Edited by Brad Thompson

Laser simulator helps avoid destroyed diodes

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LASER DIODES CAN destroy themselves in a few nanoseconds, so testing the response and stability of a feedback-stabilized laser-diode driver can be expensive. The simulator circuit in **Figure 1** shows a typical laser-diode package, which contains not only the diode, which is driven by current I_L, but also a photodiode. The laser diode's front facet emits a main beam that goes to work in the outside world, and the rear facet emits a reference beam that falls on the photodiode.

Although much weaker than the main beam, the reference beam's power is directly proportional to the main beam, as is the current, I_P, that produces the photodiode. Connecting the photodiode back to the laser-diode driver via a carefully designed amplifier completes a feedback loop that should hold the main beam power stable and constant. Ensuring that the laser diode never sustains a destructive overload under any conditions is the tricky part.

A laser diode exhibits a current threshold, or "knee," below which its emission is weak and incoherent, as is photocurrent I_P. Above the knee, laser action occurs, and the optical output and photocurrent rise linearly with increasing drive current.

A simulator must reflect these characteristics, and the circuit in **Figure 2** comprises a basic voltage-controlled current source that presents a threshold. Based on a TO-92 or E-line-packaged PNP transistor and two resistors and packaged in a blob of epoxy, the simulator takes the place of the laser diode until circuit operation is stable. It's easy to build several modules to emulate laser diodes of various ratings.

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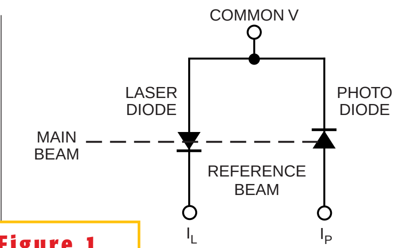

Figure 1

A P-type-laser diode assembly includes the photodiode power sensor.

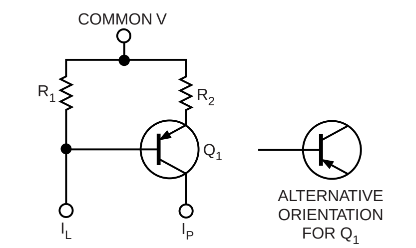

Figure 2

A laser-diode simulator requires only one transistor.

In operation, the laser driver sinks current I_L and develops a voltage, V_S, across R_1. When V_S exceeds Q_1's V_{BE}, Q_1 conducts and sources a simulated photocurrent, I_P, into the feedback-control circuitry. As I_L increases, I_P increases linearly in proportion.

As a design example, consider an average laser diode with a threshold current (I_{TH}) of 10 mA, an operating current at full optical output (I_{LMAX}) of 30 mA, and a photocurrent of 100 μA at full power. R_1 must then equal V_{BE}/I_{TH}, or 560 mV/10 mA and thus yields a value of 56Ω for R_1. Then, R_2 equals $((I_{LMAX} \cdot R_1) - V_{BE})/I_{PMAX}$, or approximately 11 kΩ. Using a value of 560 mV for V_{BE} produces the best practical relationship between I_L and I_P.

Inverting the transistor—that is, swapping Q_1's collector and emitter connections—produces a more abrupt conduction-threshold voltage of approximately 500 mV but decreases the slope of I_P versus I_L. In this example, inverting the transistor requires lowering the value of R_2 to approximately 7.5 kΩ.

The inverted-transistor circuit provides a sharper threshold and thus more realistic simulation, even though it may require some experimentation with resistor values for optimal performance. You can use almost any PNP bipolar-junction transistor for Q_1, such as a ZTX502, and decreasing R_2 by 30% renders I_P within 5% of the desired nominal value of I_P.

Note that laser diodes' characteristics vary widely, even within a batch, so using preferred-value resistors for R_1 and R_2 makes little practical difference in performance. Laser diodes exhibit a typical forward-voltage drop of about 2V, and the simulator circuit should thus drop no more voltage at full current. Also, the simulator circuit responds more slowly than a laser diode, but, if the feedback circuitry operates even more slowly, as it usually does, the simulator's slow response presents no problem.

