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Use a TL431 shunt regulator to limit high ac input voltage

Todor Arsenov, STMicroelectronics, Prague, Czech Republic

Most isolated, offline SMPSs (switched-mode power supplies), including flyback, forward, and resonant, must operate at input voltages of 90 to 260V rms. Some cases even use line-to-line voltages of 400V rms±10%, leading to increased component-voltage ratings and, thus, increased cost of the overall design. In such cases, it is preferable to use input-limiting circuits, allowing you to increase the input voltage to 440V rms without damaging the power-supply components.

The circuit in **Figure 1** limits, or clamps, input-ac voltages higher than 260V rms to levels safe for the operation of the power MOSFET in an SMPS. The circuit employs MOSFET Q$_1$ working as a 100-Hz switch and shunt-regulator IC$_1$, a TL431CZ, setting the clamped high-voltage level by divider R$_2$ and R$_4$. The circuit uses the component values shown. The clamped output voltage is 360V dc, the input voltage is 260V rms, and the maximum input voltage is 440V rms. The circuit was tested at power levels of 5 to 10W.

At an input voltage of less than 260V rms, Point C is less than 2.5V, and IC$_1$ is off, sinking the minimum off-state cathode current. Zener diode D$_2$ breaks down to 15V, ensuring a stable on-state for Q$_1$. This operation is the normal condition of Q$_1$ at input voltages lower than 260V rms. Accordingly, at these voltage levels, the circuit works as a standard full-bridge rectifier under capacitive load C$_3$.

At an input voltage of 260V rms or greater, Point C becomes higher than 2.5V, and IC$_1$ turns on, diverting and sinking the current from D$_2$. The gate-to-source voltage of Q$_1$ drops to approximately 2V, and Q$_1$ switches off. Now, no current flows to charge bulk capacitor C$_3$ even if the D$_1$ bridge-rectifier diodes are forward-biased. The rectified input-ac voltage is higher than the voltage across C$_3$, but Q$_1$ is off, the loop is interrupted, and no current flows. Accordingly, the output-dc voltage across C$_3$ gets limited because no charging current is available.

When the rectified ac-input voltage starts decreasing, it eventually hits the 2.5V threshold level of Point C, and Q$_1$ again switches on. But current does not flow because the rectifier bridge's diodes are now reverse-biased; the rectified input-ac voltage is less than the voltage across C$_3$. The voltage across C$_3$ decreases at a rate that the output-power level determines. Eventually, the voltage across C$_3$ and the rectified input-ac voltage intersect at a level when the rectifier bridge's diodes get forward-biased. Q$_1$ is still on; therefore, charging current starts flowing. A short interval follows, during which both Q$_1$ and D$_1$ conduct. The short

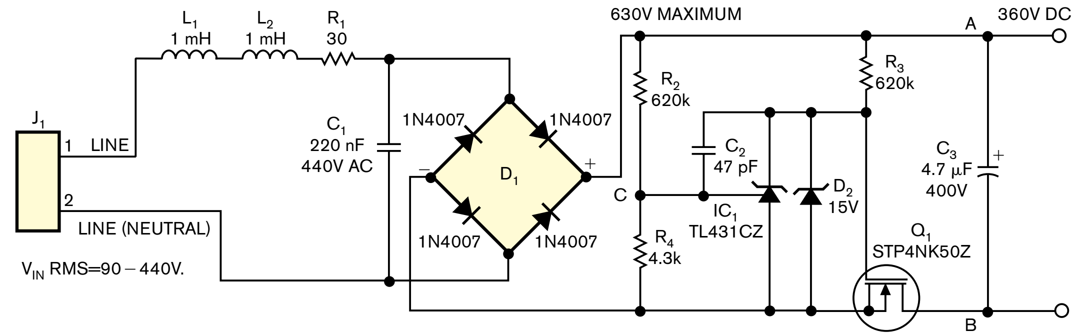

Figure 1 This simple circuit clamps input-ac voltages higher than 260V rms to levels safe for the operation of the power MOSFET in a switched-mode power supply.

charging pulses replenish the energy loss, increasing the voltage to the limited level. When the input voltage gets higher than 260V rms, Q_1 again switches off, and the whole process repeats.

