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Reliable, Efficient LED Backlighting for Large LCD Displays

by Hua (Walker) Bai

Introduction

LEDs are rapidly becoming the preferred light source for large LCD displays in computers, TVs, navigation systems, and various automotive and consumer products. LEDs offer several benefits over fluorescent tubes: high luminous efficacy (lm/W), more vivid colors, tunable white point, reduced motion artifacts, low voltage operation and low EMI. However, system engineers face certain problems associated with driving LEDs for LCD backlight applications, including effectively providing sufficient power, regulating the LED current, matching current in multiple LED strings, achieving high LED dimming ratios, and fast LED current turn on/off.

All of these issues can be easily addressed in compact and reliable circuits that use the LT3476 high current LED driver and LT3003 3-channel ballaster.

The LT3476 is a quad output, current mode DC/DC converter operating as a constant current source with up to 96% efficiency. It is ideal for driving up to 1A of current for up to eight-per-channel RGB or white LEDs (such as Luxeon III) in series. That results in a total output power of about 100W.

The LT3003 is a 3-channel LED current ballaster, which can be used to triple the number of LEDs driven by a single LT3476 channel. When LED

strings are in parallel, special care is required to ensure safe operation and accurate current matching. Otherwise, one string will almost always draw much more current and eventually be damaged. The LT3003 can be used

System engineers face a number of problems when designing LED backlights for LCD backlight applications—such as effectively providing sufficient power, accurately regulating the LED current, matching current in multiple LED strings, achieving high LED dimming ratios, and fast LED current turn on/off.

All of these issues can be easily addressed in compact and reliable circuits that use the LT3476 high current LED driver and LT3003 3-channel ballaster.

with the LT3476 or other LED drivers to regulate current in the LED strings. This is one way to reduce the per-LED current and increase brightness uniformity across a large display. For

continued on page 3

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Linear Technology in the News...

Linear Products Highlighted

Linear Technology products were honored by the readers and editors of *EE Times*, who selected three of the company's products for Ultimate Products Awards:

LTC4089 USB Power Manager

with High Voltage Switching Charger

According to *EE Times*' analog IC editor, "Linear Technology's LTC4089, a 1.2-amp battery charger and USB power manager, offers high-voltage DC input capability and 10-12 percent higher operating efficiency over competing devices. Key to the charger's operation (which succeeds the company's 4055 and 4066 first- and second-generation lithium-ion chargers) is the LTC4089's Bat-Track adaptive output control technology, which tracks the battery's voltage in order to charge it at highest efficiency."

EE Times' readers said:

- "LT does a good job of constantly improving their parts."
- "It's the wave of the future."

LTC1408 6-Channel, 14-Bit, 600ksps

Simultaneous Sampling ADC with Shutdown

EE Times' editor commented, "Simultaneous sampling is what makes these A/D converters stand out. There are only a small number of A/D converters that can do simultaneous sampling." This is a dedicated architecture for the application because it requires a lot more circuitry," said Todd Nelson, product marketing manager of LTC's mixed-signal products."

EE Times' readers said:

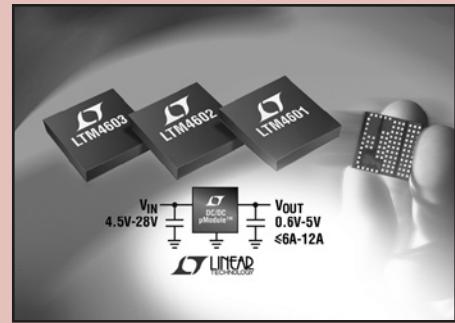
- "Another good product."
- "Small size, low power, sleep mode and simultaneous sampling are a great combination for the low power devices I design. Good subsampling performance into the low MHz range is key for me."

LT5560 0.01MHz to 4GHz Low Power Active Mixer

EE Times' RF editor stated, "Operable from VLF into the microwave regime, new mixer ICs from Linear Technology Corp. (LTC) can be impedance-matched over a range of frequencies. They're useful from 10-kHz to as high as 4-GHz. LTC's LT5560 mixers accommodate 50-ohm unbalanced signal inputs, but can also be driven differentially, including mixer injection. At higher frequencies you can match these mixers using baluns or wideband input-transformers. At very low IF (intermediate frequency) frequencies, coupling transformers can be large, however, so differential to single-ended conversion can be accomplished using an op-amp."

Linear Extends µModule Family

In November, Linear made a worldwide announcement of a significant extension of its family of µModule™ products. This announcement followed last year's announcement of the company's new LTM4600 power µModule family. These products, which feature the quality and reliability of an IC, incorporate multiple components packaged in a small-footprint package that can easily be placed on either side of a PC board. Linear's new µModule product line responds to an increasing need to provide systems designers with "plug and play" power solutions that allow them to get their end-products to market more quickly and significantly ease the design of the power portion of their systems.



The LTM4601, LTM4602 and LTM4603 each contain all the components required to build a 6A to 48A point-of-load (POL) regulator.

The new products in company's µModule family were announced worldwide via press meetings with key editors in locations as far ranging as Munich, Germany and Seoul, Korea. The new family of high voltage, high power DC/DC µModule point-of-load (POL) regulators provides new features and various output power capabilities. The LTM4601, LTM4602 and LTM4603 each contain all the components required to build a 6A to 48A point-of-load (POL) regulator, including the inductor, power MOSFETs, DC/DC controller, compensation circuitry and input/output bypass capacitors. The devices' compact 15mm × 15mm × 2.8mm LGA package protects the solution from the external environment and the modules' thermal enhancements provide highly efficient heat removal.

"The addition of these five new power µModules further enables quick and easy design of a range of power supplies," stated Don Paulus, General Manager of the Power Business Group. "Their light weight and low profile packaging allow the µModules to be soldered onto the back side of many circuit boards, making efficient use of board space, and leaving topside space for sophisticated digital ICs such as FPGAs and DDR memory. We anticipate a high level of interest in a broad range of applications."

For further information, visit www.linear.com/micromodule.

LT3476, continued from page 1

example, the 1A output LED current of the LT3476 can be safely shared by three parallel strings of LEDs when the LT3003 is added. Each string carries up to 350mA. The LT3003 guarantees 3% LED current matching.

Dimming ratio is defined as the ratio between the highest and the lowest achievable brightness of a system. A large dimming ratio is often

required for instantaneous setting of the backlight brightness according to the image information and environment in which the device is used. A large dimming ratio also helps reduce motion artifacts. Without adding components and cost, both the LT3476 and the LT3003 can achieve at least 1000:1 PWM dimming ratio with less than 5 μ s rise/fall time. Additional analog dimming is also possible.

Features

High Side Current Sensing for Versatility and Reliability

High side LED current sensing is generally more flexible than low side, in that it supports buck, boost or buck-boost configurations. High side sensing also allows for "one-wire" operation. For example, in a boost circuit with a high side sense resistor, if the LEDs are

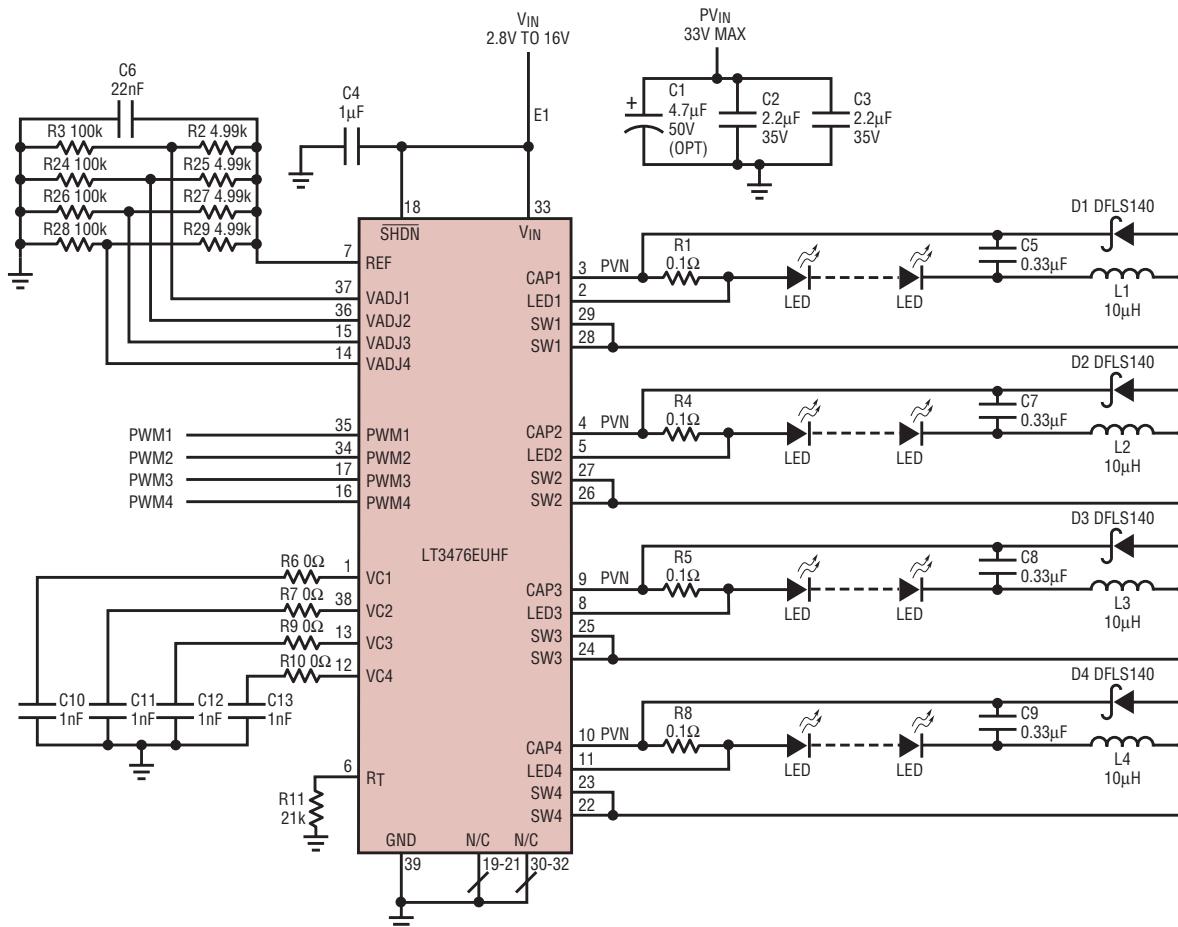


Figure 1. The LT3476 delivers 100W in buck mode

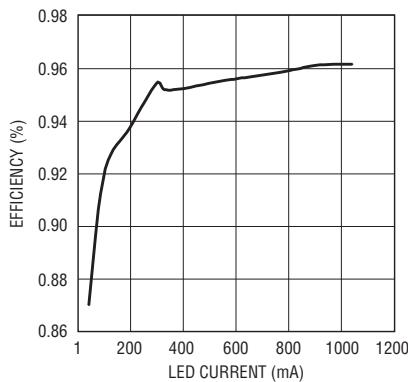


Figure 2. Efficiency of the buck mode circuit in Figure 1

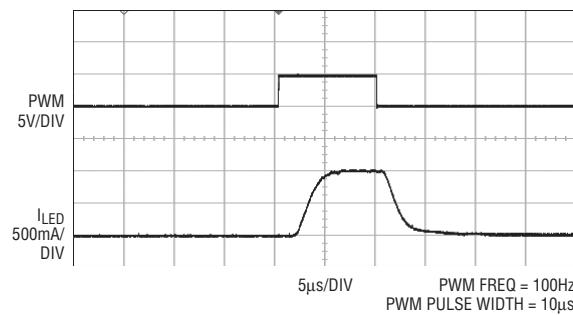


Figure 3. 1000:1 PWM dimming

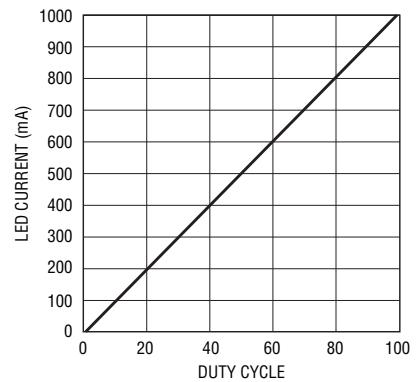


Figure 4. Average LED current vs PWM duty cycle

remote from the driver in some way, such as in a hinged laptop display, the LED current can return to the local display ground, saving a wire in the return path. Low side sensing requires an extra wire, because the LED current must return to the driver side for low noise operation. The one wire setup lowers cost and improves reliability, especially as the channels multiply in high performance displays.

Buck, Boost or Buck-Boost Operation

Because of the high side current sense scheme, the LT3476 and the LT3003 support buck, boost or buck-boost operation. In buck mode, an LT3476 circuit can achieve 96% efficiency, generating less heat and providing more reliability. For automotive applications where the LEDs must be

powered from a lead-acid battery, the LT3476 can be configured for boost mode to drive up to eight LEDs per channel. Furthermore, returning the LED current in a boost configuration to the battery enables buck-boost operation, where the input voltage can be higher or lower than the output voltage. As a result, the LT3476 and LT3003 can accept a variety of power sources.