A simulator for an N-type laser diode requires an NPN transistor and reversal of connections. More complex laser diodes may require more elaborate circuitry that includes current mirrors and additional connections. If adequate power-supply voltage is available for current-source compliance, you can connect an LED in series with the I_L lead to provide a visual indication of circuit operation. Connecting an oscilloscope across R_1 allows monitoring laser-drive and modulation currents. (In this context, "N-" and "P-type" refer not to laser-diode-device diffusions but to the polarity of the common terminal.)□

Differential driver doubles as versatile RF-switch driver

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Designed as a high-speed driver for 12-bit ADCs, the AD8137 controls SPDT GaAs (gallium-arsenide) FET-MMIC (microwave-monolithic-IC) and PIN-diode RF switches and thus provides a low-cost and versatile alternative to conventional switch drivers. This circuit achieves typical switching speeds of approximately 7 to 11 nsec, including the propagation delays of the driver and RF load.

The GaAsFET-driver circuit (**Figure 1**) converts a single-ended, 0 to 3.5V TTL signal into a complementary, 0 to −4V differential-output signal. The divider formed by the 50Ω source impedance, R_S, and the input termination, R_T, imposes a 50% signal reduction. To compensate, the circuit amplifies its input by approximately 2.3 times to yield the proper output amplitude of 4V p-p. The circuit also shifts the output level by −2V to provide the proper GaAsFET bias. **Equation 1** determines the output voltage:

$$G = \frac{R_5}{R_1} = \frac{R_6}{R_4}. \quad (1)$$

For a symmetrical output swing, gain-setting resistors R_1 and R_4 must present the same Thevenin-equivalent resistance. In **Figure 1**, R_4 increases by 20Ω over R_1. This increase compensates for the fact that source resistor R_S and termination resistor R_T combine in parallel to introduce additional resistance of 25Ω. Setting R_4 to 1.02 kΩ (the closest standard value to 1.025 kΩ) ensures that the circuit will provide approximately equal gains at the differential outputs.

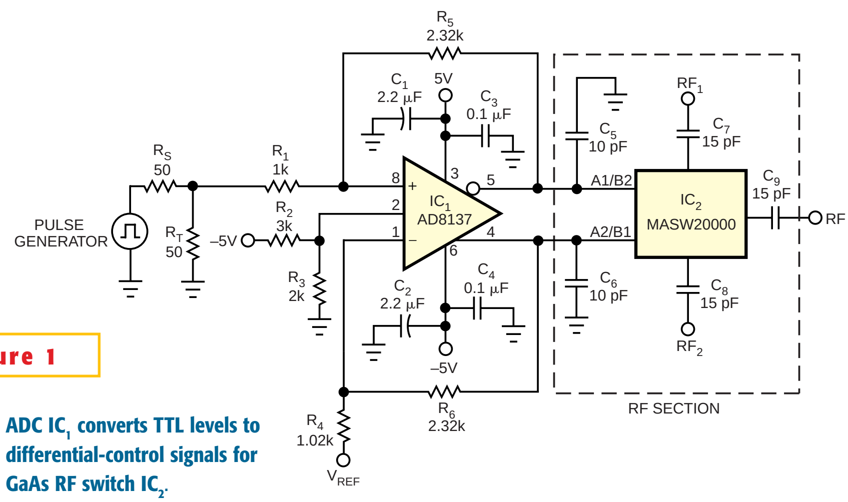


Figure 1

ADC IC_1 converts TTL levels to differential-control signals for GaAs RF switch IC_2.

The AD8137's V_{OCM} input (Pin 2) offers a convenient method of shifting the outputs' dc common-mode level. In **Figure 1**, R_2 and R_3 form a voltage divider, which sets the dc output level to −2V. Connecting the AD8137's inverting input to a reference voltage of 1.75V establishes the midpoint of the input signal and allows for proper switching of the AD8137's input stage.

