Designers of dc/dc switching power converters face the challenge of controlling EMI (electromagnetic-interference) emissions produced during normal operation. If large enough, these emissions conduct through power lines or radiate to other assemblies within a system and can compromise a system’s performance. Emission peaks typically occur at the converter’s fundamental switching frequency and gradually reduce in amplitude at each higher order harmonic, with most of the emitted energy confining itself to the fundamental and lower order harmonics. Modulating, or dithering, the power converter’s operating frequency can reduce the peak emissions by spreading EMI over a band of frequencies.

Most modern PWM controllers use an external resistor to set the operating frequency, which typically increases with decreasing resistor values. For example, the LM5020’s internal oscillator delivers a regulated 2V at its programming pin (RT), and a programming resistor connected to RT sets the current that RT delivers. The oscillator also delivers a proportional current into an internal timing capacitor (Reference 1). The period of the timing capacitor’s ramping voltage determines the oscillator’s frequency.

The external dithering circuit in Figure 1 comprises a simple stand-alone comparator-based oscillator configured to operate at approximately 800 Hz. The output state of comparator IC1 goes high upon power-up. R1, R2, and R3 set the comparator’s positive input, which initially rests at 2.9V. The voltage at capacitor C3 ramps up toward the positive threshold. When the voltage at the comparator’s negative input reaches the positive-threshold voltage, the comparator’s output switches low, which also lowers the threshold at the comparator’s positive input to 2.1V. The voltage at capacitor C3 then ramps down toward the new threshold, and, when it reaches the lower threshold voltage, the cycle repeats. The voltage across C3 approximates a triangular wave with a minimum voltage of 2.1V and a maximum of 2.9V.

To dither the LM5020 controller’s PWM-oscillator base frequency, the triangular wave generated by IC2 modulates the current from the controller’s RT pin. Resistor R5 sets the percentage of modulation dither. The right side of R5 is fixed at the RT pin’s regulated potential of 2V, and the low-frequency triangle wave coupled from IC2 through capacitor C2 drives R5’s left side. For R5 with a value of 64.9 kΩ, the peak-to-peak current through resistor R5 is approximately 12 μA. With the dither circuit disconnected, the steady-state current that RT sources is approximately 121 μA, and the 12-μA p-p dither current thus represents 10% total modulation.

An LM5020-controlled PWM flyback dc/dc converter, IC1, evaluates the dither circuit’s effectiveness. The circuit’s fundamental operating frequency is 250 kHz, which the controller’s R1

**Figure 1** A low-frequency triangular-wave oscillator modulates a PWM controller’s oscillator frequency.

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**DIs Inside**

100 Single-port pin drives dual LED

102 Network linearizes dc/dc converter’s current-limit characteristics

104 Add a Schmitt-trigger function to CPLDs, FPGAs, and applications

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Most current microcontrollers offer I/O ports that can change their functions during program execution. As outputs, the circuits can sink and source reasonably large amounts of current. This Design Idea shows three alternative methods for driving a two-pin, two-color LED from a single I/O pin.

Figure 1 illustrates one possible approach that uses external inverter IC1 to drive D1, a red/green bidirectional LED. A logic-high output on the port pin forces current through the green (upper) LED and pulls the inverter’s input high, which drives the inverter’s output low and sinks current from the green LED. A logic-low output on the port pin raises the inverter’s output high, delivering current to the red (lower) LED; the microcontroller’s output sinks current from the red LED.

To turn off both LEDs, you reconfigure the microcontroller’s port pin from sourcing or sinking current. This circuit’s primary disadvantage is that it yields no control over each LED’s brightness; instead, resistor R5 determines forward current for both LEDs.

Figure 2 presents an approach that also involves a major disadvantage. Zener diodes D3 and D4 and resistors R3 and R4 form a low-impedance voltage divider that applies VCC/2V to one end of LED D5. The value of VCC drives the selection of the zener diodes’ voltage, VZ, with lower voltage zener diodes allowing more LED current and higher voltage ones limiting maximum LED current. Given that the microcontroller’s outputs can deliver rail-to-rail voltages, the difference between VCC and V2 limits maximum forward current for both LEDs. For example, if VCC is 5V and V2 is 3V, the forward voltage across either LED is less than 2V. Once a designer selects the zener-diode voltage, only small variations in VCC can occur; otherwise, the LEDs’ brightness would fluctuate.