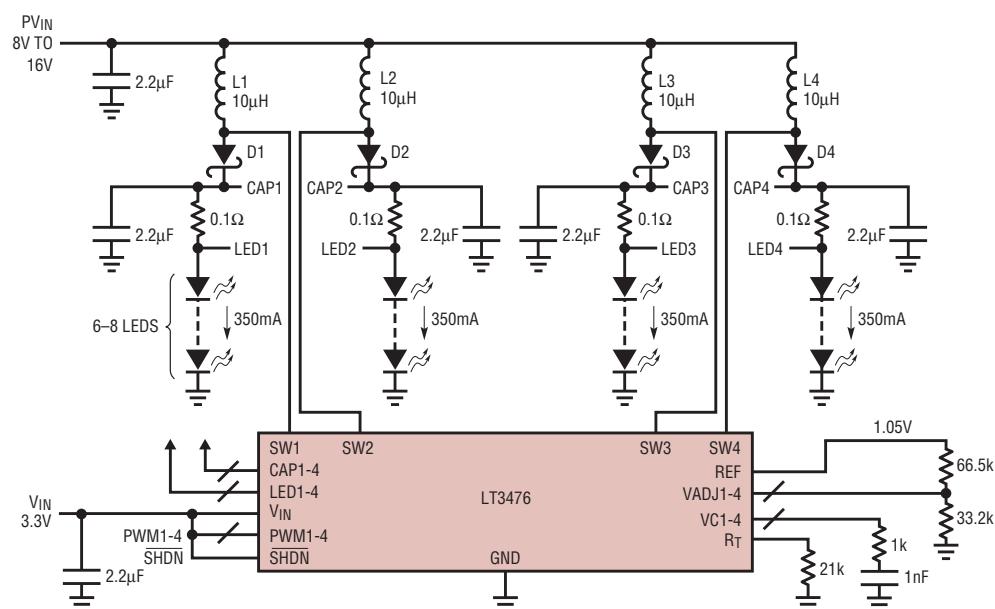


Figure 5. The LT3476 configured into a boost circuit for automotive applications

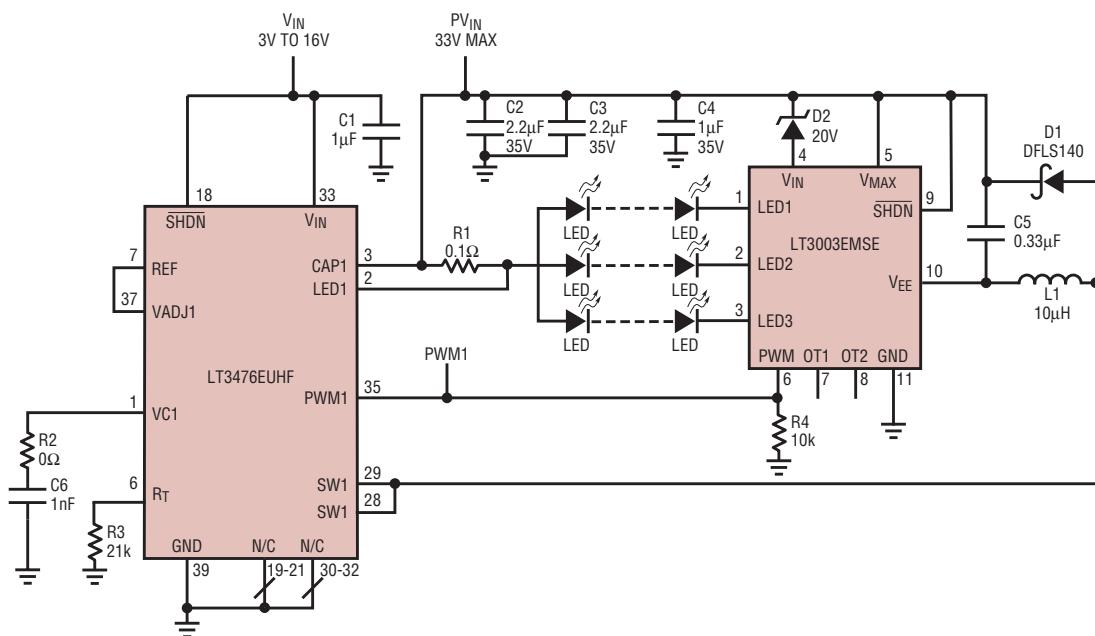


Figure 6. The LT3476 and LT3003 in buck mode

PWM and Analog Dimming

Dedicated PWM dimming circuitry inside the LT3476 and LT3003 allows a 1000:1 dimming ratio. Additional analog dimming is possible through the VADJ pins. This allows for a significant number of hues and tones, resulting in finer and more exact color definition.

Accurate Current

Monitoring and Matching

Each of the four LT3476 current monitor thresholds is trimmed to within 2.5% at the full scale of 105mV. The LT3003 drives three separate strings of LEDs at up to 350mA/string with 3% accurate current matching. Both measures result in uniform LED brightness and intensity.

Wide Range of Operating Frequencies to Match any Application

The LT3476 frequency is adjustable between 200kHz and 2MHz, allowing the user to trade off between the efficiency and the solution size.

Small Packages

The LT3476 is available in a 5mm x 7mm QFN package. The LT3003 comes in a small MS10 package. Both packages are thermally enhanced with exposed metal ground pads on the bottom of the package.

Applications

LT3476 Delivers

100W in Buck Mode

In today's large LCD TVs with LED backlights, the power requirement for driving the LEDs can be a couple hundred watts. Figure 1 shows a circuit for a high power LED driver. It is configured as a buck mode converter, delivering 100W to the LEDs from a 33V supply at 96% efficiency. Two of these circuits are enough to drive all the LEDs for a 32-inch LCD TV. For simplicity's sake only channel 1 is discussed here.

All four LT3476 channels are independent and function in the same way. When the internal power switch turns on, the SW1 pin is grounded.

The voltage crossing the inductor L1 is $P_{VIN} - V_{LED1}$, where the V_{LED1} is the voltage drop on the LED string at the given current. As a result, the inductor L1 current ramps up linearly and energy builds up. When the power switch is off, the inductor sees V_{LED1} . The energy in the inductor is discharged and transferred to the LEDs through the catch diode D1. The capacitor C5 filters out the inductor current ripple. The LED current is the average of the inductor current. Figure 2 shows the efficiency as a function of the LED current.

To change the maximum LED current, adjust the R1 value or the resistor divider values at the VADJ1 pin. The VADJ pins can be used for white balance calibration. At 100Hz PWM frequency, the PWM control of this circuit allows 1000:1 dimming as shown in Figure 3. Figure 4 shows that the PWM dimming ratio has a good linear relationship to the average LED current. Faster switch on/off time is possible if a PFET disconnect circuit with a level shifter is in series

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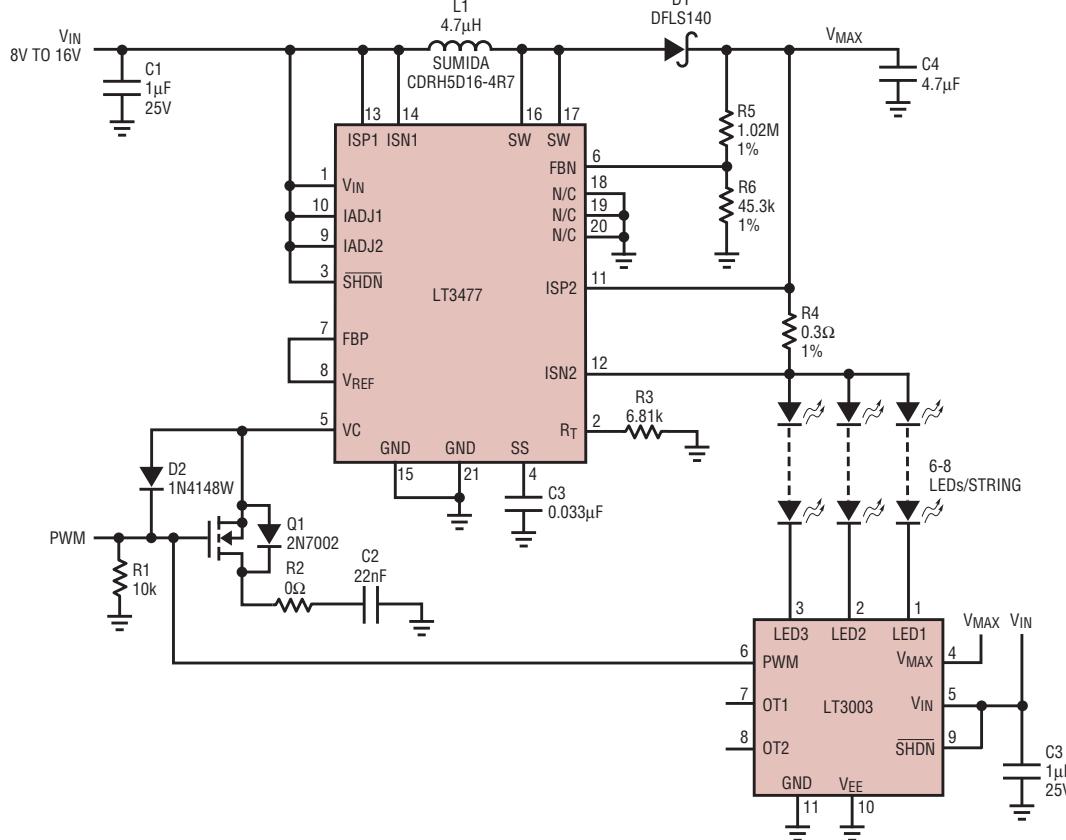


Figure 7. The LT3476 and LT3003 in boost mode

Precise Current Sense Amplifiers Operate from 4V to 60V

by Jun He

Introduction

The LTC6103 and LTC6104 are versatile, precise high side current sense amplifiers with a wide operation range. The LTC6103 is a dual current sense amplifier, while the LTC6104 is a single, bi-directional current sense amplifier—it can source or sink an output current that is proportional to a bi-directional sense voltage.

Due to the amplifiers' wide supply range (60V), fast speed (1μs response time), low offset voltage (85μV typical), low supply current (275μA/channel typical) and user-configurable gains, they can be used in precision industrial and automotive sensing applications, as well as current-overload protection circuits.

Other features include high PSRR, low input bias current and wide input sense voltage range. Both parts are available in an 8-lead MSOP.

LTC6103 Theory of Operation

Figure 1 shows a block diagram of the LTC6103 in a basic current sense circuit. A sense resistor, R_{SENSE} , is added in the load path, thereby creating a small voltage drop proportional to the load current.

An internal sense amplifier loop forces $-IN$ to have the same potential as $+IN$. Connecting an external resistor, R_{IN} , between $-IN$ and $V_{BATT,A}$ forces a potential across R_{IN} that is the same as the sense voltage across R_{SENSE} . A corresponding current

$$I_{OUT} = \frac{(I_{LOAD} + I_S)R_{SENSE}}{R_{IN}}$$

flows through R_{IN} . The high impedance inputs of the sense amplifier do not conduct this input current, so the current flows through an internal MOSFET to the OUT pin. In most application cases, $I_S \ll I_{LOAD}$, so

$$I_{OUT} \approx \frac{I_{LOAD} \cdot R_{SENSE}}{R_{IN}}$$

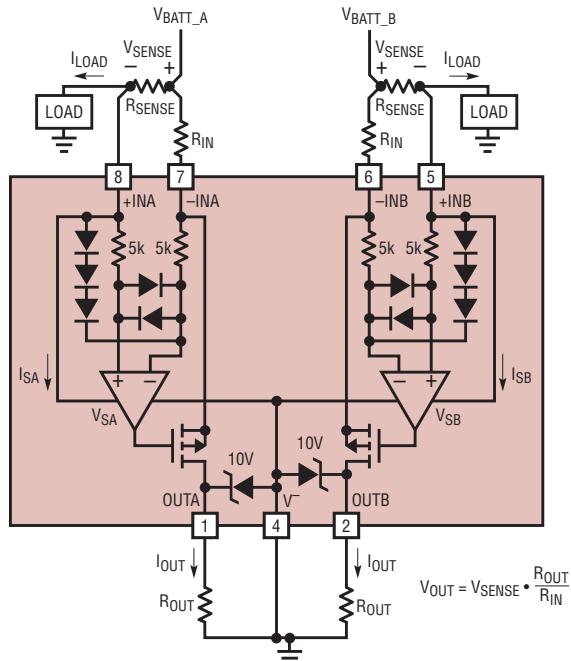


Figure 1. The LTC6103 block diagram and typical connection

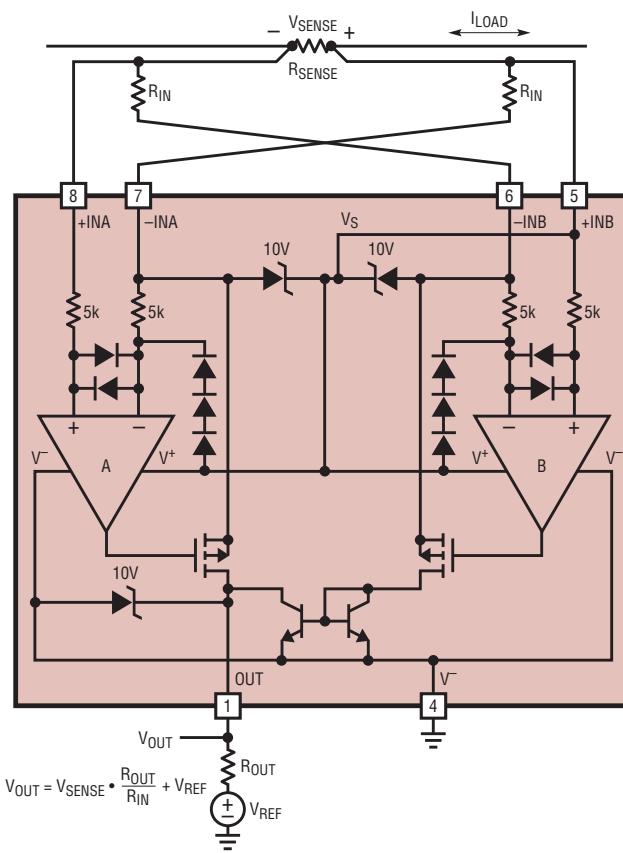


Figure 2. The LTC6104 block diagram and typical connection

The output current can be transformed into a voltage by adding a resistor from OUT to V^- . The output voltage is then

$$V_{OUT} = (V^-) + (I_{OUT} \cdot R_{OUT})$$

LTC6104 Theory of Operation

Figure 2 shows a block diagram of the LTC6104 in a basic current sense circuit.