Figure 2 shows the GaAs FET driver's turn-on switching speed of approximately 5 nsec for isolation to insertion loss—that is, 50% of the TTL input to 90% detected RF. **Figure 3** shows turn-off switching speed of approximately 11 nsec for insertion loss to isolation—that is, 50% of the TTL input to 10% detected RF.

As **Figure 4** shows, with only minor modifications, the GaAs switch-driver circuit can drive PIN-diode loads that require both positive and negative bias currents. IC_1's V_{OCM} input connects to ground to provide symmetrical outputs of 63.5V about ground and sinks and sources 10 mA of bias current. Altering

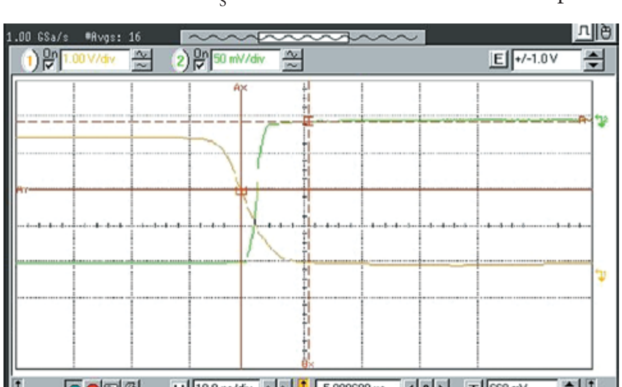


Figure 2

Vertical cursors denote GaAs switch turn-on time of approximately 5 nsec.

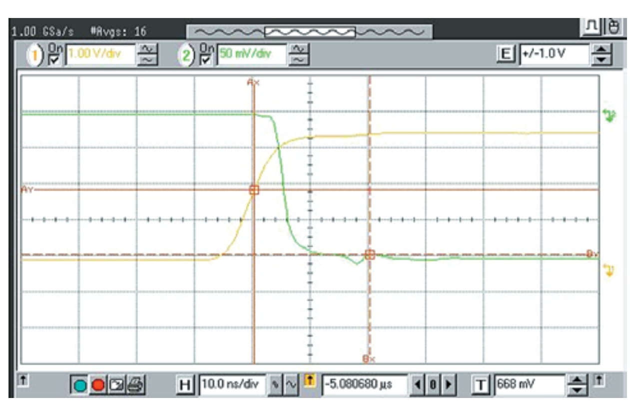


Figure 3

Turn-off time is approximately 11 nsec.

feedback resistors R_3 and R_4 to 2 kV provides an output swing of 63.5V. Resistors R_5 and R_6 set the steady-state PIN-diode current, I_{SS}, as **Equation 2** shows:

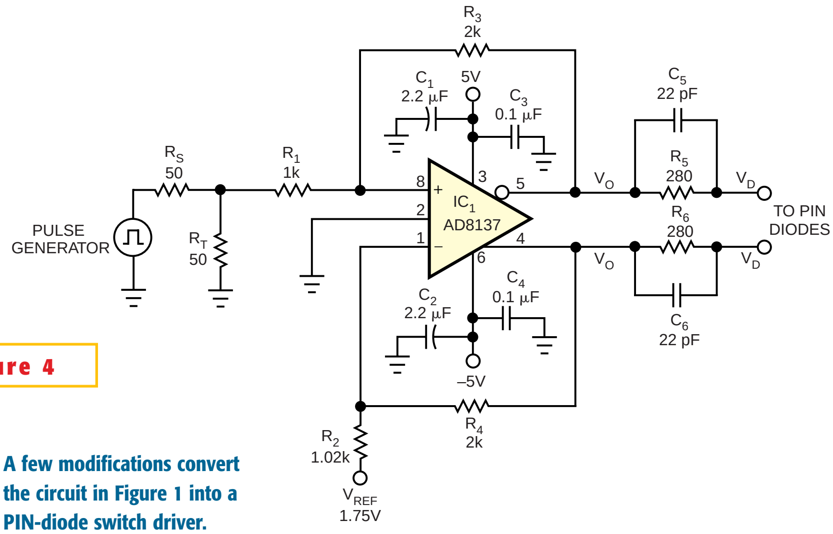
$$I=\frac{V_O-V_D}{R_5}. \quad (2)$$

Capacitors C_5 and C_6 set the spiking current I_S, which removes stored charge in the PIN diodes. You can optimize a given diode switch's response time by using the AD8137's output slew rate for dV/dt and **Equation 3**,

