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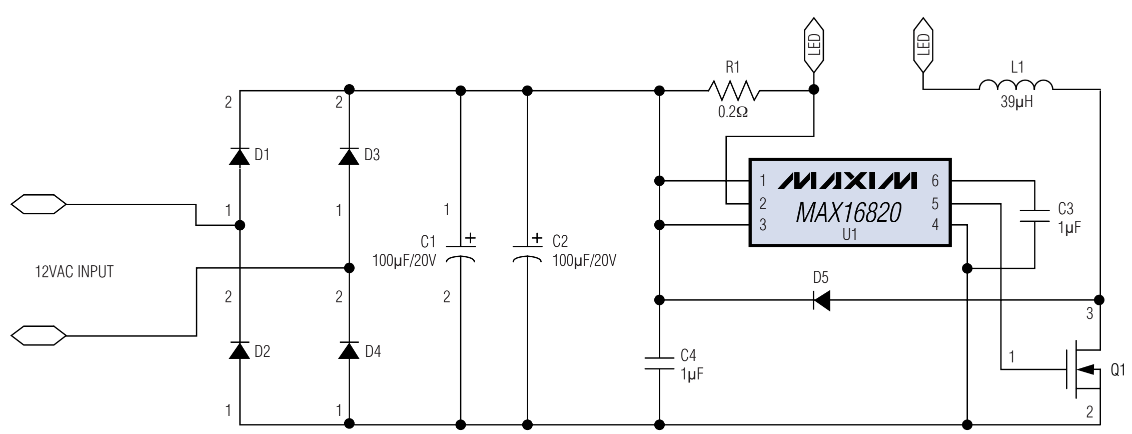

The 5W MR16 LED lamp circuit is shown using the MAX16820 LED driver. The LEDs shown are the LedEngin 5W WLEDs. *(See page 10.)*

Letter from the CEO

Maxim: A Technology Powerhouse

As we approach Maxim's 25th anniversary, it is timely to review our achievements and the advantages that they have provided to the electronics industry.

In the 80s, Maxim was known mostly for single-chip RS-232 solutions and microprocessor supervisors. In the 90s, we became a major force in power management, introducing a myriad of switching and linear supplies, and becoming the dominant supplier of the highest efficiency switchers to the notebook computer industry. Nearly everyone who reads this will have used a notebook with at least one of Maxim's power-management chips inside.

Over the years, we have developed into a large, very diverse operation with 26 major product lines, including: wireless RF, optical communications, audio, video, data conversion, and thermal management. You can see our entire list of product lines at www.maxim-ic.com. We also supply custom, large-scale integration devices to major players in very high-volume applications, such as cell phones, digital cameras, flat-panel television and PC monitors, notebook computers, and personal media players.

We began the company buying wafer production from foundries. We bought our first fab in 1989, enabling us to reduce costs, improve quality, and, most importantly, to develop exclusive processes to give our products unique performance capabilities. Now we have five fabs that provide 97% of our wafer production: San Antonio, TX; Dallas, TX; San Jose, CA; Beaverton, OR; and most recently, Irving, TX. We run over 160 proprietary wafer process technology variations in our four original fabs. This gives us tremendous potential for growth, and the confidence that we will be able to support your product delivery needs as your demands increase.

But manufacturing is not just about wafer fabrication. We have always had our own test operations, and now we have three large test facilities: one in Thailand and two in the Philippines. Test facilities in California, Oregon, and Texas are also operational, though over 98% of the job is done in our factories outside the USA.

Design, too, is spread around the world, with over 30 design centers in the USA, China, India, Japan, Korea, France, Italy, Turkey, the UK, and several other European countries. We also have sales offices in most of the industrialized world, with a total of 22 offices in 14 countries. So with all of our design, manufacturing, and sales facilities, our global operations have expanded to 18 countries.

We have distributors in 55 countries, plus we cover 15 of those countries with 6 branches of our own distribution company, Maxim Direct. Also, we supply to many other countries through our on-line sales service. Our customers include the most significant brands in consumer, computing, communications, and industrial markets. With over 35,000 customers in 55 countries, our reach is global.

Keeping the engineer in mind, we define, make, and test our own products in large quantities to very high standards. Having such a broad range of products and a global presence, Maxim is able to fulfill requirements for diverse geographical markets and numerous technical markets. We can be a partner with companies of any size, with any needs, in any location.

We are always at your service,

Tunç Doluca
President and Chief Executive Officer

Extend Battery Life in Handheld Video Systems by Reducing Video Subsystem Operating and Standby Power

Increasing numbers of portable devices, like digital still cameras, cell phones, and portable media players, are adding composite-video output connections. In such devices, a video filter amplifier follows the video digital-to-analog converter (DAC) that generates the video signal. Today's 3.3V video filter amplifiers consume 45mW of power while processing a video signal.

Battery life is an important issue with any portable equipment, and reducing the power consumption of the video system ICs is certainly preferred. As a consequence, the newest generation of video filter amplifiers, which operate from 1.8V, consume only 12mW—nearly a 70% reduction in power consumption.

Where Does All the Power Go?

In the simplest analysis, each circuit consumes power for its own operation and for driving the load. In **Figure 1**, the power supply provides total current (I_T) to the circuit, where I_Q is the quiescent current for the operational amplifier and I_L is the load current.

Power is calculated by multiplying the current by the supply voltage. The quiescent power consumption (P_Q), the load power consumption (P_L), and the total power consumption (P_T) are calculated to the first order with the following formulas:

$$P_Q = V_{DD} \times I_Q$$

$$P_L = V_{DD} \times I_L$$

$$P_T = P_Q + P_L = V_{DD} \times (I_Q + I_L)$$

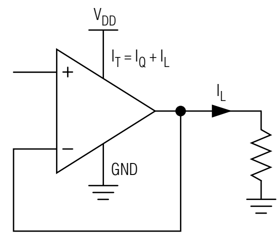

Figure 1. A single-supply operational amplifier is shown with a resistive load to ground.

To minimize actual power consumption in a real application, both P_Q and P_L must be reduced. Decreasing any combination of V_{DD}, I_Q, and I_L achieves this end.

Usually, IC data sheets provide specifications for only I_Q or P_Q; they almost never describe average power consumption with a typical signal and a typical load. P_Q is nearly useless information for any portable video filter amplifier, because the circuit is either in shutdown or fully enabled (defined as when the video filter amplifier drives a video signal into a video load). To conserve the battery, the video filter amplifier should be in shutdown if there is no video load; enabling the video filter amplifier when there is no video load wastes the battery.

Power Consumption in 3.3V Video Filter Amplifiers

When a 3.3V video filter amplifier drives a video signal into a video load, its power consumption increases, as shown in **Table 1**. Average power consumption is defined as the condition in which the video filter amplifier drives a 50% flat-field video signal into a 150Ω load to ground. The 50% flat-field signal, which appears as a gray screen on a television, is used as a proxy for a typical video signal. (The P_L depends on the picture content. A black screen requires the least power, while a white screen requires the most power.) Note how the average power consumption of the parts is quite similar, although their P_Q differs considerably.

Table 1. Average and Quiescent Power Consumption of Various Video Filter Amplifiers

Company	Part	Supply Voltage (V)	Average Current (mA)	Average Power (mW)	I_Q (mA)	P_Q (mW)	Output Style
Maxim	MAX9502	3.3	13.5	44.6	5.3	17.5	Positive DC bias
TI™	OPA360	3.3	12.2	40.1	6	19.8	Zero DC bias
Maxim	MAX9503	3.3	13.2	43.4	12	39.6	DirectDrive™

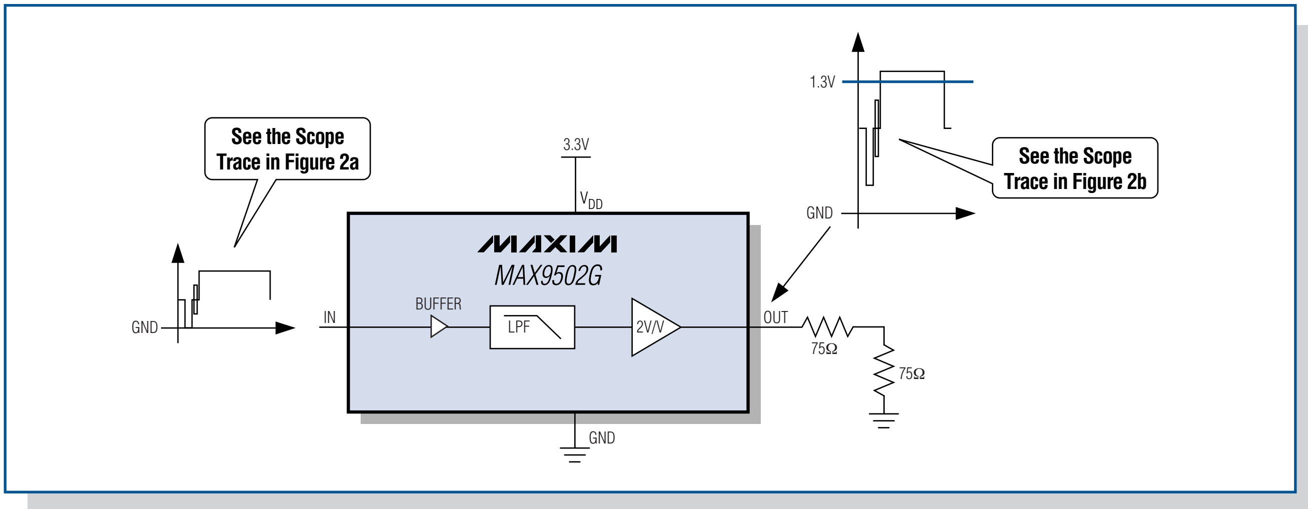

Figure 2. The MAX9502G application circuit shows the input and output of a 50% flat-field signal.