Similar to the operation of the LTC6103, the LTC6104 can transfer a high side current signal into a ground-referenced readout signal. The difference is that the LTC6104 can sense the input signal in both polarities.

Only one amplifier is active at a time in the LTC6104. If the current direction activates the "B" amplifier, the "A" amplifier is inactive. The signal current goes into the -INB pin, through the MOSFET, and then into a current mirror. The mirror reverses the polarity of the signal so that current flows into the "OUT" pin, causing the output voltage to change polarity. The magnitude of the output is

$$V_{\text{OUT}} = \frac{V_{\text{SENSE}} \cdot R_{\text{OUT}}}{R_{\text{IN}}} + V_{\text{REF}}$$

Keep in mind that the OUT voltage cannot swing below V^- , even though it is sinking current. A proper V_{REF} and R_{OUT} need to be chosen so that the designed OUT voltage swing does not go beyond the specified voltage range of the output.

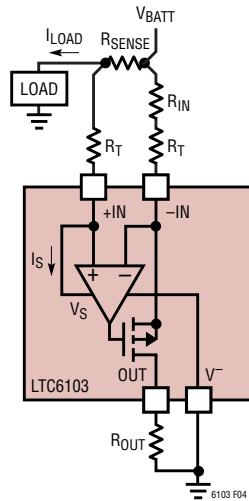


Figure 3. Error Due to PCB trace resistance

Sources of Current Sensing Error

As the output voltage is defined by

$$V_{\text{OUT}} = \frac{I_{\text{LOAD}} \cdot R_{\text{SENSE}} \cdot R_{\text{OUT}}}{R_{\text{IN}}}$$

any error of the external resistors contributes to the ultimate output error. If current flowing through the sense resistor is high, Kelvin connection of the -IN and +IN inputs to the sense resistor is necessary to avoid errors introduced by interconnection and trace resistance on the PCB.

Besides external resistors, the dominant error source is the offset voltage of the sense amplifier. Since this is a level independent error

maximizing the input sense voltage improves the dynamic range of the system. If practical, the offset voltage error can also be calibrated out.

Care should be taken when designing the printed circuit board layout. As shown in Figure 3, supply current flows through the +IN pin, which is also the positive amplifier input pin (for the LTC6104, this applies to the +INB pin only). The supply current can cause an equivalent additional input offset voltage if trace resistance between R_{SENSE} and +IN is significant. Trace resistance to the -IN terminals is added to the value of R_{IN} . In addition, the internal device resistance adds approximately 0.3Ω to R_{IN} .

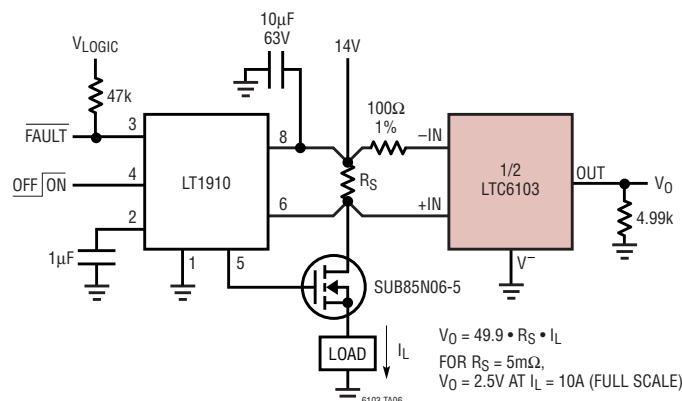


Figure 4. Automotive smart-switch with current readout

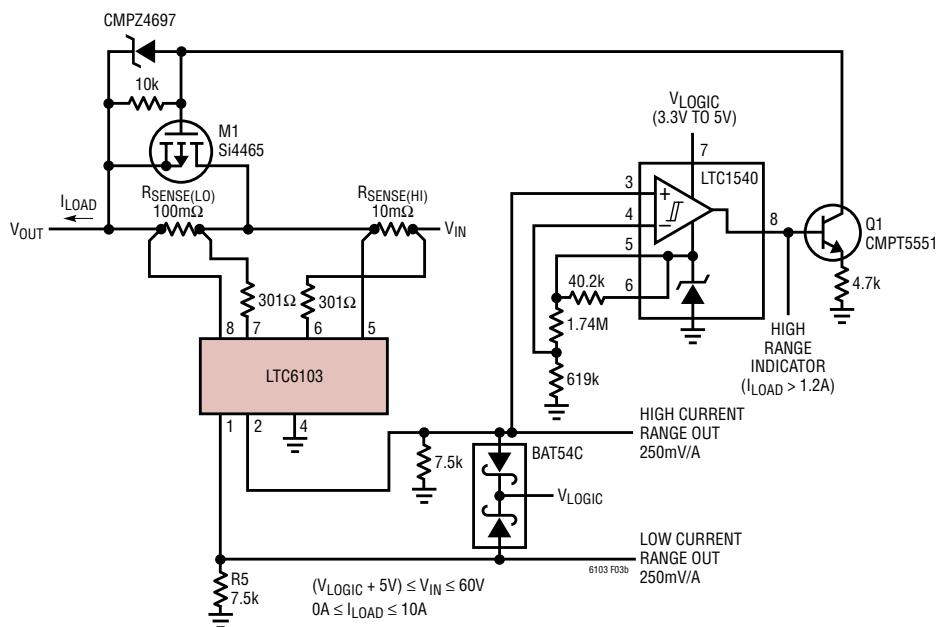


Figure 5. The LTC6103 allows high-low current ranging

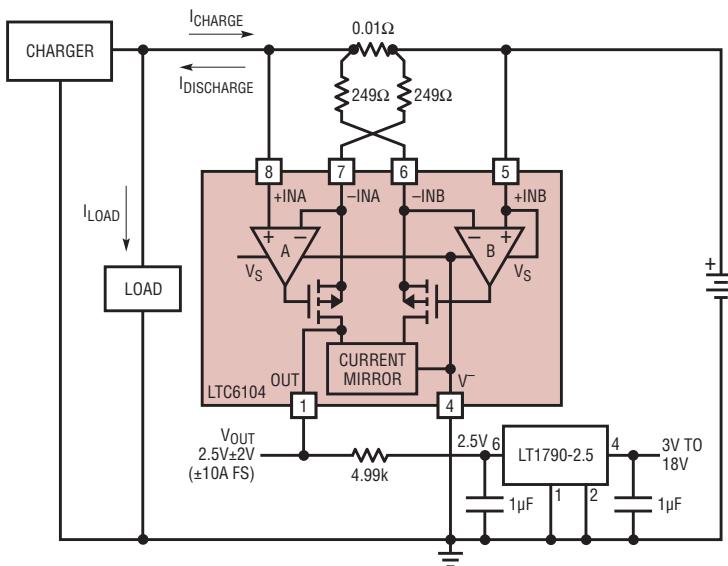


Figure 6. The LTC6104 bi-direction current sense circuit with combined charge/discharge output

Applications

The LTC6103 and LTC6104 operate from 4V to 60V, with a maximum supply voltage of 70V. This allows them to be used in applications that require high operating voltages, such as motor control and telecom supply monitoring, or where it must survive in the face of high-voltages, such as with automotive load dump conditions. The accuracy is preserved across this supply range by a high PSRR of 120dB (typical).

Fast response time makes the LTC6103 and LTC6104 the perfect choice for load current warnings and shutoff protection control. With very low supply current, they are suitable for power sensitive applications.

The gain of the LTC6103 and LTC6104 is completely controlled by external resistors, making them flexible enough to fit a wide variety of applications.

Monitor the Current of Automotive Load Switches

With its 60V input rating, the LTC6103 is ideally suited for directly monitoring currents on automotive power systems without need for additional supply conditioning or surge protection components.

Figure 4 shows an LT1910-based intelligent automotive high side switch with an LTC6103 providing an analog current indication. The LT1910 high

side switch controls an N-channel MOSFET that drives a controlled load and uses a sense resistor to provide overload detection. The sense resistor is shared by the LT6103 to provide the current measurement.

The LTC6103 supplies a current output, rather than a voltage output, in proportion to the sense resistor voltage drop. The load resistor for the LTC6103 may be located at the far end of an arbitrary length connection, thereby preserving accuracy even in the presence of ground-loop voltages.

High-Low Range Current Measurement

Figure 5 shows LTC6103 used in a multi-range configuration where a low current circuit is added to a high current circuit. A comparator (LTC1540) is used to select the range, and transistor M1 limits the voltage across $R_{SENSE(LO)}$.

Battery Charge/Discharge Current Monitor

Figure 6 shows the LTC6104 used in monitoring the charge and discharge current of a battery. The voltage reference LT1790 provides a 2.5V offset so that the output can swing above

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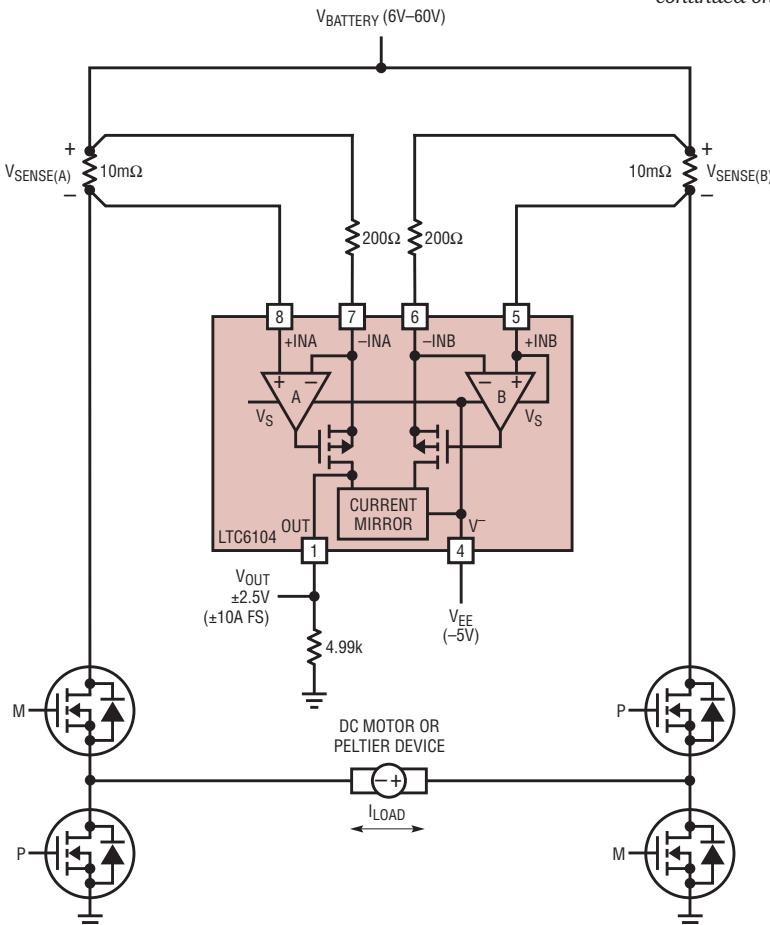


Figure 7. Current monitoring for an H-bridge application

Tiny, High Efficiency Monolithic Buck Converters are Perfect for Powering Portable Devices

by Phil Juang

Introduction

Power management for cell phones, portable media players and other battery powered handheld devices has become increasingly complex as the demand for more features and functions grows, even as the devices shrink. This trend drives an urgent need for high efficiency buck converters that both preserve battery life and take up as little board space as possible. In many cases, a monolithic DC/DC step-down regulator is the only way to meet this demand.

Linear Technology offers a complete family of synchronous, current mode, constant frequency regulators ranging in output currents from 250mA up to 8A. The LTC3410, LTC3542, LTC3547, and LTC3548 are Linear Technology's tiniest solutions for powering handheld devices, offering extremely small solution size and

Table 1. Small, low power monolithic buck converters

Part Number	Output Current	Number of Outputs	Available Packages
LTC3410	300mA	1	SC70
LTC3542	500mA	1	2mm x 2mm DFN 6-lead SOT-23
LTC3547	300mA/300mA	2	3mm x 2mm DFN
LTC3548	400mA/800mA	2	10-lead MSOP 10-lead DFN (3mmx3mm)

unmatched performance for single or dual step-down outputs requiring up to 800mA of output current.

Space-Saving Solutions Save on Battery Power

These monolithic step-down regulators save space by bringing the switching MOSFETs into the IC. They also offer

synchronous operation for high efficiency step-down regulation while eliminating the need for an external Schottky diode.

Other space saving features include:

- A high 2.25MHz operating switching frequency, which facilitates the use of small, low-profile inductors and capacitors,
- Internal compensation removes external compensation capacitors and resistors,
- Super small, low-profile packages (less than 1mm high).

To enhance light load efficiency, Linear Technology's patented power-saving Burst Mode architecture reduces unnecessary switching losses. The improved Burst Mode feature of these new products significantly reduces the output voltage ripple to only 20mV peak-to-peak when bursting. All devices draw less than 1µA during shutdown, making them perfect for use in battery-powered applications.

All of these products operate safely from supply voltages ranging from 2.5V to 5.5V, making them ideal for battery powered devices as well as applications requiring power from a USB port.

