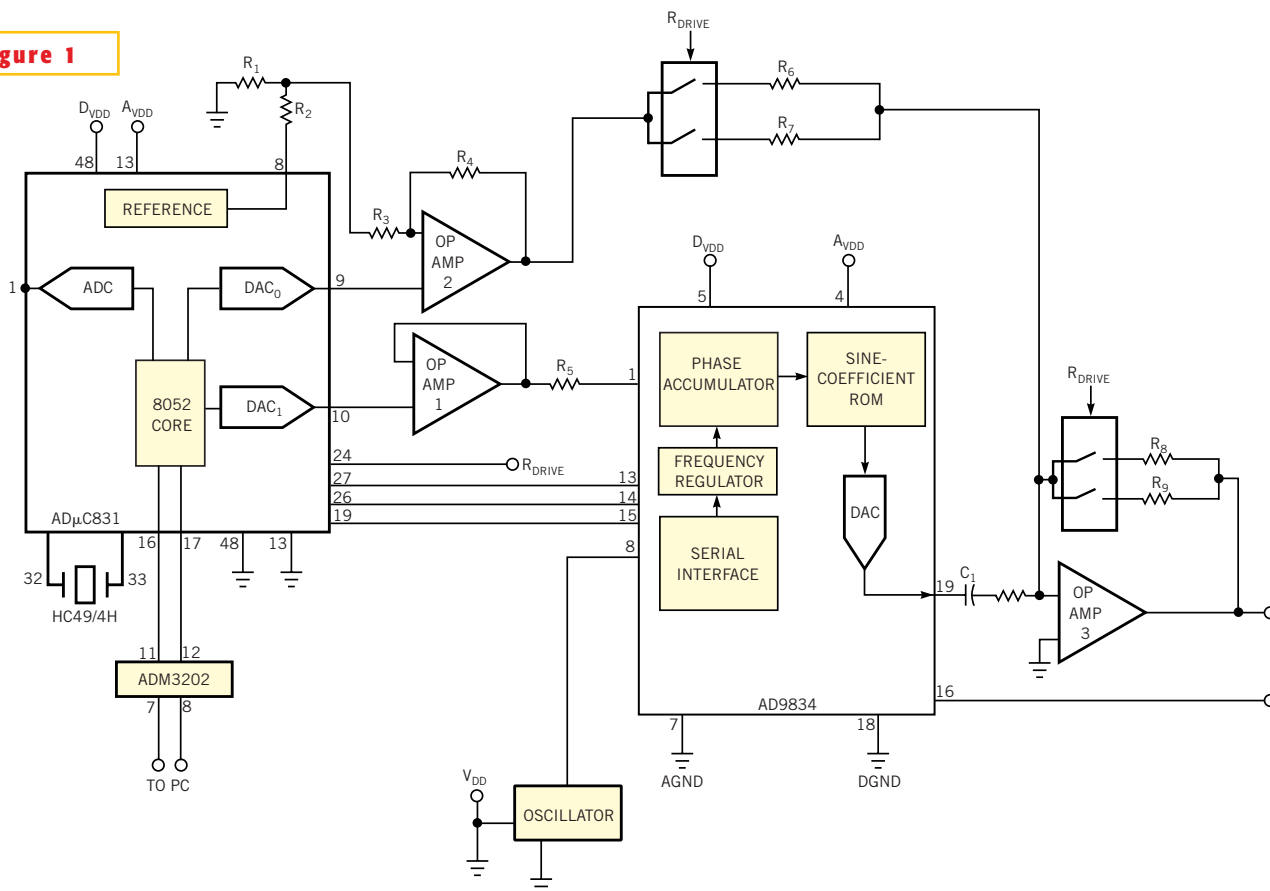
A DDS chip and a microcontroller combine to form a multiwaveform signal generator.
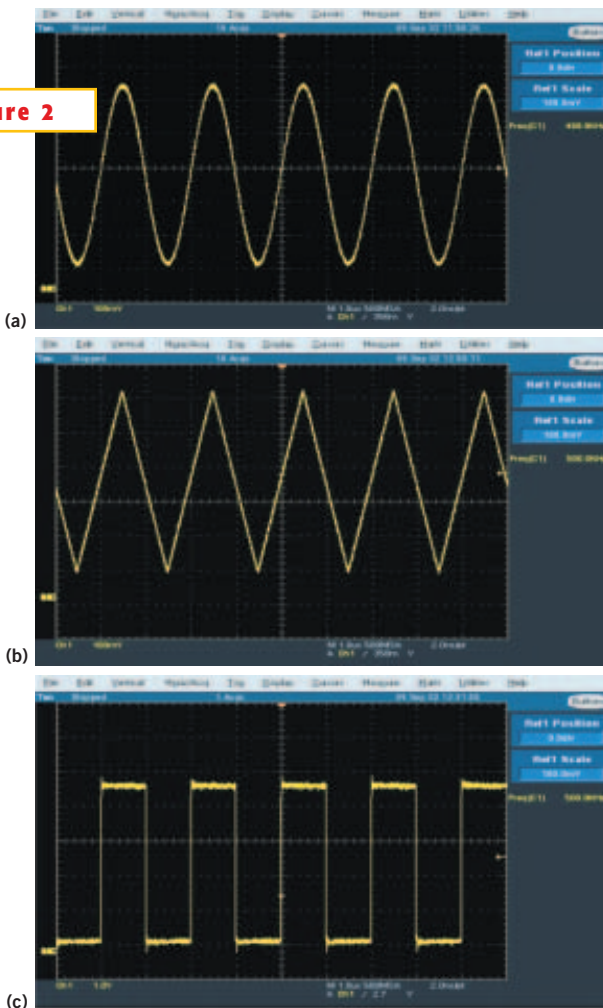
square-wave output is also available on the part. **Figure 2** shows the various waveforms available from the system. As shown in **Figure 1**, the sinusoidal/triangular output waveforms are available on the IOUT pin (Pin 19); and the square wave output is available on the SIGN_BIT_OUT pin (Pin 16). You program the DDS by writing to the frequency registers. The analog output from the part is then: $f_{OUT}=f_{MCLK}/228\times$ (frequency-register word).

