VFCs (voltage-to-frequency converters) were favorite circuits of the late analog gurus Bob Pease and Jim Williams (references 1 to 5). In tribute to them, this Design Idea reveals a circuit that provides good performance at a low price. You can obtain all of the parts for a few dollars from a local electronics shop.

The circuit has high input impedance, works with a single power supply, and connects directly to microcontrollers. The linearity error is less than 0.1% for frequencies as high as 700 kHz, and the dynamic range is 60 dB. The circuit exploits the integrator, comparator, and one-shot architecture (Figure 1). The output frequency is proportional to the input voltage: \[ f = \frac{1}{V_{CC} t_{OS}} V_{IN} \]

where \( V_{CC} \) is the power supply, 5V, and \( t_{OS} \) is the duration of the pulse that the one-shot generates, according to the equation \( t_{OS} = 0.7 \times R_{OS} \times C_{OS} \). You must filter and regulate the power supply, \( V_{CC} \). If the magnitude of the power supply changes, the slope of the calibration curve also changes. The components you use for the integrator, \( C_{INT} \) and \( R_{INT} \), do not participate in the equation so they need not be either accurate or stable. However, capacitors \( C_{INT} \) and \( C_{OS} \) must have low dielectric absorption.

You build a start-up circuit with switch \( S_1 \) and the timing network comprising \( R_1 \), \( C_1 \), and \( R_2 \). This step ensures that the circuit will oscillate with any value of input voltage. After you turn on the power supply, the switch stays closed for approximately 1 sec, keeping \( C_{INT} \) completely discharged. When the switch opens, \( C_{INT} \) starts charging by a fixed current, which the magnitude of the input voltage defines. The result is a rising ramp at the integrator’s output. When the ramp reaches 2.5V, \( IC_2 \) generates a pulse because 2.5V is the threshold level of the Schmitt trigger at the 1B input of \( IC_2 \). Because the pulse magnitude is larger than the input voltage, the current through \( C_{INT} \) reverses, and \( C_{INT} \) partially discharges (Figure 2).

When the pulse is over, the integrator starts another rising ramp, and the cycle repeats. Because of the built-in Schmitt trigger, the circuit requires no separate comparator IC. Most applica-

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**Figure 1** Three inexpensive ICs and a few passive components make a VFC with good linearity, speed, and dynamic range.

**Figure 2** Because the pulse magnitude is larger than the input voltage, the current through \( C_{INT} \) reverses, and \( C_{INT} \) partially discharges.
tions can go without any adjustment. You adjust the full-scale frequency using only the trimming potentiometer, which is part of $R_{\text{OS}}$ in Figure 1.

You can select different frequency spans (Table 1), each requiring its own values for $C_{\text{INT}}$ and $R_{\text{OS}}$. The spans have different linearity. The table shows the linearity error as a percentage of the full-scale frequency for 11 equally spaced values of the input value in a range from 2 mV to 2V.

### TABLE 1: PERFORMANCE AT DIFFERENT FREQUENCY SPANS

<table>
<thead>
<tr>
<th>Maximum frequency (kHz)</th>
<th>Duration of $t_{\text{os}}$ (μsec)</th>
<th>$R_{\text{OS}}$ value (kΩ)</th>
<th>$C_{\text{INT}}$ value (pF)</th>
<th>Linearity (% of full-scale)</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>8</td>
<td>57.2</td>
<td>400</td>
<td>±0.044</td>
</tr>
<tr>
<td>100</td>
<td>4</td>
<td>28.6</td>
<td>200</td>
<td>±0.056</td>
</tr>
<tr>
<td>200</td>
<td>2</td>
<td>14.2</td>
<td>100</td>
<td>±0.021</td>
</tr>
<tr>
<td>400</td>
<td>1</td>
<td>7.15</td>
<td>50</td>
<td>±0.031</td>
</tr>
<tr>
<td>600</td>
<td>0.67</td>
<td>4.77</td>
<td>33</td>
<td>±0.066</td>
</tr>
<tr>
<td>800</td>
<td>0.5</td>
<td>3.58</td>
<td>25</td>
<td>±0.11</td>
</tr>
<tr>
<td>1000</td>
<td>0.4</td>
<td>2.86</td>
<td>20</td>
<td>±0.42</td>
</tr>
</tbody>
</table>

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**REFERENCES**


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**Mains-driven zero-crossing detector uses only a few high-voltage parts**

Luca Matteini, Agliana, Italy

The circuit in this Design Idea generates a zero-crossing pulse off the ac mains and provides galvanic isolation. The falling edge of the output pulse happens at approximately 200 μsec before the zero crossing. You can use the circuit to safely stop the triggering of a thyristor gate, giving it time to properly turn off. The circuit generates short pulses only when the mains voltage is approximately 0V, thereby dissipating only 200 mW at 230V and a 50-Hz input.