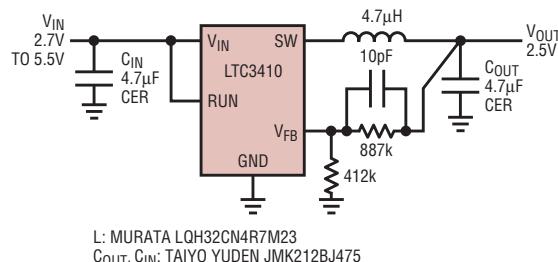


Figure 1. Tiny step-down regulator supplies 300mA with up to 96% efficiency

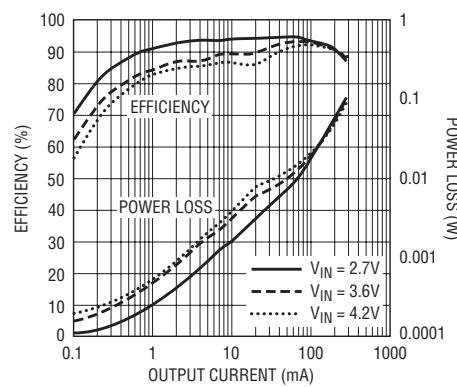


Figure 2. Efficiency and power loss of circuit in Figure 1. Burst Mode operation yields high efficiency at light loads.



Figure 3. The LTC3410 supplies 300mA without taking much space.

Fit a Complete 300mA Step-Down Regulator in 30mm² or a Dual in 100mm²

The LTC3410 is a current-mode buck converter capable of delivering 300mA of output current in a tiny, low-profile SC70 package (less than 1mm high). The converter supports output voltages as low as 0.8V. With no load on the output, the quiescent current is a mere 27µA, thereby conserving the battery power during standby. The high efficiency operation of the LTC3410 (up to 96%) ensures that little battery power is lost during normal operation. A typical application circuit is shown in Figure 1, while Figure 2 shows the efficiency and power loss graph for this circuit. This single channel DC/DC regulator takes up only 30mm² of board real estate, as shown in Figure 3.

Since handheld devices are becoming more complex and require power for several different devices, there is a growing need for multiple step-down output voltages. The LTC3547 is a synchronous buck converter with not one, but two, 300mA step-down outputs, making it the functional equivalent of two LTC3410 parts. Its

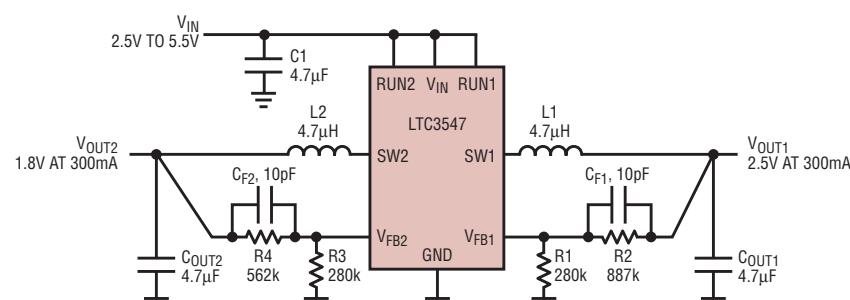


Figure 4. Dual step-down outputs with up to 96% Efficiency.

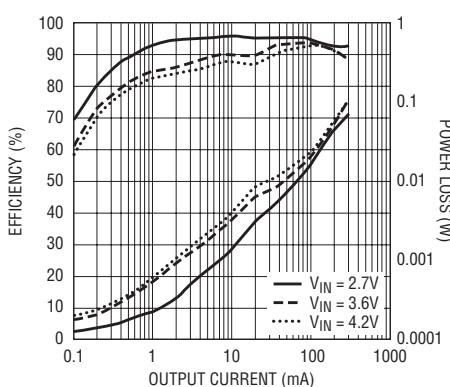


Figure 5. Efficiency and power loss of circuit in Figure 4.

tiny 3mm × 2mm DFN package makes it the smallest dual-output buck converter on the market.

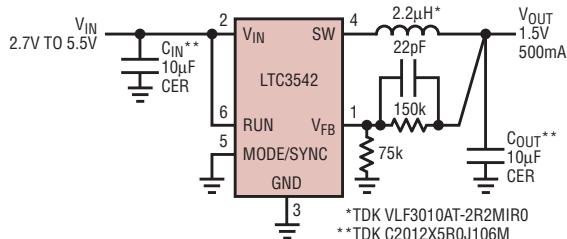


Figure 7. Single output buck converter yields 2.5V at 500mA and allows the switch frequency to be synchronized.

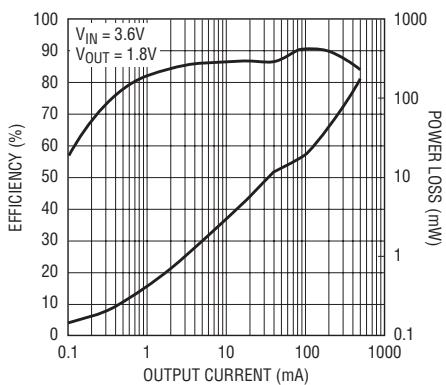


Figure 8. Efficiency and power loss of Circuit in Figure 7.

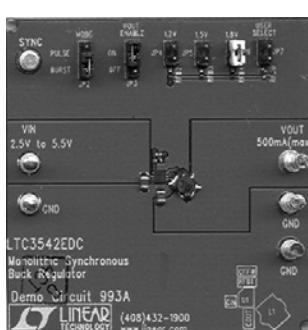


Figure 9. The LTC3542 uses few external components to generate a single 500mA supply.

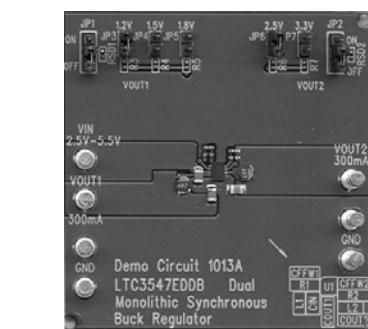


Figure 6. LTC3547 occupies only 100mm² of board space to generate two 300mA outputs.

The LTC3547 no-load quiescent current is only 40µA when both channels are enabled. Each channel can accommodate output voltages as low as 0.6V and can be enabled or disabled independently. In addition, the LTC3547 soft-start feature slowly ramps up its outputs upon start-up, reducing the initial inrush current from the supply input. The output is ramped from zero to full-scale over approximately 700µs.

Figures 4 and 5 show a typical application circuit and corresponding efficiency graph, while the photo in Figure 6 shows the LTC3547 dual buck converter circuit, which only takes up 100mm² of board space.

Linear Technology also offers LTC3410B and LTC3547B, which are identical to the LTC3410 and LTC3547, except they utilize pulse skipping mode instead of Burst Mode operation. The LTC3410B and LTC3547B are good options for noise sensitive applications which require constant-frequency operation at light loads.

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Supply Supervisor Family Accurately Monitors Multiple Voltages with Independent Undervoltage and Overvoltage Detection

Introduction

Many modern electronic systems require monitoring of the power supply levels. Some systems must know when the power supplies are present and stable before start-up. Other systems must know if the supplies deviate from safe operating conditions. Undervoltage monitoring allows a system to know when the power supplies are fully stable at start-up and prevent unreliable operation if the supply drops during normal operation. Overvoltage monitoring allows a system to know if a failure has occurred in a power supply or in a powered device causing the supply voltage to exceed a safe operating threshold. Once an undervoltage or overvoltage fault is detected, the system can then initiate housekeeping operations.

Three new power supply supervisors improve system reliability by offering accurate thresholds for both undervoltage and overvoltage monitoring. With very low part counts, any supply level can be monitored.

The LTC2914, LTC2913, and LTC2912 supervisors simultaneously monitor quad, dual, and single power supplies respectively for undervoltage and overvoltage detection with a tight 1.5% threshold accuracy over temperature. All monitors share a common undervoltage output and a common overvoltage output with a timeout period that is externally adjustable or disabled. Each monitor has input glitch rejection to ensure reliable reset operation without false or noisy triggering. Each part has two options: one with capability to latch the overvoltage output and one with capability to externally disable both outputs. The LTC2912 has a third option with latching capability on the

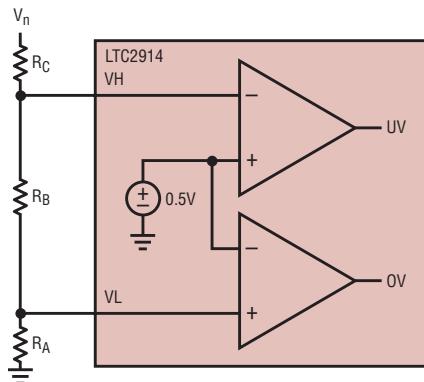


Figure 1. 3-resistor positive UV/OV monitoring configuration

overvoltage output, but with a non-inverted output. Table 1 lists the features offered for each device option.

Voltage Monitoring

Each monitored input is compared to a 0.5V threshold. When configured to monitor a positive voltage, V_n , using the 3-resistor circuit configuration shown in Figure 1, VH is connected to the high side tap of the resistive divider and VL is connected to the low side tap of the resistive divider.

The LTC2914 has polarity selection and a buffered reference allowing up to two separate negative voltages to be monitored. A three-state input pin

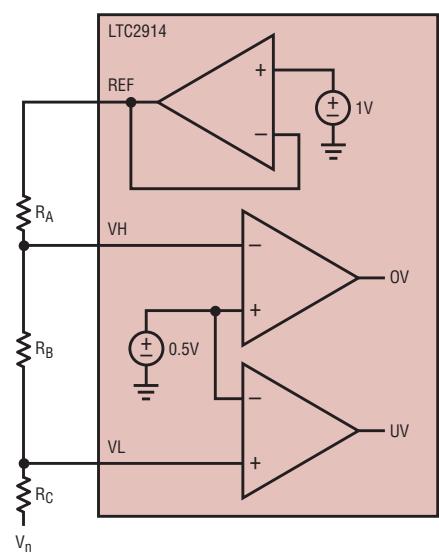


Figure 2. 3-resistor negative UV/OV monitoring configuration

selects the polarity of these two inputs without requiring any external components. If an input is configured as a negative voltage monitor, the outputs UV and OV in Figure 1 are swapped internally. The monitored voltage is then connected as shown in Figure 2. Note that VH remains connected to the high side tap of the resistive divider and VL remains connected to the low side tap.

Noise Sensitivity

In any supervisory application, noise riding on the monitored DC voltage can cause spurious faults, particularly when the monitored voltage is near the trigger threshold. A less desirable but common solution to this problem is to introduce hysteresis around the nominal threshold. However, the addition of hysteresis introduces an error term in the threshold accuracy. For example, a $\pm 1.5\%$ accurate monitor with a $\pm 1\%$ hysteresis is equivalent to a $\pm 2.5\%$ monitor with no hysteresis.

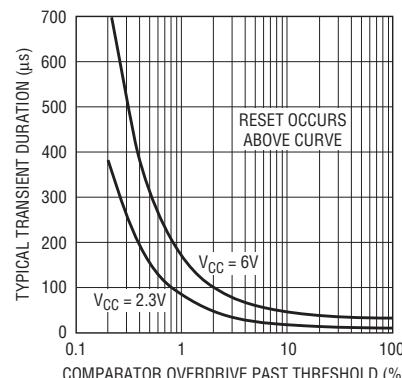


Figure 3. Transient duration vs comparator overdrive

This supervisor family solves this problem in two ways without adding hysteresis. First, each supervisor lowpass filters the output of the first stage comparator at each input. This filter integrates the output of the comparator before asserting the undervoltage or overvoltage outputs. A transient at the input of the comparator of sufficient magnitude and duration triggers the output logic. Figure 3 shows the typical transient duration versus comparator overdrive required to assert the output (overdrive shown as a percentage of the trip threshold V_{UOT}).

The second solution is the undervoltage/overvoltage timeout period t_{UOTO} . This timeout period is adjustable and holds \bar{UV} or \bar{OV} asserted after all faults have cleared. This assures a minimum output pulse width allowing a settling time delay for the monitored voltage after it has entered the valid region of operation. When any VH input drops below its threshold, the \bar{UV} pin asserts low. When all VH inputs recover above their thresholds, an undervoltage output timer starts. If all inputs are above their thresholds, the \bar{UV} pin weakly pulls high when the timer finishes. However, if any VL input rises above its threshold during this timeout period, the timer resets and restarts when all inputs again recover below their thresholds. On the LTC2912-3, the overvoltage output is not inverted and asserts high during an overvoltage fault condition.

The \bar{OV} output behaves in a similar manner. When any VL input rises above its threshold, the \bar{OV} pin asserts low. When all VL inputs recover below their thresholds, an overvoltage output timer starts. If all inputs remain

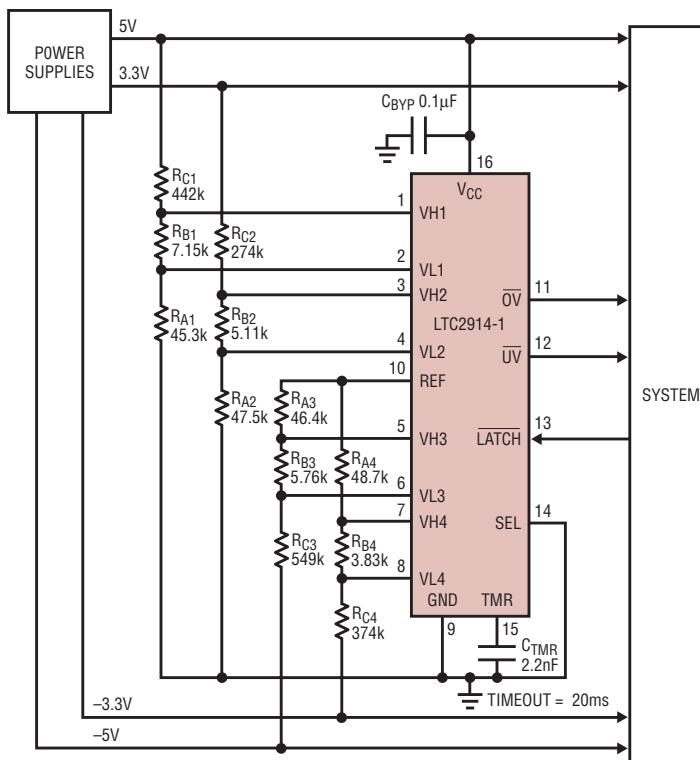


Figure 4. Dual positive and negative supply monitor

below their thresholds when the timer finishes, the \bar{OV} pin weakly pulls high. However, if any VL input rises above its threshold during this timeout period, the timer resets and restarts when all inputs again recover below their thresholds. On the LTC2912-3, the overvoltage output is not inverted and asserts high during an overvoltage fault condition.