Q_1 has small power dissipation. During every switching period, the MOSFET is on for only 450 µsec, resulting in high efficiency for this high-voltage-limiting circuit. You can use it as a MOSFET switch with the STMicroelectronics (www.st.com) SuperMesh MOSFET STP4NK50Z, which comes in a TO-220 package, but you can also use a Dpak to save space because the MOSFET is not a dissipative-voltage limiter. The current through Q_1 gets interrupted when the 50/60-Hz rectifying diodes are forward-biased. This current interruption causes ringing on the drain-to-source voltage. The clamping circuit passed the conducted EMI (electromagnetic-interference) tests, according to EN 55022 Class B, using peak and average detection. The 1-mH, 0.2A chokes, L_1 and L_2, suppress EMI. The 220-nF, 440V-ac capacitor, C_1, is a simple snubber element across the rectifying diodes of the D_1 bridge. EDN

Autozeroed amplifier with halved noise needs few components

Marián Štofka, Slovak University of Technology, Bratislava, Slovakia

The Analog Devices (www.analog.com) AD8553 autozeroed instrumentation amplifier has a unique architecture in that its two gain-setting resistors have no common junction (**Reference 1**). The first stage of the IC is a precise voltage-to-current converter, in which the first gain-setting resistor, R_1, sets the magnitude of the transconductance. The end stage of the IC is a precise current-to-voltage converter, in which the value of its feedback resistor, R_2, co-determines the overall voltage gain as $G=2(R_2/R_1)$. You can exploit the fact that the two gain-setting resistors are separate and that the input stage is a voltage-controlled current source to lower the component count in amplifiers with extreme noise-reduction demands.

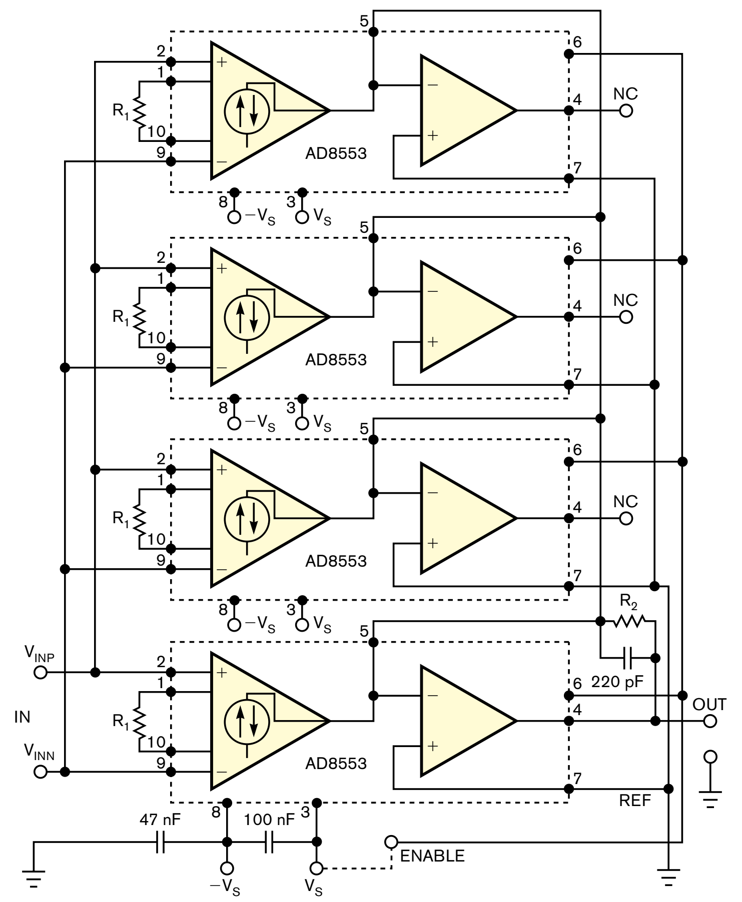

Figure 1 The unique architecture of the AD8553 instrumentation amplifier, incorporating an input-voltage-to-current converter, allows 50% noise reduction with fewer components.