The circuit charges capacitor $C_1$ up to the limit that 22V zener diode $D_3$ creates (Figure 1 and Reference 1).

You limit the input current with resistors $R_1$ and $R_5$. As the input-rectified voltage drops below the $C_1$ voltage, $Q_1$ starts conducting and generates a pulse a few hundreds of microseconds long. The coupling of $IC_1$ makes the response of $Q_1$ squarer. The rms operating voltage dictates the only requirement for $R_1$ and $R_5$. SMD, 1206-size resistors typically withstand 200V-rms operation. This design splits the input voltage between $R_1$ and $R_5$ for a total rating of 400V rms. $D_4$ limits the voltage across the bridge to 22V so that all of the sub-

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**NOTES:**

You can lower $D_3$ to an 18 to 20V part, permitting a lower voltage rating, such as 25V, for $C_1$. $R_1$ and $R_5$ must be 1206-size parts to withstand the rms voltage across them; 1206-size parts are usually rated for 200V.

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**Figure 1** This zero-crossing detector uses low-voltage parts and consumes little power.
sequent components can have lower voltage ratings. A 22V zener diode can clamp as high as 30V, so this design uses a 50V, 470-nF ceramic capacitor. Ceramic capacitors have better reliability than electrolytic or tantalum capacitors, especially at higher temperatures. If you prefer a cheaper and smaller 25V part, you can change the zener diode’s voltage to 18V and still have a good margin for safety. Use $R_4$ to limit the peak current in the LED. The primary limit on the LED current is the slope of the rectified ac input. The gradual slope doesn’t let $Q_1$ generate current spikes when it discharges $C_1$’s stored energy.

You can simulate the operation of the circuit in LTspice Version IV (Figure 2 and Reference 2). With a 230V input at 50 Hz, the simulation shows a 17-mA peak in the optocoupler LED. The simulation gives good results with inputs of 90 to 230V, both at 50 and 60 Hz. At 110V and a 60-Hz input, the LED current peak is 8.5 mA, so IC1 still works. If you need higher LED-drive currents, you can reduce the value of $R_4$, or increase the value of $C_1$.

Testing a physical circuit shows good correlation with the simulation (Figure 3). Driving the isolated output from a 5V logic supply yields a good pulse waveform (Trace 1). The mains input is fed to the scope with a 15V isolation transformer for safety (Trace 2). You can use the persistence feature of the oscilloscope to show the zero-crossing point when zooming in to the transition (Figure 4). This approach allows you to accurately measure the pulse timing relative to the input zero crossing.

**REFERENCES**


![Figure 2](image1.png) In this LTspice simulation, as the input voltage drops through 0V, the LED current makes a pulse whose edges lead and lag the crossing point. The peak optocoupler-LED current is 17 mA.

**WITH A 230V INPUT AT 50V, THE LTSPICE SIMULATION SHOWS A 17-mA PEAK IN THE OPTOCOUPLER LED.**

Figure 3 Results for a prototype circuit correlate well with the simulation.

Figure 4 You can use the oscilloscope’s persistence function to relate the exact zero-crossing point to the output-pulse timing.
You can use three discrete transistors to build an operational amplifier with an open-loop gain greater than 1 million (Figure 1). You bias the output at approximately one-half the supply voltage using the combined voltage drops across zener diode D1, the emitter-base voltage of input transistor Q1, and the 1V drop across 1-MΩ feedback resistor R2.

Resistor R3 and capacitor C1 form a compensation network that prevents the circuit from oscillating. The values in the figure still provide a good square-wave response. The ratio of R2 to R1 determines the inverting gain, which is −10 in this example.

You can configure this op amp as an active filter or as an oscillator. It drives a load of 1 kΩ. The square-wave response is good at 10 kHz, and the output reduces by 3 dB at 50 kHz. Set the 50-Hz low-frequency response with the values of the input and the output capacitors. You can raise the high-frequency response by using faster transistors and doing careful layout.
A diode ladder multiplies voltage under software control

William Grill, Riverhead Systems, Greeley, CO

The circuit in this Design Idea uses a Microchip 12F10 controller to drive a voltage multiplier ladder and a single pin to output status and to input a trigger signal you supply (Figure 1). When you trigger the signal, the software turns on a MOSFET to connect the multiplier output to a load. The microcontroller has an internal comparator with a 0.6V trip point. The circuit attenuates and feeds back the output voltage to this comparator.