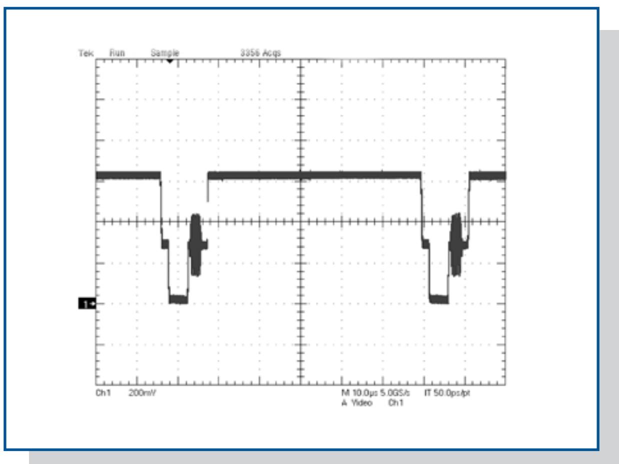

Figure 2a. This 50% flat-field waveform is input into the video filter amplifiers being considered.

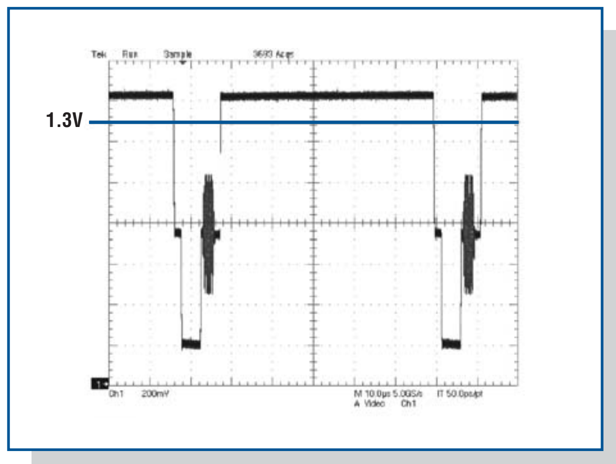

Figure 2b. In the MAX9502G's output waveform, the blue line indicates the approximate DC average of a 50% flat-field signal.

The increase in power consumption from driving a video signal into a video load greatly depends on the output style of the video amplifier. The MAX9502 outputs a video signal with a positive DC bias (see **Figure 2**). Maintaining the output signal's positive DC bias would cause an increase in overall power consumption. Therefore, the MAX9502 must source approximately 8.7mA (calculated by dividing the voltage, represented by the heavy blue line in **Figure 2b**, by 150Ω).

At its output, the OPA360 (in Table 1) can work with a SAG network, which is composed of two AC-coupling capacitors (**Figure 3**). These capacitors break the DC connection between the output and the load. As a result, the amplifier does not need to source or sink any current to maintain the bias at the output, thereby minimizing the power increase.

By using Maxim's DirectDrive technology, the MAX9503 outputs a video signal with close to zero DC bias, but does not require any AC-coupling capacitors (see **Figure 4**). This technology allows the MAX9503 to pull the output below ground because an on-chip, inverting charge pump creates a negative supply voltage. Although DirectDrive increases the P_Q, the MAX9503's average power consumption is in the same range as the MAX9502 and the OPA360 because the P_L is lower. The MAX9503 needs to source less current because the DC bias is close to ground.

The New Generation: 1.8V Video Filter Amplifiers

The MAX9509, the first part in Maxim's newest video filter amplifier family, dramatically reduces both average power consumption and P_Q, as can be seen in **Figure 5**. Its supply voltage (V_{DD}) has been reduced from 3.3V to 1.8V, which

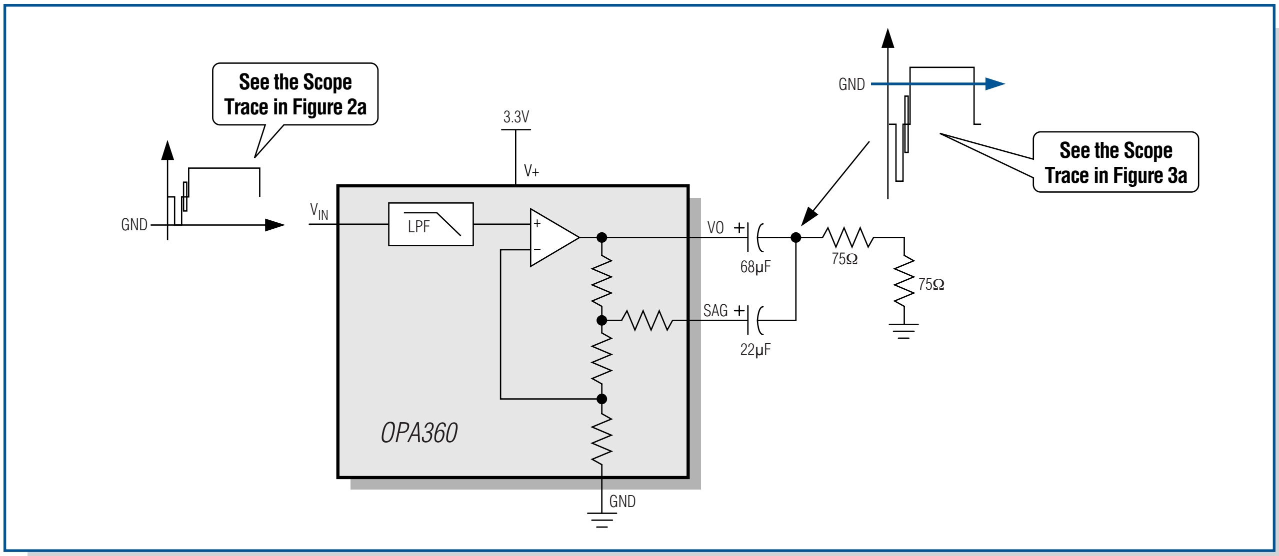

Figure 3. Given a 50% flat-field signal, the OPA360 application circuit minimizes power increase because the capacitors break the DC connection between the output and the load.

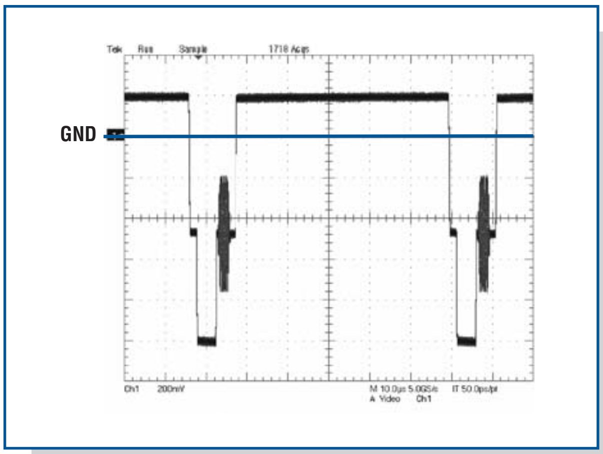

Figure 3a. The OPA360's output waveform contains a blue line indicating the approximate DC average of a 50% flat-field signal.

is the digital I/O voltage to which mobile phones are migrating. The quiescent supply current (I_Q) has also been reduced from 12mA to 3.1mA (see **Table 2**).

See the sidebar, *Circuit Considerations for 1.8V Video Filter Amplifiers*, for special issues to address when designing video circuits utilizing the new 1.8V digital I/O voltage.

DirectDrive is absolutely necessary when the video filter amplifier operates from a 1.8V supply. An amplifier with a voltage-mode output stage must swing at least $2V_{P-P}$ to output a composite video signal. A traditional amplifier operating from a single 1.8V supply does not have enough headroom to handle a $2V_{P-P}$ output signal. With DirectDrive, however, the integrated inverting charge pump creates a noisy -1.8V supply; a negative linear regulator then increases the -1.8V supply to -1V with minimal charge-pump noise. Therefore, with a supply voltage that spans -1V to +1.8V, the MAX9509 now has just barely enough headroom to output the $2V_{P-P}$ video signal.

With the combination of low supply voltage, low I_Q, and the DirectDrive output stage, the MAX9509's average power consumption (Table 2) is significantly lower than that of the 3.3V-generation devices seen in Table 1. What is remarkable is that the MAX9509's average power consumption is lower than the P_Q of the 3.3V video filter amplifiers.

One of the concerns with operating high-speed circuits at such low power is that the noise will greatly increase. This is because the circuits operate at lower current levels than normal. Noise was carefully considered during the MAX9509's design process, and this device has a very respectable peak signal-to-noise ratio (SNR) of 64dB, which is more than enough for consumer applications. To become visible on the television screen, the peak SNR needs to be around 40dB.

Having a noisy charge pump on the same die as the filter and amplifier was a major design concern. The charge pump could potentially inject switching noise into the delicate video waveform. The isolation between the MAX9509's charge pump and the video-signal-path circuitry is so effective that the charge-pump noise is not evident on a frequency sweep (**Figure 6**) and is barely noticeable in the time domain (**Figure 7**).

From the perspective of the consumer, neither wideband noise nor charge-pump noise are visible on a video screen displaying a signal output by the MAX9509.