The outputs of the DDS have 28-bit resolution, so effective frequency steps on the order of 0.1 Hz are possible to a maximum of approximately 1 MHz. **Figure 2** shows the typical waveform outputs. Two phase registers are available that allow 12-bit phase resolution. These registers phase-shift the signal by: Phase shift $=2\pi/4096\times$(phase-register word).

A 50-MHz crystal oscillator provides the reference clock for the DDS. The output stage of the DDS is a current-output DAC loaded by an external resistor. A 200Ω resistor generates the required peak-to-peak voltage range. The output is ac-coupled through capacitor C_1. The MicroConverter contains two on-chip, 12-bit DACs. DAC_1 varies the current through R_5, adjusting the full-scale cur-

**Figure 2**

(a)

(b)

(c)

Sinusoidal (a), triangular (b), and square-wave (c) waveforms are all available from the circuit in Figure 1. The 500-kHz waveforms all use a 50-MHz sampling rate.

rent of the DDS via the FSADJUST pin. The equation to control the full-scale current of the DDS DAC is: I_{OUT} (full-scale)$=18\times I\times R_5$.

DAC_0, the internal reference of the MicroConverter, and op amp 2 allow for offset control of the output voltage of the DDS. You can program this dc offset to ±10 V at 10-bit resolution. When $R_1=R_2$ and the gain of op amp 2=8, then the output of op amp 2 is: $V_{OUT}=$ (DAC output$-(V_{REF}/2))\times8$, yielding a ±10V range.