$$i=-C_5\frac{dV}{dt}, \quad (3)$$

to calculate spiking current.□

Figure 4



A few modifications convert the circuit in Figure 1 into a PIN-diode switch driver.

Pseudologarithmic thermistor signal conditioning spans wide temperature range

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Given its low cost, small size, robust construction, accuracy, versatility and sensitivity, it's no wonder that the thermistor rates as one of the most popular temperature sensors available. However, in some applications, a thermistor can exhibit too much sensitivity for wide-range temperature measurements. For example, when you combine a thermistor's radically nonlinear exponential resistance-versus-temperature-response curve with a linear signal conditioner (**Reference 1**), the resultant graph resembles a difficult-to-characterize response, as Trace A shows (**Figure 1**).

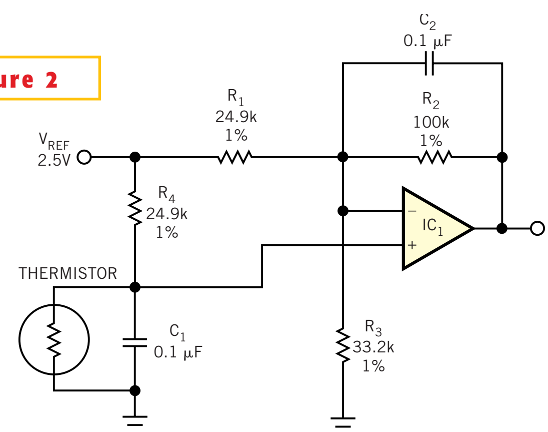
Note that most of the thermistor's resistance range crowds into a small span of temperatures at the lower limit of the range. As Curve B in **Figure 1** shows, the change of resistance per degree of temperature change looks exaggerated at low temperatures. As temperature increases, the resolution diminishes and may become inadequate at the upper end of the temperature scale.

In contrast, the signal-conditioning circuit in **Figure 2** mitigates the thermistor's inherent nonlinearity by generating a compensating, pseudologarithmic response function that's Curve C in **Figure 1** represents. The following **equation** relates the circuit's ±10V output span for an ADC ±10V input span to thermistor resistance: Thermistor resistance$=R_4\times(V_O+10)/(10-V_O)$.

Curve D in **Figure 1** shows the resultant resolution curve. Maximum resolution occurs at a temperature that corresponds to a thermistor resistance equal to R_4 and an output voltage of 0V. Although selection of this value optimizes resolution in the middle of the measurement range, the thermistor's nominal resistance is relatively noncritical. You can also select a different value of reference voltage, V_{REF}, to shift the ±10V output span to meet other requirements. For best performance, you can share the same reference source for V_{REF} and for the measurement system's ADC, which makes measurements ratiometric and thus insensitive to reference-voltage drift.

In addition to resolution, span, and

Figure 2



This pseudologarithmic thermistor signal conditioner uses a single operational amplifier and a few passive components.

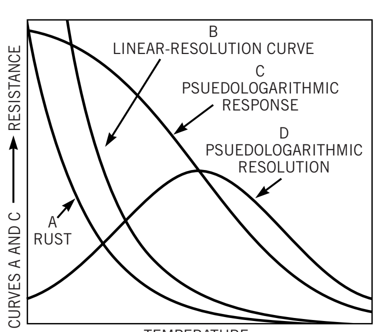


Figure 1 These resistance-versus-temperature curves highlight the improved resolution available in a pseudologarithmic circuit.

reference-drift considerations, thermistors—and, for that matter, all temperature sensors—experience self-heating effects. The excitation current produces ohmic power that causes these effects. The limited thermal mass of miniature thermistors provides limited heat dissipation—often only a few microwatts per degree—and that dissipation can produce especially severe temperature shifts.