The value of capacitor, C_{TMR} , needed for a particular timeout period, t_{UOTO} , is determined by:

$$C_{TMR} = t_{UOTO} \cdot 115 \cdot 10^{-9} \left(\frac{F}{s} \right)$$

where t_{UOTO} is the desired timeout period in seconds.

OV Latch

On each part option with latching capability, the \bar{OV} pin latches low (high for the LTC2912-3) if an overvoltage condition is detected while the LATCH pin is held low. The latch is cleared by pulling the LATCH pin high. If all overvoltage conditions clear while LATCH is held high, the latch is bypassed and the \bar{OV} pin behaves the same as the \bar{UV} pin with a similar timeout period. If LATCH is pulled low

Table 1. Supervisor family feature options

	UV Inputs	OV Inputs	1V Reference	Polarity Selection	OV Latch	Output Disable	Active High OV Output
LTC2914-1	4	4	✓	✓	✓		
LTC2914-2	4	4	✓	✓		✓	
LTC2913-1	2	2			✓		
LTC2913-2	2	2				✓	
LTC2912-1	1	1			✓		
LTC2912-2	1	1				✓	
LTC2912-3	1	1			✓		✓

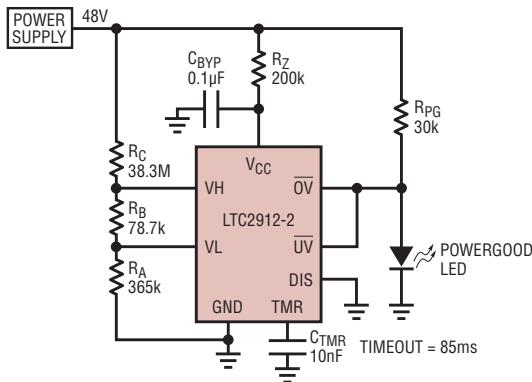


Figure 5. 48V supply powergood monitor

while the timeout period is active, the \overline{OV} pin latches as before.

Margin Disable

When margining the power supplies, part options with the margin disable function allow the \overline{UV} and \overline{OV} outputs to be disabled via the DIS pin. Pulling DIS high forces both outputs to remain weakly pulled high, regardless of any

faults that occur on the inputs. However, if an undervoltage lockout (UVLO) condition occurs, \overline{UV} asserts and pulls low while bypassing the timeout function. \overline{UV} pulls high as soon as the UVLO condition is cleared.

Shunt Regulator

Each part has an internal shunt regulator. The V_{CC} pin operates as a direct

3-Step Design Procedure

The following 3-step design procedure determines the appropriate resistances to obtain the desired undervoltage and overvoltage thresholds for the positive voltage monitoring circuit in Figure 1 and the negative voltage monitoring circuit in Figure 2.

V_n is the desired nominal operating voltage to be monitored, I_n is the desired nominal current through the resistive divider, V_{OV} is the desired overvoltage threshold, and V_{UV} is the desired undervoltage threshold.

For negative supply monitoring, to compensate for the 1V reference shown in Figure 2, 1V must be subtracted from V_n , V_{OV} , and V_{UV} before using each in the following equations.

1. Choose R_A to obtain the desired overvoltage threshold.

R_A is chosen to set the desired threshold for the overvoltage monitor.

$$R_A = \left| \frac{0.5V}{I_n} \bullet \frac{V_n}{V_{OV}} \right| \quad (1)$$

2. Choose R_B to obtain the desired undervoltage threshold.

Once R_A is known, R_B is chosen to set the desired threshold for the undervoltage monitor.

$$R_B = \left| \frac{0.5V}{I_n} \bullet \frac{V_n}{V_{UV}} \right| - R_A \quad (2)$$

3. Choose R_C to Complete the Design.

Once R_A and R_B are known, R_C is determined by:

$$R_C = \left| \frac{V_n}{I_n} \right| - R_A - R_B \quad (3)$$

If any of the variables V_n , I_n , V_{OV} , or V_{UV} change, then each step must be recalculated.

supply input for voltages up to 6V. For V_{CC} voltages higher than 6V, the V_{CC} pin operates as a shunt regulator and must have a resistance R_Z placed between it and the supply.

Dual Positive and Negative Supply Monitor Example

Consider a complex multiple supply system with +5V, +3.3V, -5V, and -3.3V supplies. Both the positive and negative 5V supplies have a +10%/-5% safe operating range. Both the positive and negative 3.3V supplies have a ±5% tolerance to maintain system specifications. The overvoltage detection on all supplies must latch in its fault condition to allow the system to perform necessary housekeeping. Each resistive divider string must have a nominal 10µA current. A 20ms timeout period is required on the outputs.

The LTC2914-1 is a good match to meet to these system requirements. This allows all four supplies to be monitored using a single device, and allows the overvoltage fault output to latch until the system is ready. Figure 4 shows the complete four supply monitoring system. The 2.2nF C_{TMR} capacitor implements a 20ms timeout at both outputs.

±5V Supply Monitoring

R_A is obtained by following Equation (1) of the "3-Step Design Procedure" (see sidebar).

For the +5V supply,

$$\begin{aligned} R_{A1} &= \left| \frac{0.5V}{I_{n1}} \bullet \frac{V_{n1}}{V_{OV1}} \right| \\ &= \left| \frac{0.5V}{10\mu A} \bullet \frac{5V}{5.5V} \right| \\ &\approx 45.3k\Omega \end{aligned}$$

For the -5V supply,

$$\begin{aligned} R_{A3} &= \left| \frac{0.5V}{I_{n3}} \bullet \frac{V_{n3} - 1V}{V_{OV3} - 1V} \right| \\ &= \left| \frac{0.5V}{10\mu A} \bullet \frac{-5V - 1V}{-5.5V - 1V} \right| \\ &\approx 46.4k\Omega \end{aligned}$$

R_B is obtained by following Equation (2) of the "3-Step Design Procedure."

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Improve that Mobile Phone Camera: Replace the Anemic LED Flash with a Xenon Flashlamp and a Tiny Photoflash Capacitor Charger

by Wei Gu

Introduction

Xenon flashlamps and LEDs are two practical choices for compact camera flash lighting. In general, a flashlamp makes a better flash. Its light output can be hundreds of times greater than that of an LED, and its spectral quality is well suited to photography. LEDs typically take less space, which makes them popular for mobile phones, PDAs and other compact applications. If a flashlamp system could be shrunk to a small enough size, then it would be possible to significantly improve the performance of the cameras in mobile phones and other compact products.

The LT3585 is an integrated photoflash capacitor charger that makes it possible to fit a flashlamp into a mobile phone. Its IGBT driver has two output pins, offering individual speed control of the turn-on and turn-off of the IGBT. Four LT3585 versions, each with different primary current limits, offer the flexibility to trade-off between input current and charge time. The LT3585-0 has a primary current limit of 1.2A, whereas the LT3485-3, LT3485-2, and LT3485-1 have current limits of 1.7A, 0.85A and 0.55A respectively. Additionally, input current can be further lowered by adjusting the voltage on the CHRG/IADJ pin to extend battery life.

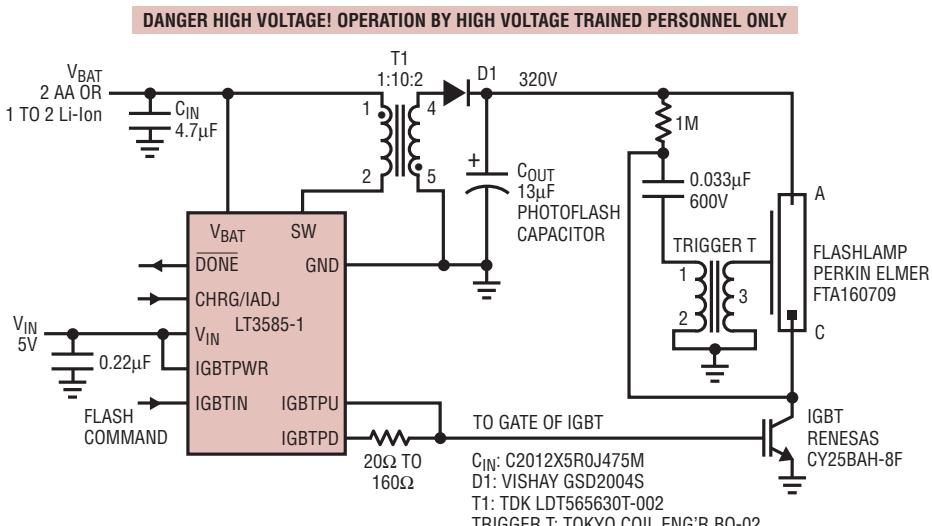


Figure 1. Complete photoflash capacitor charging circuit

If a flashlamp system could be shrunk enough, it would be possible to significantly improve the performance of the cameras in mobile phones and other compact products. The LT3585 does just that.

Figure 1 shows a typical LT3585 photoflash capacitor charging circuit which requires very few external components. Each part contains an

on-chip high voltage NPN power switch. Output voltage detection is completely contained within the part, eliminating the need for any discrete zener diodes or resistors. The output voltage can be adjusted by simply changing the turns ratio of the transformer. When operated from a 4V power source, this circuit charges a 13μF capacitor to 320V in 1.08 seconds as shown in Figure 2.

The capacitor charging circuit uses the flyback topology operating in boundary conduction mode. A low-to-high transition on the CHRG/IADJ enables the part and the switching starts. The internal power switch turns on and the current in the transformer's primary begins ramping up until it reaches the current limit. The power switch then turns off and all the energy stored in the transformer is delivered to the output cap. As the secondary current decreases to zero, the voltage on the SW pin drops to V_{BAT} or lower. When the SW pin voltage is a DCM comparator trip voltage above V_{BAT} , the part commands the power switch

Table 1. Feature Comparison of the LT3585 and LT3485

	LT3585	LT3485
Minimum V_{BAT} (V)	1.5	1.8
Two Pins for IGBT Driver	LT	
Adjustable Input Current	LT	
V_{OUT} Monitor		LT
Package	3mm × 2mm DFN 10L	3mm × 3mm DFN 10L

to turn on again. This cycle continues, delivering power to the output.

Output voltage detection is accomplished through the primary side sensing. When the reflected output voltage reaches the V_{OUT} comparator threshold voltage, the power delivery halts and the DONE pin is pulled low. As shown in Figure 1, a flash command pulse on IGBTIN drives the IGBT gate through the use of internal drive circuitry. When the IGBT is turned on, the trigger transformer generates several kilovolts along the glass envelope of the Xenon bulb to ionize the gas and form a low impedance path within the bulb. The energy stored in the photoflash capacitor quickly flows through the bulb, producing intense light. Low level flash events, such as red-eye reduction, are facilitated by multiple short duration flash input commands.

Flexible IGBT Driver Circuit

Special attention needs to be paid to the turn-on and turn-off durations applied to the gate of the IGBT. The turn-off speed is critical to the safe operation of the IGBT. The IGBT gate is a network of resistors and capacitors. When the gate terminal is pulled low, the capacitance closest to the terminal goes low but the capacitance further from the terminal remains high. This causes a smaller portion of the IGBT to handle a larger portion of the current, which can damage the IGBT. The pull-down circuitry needs to pull down slower than the internal RC time constant in the gate of the IGBT. For example, the datasheet of Renesas's IGBT CY25BAH-8F states that peak reverse gate current during turn-off must not exceed 25mA. This is easily accomplished with a resistor placed in series with the driver output pin. However, this resistor slows down the rise time, and the trigger circuitry might not have a fast enough edge to create the required 4kV pulse along the glass envelope of the Xenon bulb.

The LT3585 solves this problem by providing two output pins for the IGBT driver as shown in Figure 1. The IGBTUP pin is used to pull the gate

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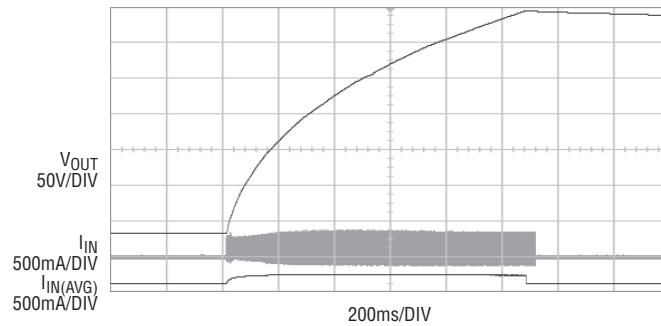


Figure 2. Output voltage and input current waveforms

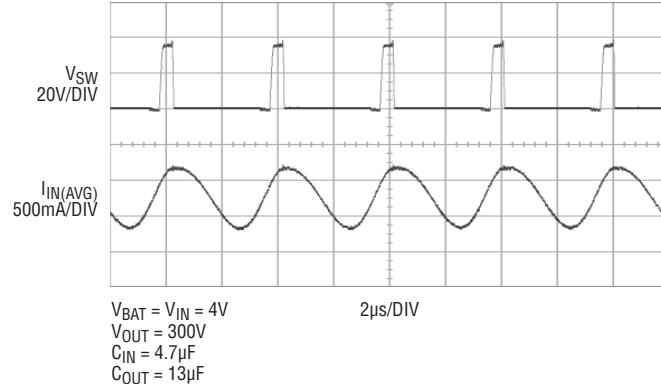


Figure 3. Input current ripple in normal input current mode.