Resistors R_6 through R_9 allow for control of gain through op amp 3. The gain of the op amp is a function of resistor switching, which you enable using the R_{DRIVE} pin available on the MicroConverter. This operation allows for an effective programmable-output amplitude of approximately ±10V p-p. Thus, the circuit allows for programmable sinusoidal and triangular waves, including dc offsets, and the ability to set peak-to-peak amplitude of approximately ±10V. The square wave output on the SIGN_BIT_OUT pin has 0 to 5V amplitude. For low-frequency operation, a lowpass filter normally serves to filter reference-clock frequencies, spurs, and other images. For applications in which the output signal needs amplification, you should use a narrowband filter to filter out unwanted noise before the gain stage. A third-order filter would be good enough to remove most of the unwanted noise. **Figure 3** shows a typical spectral plot of the output. Applications for this circuit range from signal-waveform generation to digital modulation. You can use the system in frequency-sweeping and -scanning applications and in resonance applications that use the frequency as an excitation signal to determine circuit resonance. Another useful application is as a reference oscillator for a PLL system.

Figure 3
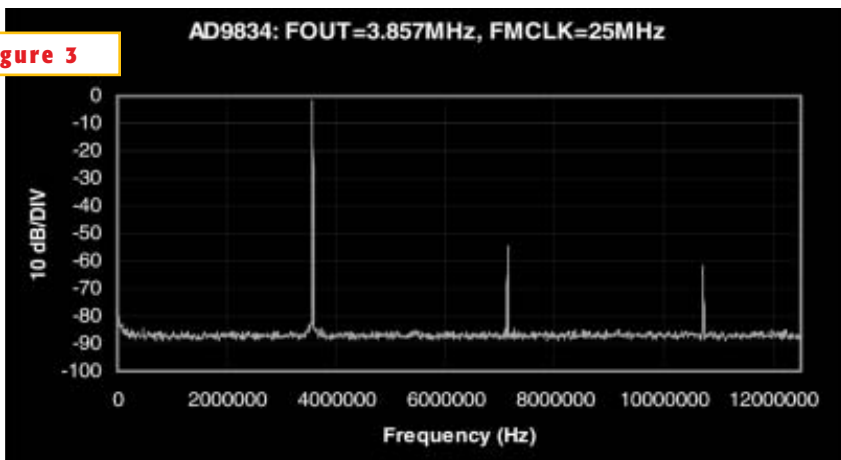
This spectral plot shows the fundamental, second harmonic, and third harmonic for a 3.857-MHz signal.

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48-input×16-output crosspoint IC eliminates the need for multiplexer amps

Mike Hess, Maxim Integrated Products, Sunnyvale, CA

CROSSPOINT SWITCHES are ideal for use in video-security systems, which accept multiple camera inputs while providing playback and multiple loop-through to multiple monitors. To provide video loop-through or monitor outputs, these systems often require additional multiplexer amplifiers that can drive standard video loads. Thus, one or more external multiplexer amps often follow a crosspoint-matrix switch. As an alternative, you can employ a 32×16 nonblocking crosspoint-matrix switch, whose 16 2-to-1 multiplexers eliminate the need for extra multiplexer amps (**Figure 1**). The internal 2-to-1 multiplexers appear before the video-output buffers, which can each directly drive a standard video load. That configuration eliminates the need for external multiplexer amps and their associated cost, space requirement, and power consumption.

The MAX4358 IC is a fully buffered, 32-input×16-output nonblocking crosspoint switch that includes 16 additional buffered analog-video inputs (OSDFILL) intended for the insertion of OSD (on-screen-display) information. ("Nonblocking" means that the IC can route any input to any output.) The 16 fully buffered OSDFILL analog inputs are identical to the 32 inputs of the buffered crosspoint-switch matrix, so the 16 additional video inputs can implement a single-chip, 48-input×16-output crosspoint-switch matrix. The output buffers feature programmable gain (A_V=1V/V or 2V/V). The programmability allows versatility in routing short video traces or driving video-transmission lines. Operating from dual ±3 to ±5V supplies or a single 5V supply, the device reduces power consumption by as much as 60% versus standard ±5V-only ICs.