You can use more amplifiers to reduce noise in two ways. First, assume that the sources of random noise in the amplifiers are mutually independent. Further, assume that the noise obeys a gaussian distribution. When averaging the outputs of classic voltage amplifiers, you can reduce the noise to a fraction of $1/\sqrt{N}$ by using N amplifiers and three times as many resistors (**Reference 2**). The internal structure of the AD8553 allows you to use just N+1 resistors for an almost-unlimited number of ICs operating in parallel. By paralleling the respective input pins of more ICs, the connected internal voltage-to-current sources easily operate in parallel (**Figure 1**). The microvolt-range input-voltage-offset mismatch at paralleled input pins of several ICs is harmless here because the output resistances of the voltage-to-current converters are theoretically infinite.

The net result of paralleling N input stages is that they output current of $N(V_{INP}-V_{INN})/(2R_1)$, or N times that of a single IC. You use only one of the current-to-voltage stages of the N ICs. That stage's feedback resistor has the value of R_2/N, where R_2 is the value for a desired voltage gain of A_V in a single IC. Because the primary source of noise in an amplifying IC is its input stage, you can assume that the standard deviation of the random component of output current of the paralleled-N voltage-to-current converters is $\sigma_{NI}=\sigma_I\times\sqrt{N}$, where σ_I is the standard deviation of the random component of output current of a voltage-to-current converter. These results differ from those in **Reference 2**, in which the authors perform noise reduction by averaging multiple voltages. On the other hand, the deterministic part of current at the common output of the voltage-to-current converters in **Figure 1** has the value of N times that of the single IC. The following **equation** calculates the RSNR (relative signal-to-noise ratio), which you define as the output current over the standard deviation of output noise: $RSNR_N=(N\times I)/(\sigma_I\times\sqrt{N})=\sqrt{N}\times RSNR_I$. It means that, in effect, the noise of the circuit has decreased to a fraction of $1/\sqrt{N}$ compared with that of a single IC. **EDN**

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Buck regulator controls white LED with optical feedback

Dhananjay V Gadre, Netaji Subhas Institute of Technology,
New Delhi, India

There is much interest in LED-based lighting due to the availability of high-power, high-efficiency white—and other-color—LEDs (**Reference 1**). Because an LED is a current-controlled device, typical control circuits regulate the current through the LED to maintain uniform intensity. To optimize available power, users often operate the LEDs with a switching-converter circuit—either a buck or a boost converter—depending on the input-dc voltage. **Figure 1** illustrates the configuration of typical buck- and boost-converter white-LED-driver circuits. Adding the resistance, R, in series with the white LED sets the current through the LED. The value of the resistance depends on the desired LED current and the feedback voltage that the buck/boost converter requires. For example, the required resistance is 12Ω for a 100-mA average current through the LED and a 1.23V feedback voltage.

To reduce the power dissipated in the series resistance, engineers often employ the circuit configurations in **Figure 2**. In this configuration, the amplifier's gain reduces the power dissipated in the series resistor by a factor equal to the gain (**Reference 2**).

The circuit configurations in **figures 1** and **2** work well in regulating

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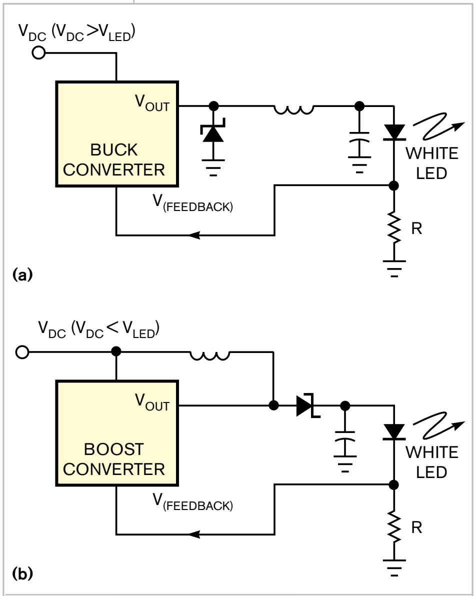

Figure 1 Depending on the input voltage, buck (a) or boost (b) regulators drive high-intensity white LEDs.