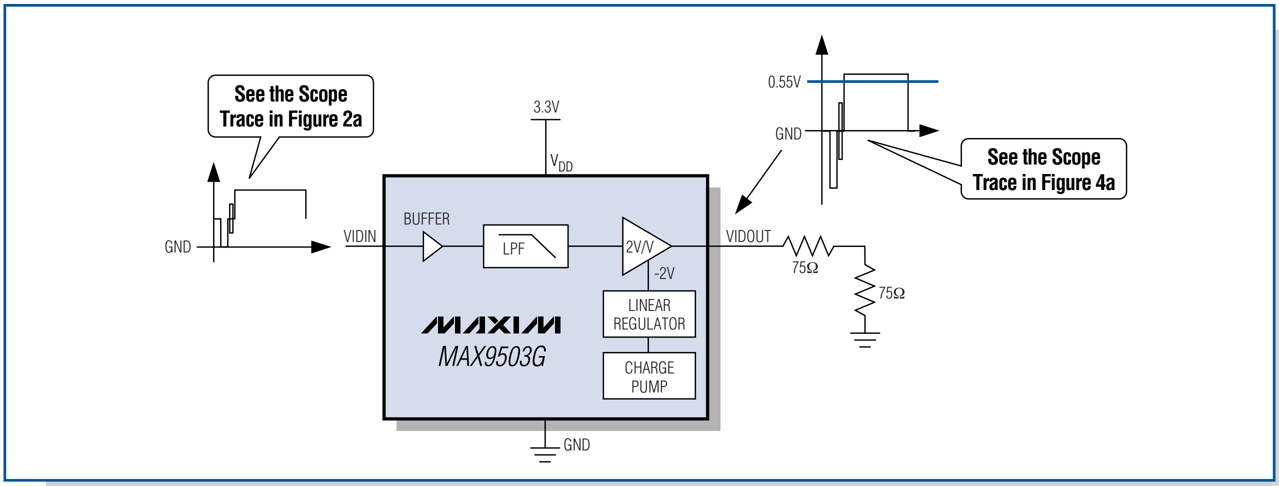

Figure 4. A 50% flat-field signal is processed through the MAX9503G application circuit.

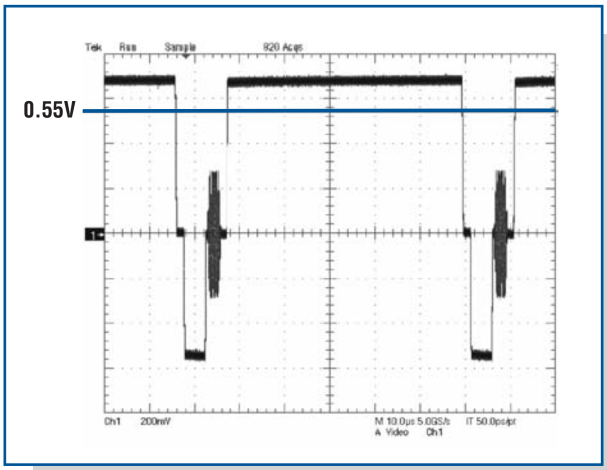

Figure 4a. The MAX9503G's output waveform has a blue line indicating the approximate DC average of a 50% flat-field signal.

Future Direction for Low-Power Video Filter Amplifiers

While recent developments have been made in low-power video filter amplifiers, IC designers still have work to do.

Consider video-load detection. If, for example, a video filter amplifier could electrically detect the load and provide load status to the microcontroller operating a system, then the video output circuitry could be turned on only when a valid video load was present. As a result, the system could manage video power more intelligently. The current method of video-load detection is to turn on the video output circuitry upon the mechanical detection of a jack insertion. This could waste battery power if the other end of the cable were not actually plugged into the jack of a television or other video monitor. A side benefit of electrical video-load detection is that it needs only a standard connector instead of a connector with mechanical jack sense, which adds cost and increases space compared to a standard connector.

Low power consumption has always been important in portable devices, but it is becoming more important even for wall-powered devices because of higher energy costs and concerns about global warming. Hence, the trend is for more intelligent power management to be integrated into analog chips. For the video filter amplifier, not only must it be low power, but it must have video-load detection, video-input detection, and control circuitry to cycle through operating modes. The greatest challenge will be to add intelligent power management without significantly increasing cost, as video chips are mostly used in competitively priced consumer electronics.

†U.S. Patent #7,061,327

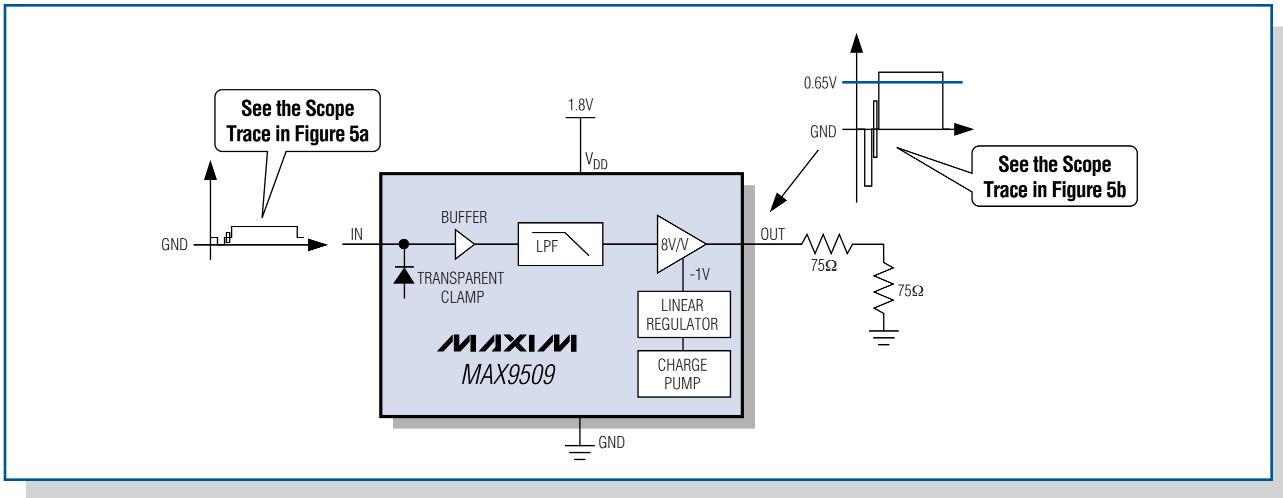

Figure 5. The MAX9509 1.8V application circuit processes a 50% flat-field signal, showing significantly reduced power consumption.

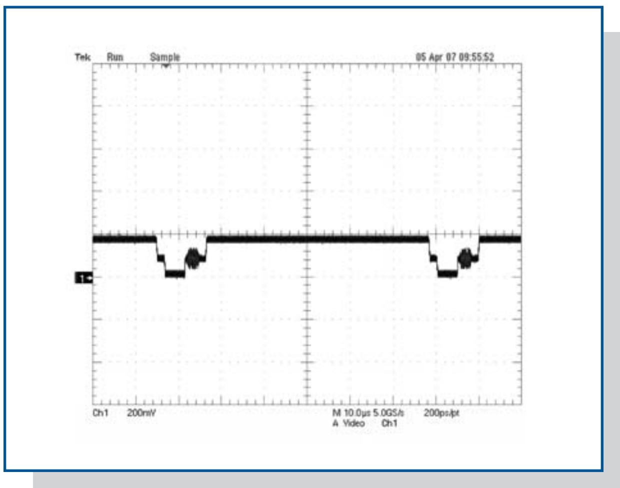

Figure 5a. A 50% flat-field waveform is input into the MAX9509; it has one-quarter the amplitude of the waveform used in Figure 2a.

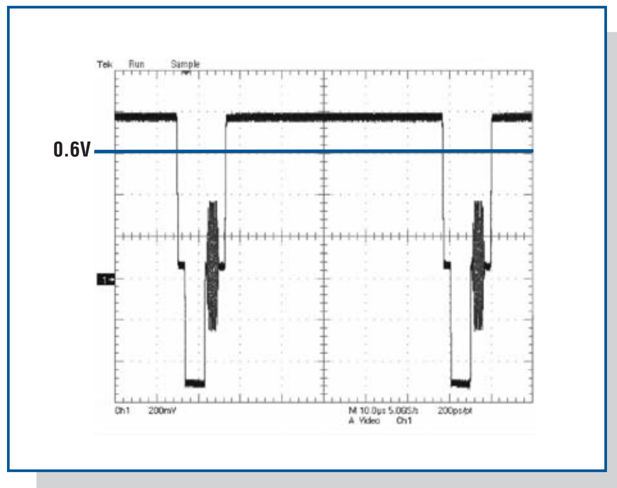

Figure 5b. In the MAX9509's output waveform, the blue line indicates the approximate DC average of a 50% flat-field signal.

Table 2. Average and Quiescent Power Consumption of the MAX9509

Company	Part	Supply Voltage (V)	Average Current (mA)	Average Power (mW)	I_Q (mA)	P_Q (mW)	Output Style
Maxim	MAX9509	1.8	6.5	11.7	3.1	6	DirectDrive

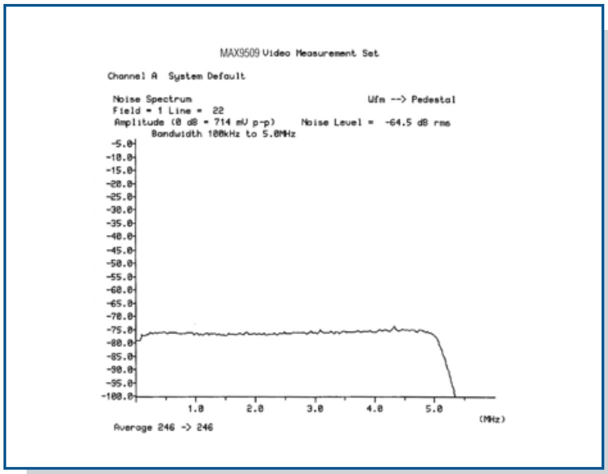

Figure 6. Charge-pump frequency spikes are not discernable when measuring the MAX9509's noise vs. frequency.