Using discrete components, another circuit offers an inexpensive approach that avoids the other circuits’ disadvantages (Figure 3). When the microcontroller’s output port goes high, current flows through the green (upper) LED, R1, D2, and FET Q2, which the port’s high level turns on. When the microcontroller’s output port goes low, transistor Q1 turns on and delivers current to the port pin through R2 and the red (lower) LED. The circuit operates symmetrically because silicon diode D2’s forward-voltage drop is present.
Network linearizes dc/dc converter’s current-limit characteristics

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Recently announced versions of integrated step-down dc/dc converters have eliminated the requirement for a high-side current-sense resistor by sampling the voltage drop across an external, low-side, MOSFET synchronous rectifier. This topology eliminates the sense resistor’s cost and pc-board-space requirement and also provides a modest increase in circuit efficiency. However, the MOSFET’s highly temperature-dependent on-resistance dominates the current-limit value. Fortunately, certain newer dc/dc converters, such as Maxim’s MAX1714, allow external adjustment of the current-limit threshold. The circuit in Figure 1 shows how a thermistor applies temperature compensation to the circuit’s output-current limit.

The MAX1714’s linear current-limit (\(I_{\text{LM}}\)) input range at Pin 6 of IC\(_1\) spans 0.5 to 2V, which corresponds to current-limit thresholds of 50 to 200 mV, respectively. For the default current-limit setting, 100 mV, the circuit imposes a 7.5A current limit at 25°C. However, Figure 2 shows that the current limit varies from 9A at \(-40°C\) to 6A at 85°C. To design the temperature-compensation network, begin by breadboarding the circuit and using an external power supply to vary the MAX1714’s current-limit input voltage such that the output-current-limit value remains constant. You repeat the

and the LED-driver circuit’s \(V_{CC}\).

Replace \(R_2\) with two resistors connected in series between Q\(_1\)’s emitter and D\(_1\)’s anode. Connect the midpoint of the two resistors to the LEDs.

With the microcontroller’s port pin configured as an “input with pullup,” the port delivers a small current to the green LED. However, pullup-resistor values of 22 k\(\Omega\) or higher do not cause misleading light output from LEDs in the off-state. When the input signal from the port pin floats—that is, with \(V_{CC}\) at 5V and the port configured as an input with no pullup resistor—the circuit draws no additional current, and the quiescent current, which \(R_1\) determines, averages less than 100 \(\mu\)A.
Thanks to its internal hysteresis, the highly useful Schmitt-trigger circuit accepts a low-slew-rate input signal and produces a clean, glitch-free output transition. Unfortunately, user-programmable logic devices, such as CPLDs and FPGAs, generally offer no direct method of synthesizing Schmitt-trigger gates and buffers. This Design Idea shows how a few external components and some VHDL code can implement a Schmitt trigger and put it to work in several useful applications.

To create an equivalent of the basic Schmitt-trigger buffer, you use two external resistors to create positive feedback around a buffer (Figure 1a and b). You can also use four external resistors to set two threshold levels around an R-S flip-flop (Figure 1c).

The following equations, respectively, describe the basic Schmitt trigger’s positive- and negative-threshold levels:

\[ V_+ = \frac{R_1}{R_2} V_{CC} + V_{TH} \left( 1 - \frac{R_1}{R_2} \right) \]

\[ V_- = V_{TH} \left( 1 - \frac{R_1}{R_2} \right) \]

In these equations, \( V_{TH} \) represents the input-voltage threshold of the CPLD/FPGA device, and \( V_{CC} \) is its power-supply voltage.

Based on the equivalent Schmitt-trigger circuit in Figure 1b, the low-cost resistance-capacitance oscillator in Figure 2 requires four external passive components. Resistor \( R \) and capacitor \( C \) set the circuit’s oscillation frequency. Note that the resistance values of \( R_1 \) and \( R_2 \) are measured at 10°C intervals over the circuit’s operating-temperature range.

