design ideas

Edited by Bill Travis

Current-sensing scheme improves PFC on/off sequences

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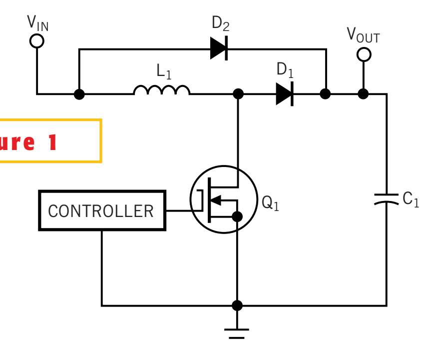
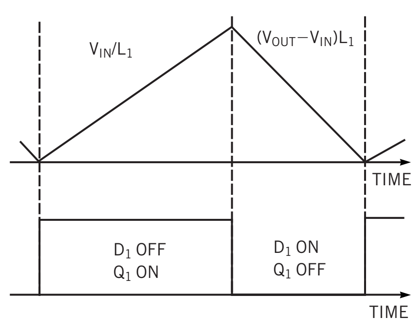
PFC (power-factor-correction) preconverters typically use the step-up, or boost, configuration, because this type of converter is relatively easy to implement (**Figure 1**). However, this topology requires the output voltage to be higher than the input voltage. When this condition is not the case—for example, with on/off sequences or under load conditions—some inrush current flows through the boost inductor and diode to abruptly charge the output capacitor. For instance, before start-up, the output capacitor discharges. When you plug in the PFC stage, the output capacitor attempts to charge resonantly to twice V_{IN}. During this sequence, the current can largely exceed the levels obtained during normal operation. Too often, these uncontrolled inrush currents make PFC-stage designers nervous during on/off reliability tests. Except for On Semiconductor's (www.onsemi.com) MC33260, which monitors the entire loop current, available controllers cannot detect this overcurrent state. These controllers may turn on the power switch while a huge and potentially destructive current flows through the inductor.

Figure 1



In this classic boost-converter topology, inrush currents can unduly stress the power-MOSFET switch.

Figure 2



Boost-converter operation occurs in two phases.

In the boost structure of **Figure 1**, the controller turns the power switch, Q_1, on and off to control the L_1 inductor current. **Figure 2** illustrates two phases:

- Switch Q_1 is on. The boost structure's input voltage (V_{IN}, the rectified ac-line voltage) appears across the inductor, L_1, which charges linearly.
- Switch Q_1 is off. The diode, D_1, turns on and drives the inductor current toward the output capacitor, C_1. The inductor current ramps down with a slope equal to $(V_{OUT}-V_{IN})/L_1$. V_{OUT} must be higher than V_{IN} to properly discharge the inductor.

If V_{IN} exceeds V_{OUT}, an inrush current flows through L_1 and D_1 and charges output capacitor C_1. Designers generally place a diode, D_2, between V_{IN} and V_{OUT}. This diode conducts a major part of the inrush current, thus improving the safety of the first power-switch turn-on. However, when the output voltage is in the neighborhood of V_{IN}, the current that ramps up during this first switch-on time generally cannot significantly decay during the off-time. As a consequence, the following turn-on operation may occur while the inductor is still charged. Moreover, if V_{OUT} needs several switching cycles to significantly charge (for example, under heavy load conditions), the power MOSFET faces a succession of stressful turn-ons that may jeopardize the circuit's reliability.

When you cannot use the MC33260, you can use the circuit in **Figure 3** to improve the reliability of the PFC stage. You can test this configuration using the MC33262. Typically, the current-sense resistor, R_1 connects between the power MOSFET's source and ground, and the