In the circuit in **Figure 2**, the power dissipated due to thermistor's self-heating reaches a maximum at midspan—that is, when output voltage is 0V and equals: $V_{REF}^2/(4\times R_4)=63\ \mu W$ for the component values illustrated. For all but the smallest thermistors, this level of self-heating produces acceptably low fractional-degree errors.□

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1. Woodward, Steve, "Optimize linear-sensor resolution," *EDN*, March 7, 2002, pg 131.

RF-telemetry transmitter features minimal parts count

Francis Rodes, Eliane Garnier, and Olivier Chevalerias, ENSEIRB Talence, France

REQUIREMENTS FOR portable, short-range telemetry systems frequently include low power consumption, small size, and low cost. The circuit in **Figure 1** meets these criteria and uses only three off-the-shelf ICs and a few passive components. Although dedicated to conditioning the low-level signal a strain-gauge bridge produces, the circuit can operate with almost any resistive transducer based on a Wheatstone bridge. The circuit comprises a VFC (voltage-to-frequency converter) that produces a PPM (pulse-position-modulation) output, and an OOK (on/off-keyed) RF transmitter. Based on direct digitization, the VFC in **Figure 1** comprises IC_1, IC_2, IC_{3A}, and IC_{3B}. A Wheatstone bridge containing strain gauge R_X produces an output of approximately 5 mV. An integrator stage comprising C_6 and IC_1, a Linear Technology LTC1250 offset-compensated, low-drift operational amplifier, connects directly across the bridge. (Note that the value of the remaining resistors in the bridge depends on the application.)

To yield a ratiometric conversion, the voltage applied to the bridge, V_{COMP}, varies with power-supply voltage and equals the difference between the two threshold voltages of a Schmitt-trigger circuit. In **Figure 1**, the Schmitt trigger comprises a Maxim MAX9075 comparator, IC_2; a CMOS inverter, IC_{3A}; and the positive feedback network comprising R_1, R_2, and C_1. **Equations 1** and **2** define the Schmitt trigger's high, V_{TH}, and low, V_{TL}, threshold voltages:

$$V_{TH}=V_{DD}\frac{R_1}{R_1+R_2}. \quad (1)$$

$$V_{TL}=V_{DD}\frac{R_2}{R_1+R_2}. \quad (2)$$

To understand the circuit's operation, assume that the comparator's output is high and, consequently, the inverter's output is low. Also assume that the series combination of the strain gauge, R_X, and the trimmer, R_T, is always equivalent to $R(1+X)>R$. That is, for linear operation, the sensor's resistance variation represents a small fraction of the arm resistance.

Under these conditions, the noninverting input of IC_1 biases to $V_{DD}/2$, and the Wheatstone bridge's active arm drives a positive current, I_1, into the summing node of IC_1. This current causes the integrator's output, V_{OI}, to ramp down toward the low threshold voltage, V_{TL}, of the Schmitt trigger.

When $V_{OI}=V_{TL}$, the comparator's output goes to zero, and the inverter's output consequently rises to V_{DD}. This action inverts the direction of the integrator's input current, causing the in-