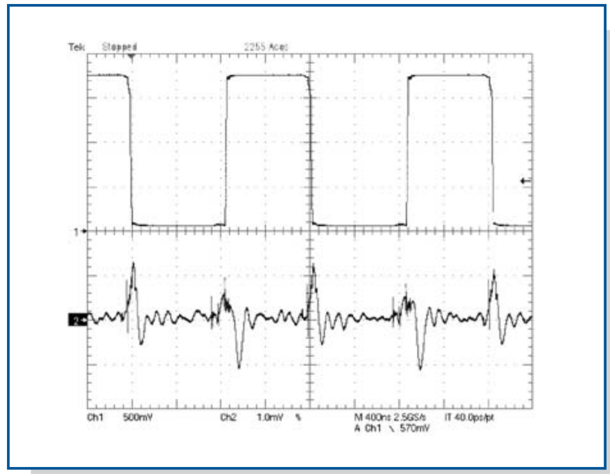

Figure 7. The MAX9509's output (bottom trace) is measured vs. time, taken with respect to a $1V_{P-P}$ video signal. The spikes are $1.4mV_{P-P}$. The top trace is the voltage on the top plate of the charge pump's flying capacitor.

Circuit Considerations for 1.8V Video Filter Amplifiers

When designing a low-power 1.8V video filter amplifier, special considerations must be made. Choose bias current values wisely to allocate the supply current where it is most productive. Careful layout techniques result in smaller parasitics and good device matching. Finally, carefully question and analyze the current used in all branches of the circuit. These efforts optimize power consumption, reducing the bandwidth of the circuitry to only what is required to maintain the necessary frequency response and video performance.

The power consumption of the MAX9509 is lower than in previous generations through careful circuit design and by taking advantage of Maxim's advanced BiCMOS process technology. All of the circuits from the previous generation of video designs were analyzed and optimized to result in the lowest power consumption while still maintaining adequate performance for the intended applications. For example, the number of times that a bias current is mirrored between supply rails in the MAX9509 was reduced; use of the generated negative rail was also minimized. Additionally, proprietary circuitry was applied to eliminate distortions that would otherwise be introduced by running the amplifier on such lean currents.

The MAX9509's power consumption was also reduced by Maxim's advanced analog process technology, which permits the optimal selection of components (e.g., bipolar vs. MOS) for the particular video signal path. The MAX9509's 5-pole filter eliminates an extra biquad filter stage needed for previous generations of amplifiers with 6-pole filters (**Figures 8** and **9**). The difference in filter specifications between a 5-pole and 6-pole filter is minimal for consumer applications, and the elimination of a biquad filter stage enables greater than 10% reduction in overall supply current.

Through careful partitioning of the filter and amplifier circuitry, the requirements of each block in the signal path can be optimized to use less total current for a given set of system specifications. For example, to achieve the gain of 8 in the MAX9509, a gain-of-4 preamplifier is used within the filter. Therefore, only a gain of 2V/V is required in the final video amplifier (Figure 9), thus reducing the requirements, and hence the power needs, of the final video amplifier. The overall power consumption of both amplifiers is low and optimized for the function being performed.

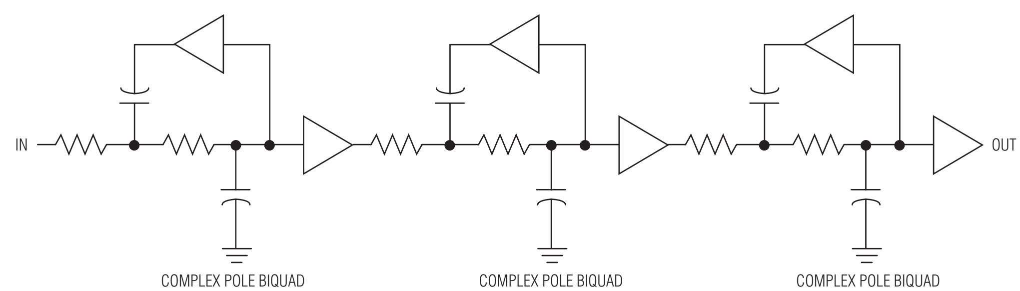

Figure 8. Previous generations of video filter amplifiers had 6-pole filters.

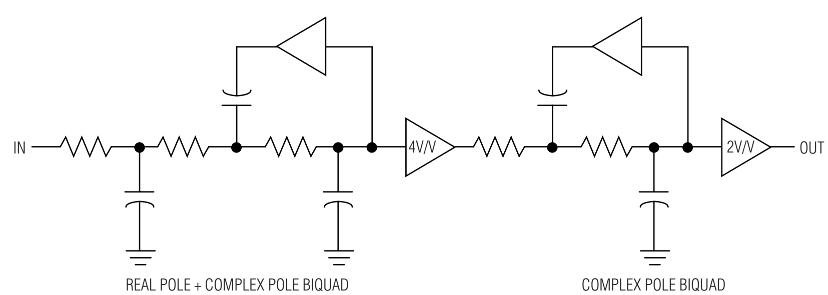

Figure 9. The MAX9509 has only a 5-pole filter, which eliminates a biquad filter stage, thus providing a 10% reduction in overall supply current.

Replace Inefficient MR16 Halogen Lamps with LEDs

Halogen MR16 lamps are widely used in professional store and home decorative lighting applications, though they have several disadvantages that limit their potential. The power dissipation of the most commonly used halogen MR16 lamps ranges from 10W to 50W, and their light output ranges from 150 lumens (lm) to 800lm. That equates to an efficacy of about 15lm/W, or a luminous efficiency of 15%. The lifetime of a typical halogen bulb is limited to about 2000hrs. Additionally, the filament should not be exposed to high levels of vibration to prevent the bulb from failing prematurely.

Today's LED technologies offer an MR16-compatible, solid-state, cost-effective alternative to halogen lamps. For example, the latest generation of 5W (single-chip, 4mm x 4mm package) and 10W (four-chip, 7mm x 7mm package) high-power LEDs from LedEngin™ generate typical efficacies of 45lm/W at 1000mA with a junction temperature (T_J) of +120°C. Under actual operating conditions, these specifications equate to typical lumen output levels of 155lm (at 1000mA, T_J = +120°C) for the 5W package and 345lm (at 700mA, T_J = +120°C) for the 10W package. When these LEDs perform at the same brightness level as halogen bulbs, power dissipation can be reduced by about 50%. In addition, LedEngin predicts a remarkable lumen maintenance of 90% (at 100khr, T_J = +120°C), thus eliminating the need for bulb replacement throughout the life of the product.

MR16 LED Reference Design

For the MR16 LED reference design shown in **Figure 1**, Maxim selected the LedEngin 5W white LEDs (WLEDs) to demonstrate the 1000mA drive capabilities of the MAX16820. **Tables 1** and **2** detail the parts list and electrical specifications of the MR16 reference design, which has a 12VAC ±10% input voltage typical of most MR16 applications.

Table 1. Parts List for the 5W MR16 LED Lamp Driver Circuit

Designation	Description
D1–D4	Rectifier diodes 1N4001
C1, C2	100µF/20V tantalum capacitors or one 220µF/25V electrolytic capacitor
C4	1µF/25V ceramic capacitor
R1	0.2Ω ±1% sense resistor IRC LRC-LR1206LF-01-R200-F
C3	1µF/6.3V ceramic capacitor
Q1	MOSFET FDN359BN
D5	Freewheeling diode FBR130
U1	MAX16820
L1	39µH/1.2A buck inductor Sumida CDRH6D38NP-390NC

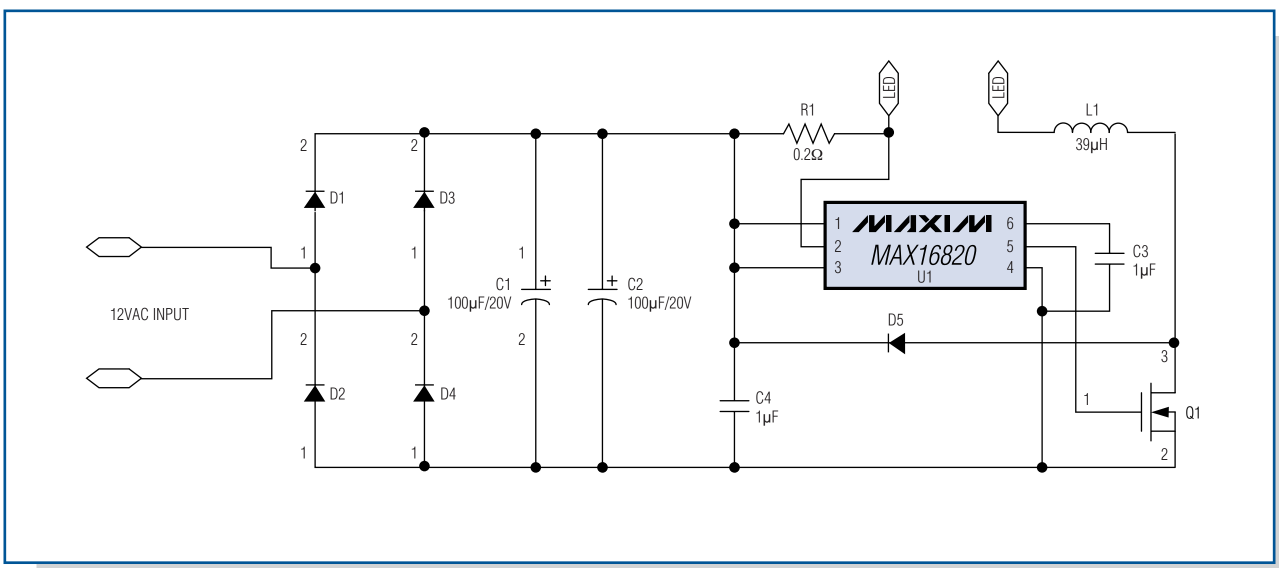

Figure 1. The 5W MR16 LED lamp circuit is shown using the MAX16820 LED driver. The LEDs shown are the LedEngin 5W WLEDs.