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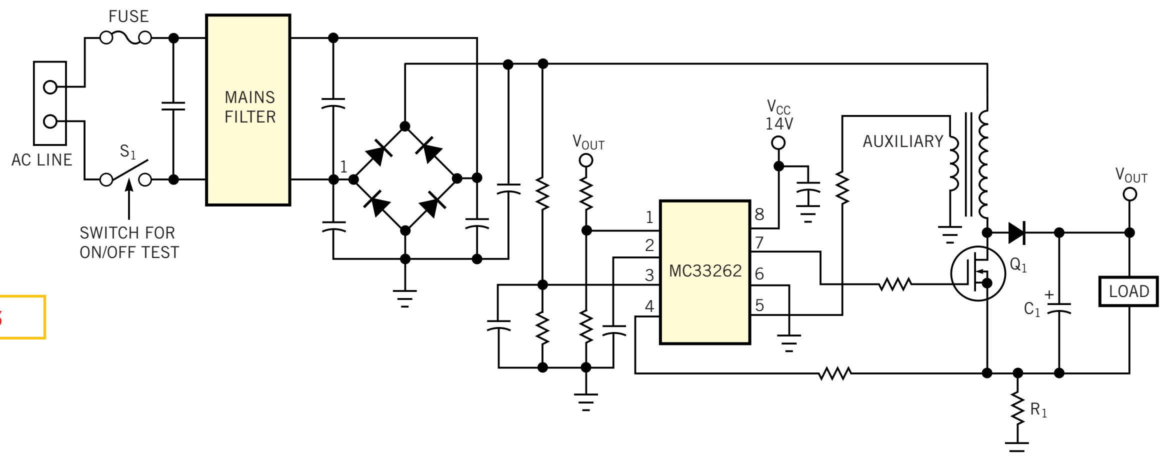

Figure 3

In this circuit, a modification in the placement of the current-sensing resistor increases the robustness of the PFC preconverter.

negative terminal of the output capacitor, C_1, connects to ground. As a result, R_1 senses only the power-switch current. In the modified circuit of **Figure 3**, the output capacitor's charging current also passes through R_1. As a result, the sense resistor senses the entire inductor current. The MC33262 keeps the power switch off as long as the sensed current is higher than the setpoint that an internal multiplier establishes. When the sensed current is below this setpoint, the MOSFET turns on upon core-reset detection. If detection is impossible, a 600-μsec watchdog timer reactivates the power switch. In the case of on/off sequences, the core-reset information is generally unavailable, and the MOSFET turns on in the following cases:

- 600 μsec after the preconverter switches on, regardless of the inductor current, in the traditional application, or
- once the coil current measures lower than setpoint in the modified application.

Figure 4 clearly shows that no switching takes place when the input current is high, as long as the current is higher than the setpoint. In **Figure 3**'s example, the situation is even better. In effect, the 600-μsec timer delays the power switch's turn-on even after the current-sensing block allows the turn-on, so the MOSFET finally switches on when the inductor current is zero. You can see that the modified application schematic increases the power dissipated in the current-sense resistor. In effect, the losses are $P_{MODIF}=\frac{1}{6}R_{SENSE}(I_{PKMAX})^2$, where R_{SENSE} is the current-sense resistor (R_1 in **Figure 3**), and I_{PKMAX} is the maximum inductor peak current (obtained at the top of the sinusoid). In this application, you can compute the losses using the following equation:

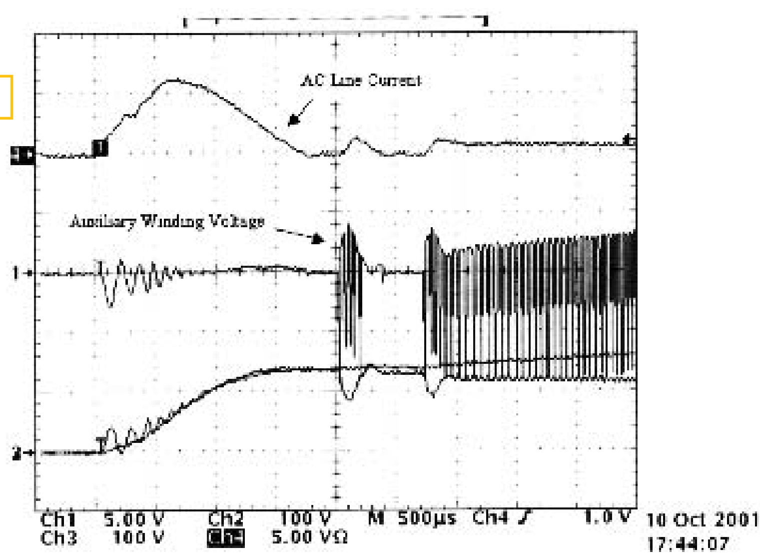

Figure 4

No switching occurs when the input current is high; a 600-μsec timer delays the power switch's turn-on.