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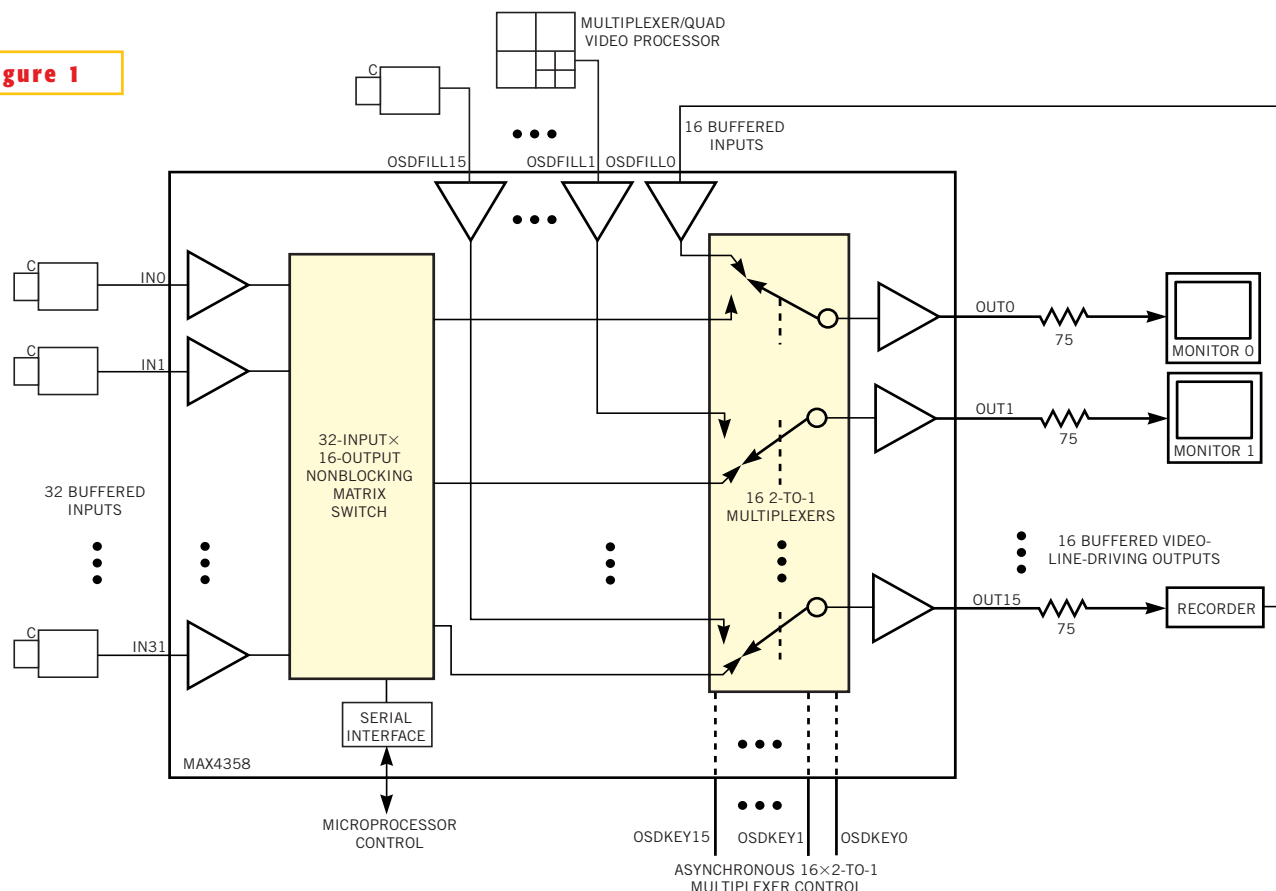
Internal buffers allow this IC to implement a 48-input×16-output matrix switch without the need for external multiplexer amps.

Transmission line tests 1-kW device using only 100W

David Cuthbert, Meridian, ID

A 13.56-MHz, ISM (industrial, scientific, and medical)-band RF-measurement device had the test requirement of a 50-hour, 1-kW burn-in. The device under test needed to be simultaneously stressed at an RF potential and RF current equivalent to 1 kW, but the only spare RF source at hand was a 100W RF generator. Besides, saving energy seemed important. The circuit in **Figure 1** develops 1 kW from the 100W RF source by storing energy in a transmission line. The circuit comprises two 43° sections (approximately 6 feet) of RG-213 coaxial cable with UHF connectors at each end. The device under test, which has an electrical length of 4°, connects between the two lines, T_1 and T_2, making the total line length 90°. An off-the-shelf amateur-radio antenna