Listing 1, which is available at www.edn.com/111201dia, shows the controller-based software, which stops the oscillator, driving the voltage multiplier when the internal comparator indicates that the output voltage has reached an upper limit. This circuit works in a wireless-monitor design, increasing the voltage, power, and range of a small periodic transmitter. It can provide 12 to 15V and 9 to 11 mA.

Processing begins when power is applied. The controller qualifies its Port 3 input on Pin 4. When at a logic high, logic is true, and the software code generates complementary PWM outputs on ports 4 and 5, which are pins 3 and 4, respectively.

These oscillations charge the ladder network. The controller outputs a low on the Port 2/Pin 5 status line, indicating that charging is under way. You choose the ratio of R1 and R2 so that the center node of the ladder is at 0.6V when the output voltage reaches the desired value. When the output reaches the final value, the controller puts the status pin in tristate mode, and the 20-kΩ resistor pulls the pin up to the power-rail voltage. Port 2 on Pin 5 then becomes an input.

When you pull this pin low, the microcontroller asserts Port 1 and Pin 6 high, turning on the P-channel MOSFET through Q2, and applies the output voltage on C4 to the load. Meanwhile, Port 1 and Pin 6 go high, shifting the lower pin of output capacitor C4 from ground to the power rail and adding a few volts to the output of the voltage ladder.

The program drives the complementary outputs at pins 2 and 3 with a 700-μsec PWM period with a 50% duty cycle. You can change the software code to vary these parameters. The controller has an internal 4-MHz oscillator and supports a user-settable reference block. The code continues to monitor the enable pin, C4’s voltage feedback, and the pump operation during the discharge to the load. You must set certain bits in the processor configuration for this code to work (Figure 2).

![Figure 1](Fig 1.eps) This circuit boosts 3V to a regulated 12V and connects the multiplied voltage to a load under software control.

![Configuration Bits](Fig 2.eps) Set these configuration bits for proper operation of the software code in the microcontroller.
A circuit that properly charges sealed lead-acid batteries ensures long, trouble-free service. Fig 1 is one such circuit; it provides the correct temperature-compensated charge voltage for batteries having from one to as many as 12 cells, regardless of the number of cells being charged.

The Fig 1 circuit furnishes an initial charging voltage of 2.5V per cell at 25°C to rapidly charge a battery. The charging current decreases as the battery charges, and when the current drops to 180 mA, the charging circuit reduces the output voltage to 2.35V per cell, floating the battery in a fully charged state. This lower voltage prevents the battery from overcharging, which would shorten its life.

The LM301A compares the voltage drop across R₁ with an 18-mV reference set by R₂. The comparator’s output controls the voltage regulator, forcing it to produce the lower float voltage when the battery-charging current passing through R₅ goes below 180 mA. The 150-mV difference between the charge and float voltages is set by the ratio of R₄ to R₅. The LEDs show the state of the circuit.

Temperature compensation helps prevent overcharging, particularly when a battery undergoes wide temperature changes while being charged. The LM334 temperature sensor should be placed near or on the battery to decrease the charging voltage by 4 mV/°C for each cell. Because batteries need more temperature compensation at lower temperatures, change R₅ to 30Ω for a TC of −5 mV/°C per cell if your application will see temperatures below −20°C.

When the circuit charges more than six cells, the additional voltage across the LM334 increases self-heating, so use a small heat sink and increase the resistance of R₅. Likewise, use higher resistances in series with the LEDs to avoid overloading the LM301A.

The charger’s input voltage must be filtered dc that is at least 3V higher than the maximum required output voltage: approximately 2.5V per cell. Choose a regulator for the maximum current needed: LM371 for 2A, LM350 for 4A, or LM338 for 8A. At 25°C and with no output load, adjust R₇ for a V_OUT of 7.05V, and adjust R₈ for a V_OUT of 14.1V.

Figure 1 This circuit charges lead-acid batteries by applying 2.5V per cell (at 25°C) and floats them at 2.35V when they are fully charged. Use an LM371 regulator for a 2A rating, an LM350 for 4A, or an LM338 for 8A.

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Charger extends lead-acid-battery life
Fran Hoffart, National Semiconductor Corp, Santa Clara, CA

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