Table 2. Electrical Specifications for the 5W MR16 LED Lamp Driver Circuit

V_{IN} (min)	10.8VAC
V_{IN} (max)	13.2VAC
V_{LED} (min)	5V
V_{LED} (max)	3.1V
I_{LED}	1A
I_{LED} Tolerance	±15%
Open-LED Protection	Yes
Shorted-LED Protection	Yes

The MAX16820 has been specifically designed for LED driver applications targeting, among others, LED-based MR16s. The device was, therefore, the logical choice for the MR16 LED lamp circuit. The MAX16820 is available in a very small, 6-pin TDFN package, operates over a 4.5V to 28V input voltage range, and can drive external, cost-effective MOSFETs for a broad range of LED current-drive capabilities. It is specified over the wide, automotive operating temperature range (-40°C to +125°C), which allows the MAX16820 to be safely operated in the high-temperature environment of the MR16 light fixture. While the MAX16820 can control power levels up to 25W or even higher, its 2MHz (typ) switching frequency requires only small external inductors and capacitors, which allows the driver circuit to be placed in the MR16 fixture.

Figure 1 shows a 5W MR16 LED lamp driver composed of a rectifier bridge (D1–D4), 100µF filter capacitors (C1 and C2), and a buck converter circuit. The buck LED converter is composed of the MAX16820, buck inductor (L1), power MOSFET (Q1), freewheeling diode (D5), and sense resistor (R1).

5W high-brightness LEDs (HB LEDs) require 1A of drive current. The buck LED driver is designed to output 1A DC current. The hysteretic control method is used to control this buck inductor current which, in turn, provides the LED with its 1A current requirement. The hysteretic control implemented in MAX16820 results in a simple and very robust driver, delivering 5% LED current accuracy.

To make a 5W HB LED run at a constant 1A current for the entire line-frequency period, DC bus filter capacitors are added to limit the DC bus voltage ripple. The total capacitance should be at least 200µF provided by tantalum or electrolytic DC capacitors with a 220µF/25V rating for low cost.

To keep the accuracy of the output current high enough, the inductor current's maximum ΔI/ΔT should be limited to less than 0.4A/µs. As shown in Figure 1, the maximum voltage drop on the inductor is V_{L1MAX}. The following equations can be used to calculate the value of the inductor L1:

$$V_{L1MAX} - V_{AC\_IN} \text{ x } (1 + \delta) \text{ x } \sqrt{2} - V_O \qquad \text{Eq. 1}$$

$$L1 = \frac{V_{L1MAX}}{\Delta I / \Delta T} \qquad \text{Eq. 2}$$

For V_{AC_IN} = 12V, δ = 10%, and V_O = 3.6V, L1 must be greater than 37µH. Therefore, 39µH is the standard value chosen for L1, where δ is the allowed AC-input-variation percentage and V_O is LED forward voltage.

The design was tested using a LedEngin 5W WLED-based MR16 light fixture; **Figure 2** shows the setup. The bench-test waveforms for this design are shown in **Figures 3** through **6**. The input voltage is 12VAC (nom), and the output current ripple is approximately 10%.

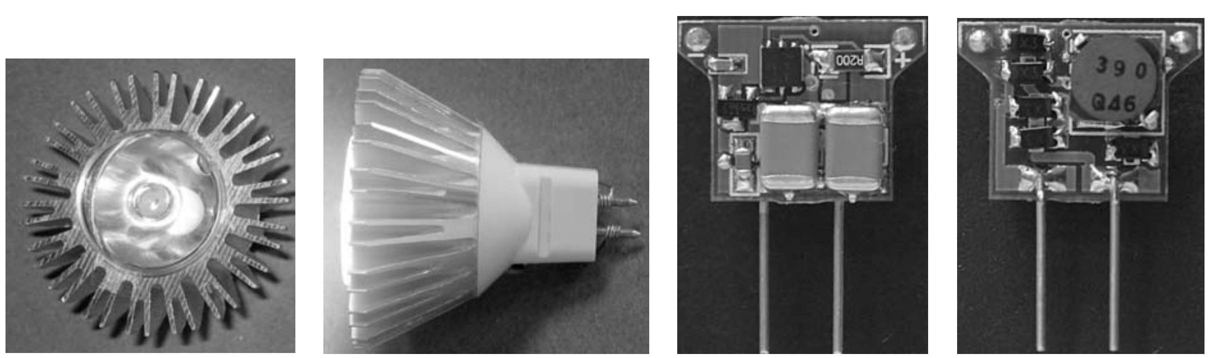

Figure 2. The LedEngin LED-based MR16 lamp has a very unique heatsink for dissipation of heat into the air. The MAX16820-based lamp driver board is placed just behind the heatsink.

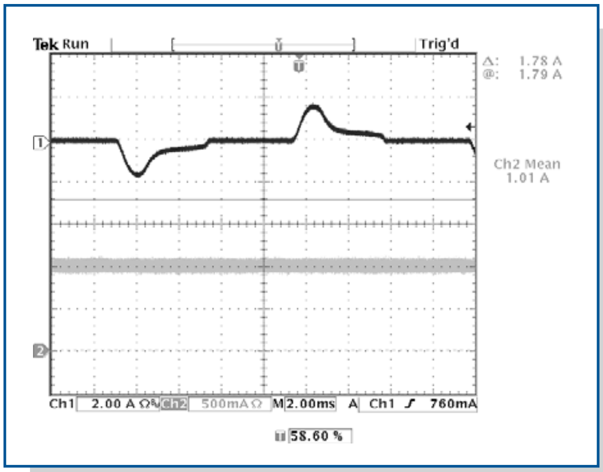

Figure 3. The first MR16 reference design bench test has the input AC current as CH1, and the output DC current as CH2.

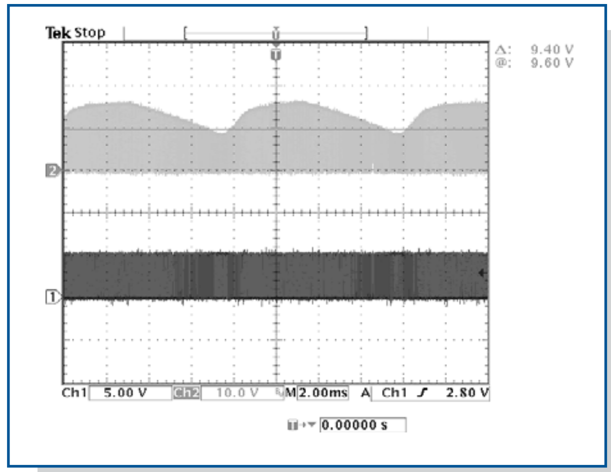

Figure 5. In this bench test, CH1 is the MOSFET gate-driver voltage envelope, and CH2 is the drain-source voltage envelope.

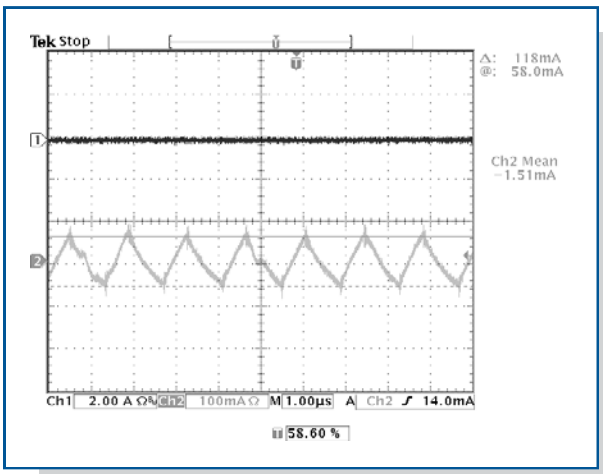

Figure 4. This detailed waveform has the output current ripple as CH2.

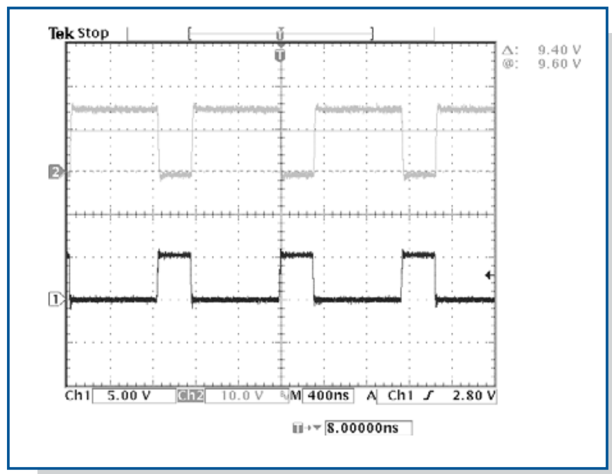

Figure 6. These detailed waveforms show the MOSFET gate driver as CH1, and the drain-source voltage as CH2.