$$P_{TYP}=\frac{1}{6}R_{SENSE}(I_{PKMAX})^2\left(1-\frac{1.2V_{AC}}{V_{OUT}}\right), \quad (1)$$

where V_{AC} is the rms input voltage and V_{OUT} is the output voltage. The relative increase in dissipation then conforms to the following equation:

$$\frac{P_{MODIF}-P_{TYP}}{P_{TYP}}(\%)=\frac{120V_{AC}}{V_{OUT}-1.2V_{AC}}. \quad (2)$$

From **Equation 2**, you can determine that if $V_{AC}=90V$ and $V_{OUT}=400V$, the dissipation increases by 36%. If $V_{AC}=180V$ and $V_{OUT}=400V$, the dissipation increases by 117%. The simple modification of the sensing resistor's location significantly increases the robustness of the PFC preconverter during on/off tests at the price of a reasonable increase of the power dissipation in the sensing resistor.

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Reset generator uses “fleapower”

Philip Simpson, Maxim Integrated Products, Reading, UK

When a processor-controlled device must operate reliably, designers often choose to periodically reset the processor rather than rely on a watchdog configuration. In low-power systems, this periodic-reset circuit can consume a large part of the system’s current budget or may fail to operate at low voltages. The circuit in **Figure 1** generates a low-going reset pulse of 100-μsec duration. The circuit consumes less than 1-μA operating current and operates from 1.8 to 5V supplies with little variation in the output period. The circuit is an adaptation of a normal relaxation oscillator. The circuit has a differentiator and diode clamp on the output to generate the 100-μsec low-going pulse. You can adjust the period of the output waveform by varying R_1, C_1, or both. You can adjust the pulse width of the low-going reset pulse by varying R_P, C_P, or both, or you can change the polarity by repositioning D_1. **Figure 2a** shows the comparator’s output waveform, which has a period of approximately 1.3 sec. The period varies from 1.308 sec with a 4.5V supply to 1.306 sec with a 1.8V supply. **Figure 2b** shows details of the low-going reset pulse, which takes the shape of the output of a normal relaxation oscillator. The reset pulse is 100 μsec wide at its 30% point on the exponential curve.

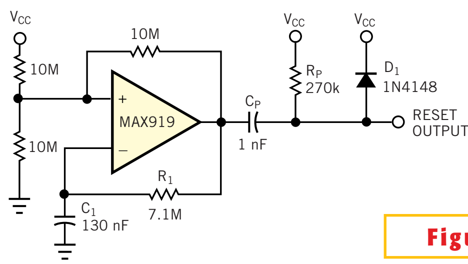

Figure 1

This reset circuit consumes less than 1 μA and delivers a 100-μsec-wide reset pulse every 1.3 sec.

The 350-nA supply current, the 1.8 to 5.5V supply range, and the SOT-23 package make the MAX919 ideal for this application. Measurements for the circuit reveal lower than 1-μA operating current. This low consumption would allow the circuit to operate from a single AA lithium cell for 250 years. With judicious component choice, the circuit can generate periods from milliseconds to minutes. To ensure good temperature stability, you should use metal-film types for R_1 and R_P and NP0 types for C_1 and C_P. Assuming a reset-logic threshold of 30% of the supply rail, you can use the following formulas to adjust the output pulse width and period: pulse width$\simeq 0.36R_PC_P$, and period$\simeq .4R_1C_1$.

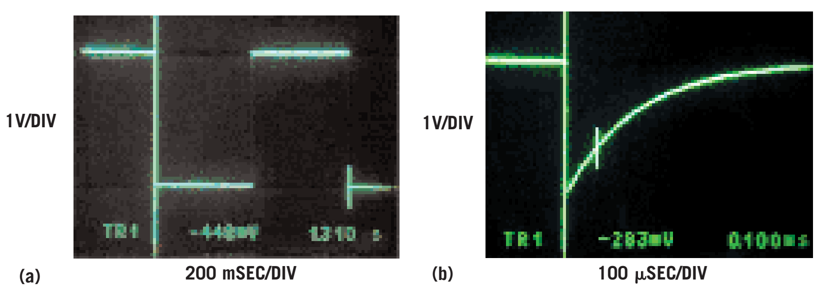

Figure 2

The comparator in Figure 1 produces a square wave with a period of 1.3 sec (a); the output differentiator yields a low-going pulse of 100-μsec width at its 30% point (b).