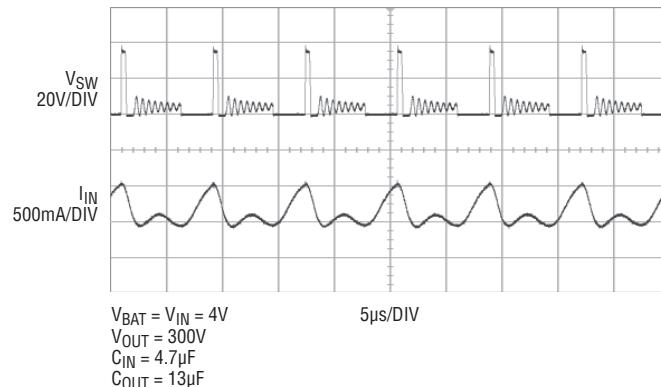


Figure 4. Input current ripple in reduced input current mode.

Table 2. Input current from battery
($V_{OUT} = 300V$, $V_{IN} = 4.7\mu F$, $C_{OUT} = 13\mu F$, $V_{BAT} = V_{IN} = 4V$)

C_{IN}	Normal Mode		Reduced Mode	
	Peak (mA)	Avg (mA)	Peak (mA)	Avg (mA)
TDK, 4.7μF, 0805 C2012X5R0J475M	655	246	545	121
TDK, 10μF, 0805 C2012X5R0J106M	640	242	515	117
TDK, 22μF, 0805 C2012X5R0J226M	515	241	390	118
TDK, 22μF, 1206 C3216X5R0J226M	420	242	310	119

High Performance, Feature-Rich Solutions for High Voltage DC/DC Converters

by Kevin Huang

Introduction

In distributed power systems, efficient DC/DC conversion circuits that can handle high voltages at substantial load currents are increasingly necessary. The LT3845 and LT3844 DC/DC controllers offer simple and feature-packed solutions that meet these requirements. The LT3845 is a synchronous controller mainly targeting step down applications, while the LT3844 is a single-switch controller that can be used in step-down, step-up, inverting and SEPIC converter topologies. Both controllers are capable of offering high efficiencies over a wide input voltage range (4V–60V) and a wide range of load currents.

LT3844 and LT3845 Features

The LT3845 and LT3844 use current-mode architectures with a user programmable 100kHz to 500kHz switching frequency. The operating frequency can also be synchronized to an external clock for noise-sensitive applications. An internal high voltage bias regulator allows for simple start-up and biasing. To increase supply

efficiency and lower power dissipation in the IC, it can be back driven by the output.

Selectable Burst Mode operation and a reverse inductor current inhibit feature (LT3845) maximize efficiencies during light-load and no-load conditions, making these controllers ideal for use in applications with supply maintenance requirements. A precision shutdown pin threshold allows for easy integration of input supply undervoltage lockout (UVLO) using a simple resistor divider. Quiescent currents are reduced to less than 10 μ A while the IC is in shutdown.

The LT3844 and LT3845 employ continuous high side inductor current sensing using an external sense resistor. If the inductor current exceeds the maximum current sense threshold, pulse skipping occurs. The current limit is unaffected by duty cycle. Both controllers incorporate a soft-start that controls the slew rate of the converter output voltage during start-up to reduce supply inrush currents and output voltage overshoot.

The gate drivers of the LT3844 and LT3845 are capable of driving large, low $R_{DS(ON)}$, standard level, N-channel MOSFETs without the need for a gate drive buffer. The driver of the LT3844 uses a bootstrapped supply rail which allows it to drive either a high side MOSFET, as found in buck converters, or a low side MOSFET, as found in boost converters. The synchronous controller, LT3845, also employs a bootstrapped supply rail for the main switch MOSFET driver.

Onboard Regulator

The LT3844 and LT3845 eliminate the need for an external regulator or a slow-charge hysteretic start scheme through the integration of an 8V linear regulator. This regulator generates V_{CC} , the local supply that runs the IC, from the converter input V_{IN} .

The onboard regulator can operate the IC continuously, provided the input voltage and/or FET gate charge currents are low enough to avoid excessive power dissipation in the part. Common practice uses the onboard regulator during start-up and then back drives

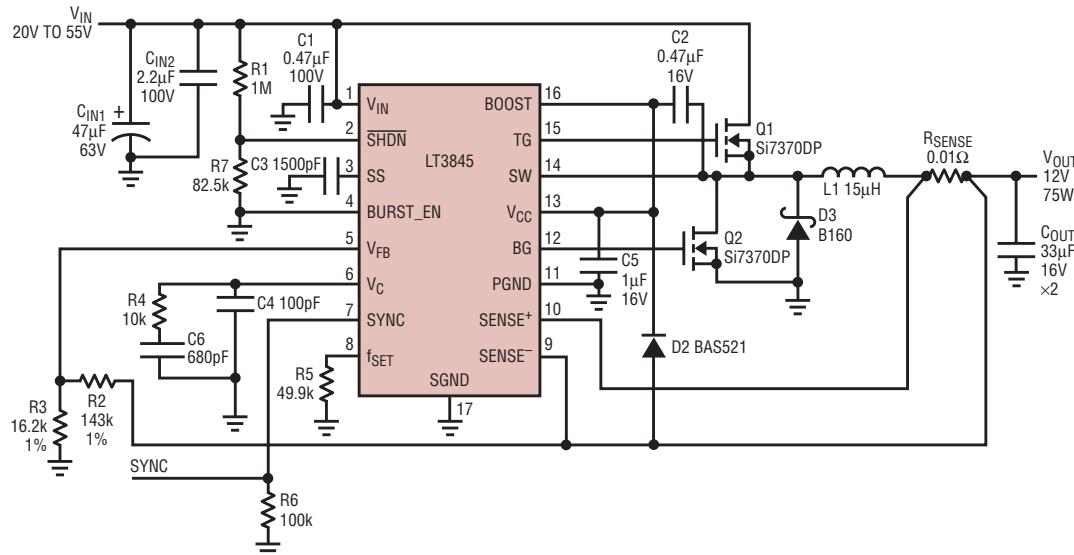


Figure 1. A 20V–55V to 12V, 75W DC/DC converter using the LT3845, featuring Burst Mode operation, reverse current inhibit and input undervoltage lockout

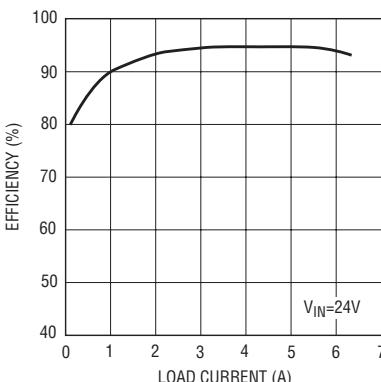


Figure 2. Efficiency of the converter in Figure 1

the V_{CC} pin above its 8V regulated voltage during operation. This reduces the power dissipation in the IC and increases converter efficiency.

The LT3844 and LT3845 have a start-up requirement of V_{IN} of about 7.5V. This ensures that the onboard regulator has ample headroom to bring the V_{CC} pin above its UVLO threshold. If V_{CC} is maintained using an external source, such as the converter output, these controllers can continue to operate with V_{IN} as low as 4V.

Burst Mode Operation

Both the LT3844 and LT3845 support low current Burst Mode operation to maximize efficiency during light load conditions. Burst Mode® operation is

enabled by shorting the BURST_EN pin to SGND, and can be disabled by shorting BURST_EN to either V_{FB} or V_{CC}.

When the peak switch current is below 15% of the programmed current limit, Burst Mode function is engaged. During the Burst interval, switching ceases and all internal IC functions are disabled, which reduces V_{IN} pin current to 20µA and reduces V_{CC} current to 100µA. If no external drive is provided for V_{CC}, all V_{CC} bias currents originate from the V_{IN} pin, giving a total V_{IN} current of 120µA. An internal negative-excursion clamp on the V_C pin is set 100mV below the switch disable threshold, which limits the negative excursion of the pin voltage during the Burst interval. This clamp minimizes converter output ripple during Burst Mode operation.

Reverse Current Inhibit of the LT3845

In addition to Burst Mode operation, the LT3845 offers a reverse-current inhibit feature, which also works to maximize efficiency during light load conditions. This mode of operation prevents negative inductor current, and is sometimes called "pulse-skipping" mode. This feature is always

enabled with Burst Mode operation when the BURST_EN pin is connected to ground. The reverse-current inhibit feature can also be enabled without Burst Mode operation by connecting the BURST_EN pin to the V_{FB} pin.

When reverse-current inhibit is enabled, the LT3845 sense amplifier detects inductor currents approaching zero and disables the synchronous switch for the remainder of that switch cycle, simulating the light-load switching characteristics of a non-synchronous converter. Reverse-current inhibit reduces losses associated with inductor ripple currents, improving conversion efficiencies with loads that are less than half of the peak inductor ripple current.

Precision Shutdown Threshold

Both the LT3844 and LT3845 have a precision-threshold shutdown feature, which allows use of the SHDN pin for analog monitoring applications, as well as logic-level controlled applications.

Input supply voltage undervoltage lockout for sequencing or start-up over-current protection is easily achieved by driving the SHDN pin with a resistor divider from the V_{IN} supply. The resistor divider is set such that the SHDN pin sees 1.35V when V_{IN} is

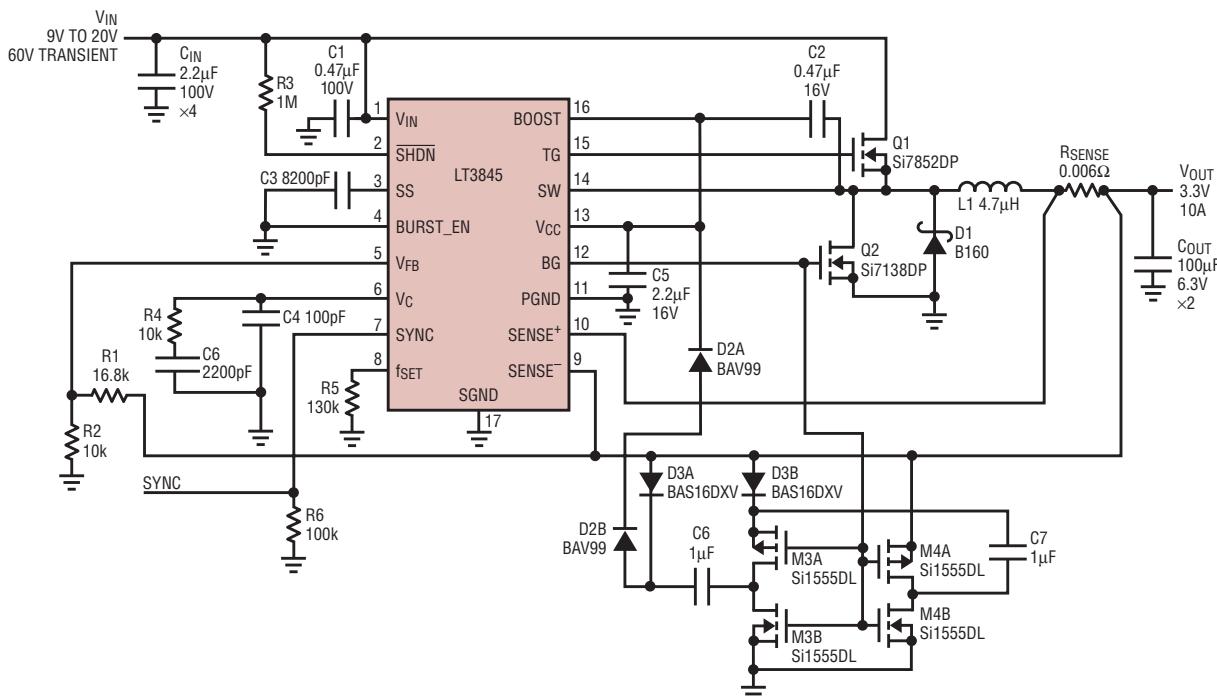


Figure 3. This 9V-20V to 3.3V/10A DC/DC converter using the LT3845 is capable of withstanding 60V transients.

at the desired UVLO rising threshold voltage. The **SHDN** pin has 120mV of input hysteresis, which allows the IC to resist almost 10% of input supply droop before disabling the converter. The **SHDN** pin has a secondary threshold of 0.7V, below which the IC operates in an ultralow current shutdown mode, reducing the supply current to less than 10 μ A. The shutdown function can be disabled by connecting the **SHDN** pin to V_{IN} through a large value pull-up resistor.

Continuous High Side Inductor Current Sensing

The LT3844 and LT3845 use a wide common mode input range current sense amplifier that operates from 0V to 36V. This current sense amplifier provides continuous inductor current sensing via an external sense resistor. A continuous inductor current sensing scheme does not require blanking intervals or a minimum on-time to monitor current, an advantage over schemes that sense switch current.

The sense amplifier monitors inductor current independent of switch state, so the main switch is not enabled unless the inductor current is below what corresponds to the V_C pin voltage. This turn-on decision is performed at the start of each cycle, and individual switch cycles will be skipped should an over-current condition occur. This eliminates many of the potential over-current dangers caused by minimum

on-time requirements, such as those that can occur during start-up, short-circuit, or abrupt input transients.

Soft Start

Both controllers employ a soft-start scheme that controls the slew rate of the DC/DC converter output voltage during start-up. A controlled output voltage ramp minimizes output voltage overshoot, reduces inrush current from the V_{IN} supply, and facilitates supply sequencing.

A capacitor, C_{SS} , connected from the C_{SS} pin to SGND, programs the slew rate. The capacitor is charged from an internal 2 μ A current source producing a ramped voltage. The capacitor voltage overrides the internal reference to the error amplifier. The soft-start circuit is disabled once the C_{SS} pin voltage has been charged to 200mV above the internal reference of 1.231V.