Figure 4 shows that, with a 200µF DC filter capacitor, the DC-bus voltage ripple is 8.5V. The MAX16820-based hysteretic mode control is shown to have very good line-regulation performance. The output LED current has minimal variation as a result of the input-bus voltage. For the 5W MR16 LED lamp driver, the bench tests show that the AC-input ripple and variation can be more than 8.5V, while the output LED current is regulated to a constant 1A current.

The MR16 lamp driver PCB shown in **Figure 7** consists of two layers. All components are on both top and bottom layers, including the two AC-input connection pads and the two DC-output connection pads (labeled LED+ and LED-).

In HB LED applications, it is best to limit the junction temperature of the 5W LedEngin LED to less than +120°C when long-term lumen-maintenance performance of 90% after 100khr is required. Heatsinking is a low-cost solution to transfer the heat generated in the LED junction to the air. The 5W MR16 LED lamp has a heatsink to dissipate 5W of LED power. The 5W MR16 LED lamp driver PCB is mounted on the backside of the 5W MR16 LED lamp's heatsink.

Noteworthy is the unique heatsink design of the 5W MR16 LED lamp's assembly. Unlike in halogen-based assemblies where the lamp heat is primarily radiated to the environment, in LED-based designs the heat is conducted to the heatsink (such as the one shown in Figure 2) and then transferred to the surrounding air through convection.

Conclusion

When compared to other, lower power (1W and 3W) LED solutions, the high-power, 5W MR16 LED reference design significantly increases the amount of usable light.

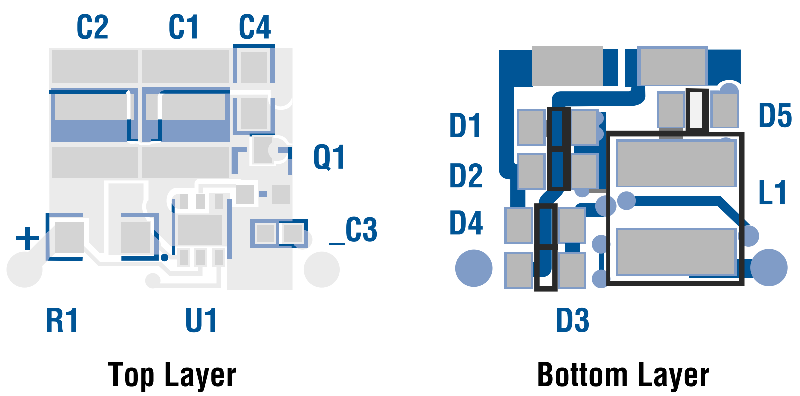

Figure 7. The LED+ and LED- DC-output connection pads can be seen on the 5W MR16 LED lamp driver PCB silk screen (top and bottom layers).

Therefore, this design eliminates the need for multiple emitter solutions required to meet the MR16 performance levels of a 10W halogen solution.

An Introduction to Switch-Mode Power Supplies

Considering the multiple DC voltage levels required by many electronic devices, designers need a way to convert standard power-source potentials into the voltages dictated by the load. Voltage conversion must be a versatile, efficient, reliable process. Switch-mode power supplies (SMPSs) are frequently used to provide the various levels of DC output power needed for modern applications, and are indispensable in achieving highly efficient, reliable DC-DC power-conversion systems.

Why SMPS?

The majority of electronic DC loads are supplied from standard power sources. Unfortunately, standard source voltages may not match the levels required by microprocessors, motors, LEDs, or other loads, especially when the source voltage is not regulated. Battery-powered devices are prime examples of the problem: the typical voltage of a standard Li+ cell or NiMH stack is either too high/low or drops too far during discharge to be used in conventional applications.

Versatility

Fortunately, the versatility of SMPSs solves the problem of converting a standard source voltage into a usable, specified output voltage. There are numerous SMPS topologies, which are classified into fundamental categories—these power supplies step up, step down, invert, or even step up and down the input voltage. Unlike linear regulators, which can only step down an input, SMPS are attractive because a topology can be selected to fit nearly any output voltage.

Customization

Additionally, modern SMPS ICs are designed with varying levels of integration, allowing the engineer to choose among topologies with more or less of the standard SMPS features brought into the IC. In doing so, manufacturers ease the design burden for commonly used, application-specific power supplies or offer the engineer basic SMPS ICs for custom projects, thereby enhancing the versatility of these widely used devices.

Efficiency

Engineers also face the other common problem of how to convert DC power efficiently. For instance, it is often required to step down an input voltage to achieve a lower output voltage. A simple solution implements a linear regulator, as this device requires only a few capacitors and adequate thermal management. However, where such simplicity ends, inefficiency begins—even to unacceptable levels if the voltage differential is large.

The efficiency of a linear regulator is directly related to the power dropped across its pass transistor. This power drop can be significant because dissipated power is equal to $I_{LDO} \times (V_{IN} - V_{OUT})$. For example, when stepping down a 100mA load from a 3.6V battery to a 1.8V output, 0.18W is dropped across the linear regulator. This power drop yields a low 50% efficiency, which reduces battery longevity by 50% (assuming ideal operation).

Understanding this efficiency loss, the dutiful engineer is driven to achieve an improved solution, and here is where the SMPS excels. A well-designed SMPS can achieve 90% efficiency or more, depending on load and voltage levels. As in the previous example, using the step-down SMPS of **Figure 1** instead of a linear regulator, 90% efficiency is observed. This is an efficiency improvement of 40% over the linear regulator. The advantage of the step-down SMPS is apparent, and similar or better efficiencies are observed in other SMPS topologies.

Although high efficiency is the principal advantage with SMPS designs, other benefits naturally occur as a direct result of minimizing power loss. For example, a reduced thermal footprint is observed in the SMPS when compared to its less efficient counterparts. This benefit equates to reduced thermal-management requirements. Also, more importantly, lifetime increases due to improved reliability, because components are not subjected to excessive heat, as they would be in a less efficient system.

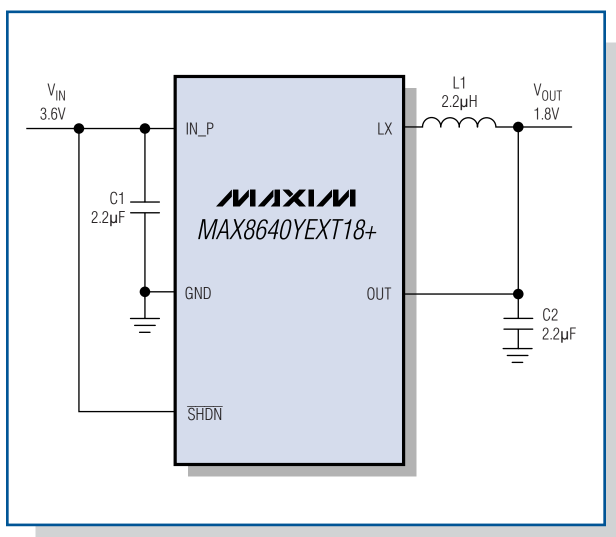

Figure 1. The MAX8640Y is used within a simple, step-down SMPS circuit.

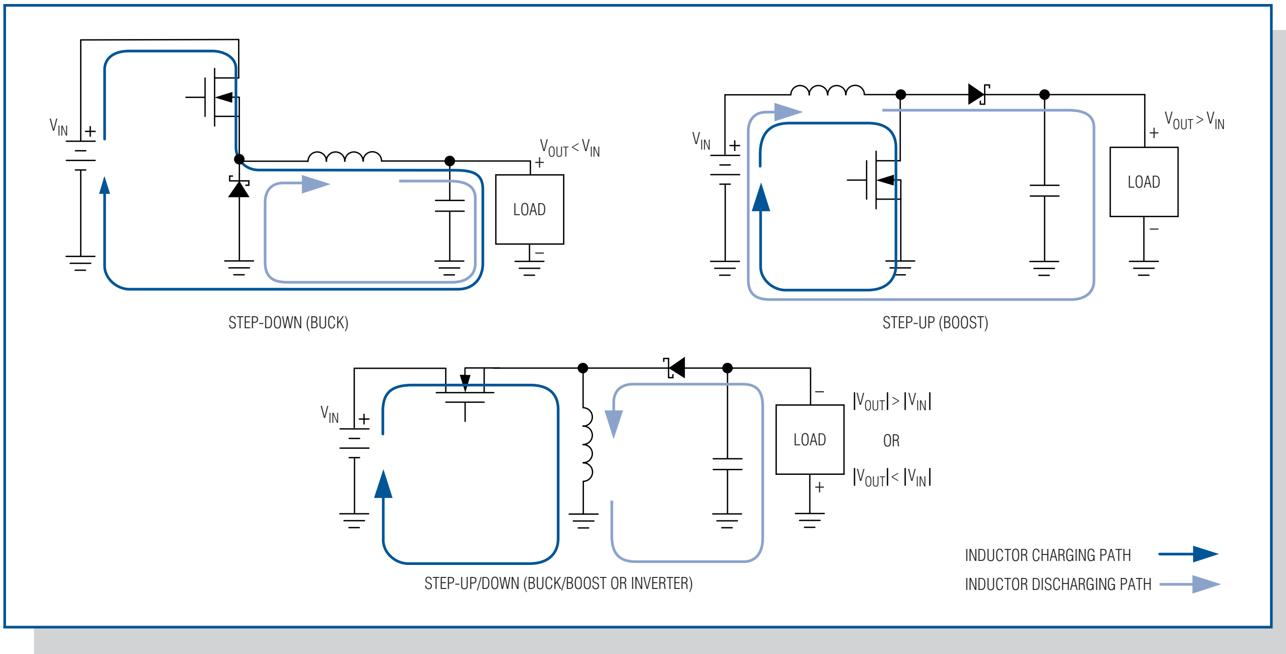

Figure 2. Buck, boost, and buck-boost compose the fundamental SMPS topologies.