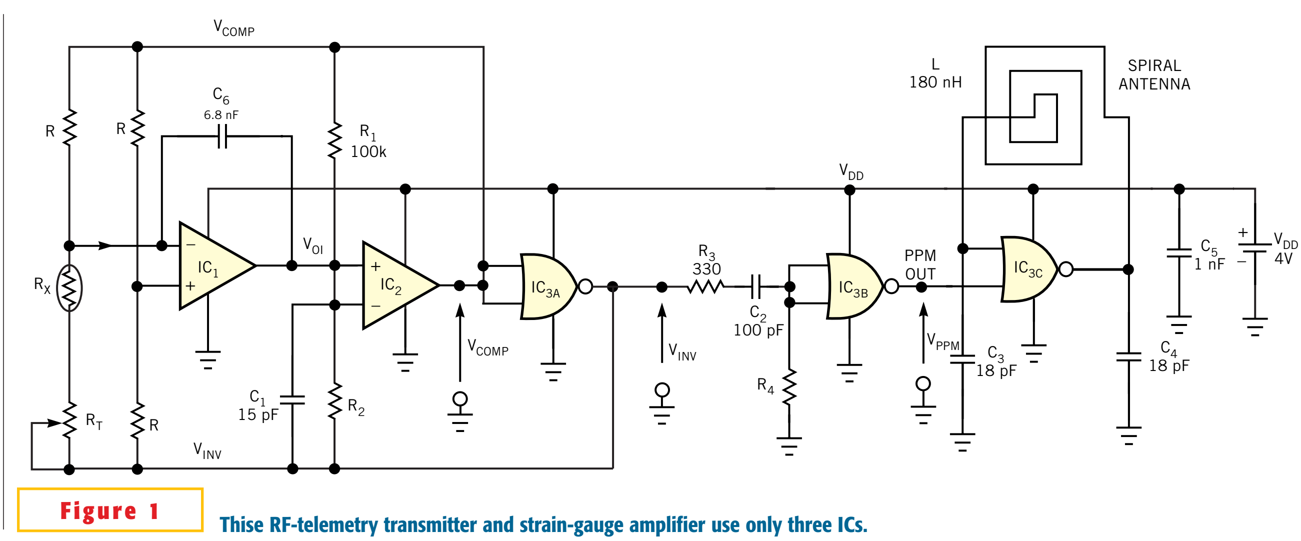

Figure 1

Thise RF-telemetry transmitter and strain-gauge amplifier use only three ICs.

tegrator's output to ramp upward to the Schmitt trigger's high threshold voltage. Finally, when $V_{OI} = V_{TH}$, the comparator's output goes high, and the above sequence repeats indefinitely, producing a free-running oscillation in which the integrator's output ramps up and down between the threshold voltages of the Schmitt trigger (Trace 1 in **Figure 2**). Meanwhile, the comparator's output and the inverter's output deliver two square waves with a 50%-duty-cycle ratio (traces 2 and 3, respectively, in **Figure 2**) that drive the bridge.

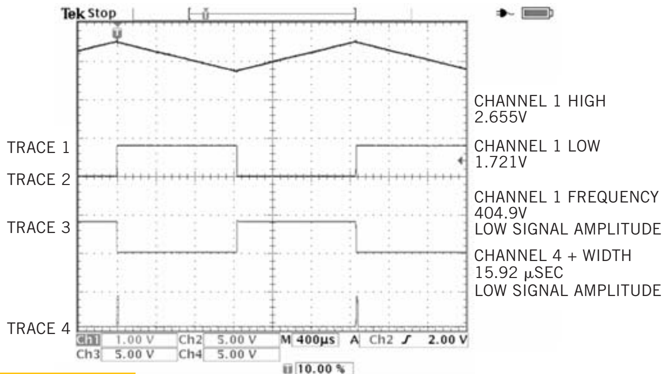

Figure 2 This oscilloscope photo shows circuit's internal waveforms: Trace 1 is the integrator's output voltage, Trace 2 is the comparator's output voltage, Trace 3 is the inverter's output voltage, and Trace 4 is the PPM-output voltage.

To produce the PPM signal (Trace 4 in **Figure 2**), the inverter's output drives a monostable circuit comprising a second inverter, IC_{3B}, and timing components R_4 and C_2, which produces a 15-µsec-wide pulse. Current-limiting resistor R_3 prevents latch-up of IC_3, and R_2 and C_4 set the output-pulse width.