In normal operation, the C_{SS} pin voltage is clamped to a diode drop above the V_{FB} pin voltage. During a V_{IN} undervoltage lockout, V_{CC} undervoltage lockout or **SHDN** undervoltage lockout event, the C_{SS} pin voltage is discharged with a 50 μ A current source to retrigger a soft start. The soft-start circuit also takes control of the output voltage slew rate once the V_{FB} pin voltage has exceeded the slowly ramping C_{SS} pin voltage, reducing the output voltage overshoot during a short circuit recovery.

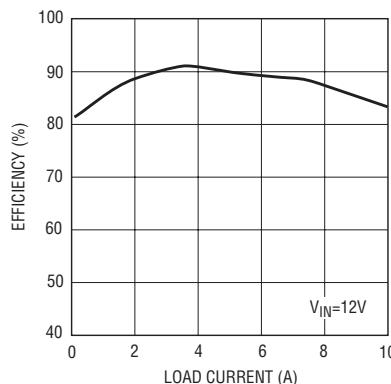


Figure 4. Efficiency of the circuit in Figure 3

20V–55V to 12V, 75W DC/DC Converter with the LT3845

Figure 1 shows a 20V–55V to 12V, 75W converter configured for Burst Mode operation, reverse current inhibit and input undervoltage lockout. Power for the IC is obtained directly from V_{IN} through the LT3845's internal V_{CC} regulator at start-up.

When the converter output comes up, D2 pulls V_{CC} above regulation, disabling the internal regulator and providing a current path from the converter output to the V_{CC} pin. Using output-generated power in high input voltage converters results in significant reduction of IC power dissipation and increases overall conversion efficiency. The **BURST_EN** pin is tied to the ground to enable Burst Mode operation and reverse current inhibit operation to achieve high efficiency at light load. Figure 2 shows the conversion efficiency for this DC/DC converter.

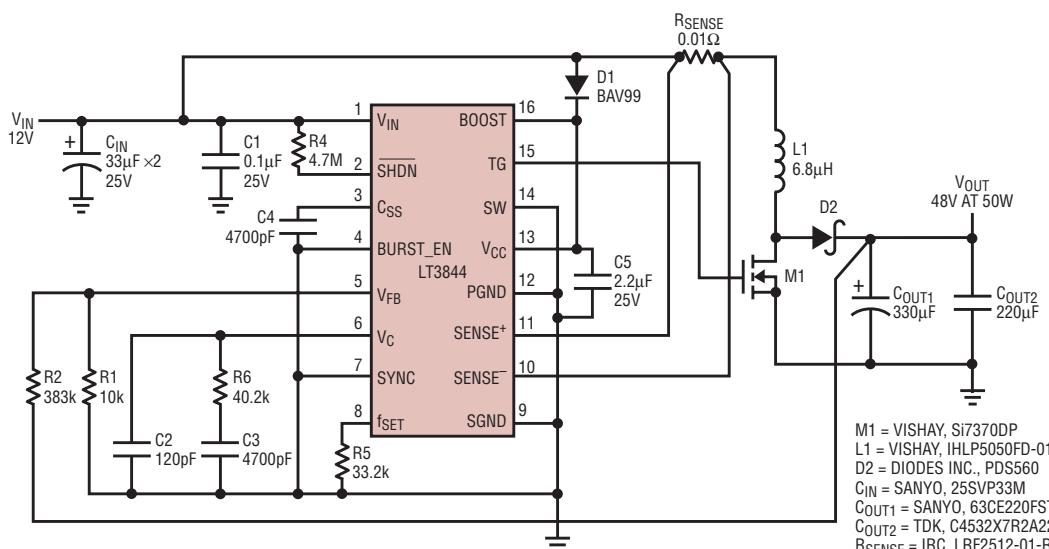


Figure 5. A 9V–16V to 48V, 50W boost converter using the LT3844, featuring Burst Mode operation

V_{IN} UVLO is programmed via a resistor divider to enable the LT3845 at 90% of the specified low end of the V_{IN} range, or 18V, which corresponds to 1.35V on the SHDN pin. The SHDN input has 120mV of hysteresis, so the converter is disabled if V_{IN} drops below 16V.

9V–20V to 3.3V, 10A DC/DC Converter with 60V Transient

In LT3845 and LT3844 converter applications with output voltages in the 9V to 20V range, back-feeding V_{CC} from the converter output is accomplished by connecting a diode from the output to the V_{CC} pins. Outputs lower than 9V require step-up techniques to generate back-feed voltages greater than the V_{CC} regulated output.

The 9V–20V to 3.3V 10A DC/DC converter shown in Figure 3 uses two Si1555DLs (M3, M4) to create a charge pump tripler that steps up the output voltage. This simple tripler uses the synchronous gate drive (BG pin) as a control signal. In typical automotive battery-voltage applications, high volt-

age line transients, such as during a load-dump condition, must be accommodated. The converter can operate through intermittent high-voltage excursions up to 60V. The switching frequency can be synchronized to an external clock from 150kHz to 250kHz. Figure 4 shows the conversion efficiency at 200kHz switching.

9V–16V to 48V, 50W Boost Converter with the LT3844

Figure 5 shows a 9V–16V to 48V, 50W boost converter with LT3844. LT3844 is a single switch controller that can be used for various topologies, and this application shows the versatility of the LT3844 by configuring it to control a battery powered boost converter application. Because the typical line voltage is moderate, the LT3844 can operate directly from the internal V_{CC} regulator without excessive power dissipation. This converter design is programmed to operate at a 400kHz switching frequency. Figure 6 shows the converter efficiency versus load current.

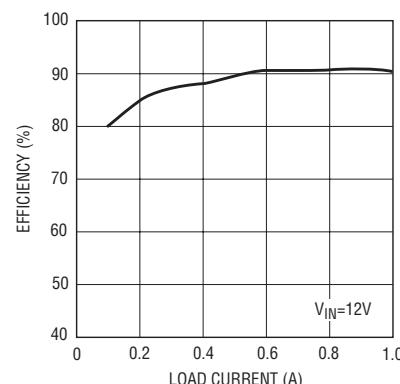


Figure 6. Efficiency of the converter in Figure 5

Conclusion

The LT3844 and LT3845 are feature packed controllers that produce high voltage DC/DC converter solutions with few external components and high efficiencies over wide load ranges. The integrated start-up regulator facilitates true single-supply operation and Burst Mode operation improves efficiency at light loads. The programmable operating frequency and synchronization functions offer extra flexibility in the DC/DC converter designs. □

LTC3585, continued from page 15
of the IGBT up. This should be done quickly to guarantee proper Xenon flashlamp ignition. The IGBTPD pin is pinned out separately to allow for greater flexibility in choosing a series resistor between the pin and the gate of the IGBT. This resistor is used to slow down the turn off of the IGBT for safe operation without affecting the pull-up transition time.

The LT3585 is similar to the LT3485, but its two drive pins give it more flexibility in terms of IGBT control. Table 1 shows the major functional differences between these two parts.

Adjustable Input Current Led to Longer Battery Life

Lithium-ion batteries are commonly used in cameras and mobile phones because of their high energy density. Aging of the battery leads to an increase in the internal resistance caused by oxidation. An aged battery may not be able to deliver the stored energy due to this increased cell resistance—even

when fully charged. For this reason, reducing the load current can keep a battery alive whereas the IR drop caused by a larger load current would cause the battery output voltage to drop below minimum operating levels.

By applying a voltage greater than 1.1V to the CHRG/IADJ pin and then floating the pin, the input current of an LT3585 circuit can be lowered by approximately 50%. When the CHRG/IADJ pin is floated, an internal circuitry drives the voltage on the pin to 1.28V. This allows a single I/O port pin, which can be three-stated, to enable or disable the part as well as place the part into the input current reduction mode. In the reduced input current mode, a time delay is added before the power switch is turned on, effectively reducing the switching frequency. Since the energy delivered to the output is still the same in each switching cycle, the input power decreases with the switching frequency. Thus, when the input voltage remains

the same, the average input current decreases.

Figures 3 and 4 show the input current waveform in normal and reduced input current mode (before C_{IN}). Table 2 shows the peak and average current with different input capacitors in the circuit shown in Figure 1. The lower peak current also extends the battery life, but the size and capacitance of the input capacitor are constrained by the board space.

Conclusion

The LT3585 provides a space-saving, simple and efficient capacitor charging solution that saves battery life, design time and cost. The two output pins of the IGBT drive offer individual control of the IGBT. The four current limit options in the LT3585 family allow for a balance between input current and charge time. Additionally, the input current can be further lowered by adjusting the voltage on the CHRG/IADJ pin. □

OLED Driver Has Low Ripple, Small Footprint and Output Disconnect

Introduction

OLED displays are increasingly popular in cellular phones, PDAs, MP3 players, portable games and a host of other hand-held devices. The OLED driver for these products must be small and efficient to fit in the tight board spaces and preserve battery life. The LT3494 delivers both, along with low output noise, a true output disconnect and full dimming control.

The LT3494 is a monolithic converter featuring an integrated high performance NPN power switch, Schottky diode, feedback resistor, and output disconnect circuitry in a tiny 8-lead 3mm × 2mm DFN package. The LT3494 has a typical switch current limit of 180mA. Its quiescent current is a low 65µA, which is further reduced to under 1µA in shutdown. The LT3494 is optimized for driving Passive Matrix OLED displays. Figure 1 shows a typical schematic for the LT3494.

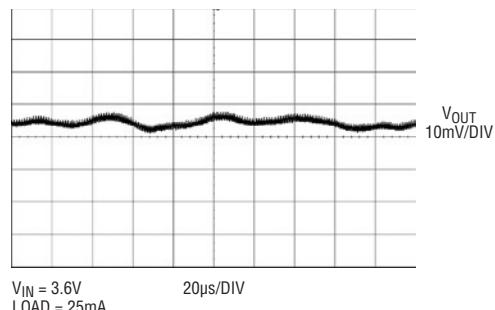


Figure 3. Output ripple for the LT3494 boost converter

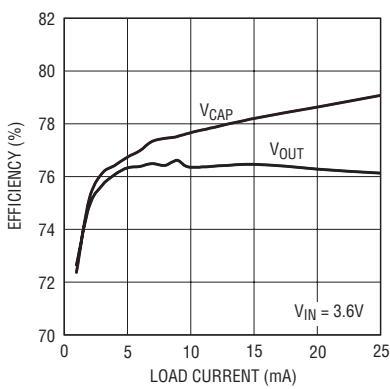


Figure 4. Efficiency for the LT3494 boost converter

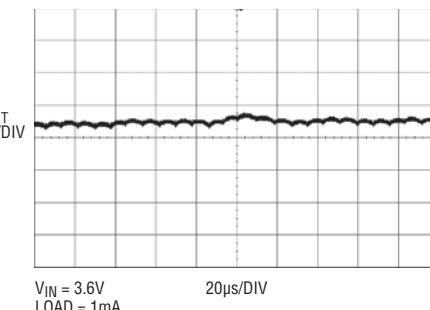


Figure 3. Output ripple for the LT3494 boost converter

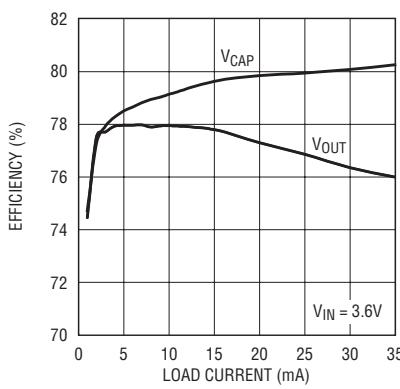


Figure 5. Efficiency for the LT3494A boost converter

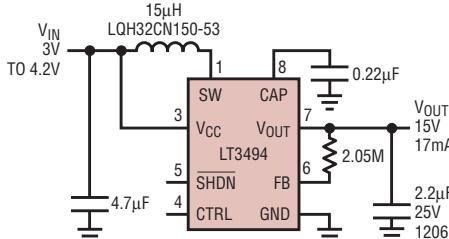


Figure 1. Single-cell Li-Ion input boost converter to 15V at 17mA

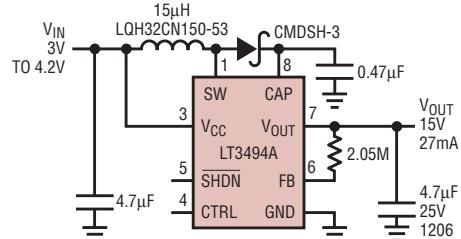


Figure 2. Single-cell Li-Ion input boost converter to 15V at 27mA

Low Output Ripple

The LT3494 converter uses a novel control technique that keeps output ripple low and the switching frequency non-audible over the entire load range, a feature highly desired when driving OLEDs. Typically, OLED drivers use a DC/DC converter that runs at a fixed frequency, which gives poor efficiency at light loads, or they run in discontinuous or pulse skipping mode to improve light load efficiency,

but these modes of operation produce higher ripple, and possibly audible noise. The LT3494 instead adjusts the frequency depending on the load. As the load decreases, the switching frequency pulls back to a minimum of around 50kHz—well above the audible noise spectrum. Figure 3 shows output ripple for the LT3494 at two load conditions.

Output Disconnect

In a standard boost regulator, the inductor and Schottky diode provide a DC current path from the input to the output, even when the regulator is not switching. Any load at the output when the chip is shut down can continue to drain the input source. This is addressed in the LT3494 by using a PMOS switch that eliminates the DC pass between input and output in shut down. This feature prevents any leakage current from the battery to OLEDs when the converter is shut down.