SMPS Topologies and Conversion Theory

As mentioned in the previous section, SMPSs can convert a DC input voltage into a different DC output voltage, depending on the circuit topology. While there are numerous SMPS topologies used in the engineering world, three are fundamental and seen most often. These topologies (seen in **Figure 2**) are classified according to their conversion function: step-down (buck), step-up (boost), and step-up/down (buck-boost or inverter). The inductor charge/discharge paths included in the Figure 2 diagrams are discussed in the following paragraphs.

All three fundamental topologies include a MOSFET switch, a diode, an output capacitor, and an inductor. The MOSFET, which is the actively controlled component in the circuit, is interfaced to a controller (not shown). This controller applies a pulse-width-modulated (PWM) square-wave signal to the MOSFET's gate, thereby switching the device on and off. To maintain a constant output voltage, the controller senses the SMPS output voltage and varies the duty cycle (D) of the square-wave signal, dictating how long the MOSFET is on during each switching period (T_S). The value of D, which is the ratio of the square wave's on-time to its switching period (T_{ON}/T_S), directly affects the voltage observed at the SMPS output. This relationship is illustrated in equations 4 and 5.

The on and off states of the MOSFET divide the SMPS circuit into two phases: a charge phase and a discharge phase, both of which describe the energy transfer of the inductor (see the path loops in Figure 2). Energy stored in the inductor during the charging phase is transferred to the output load and capacitor during the discharge phase. The capacitor supports the load while the inductor is charging and sustains the output voltage. This cyclical transfer of energy between the circuit elements maintains the output voltage at the proper value, in accordance with its topology.

The inductor is central to the energy transfer from source to load during each switching cycle. Without it, the SMPS would not function when the MOSFET is switched. The energy (E) stored in an inductor (L) is dependent upon its current (I):

$$E = \frac{1}{2} \text{ x } L \text{ x } I^2 \qquad \text{Eq. 1}$$

Therefore, energy change in the inductor is gauged by the change in its current (ΔI_L), which is due to the voltage applied across it (V_L) over a specific time period (ΔT):

$$\Delta I_L = \frac{V_L \text{ x } \Delta T}{L} \qquad \text{Eq. 2}$$

The (ΔI_L) is a linear ramp, as a constant voltage is applied across the inductor during each switching phase (**Figure 3**). The inductor voltage during the switching phase can be determined by performing a Kirchoff's voltage loop, paying careful attention to polarities and V_{IN} / V_{OUT} relationships. For example, inductor voltage for the step-up converter during the discharge phase is $-(V_{OUT} - V_{IN})$. Because $V_{OUT} > V_{IN}$, the inductor voltage is negative.

During the charge phase, the MOSFET is on, the diode is reverse biased, and energy is transferred from the voltage

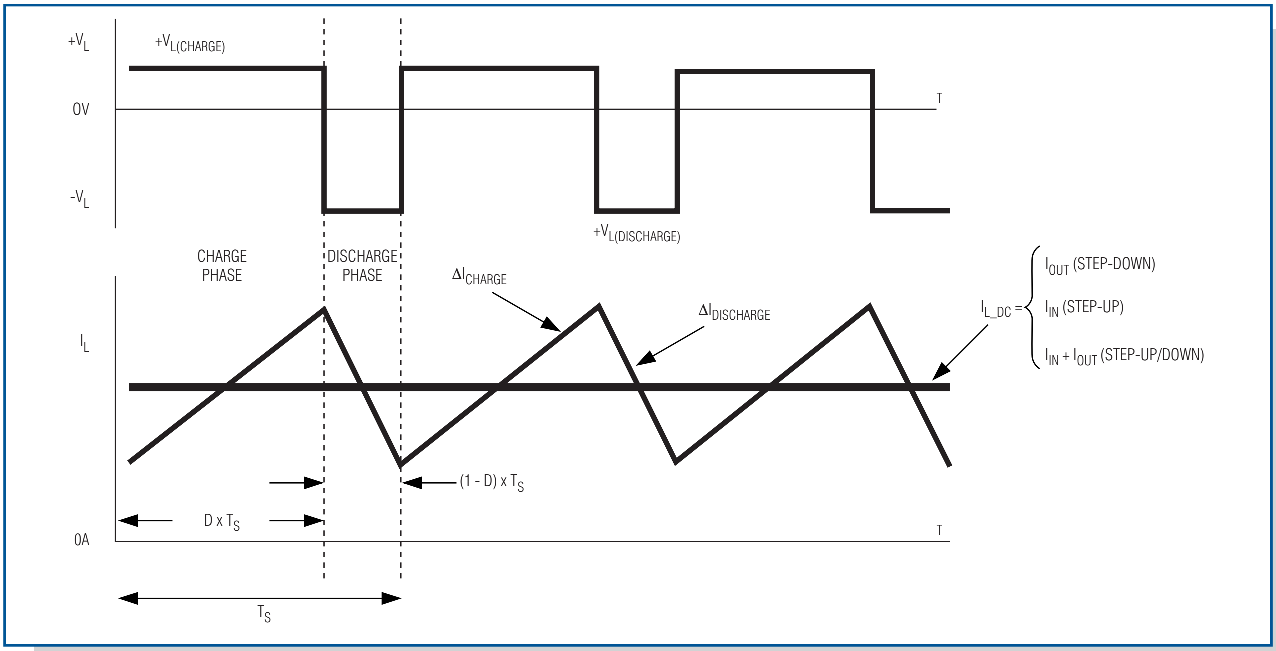

Figure 3. Voltage and current characteristics are detailed for a steady-state inductor.

source to the inductor (Figure 2). Inductor current ramps up because V_L is positive. Also, the output capacitance transfers the energy it stored from the previous cycle to the load in order to maintain a constant output voltage.

During the discharge phase, the MOSFET turns off, and the diode becomes forward biased and, therefore, conducts. Because the source is no longer charging the inductor, the inductor's terminals swap polarity as it discharges energy to the load and replenishes the output capacitor (Figure 2). The inductor current ramps down as it imparts energy, according to the same transfer relationship given previously.

The charge/discharge cycles repeat and maintain a steady-state switching condition. During the circuit's progression to a steady state, inductor current builds up to its final level, which is a superposition of DC current and the ramped AC current (or inductor ripple current) developed during the two circuit phases (Figure 3). The DC current level is related to output current, but depends on the position of the inductor in the SMPS circuit.

The ripple current must be filtered out by the SMPS in order to deliver true DC current to the output. This filtering action is accomplished by the output capacitor, which offers little opposition to the high-frequency AC current. The unwanted output-ripple current passes through the output capacitor, and maintains the capacitor's charge as the current passes to ground. Thus, the output capacitor also stabilizes the output voltage. In nonideal applications, however, equivalent series resistance (ESR) of the output capacitor causes output-voltage ripple proportional to the ripple current that flows through it.

So, in summary, energy is shuttled between the source, the inductor, and the output capacitor to maintain a constant output voltage and to supply the load. But, how does the SMPS's energy transfer determine its output voltage-conversion ratio? This ratio is easily calculated when steady state is understood as it applies to periodic waveforms.

To be in a steady state, a variable that repeats with period T_S must be equal at the beginning and end of each period. Because inductor current is periodic due to the charge and discharge phases described previously, the inductor current at the beginning of the PWM period must equal inductor current at the end. This means that the change in inductor current during the charge phase (ΔI_{CHARGE}) must equal the change in inductor current during the discharge phase ($\Delta I_{DISCHARGE}$). Equating the change in inductor current for the charge and discharge phases, an interesting result is achieved, which is also referred to as the volt-second rule:

$$\left|\Delta I_{CHARGE}\right| = \left|\Delta I_{DISCHARGE}\right|$$

$$\left|\frac{V_{L(CHARGE)} \text{ x } D \text{ x } T_s}{L}\right| = \left|\frac{V_{L(DISCHARGE)} \text{ x } (1 - D) \text{ x } T_s}{L}\right| \quad \text{Eq. 3}$$

$$\left|V_{L(CHARGE)}\right| \text{x } D \text{ x } T_s = \left|V_{L(DISCHARGE)}\right| \text{x } (1 - D) \text{ x } T_s$$

Simply put, the inductor voltage-time product during each circuit phase is equal. This means that, by observing the

SMPS circuits of Figure 2, the ideal steady-state voltage-/current-conversion ratios can be found with little effort. For the step-down circuit, a Kirchhoff's voltage loop around the charge phase circuit reveals that inductor voltage is the difference between V_{IN} and V_{OUT}. Likewise, inductor voltage during the discharge phase circuit is $-V_{OUT}$. Using the volt-second rule from equation 3, the following voltage-conversion ratio is determined:

$$\left|V_{IN} - V_{OUT}\right| \times D = \left|-V_{OUT}\right| \times (I - D)$$

$$\frac{V_{OUT}}{V_{IN}} = D \qquad \text{Eq. 4}$$

Further, input power (P_{IN}) equals output power (P_{OUT}) in an ideal circuit. Thus, the current-conversion ratio is found:

$$P_{IN} = P_{OUT}$$

$$I_{IN} \times V_{IN} = I_{OUT} \times V_{OUT} \qquad \text{Eq. 5}$$

$$\frac{I_{IN}}{I_{OUT}} = \frac{V_{OUT}}{V_{IN}} = D$$

From these results, it is seen that the step-down converter reduces V_{IN} by a factor of D, while input current is a D-multiple of load current. **Table 1** lists the conversion ratios for the topologies depicted in Figure 2. Generally, all SMPS conversion ratios can be found with the method used to solve equations 3 and 5, though complex topologies can be more difficult to analyze.