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Voltage-to-current converter drives white LEDs

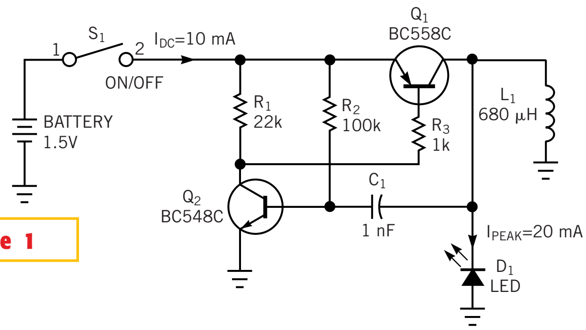
Susanne Nell, Breitenfurt, Austria

You sometimes need to drive a white LED from one 1.5V battery. Unfortunately, the forward voltage of a white LED is 3 to 4V. So, you would need a dc/dc converter to drive the LED from one battery. Using the simple circuit in **Figure 1**, you can drive one white LED or two series-connected green LEDs, using only a few components. The circuit is a voltage-to-current converter, which converts the battery voltage to a current that passes through the LED. You can adjust this current and, thus, the brightness of the LED, by varying resistor R_3. If you turn on switch S_1, resistor R_2 feeds base current to transistor Q_2. Q_2 turns on, and its collector current, via R_3, turns on Q_1. Now, the current through inductor L_1 increases. The slope of the increase is a function of the value of L_1 and the battery voltage. The current through L_1 increases until it reaches a maximum value, which depends on the gain of Q_1. Because the value of R_3 sets the base current drawn from Q_1, Q_1's collector current is also limited.

Once the current through L_1 reaches its maximum value, the slope of the current through L_1 changes. At that instant, the voltage on L_1 switches to a negative polarity forced by the changed slope. This negative voltage traverses capacitor C_1 and turns off Q_2, which in turn turns off Q_1. The negative voltage on L_1 increases until it reaches the forward voltage of the LED. The peak current through inductor L_1 now flows through the LED and decreases to zero. Now, Q_2 switches on again, via the current through R_2, and the cycle starts again. By adjusting resistor R_3, you can set the peak current through L_1 and the peak current through the LED. The brightness of an LED is a linear function of the current through the LED. So, adjusting the value of R_3 also adjusts the brightness of the LED.

It doesn't matter which LED you use; the forward voltage on the LED always increases until the peak current through L_1 flows through the LED. Different forward voltages of the LEDs yield different on-times (duty cycles) but the same peak current through the LED. With the values shown in **Figure 1**, the circuit oscillates at a frequency of approximately 30 kHz and delivers a 20-mA peak current through the LED. The duty cycle depends on the ratio of the battery voltage to the forward voltage of the LED. One advantage of this circuit is that it requires no series-limiting resistor for the LED. The peak current through the LED is a function of the value of R_3 and the gain of Q_1.

Figure 1



You can eschew expensive dc/dc converters by using this inexpensive circuit to drive a white LED from a single battery cell.

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Use time-domain analysis of Zobel network

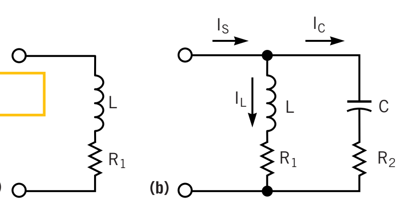
Noël Boutin, Université de Sherbrooke, PQ, Canada

A Zobel network is useful in making a reactive load appear as a pure resistance to a driving source prone to stability problems (**Reference 1**). A typical situation is an audio power amplifier driving a loudspeaker, modeled at first approximation as an inductance and a series resistor (**Figure 1a**). The addition of a series R_2C network in parallel with the series R_1L network forms a Zobel network (**Figure 1b**). If you select the proper values of R_2 and C, the driving source sees a purely resistive load. **Reference 2** discusses the computation of the total impedance, Z_L, of the Zobel network:

$$Z_L=\frac{R_1\left[s^2\left(LC\frac{R_2}{R_1}\right)+s\left(R_2C+\frac{L}{R_1}\right)+1\right]}{s^2(LC)+sC(R_2+R_1)+1},$$

from which you find that the following conditions must prevail: $R_2=R_1$ and $C=L/R_1^2$. Designing the Zobel network in the time domain, rather than in the transformed-s domain, yields an easier way to arrive at the same result. Moreover, the method provides a better understanding of the reasons why the driver sees a purely resistive load with a Zobel network.