To model the VFC's transfer function, calculate period T_X of the comparator's output. Due to the symmetry of IC_1's output, this period is twice the time the integrator's output takes to ramp linearly between the two threshold voltages of the Schmitt trigger. Consequently, you can express t_X as in **Equation 3**:

$$t_X = 2\frac{V_{TH}-V_{TL}}{\frac{dV_{OI}}{dt}}, \qquad (3)$$

where dV_{OI}/dt is the slope of the ramp at the integrator's output. Assuming that the integrator's input current is constant during one period, **Equation 4** gives the slope:

$$\frac{dV_{OI}}{dt} = \frac{I_I}{C}. \qquad (4)$$

After applying a Thevenin transformation of the bridge's active arm, you can express the integrator's input current as

$$I_I = \frac{V_{DD}}{2R} \times \frac{X}{1+X}. \qquad (5)$$

Replacing V_{TL}, V_{TH}, and dV_{OI}/dt in **Equation 3** with the respective expressions from **equations 1**, **2**, and **4** finally yields the frequency in **Equation 6**.

$$f_X = \frac{1}{t_X} = \frac{1}{4RC} \times \frac{R_1+R_2}{R_1-R_2} \times \frac{X}{1+X}. \qquad (6)$$

Knowing that $X<<1$, you can approximate the PPM's output frequency by **Equation 7**:

$$f_X \approx \frac{1}{4RC} \times \frac{R_1+R_2}{R_1-R_2} \times X. \qquad (7)$$

This transfer function highlights the VFC's three most important features: that the modulation frequency, f_X, is directly proportional to the relative variation of the bridge's sensor resistance, R_X; that the modulation frequency is independent of the power-supply, V_{DD}; and, therefore, that this PPM converter is ratiometric. This feature is attractive for any portable-system application in which the supply voltage decreases as the battery ages. The modulation frequency is independent of the PPM pulse width, which eliminates the source of error you typically encounter with single-slope VFCs.

Using the components values shown in **Figure 1**, the converter's frequency span is: $200\text{ Hz}<F_X<600\text{ Hz}$ for a relative variation of the strain-gauge resistance: $-4.2\times10^{-3}<\Delta R/R<4.2\times10^{-3}$. Consequently, with zero force applied to the strain gauge, $\Delta R/R=0$, and the converter produces a 400-Hz modulation frequency. The waveforms in **Figure 2** correspond to this case.

To transmit data over a distance of a few meters, the PPM signal modulates an 80-MHz OOK RF transmitter. This transmitter comprises an interruptible Colpitts oscillator based on a very-high-speed CMOS NOR gate, IC_{3C}; a 74VHC02; and a tank circuit comprising two identical feedback capacitors, C_3 and C_4; and a square-spiral pc inductor, L. To obtain reliable oscillation start-up, set $C_3=C_4=C$. Neglecting the effects of stray capacitances, you can calculate the output frequency of the Colpitts oscillator via **Equation 8**:

$$f_C = \frac{1}{2\pi\sqrt{L\frac{C}{2}}}. \qquad (8)$$

The values of L, C_3, and C_4 shown in **Figure 1** produce a carrier frequency, f_C, of approximately 80 MHz. Inductor L_1 also doubles as the transmitter's antenna, and its characteristics of eight turns in an 8×8-mm pc-board footprint stem from a process that ensures that the transmitter's radiated power never exceeds 250 nW at 80 MHz.

According to European Telecommunication Standard I-ETS 300 220, the transmitter requires no license and can operate at any carrier frequency within the 74- to 87.5-MHz band. Consult applicable regulations for unlicensed transmitter operation in your locality.

At a transmitted power of less than 250 nW, an AM receiver with a tangential sensitivity of 1 µV provides a reception range as long as 10m, which is sufficient for many indoor-telemetry applications. At a maximum PPM frequency of 600 Hz, the current drain of the circuit in **Figure 1** is approximately 2 mA at a supply voltage of 4V.□

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