Table 1. SMPS Conversion Ratios

Topology	Voltage-Conversion Ratio	Current-Conversion Ratio
Step-down	$V_{OUT} / V_{IN} = D$	$I_{IN} / I_{OUT} = D$
Step-up	$V_{OUT} / V_{IN} = 1 / (1 - D)$	$I_{IN} / I_{OUT} = 1 / (1 - D)$
Step-up/down	$V_{OUT} / V_{IN} = D / (1 - D)$	$I_{IN} / I_{OUT} = D / (1 - D)$

Disadvantages and Tradeoffs of SMPSs

Of course, the high efficiency afforded by SMPSs is not without its penalties. Perhaps the most often cited issue regarding switch-mode converters is their propensity to radiate electromagnetic interference (EMI) and conduct noise. Electromagnetic radiation is caused by the fast transitions of current- and voltage-switching waveforms that exist in SMPS circuits. Rapidly changing voltages at the inductor node cause radiated electric fields, while fast-switching currents of the charge/discharge loops produce magnetic fields. Conducted noise, however, is propagated to input and output circuits when SMPS input/output capacitances and PCB parasitics present higher impedances to switching currents. Fortunately, good component placement and PCB layout techniques can successfully combat EMI and reduce noise.

SMPSs also can be quite complex and require additional external components, both of which can equate to an increase in overall cost of the power supply. Fortunately, most SMPS IC manufacturers provide detailed literature not only about device operation, but also about choosing correct external components. Additionally, the high levels of integration in modern SMPS ICs can reduce the number of external components required.

Despite these issues, SMPSs are widely used in numerous applications. The disadvantages can be managed, and the efficiency and versatility gained from their use is very desirable, and often required.

DESIGN SHOWCASE

Low-Cost Power-Supply Sequencer

Most point-of-load DC-DC converters can be powered up sequentially by wiring the power-good output of one converter to the enable input of the next. That approach works well for simple designs, but it does not satisfy a requirement of many modern microprocessors and DSPs—that the power-supply rails be sequenced in reverse order during power-down. Various vendors provide programmable-sequencing ICs for that purpose, but these parts are usually too expensive for cost-sensitive applications.

An alternative to programmable-sequencing ICs, the circuit of **Figure 1** can sequence and monitor four power-supply rails economically and effectively. Four DC-DC power supplies individually provide the application circuit with 3.3V, 2.5V, 1.8V, and 1.2V, respectively. A quad supervisor circuit (U1) monitors each rail and generates the master power-OK (POK) signal. Also during power-up, U1 ensures that the next supply in the sequence does not turn on until the preceding supply voltage is valid. A second quad supervisor (U2) creates the power-up and power-down sequences using an RC circuit consisting of R1, R2, R3, and C1. External resistive dividers are not necessary, because each supervisor has internally set thresholds.

A power-up sequence is initiated by connecting the power on/off signal to the 5V input, which causes C1 to charge through R2. As the capacitor voltage slowly exceeds 1.2V, then 1.8V, 2.5V, and 3.3V, each corresponding U2 output floats, thereby allowing the power supplies to turn on in a

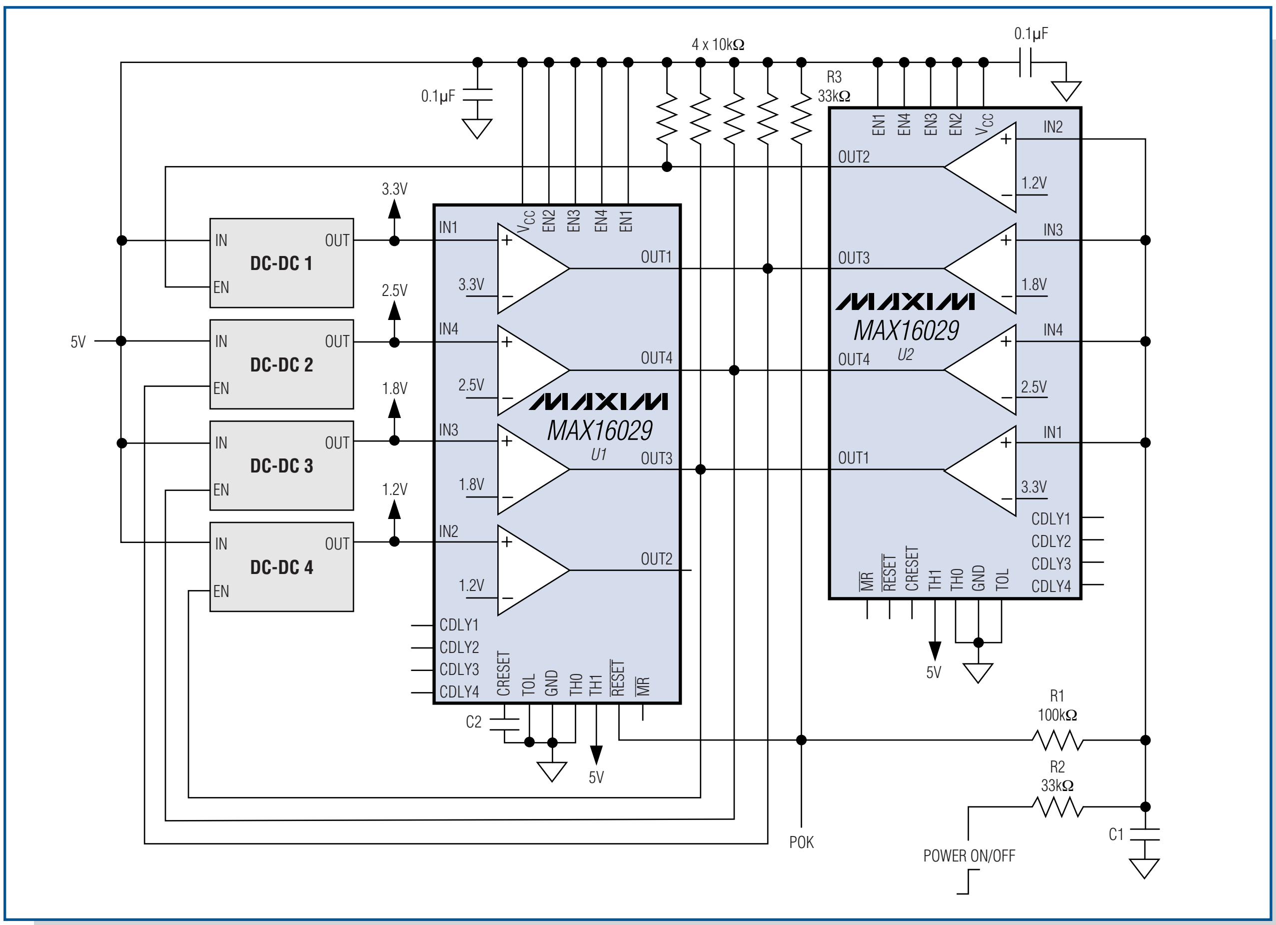

Figure 1. Using inexpensive ICs, this circuit first applies the four supply voltages in a specified order at power-up, then applies them in reverse order at power-down.

prescribed sequence. After all four supplies are on, the POK signal asserts following a timing delay set by C2.

To monitor the supply rails, allow the power on/off signal to float high. The POK signal then maintains the C1 voltage through R1 and R3, and keeps the power supplies on. In response to a fault, POK de-asserts rapidly, which discharges C1 through R1 and shuts off all the supplies. To power down, connect the power on/off signal to ground. C1 discharges through R2 and R1 when POK de-asserts, turning off each power supply in reverse order (**Figure 2**).

A similar article appeared in the December 11, 2006 issue of EDN.

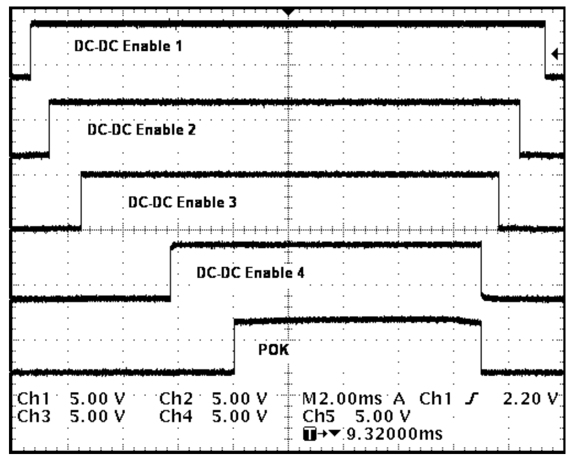

Figure 2. Beginning with DC-DC converter 1, the circuit of Figure 1 switches on three additional converters in sequence and generates a POK signal. Pulling the circuit's on/off input low removes the POK signal and switches off all four converters in reverse order.