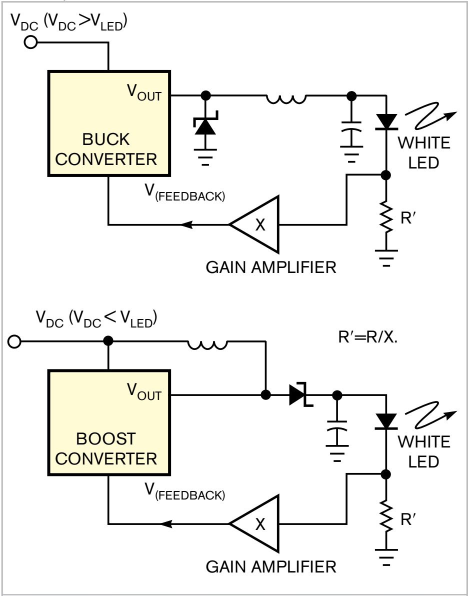

Figure 2 In these circuits, an amplifier gain reduces the power dissipated in the series resistor by a factor equal to the gain.

the current through the LED, provided that the ambient temperature remains constant. However, white and other-color LEDs exhibit significant variation in luminosity as a function of temperature (**references 2** and **3**). Typical figures for variation in luminosity range from 40 to 150% for a 100°C change in temperature. Thus, if you expect the ambient temperature to vary, regulating only the current through the LED is an inefficient way to control the LED. An alternative is to use optical feedback to control the LED (**Reference 3**).

However, rather than use an expensive light sensor and amplifier circuit, you can use a suitable LED as a light sensor (**Reference 4**). **Figure 3** illustrates a controller for a white LED using an inexpensive buck-regulator IC, an adjustable LM2575. A 3-mm red LED in a transparent package senses the light from a 10-mm white LED. The white-LED spectrum is wide enough to excite the red LED as a sensor. For a test current of 60 mA through the white LED, the red-LED-sensor voltage is approximately 40 mV. Because the circuit uses the red-sensor LED's voltage as a feedback to the buck regulator, you must use an amplifier with a gain of approximately 30 because the internal reference voltage of the LM2575 buck regulator is 1.23V. Resistors R_1, R_2, and R_3 control the gain of the amplifier, which comprises an inexpensive LM358 dual op amp. The input-dc voltage powers the op amp. Resistors R_1, R_2, and R_3 have values of 270, 560, and 10 kΩ, respectively. Because R_2 is a variable resistor, changing its setting changes the gain and, thus, the current through the white LED. Thus, R_2 acts as brightness control. The amplifier gain ranges from 28 to 84, depending on the setting of R_2.