Figure 1
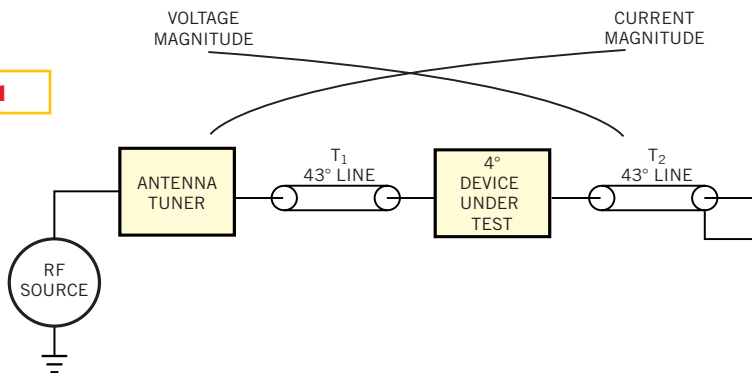
As if by magic, this setup allows you to apply 1 kW to a device using only a 100W generator.

tuner matches the 50Ω RF generator to the line's input impedance.

Circuit operation is simple. The RF energy connects to the line's input through the antenna tuner. The energy travels to

the shorted end of the line, where it is reflected and travels back to the input. The reflected energy then reflects off the conjugate match, which the antenna tuner provides, and combines with the next

half-cycle of input energy to flow toward the short again. This process continues with the stored energy continuing to build until the circuit losses equal the generator power. Considering the circuit operation in terms of impedances reminds you that the input impedance of a shorted, lossless, 90° transmission line is infinite. At the shorted end of the line, V/I is zero, and, at the input of the line, V/I is infinite. At the center, where the device under test is located, the magnitude of V/I is equal to the characteristic impedance of the transmission line—50Ω in this case. The RF voltage and current are 90° out of phase, but that fact does not affect the burn-in of the device.

Consider how much line input power is required to develop 1 kW at the device under test. The loss of each 6-foot section of RG-213 is 0.025 dB, and the device-under-test loss is 0.05 dB. The loss for a wave traveling down the line is therefore 0.1 dB. The return loss, R_L, is twice this amount, or 0.2 dB, because the wave must travel down the line to the short circuit and return to the

source. Now, you can calculate the reflected power, P_R, for an incident power, P_{IN}, of 1000W using the following formula: $P_R = (P_{IN})10^{(RL/10)} = (1000)10^{(-0.2/10)} = 955W$.

So, when 1000W flows down the line, 955W returns to the input. The line input power required is equal to the incident power minus the reflected power, which is $1000-955$, or 45W. Because the line loss and device-under-test loss are both 0.05 dB, half of the 45W loss is dissipated in the coax, and half is dissipated in the device under test. The measured antenna-tuner loss is 40W, which makes the total circuit loss 85W. You can determine the line's input impedance by calculating the line-input complex reflection coefficient (Γ) and solving for the input impedance using $\Gamma = 10^{(RL/20)} = 10^{(-0.2/20)} = 0.9772$, and

$$Z_{IN} = \frac{Z_{LINE}(\Gamma + 1)}{1 - \Gamma} = \frac{50(0.9772 + 1)}{1 - 0.9772} \approx 4.3 \text{ k}\Omega \text{ .}$$

The antenna tuner must match the 50Ω generator's output impedance to the 4.3-kΩ line-input impedance. You can confirm the circuit's operation by measuring the amplitude and phase of the device's voltage and current. The measurement uses an oscilloscope with voltage and current probes. A power meter at the device under test measures forward and reflected power of 1 kW. Because of the high circuit Q, you'll find that adjusting the RF-source frequency to obtain circuit resonance is easier than trimming cable lengths to obtain circuit resonance. The primary limiting factors with this circuit are the temperature rise of the coaxial cable and the losses incurred in the impedance-matching circuit. Coax having lower loss allows you to achieve a higher "power multiplication" and higher device-under-test power.

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