Figure 1



A first-approximation model of a loudspeaker is an inductance and a resistor in series (a); the addition of a series RC network (b) makes the speaker look purely resistive to the driving source.

Without loss of generality, let the driving source be an ideal V-volt step-function voltage source. If the load were purely resistive, the source current, I_S, would also be a step function. In the absence of the series R_2C network, the current flows only through the series R_1L load, starting from a zero value and exponentially increasing toward a final value. The time constant in this case is $\tau_1=L/R_1$. For the source to supply a step current, you must add another branch that draws a current, I_C, such that it compensates for the slow-rising load current, I_L. Adding the series R_2C network meets that requirement. The current, I_C, flowing through that network is instantaneously equal to V/R_2 and then decreases exponentially to zero with a time constant $\tau_2=R_2C$. For the sum of the current, I_L, flowing through the series R_1L network and the current I_C flowing through the series R_2C network to yield a step current I_S, R_2 must equal R_1 and τ_2 must equal τ_1. That is, $R_2=R_1$ and $R_2C=L/R_1$.

References

1. Zobel, OJ, "Distortion Correction in Electrical Circuits with Constant-Resistance Networks," *Bell Systems Technical Journal*, July 1982, pg 438.

2. Albean, D, "Zobel network tames reactive loads," *EDN*, Dec 21, 1995, pg 82.

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Circuit improves further on first-event detector

Vasiliy Borodai, Zaporozhje, Ukraine

The circuits in **Figure 1** have certain advantages over those in a previous Design Idea (**Reference 1**). The first-event detector with autoreset (**Figure 1a**) consists of N sets of monostable multivibrators, using 4001 logic circuits with LEDs attached. After any player (1 through N) presses a pushbutton, the corresponding monostable multivibrator switches on, and its associated LED lights. The voltage at Point A changes to a level of nearly 2V (the forward voltage of the LED). After that instant, no other player can change the situation by pressing a pushbutton, because, to switch, the monostable multivibrators need a voltage exceeding 4V. After an interval of 0.5RC (nearly 15 sec for R=1 MΩ and C=33 μF), the monostable multivibrator returns to its previous state, and the circuit is ready for its next first-event detection. **Figure 1b** shows another first-event detector with autoreset, using 4011 logic ICs. This circuit is a mirror image of the circuit in **Figure 1a** and operates similarly.

Reference

1. Arendt, Lawrence, "Circuit improves on first-event detection," *EDN*, Aug 16, 2001, pg 106.

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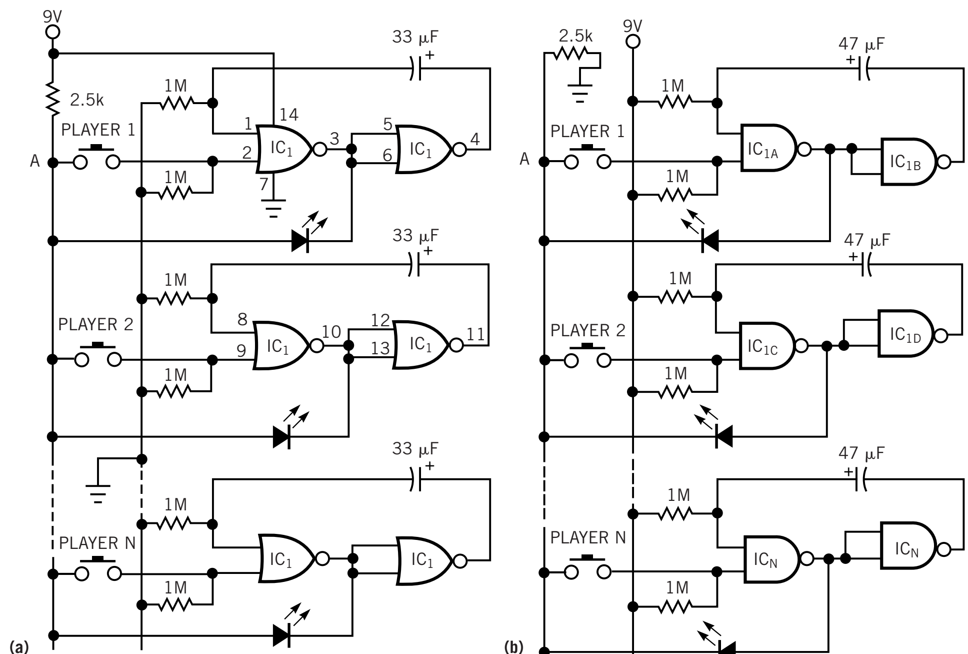
Figure 1



Monostable multivibrators and LEDs form the heart of these first-event detectors with autoreset.