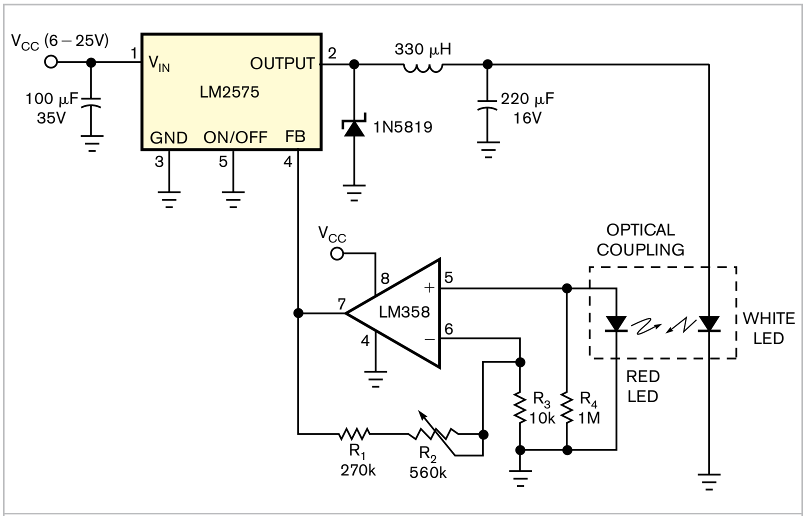

Figure 3 Use an inexpensive buck regulator and a red LED as a sensor for optical feedback to control the intensity of a white LED.

The red LED as a sensor mounts on the side of the white LED itself, thereby using only a fraction of the emitted light from the white LED. File the 3-mm red LED's top to get a flat surface, and then use a drop of superglue to secure the 3-mm red LED onto the side of the white LED.

The LM2575 buck regulator works by changing its duty cycle to regulate the output voltage. If the white-LED output light falls because of increased temperature, the red-LED sensor's voltage falls proportionally. The output of the red-LED sensor connects to the feedback input (Pin 4) of the regulator IC, and, in response, the regulator IC increases the duty cycle of the output voltage you apply to the white LED, thus stabilizing the light. In case of a decrease in ambient temperature, the white-LED light increases, and the regulator reduces the output voltage, which stabilizes the white-LED light.

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Routines directly measure microcontroller-bus clock

Kerry Erendson, Bulova Technologies

The Freescale HC08 and newer HCS08 microcontroller families have versatile peripheral modules. Their clock generators are no exceptions. They range from the internal clock, which frees I/O pins, to external crystals or oscillators. Once you select the timing source, you have many options for controlling the final bus frequency. For instance, connecting a 32,768-Hz crystal to an MC9S08GB microcontroller allows you to use the FLL (frequency-locked loop) to generate many bus frequencies as high as 18.874 MHz.

Selecting the source, the divisors, and the FLL settings allows versatility but also can get complicated.

Once you write the bus-clock-initialization routine, you may want to verify that the bus is running at the speed you intend before moving on to the rest of the project. This Design Idea presents routines that output a square wave at exactly one-tenth the bus speed on any I/O port (**listings 1** and **2**). Just connect a frequency counter to this pin, and it will display your bus frequency. All you have to do is move the decimal point one place to the right. Once you verify the bus speed, you can confidently write the timer, serial-I/O, and other clock-dependent routines.

You need to write code only to first disable interrupts and disable the COP (common on-chip processor). In your bus-clock-initialization routine, be sure to initialize the I/O port you want to use as an output. Then, just jump to the toggle clock, which outputs the bus frequency divided by 10 until power-down. This Design Idea uses PB0 in the HC08 version and PD0 in the HCS08 version. You can use any available I/O port by altering the first line to identify the port and the second line to choose a bit. Also, this Design Idea names ports with the older notation PB, instead of today's more fashionable PTB. **EDN**

LISTING 1 CODE FOR HC08

```
;TOGCLK - toggle PB0 at 1/10th the bus clock freq. (square wave)
;(NEVER ENDS)

TOGCLK   LDHX #PB          ;put 16-bit address of PB in H:X
         LDA #$01          ;make whatever bits in PB that will toggle=1
TOG01    CLR ,X            ;2
         NOP               ;1
         NOP               ;1
         NOP               ;1
         STA ,X            ;2
         BRA TOG01         ;3
```

LISTING 2 CODE FOR HCS08

```
;TOGCLK - toggle PD0 at 1/10th the bus clock freq. (square wave)
;(NEVER ENDS)

TOGCLK   LDHX #PD          ;put 16-bit address of port PD in H:X
         LDA #$01          ;make whatever bits in PD that will toggle=1
TOG01    STA ,X            ;2
         NOP               ;1
         CLR ,X            ;4
         BRA TOG01         ;3
```