To compensate for IC’s temperature variation, you can select from among several possible resistor-thermistor-network topologies. First, you need to select a suitable thermistor and characterize its resistance-versus-temperature variation. Because the MAX1714’s current-limit input pin feeds a relatively high input-impedance voltage-follower stage, this thermistor requires a high nominal resistance of 100 kΩ. Resistance-versus-temperature characteristics of inexpensive thermistors exhibit considerable nonlinearity, but one relatively simple approach to linearization involves paralleling the thermistor with a fixed resistor equal to the thermistor’s nominal resistance (Reference 1). In the network of Figure 1, \( R_i \) linearizes the thermistor, and \( R_2 \) and \( R_3 \) respectively set the slope and intercept of the current-limit-voltage-versus-temperature-characteristic curve.

To arrive at optimal values for \( R_2 \) and \( R_3 \), we prepared a spreadsheet incorporating the original current-limit-voltage-versus-temperature data and added columns for each of the network’s resistors, plus the thermistor specification sheet’s resistance-versus-temperature data. While observing the circuit’s temperature-versus-voltage transfer function, we varied the spreadsheet’s values for \( R_i \) and \( R_2 \) until the transfer function best approximated the measured current-limit-voltage-versus-temperature data. Finally, we constructed the circuit and tested it over the temperature range and noted that it yielded a reasonably flat response.

The curvature of the corrected output characteristic of Figure 2 (red trace) is intrinsic to the thermistor. Though not perfectly flat, the corrected curve represents a great improvement over the original (black trace) and is sufficient to meet the original design goal. You can achieve more precise compensation by selecting a different thermistor or by incorporating multiple thermistors.

**REFERENCE**

and R₂ must be larger than that of R. Listings 1 and 2 contain the circuit’s VHDL implementation and RTL architecture, respectively.

In Figure 3, an open-collector buffer provides the trigger for the basic Schmitt-trigger-retriggerable monostable circuit by discharging timing capacitor C. The circuit’s output pulse width approximately equals the time constant RC. Listing 3 shows the VHDL implementation and RTL architecture, respectively.

You can convert the retriggerable monostable into the nonretriggerable monostable in Figure 4 by using an open-collector NAND gate to discharge timing capacitor C. As long as the circuit’s output remains high during the timing interval, the system locks out external triggers. As in the previous circuit, the output pulse width approximately equals the time constant RC. Listing 4 contains the VHDL and RTL codes.

You can use the basic CPLD buffer-with-feedback circuit to provide hysteresis for a contact-debouncing circuit. In Figure 5, resistor R₄ provides contact-cleaning current, and R₁ and C form a low-pass filter to reduce noise that contact bounce generates. Component values vary depending on the application.

```
Entity Oscillator is
Port ( A : in std_logic;
      B : in std_logic;
      OUT : out std_logic
    );
end Oscillator;

architecture RTL of Oscillator is
begin
  A <= B;
  OUT <= not A;
end RTL;

Entity Monostable is
Port ( A : in std_logic;
      B : in std_logic;
      Trigger : in std_logic;
      C : out std_logic;
      OUT : out std_logic
    );
end Monostable;
architecture RTL of Monostable is
begin
  A <= B;
  OUT <= not A;
  C <= '0' when Trigger='1' else '2';
end RTL;

Port ( A : in std_logic;
      B : in std_logic;
      Trigger : in std_logic;
      C : out std_logic;
      OUT : out std_logic
    );
end Monostable;
architecture RTL of Monostable is
begin
  A <= B;
  OUT <= not A;
  C <= '0' when Trigger='1' and A='0' else '2';
end RTL;
```

Figure 2 Add a resistor and capacitor to a basic Schmitt-trigger buffer to form a low-cost oscillator.

Figure 3 An active-low buffer holds timing capacitor C discharged in this version of a retriggerable monostable multivibrator based on the Schmitt-trigger buffer.

Figure 4 A NAND gate locks out trigger pulses, forming a nonretriggerable, monostable circuit.

Figure 5 Use a Schmitt trigger to debounce switch contacts.