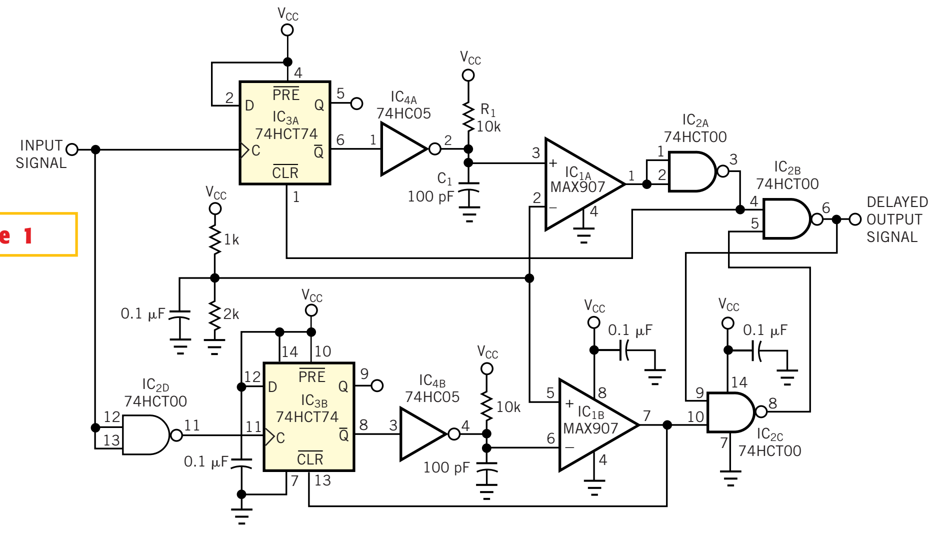
Delay line has wide duty-cycle range

John Guy, Maxim Integrated Products, Sunnyvale.CA

Today's digital delay lines can process pulses no shorter than their delay times, and that restriction confines the devices to applications in which the duty cycle remains near 50%. A limited range of available delays (2 to 100 nsec per tap) further limits their use. Longer delay is available with one-shot multivibrators of standard digital-logic families, but those devices do not retain duty-cycle information. As an example, a PWM control circuit (**Figure 1**) must handle relatively long delays while retaining information about the input duty cycle. The upper half of this dual-path, precision one-shot works on the input signal's rising edge. The rising edge triggers the D flip-flop, IC_{3A}, to drive IC_{4A}'s input low. IC_{4A} has an open-drain output; the output therefore rises exponentially according to the single R_1C_1 time constant. IC_{1A} compares the output with a dc voltage equal to 67% of V_{CC}, producing a conveniently scaled delay equal to R_1C_1.

The output of comparator IC_{1A} drives the set input of an RS latch (IC_{2B} and IC_{2C}). It also feeds back to the input flip-flop, thereby resetting the flip-flop in anticipation of the next rising edge. The lower half of the circuit in **Figure 1** works in a similar fashion, but it triggers on the input's falling edge and drives the reset input of the RS latch. You can test the circuit with a 100-kHz input signal and a nominal delay of 1 μsec. When the input duty cycle varies from 10 to 90% (limits imposed by test equipment), the duty-cycle error is less than 0.1%. You can obtain this performance with unmatched components. The circuit produces accurate pulse widths for pulses as narrow as 20 nsec. To ensure accuracy, the timing capacitors should be NP0 types with 5% tolerance, and the resistors should be 1% accurate. The MAX907 comparator from Maxim (www.maxim-ic.com) provides the high input impedance, high precision, and low propagation delay the circuit requires. For most applications, 74-HC/HCT logic is fast enough to minimize propagation-delay errors. Note the inclusion of a NAND gate connected as an inverter, IC_{2A}, which enhances accuracy by equalizing the propagation delays in each channel.

Figure 1



Based on a precision dual comparator, this delay line generates accurate duty cycles.

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