Integrated Solution Reduces Footprint Size and Component Count

As shown in Figure 1, only three capacitors, one resistor and one inductor are used in a typical LT3494-based converter. That is due to the integration of the switch, Schottky diode, bottom-side feedback resistor and output disconnect circuitry. In situations where efficiency is more important

continued on page 24

Hot Swap Controller Controls Power to Two PCI Express Slots

by CY Lai

Introduction

PCI Express is a new I/O technology that has been developed for desktop, mobile, server and communications platforms to increase system performance. PCI Express is rapidly replacing the older PCI standard in high availability systems such as those in the telecom, air-traffic control, and real-time-transaction processing.

In these systems, zero down time is paramount, so hardware exchanges during upgrades and maintenance must be performed on a powered system. Ideally, live hardware insertion and removal, or hot swapping, does not disturb the data and power buses of the system.

The PCI Express power supply bus consists of two main power rails, at 12V

and 3.3V and an auxiliary power rail at 3.3V—all available to add-in cards via slots on the system board (see Table 1). The LTC4242 Hot Swap controller enables hot plug functionality and fault isolation on the power bus for two PCI Express slots. To save space, the LTC4242 incorporates two low on-resistance, current-limited, robust on-board power FETs for the 3.3V_{AUX}

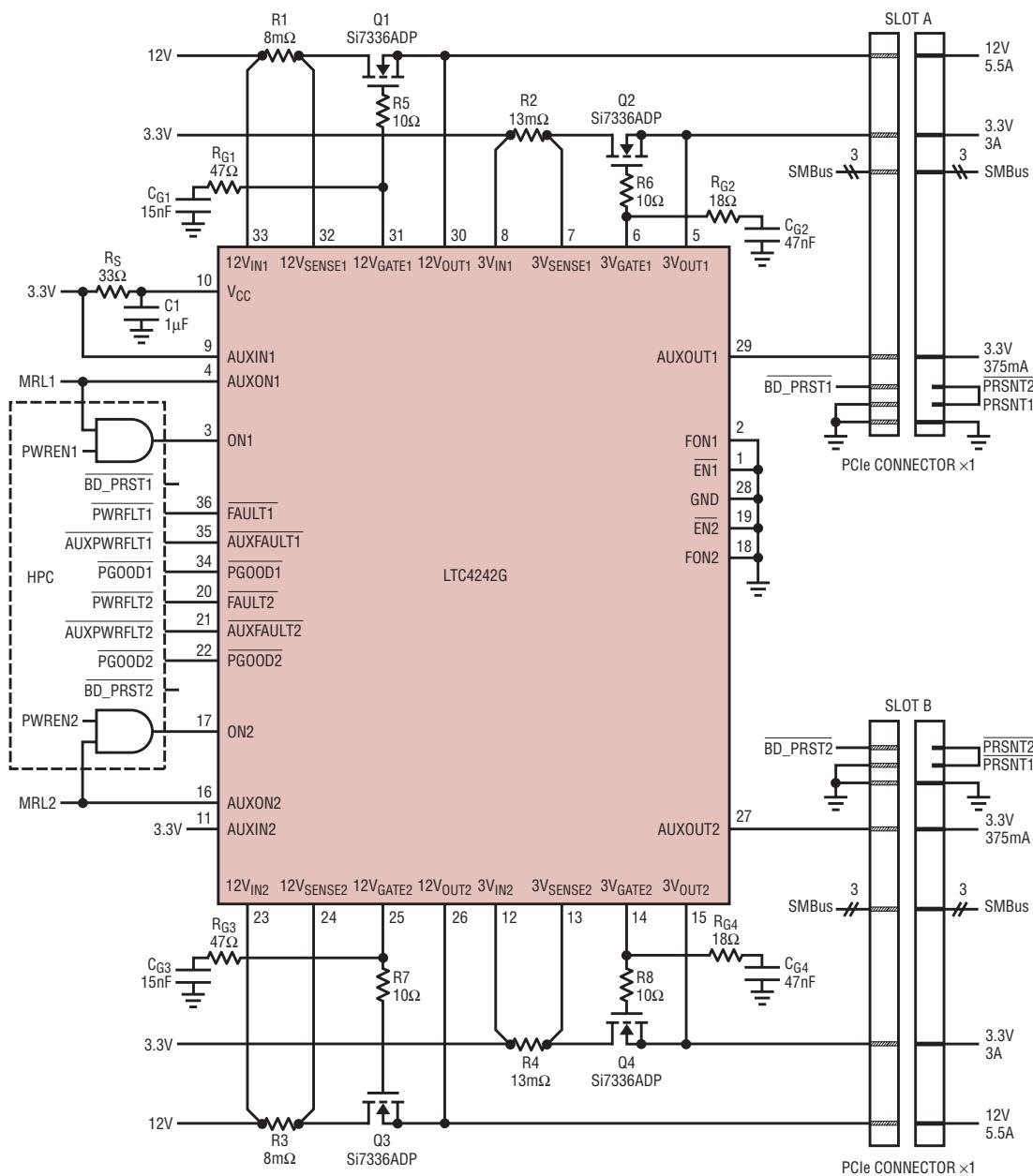


Figure 1. A typical PCI Express Hot Swap application, where the hot swapping events are controlled by the hot plug controller

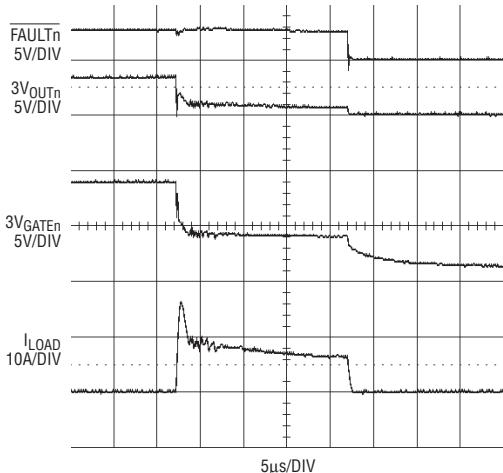


Figure 2. The 3.3V output is shorted without load capacitance. The current is brought under control by the current limit amplifier. After 20μs, the circuit breaker trips and FAULT pulls low.

power rail. An internal thermal shutdown circuit provides another level of protection for the FETs.

In a typical application, the LTC4242 uses four external N-channel pass transistors in addition to the two integrated FETs to isolate the add-in cards from the system when they are first inserted (Figure 1). When the system's Hot Plug Controller (HPC) senses that the add-in cards are seated correctly in the slot, it instructs the hot swap controller to apply power. Power is ramped gradually to minimize disturbance to the system. The LTC4242 continues to monitor for power path faults after the power-up process.

In Figure 1, four N-channel pass transistors, Q_{1-4} , and the integrated FETs control the application of power to two hot swappable cards. The series sense resistors, R_{1-4} , allow the LTC4242 to measure the load current in the power paths. Resistors R_{5-8} suppress self-oscillations in Q_{1-4} ; R_S and C_1 form a lowpass filter that ensures a stable supply to the part when the system power supply momentarily dips; and C_{G1-G4} control the inrush current on the 12V/3.3V power rails. C_{G1-G4} and R_{G1-G4} also form the compensation networks for the current limit loops.

The HPC enables the power to the add-in cards by pulling the ON and AUXON pins high. When the FON pins are pulled high, the pass transistors are turned on unless there is a thermal shutdown or undervoltage at the

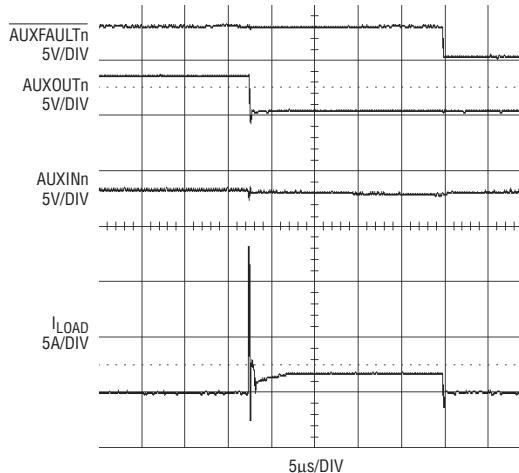


Figure 3. The 3.3V_{AUX} output is shorted without load capacitance. The current is brought under control by the current limit amplifier. After 22μs, the circuit breaker trips and AUXFAULT pulls low.

V_{CC} pin. This allows the user to pulse higher than normal current to the add-in cards to locate faulty parts or connections during a diagnosis.

Inrush Current Control

External capacitors, C_{G1-G4} , are connected from the GATE pins to ground to limit the inrush current by slewing the GATE voltage. With GATE pull-up current of 9μA, the GATE slew rate is given by:

$$\frac{dV_{GATE(n)}}{dt} = \frac{9\mu A}{C_{ISS} + C_{G(n)}}$$

where C_{ISS} is the external MOSFET's gate input capacitance. The inrush current flowing into the load capacitor, C_{LOAD} , is limited to:

$$I_{INRUSH} = C_{LOAD} \cdot \frac{dV_{GATE}}{dt}$$

For a 75W slot (see Table 1) with $C_{LOAD(12V)} = 2000\mu F$, $C_{LOAD(3.3V)} = 1000\mu F$, $C_{G1} = 15nF$, $C_{G2} = 47nF$ and $C_{ISS} = 3nF$, $I_{INRUSH(12V)} = 1A$ and $I_{INRUSH(3.3V)} = 0.18A$. The inrush current must be kept below the circuit breaker trip threshold to ensure successful start-up.

For the internal FET, an internal soft start circuit slews the gate, such that the inrush current is:

$$I_{INRUSH(3.3VAUX)} = SR \cdot C_{LOAD(3.3VAUX)}$$

where SR is the 3.3V_{AUX} output rising slew rate. From Table 1, with $C_{LOAD(3.3VAUX)} = 150\mu F$ and the internal

$SR = 1.2V/ms$, $I_{INRUSH(3.3VAUX)} = 0.18A$. $C_{LOAD(3.3VAUX)}$ must be chosen such that the inrush current doesn't exceed the circuit breaker trip threshold of 550mA.

Overcurrent Protection

The main power rails circuit breaker trip sense voltage is 50mV with a 10% tolerance. For the internal FET, its circuit breaker trips when the load current exceeds 550mA and has a 30% tolerance. The response time of the circuit breakers are internally fixed at 20μs. When the 20μs expires after an overload condition, the power switches are immediately turned off to disconnect the add-in cards from the system supply. The FAULT and AUXFAULT pins are pulled low to indicate an overcurrent fault has occurred on either the main or 3.3V_{AUX} power rails. The ON and AUXON pins must be pulled below 0.6V to reset the internal fault latches. Another way to reset the part is to cycle the power supplies below the UV level.

In addition to the circuit breaker, the LTC4242 includes a fast analog current limit amplifier to offer dual-level protection on each of the power rails. The amplifier is compensated for stability by the RC network at the GATE pins, which are servoed to limit the voltage drop across the sense resistors, R_{1-4} , to 100mV. In the event of a severe overload, the load current may overshoot as Q_{1-4} initially have large gate overdrive. The gates of Q_{1-4}

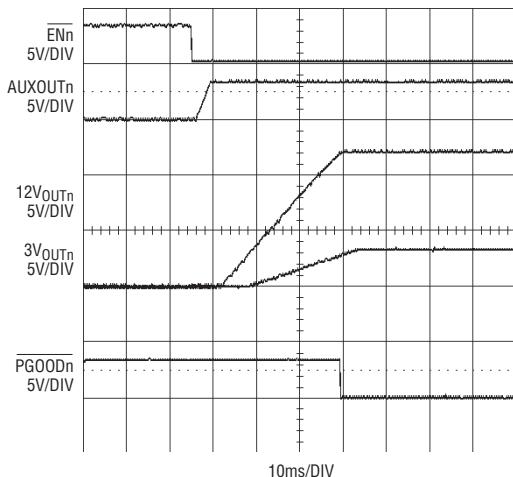


Figure 4. A typical power-up sequence.

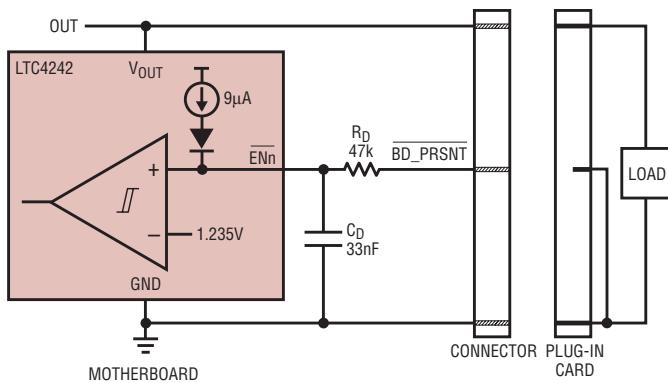


Figure 5. An RC network is connected from the \overline{EN} pin to the $\overline{BD_PRSNT}$ pin to generate plug-in debounce.

are quickly discharged by a 250mA pull down to the OUT pins, followed by the analog current limit amplifier response. R_{5-8} allows the voltage on C_{1-4} to be higher than the GATES immediately after the strong discharge, which could aid gate recovery by providing an alternative-charging path to pull up the GATES.

Figure 2 shows the output of the 3.3V rail being shorted into a 0.1Ω load without any load capacitance.

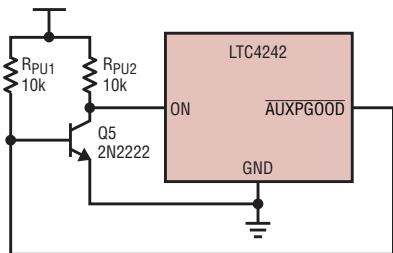


Figure 6. The main supplies are enabled by the $3.3V_{AUX}$ supply, which is used for board management and control functions.

is turned off and \overline{FAULT} pulls low. In another scenario, the output of the $3.3V_{AUX}$ rail is shorted into a $30m\Omega$ load without any load capacitance, and the fault current is swiftly limited. (Figure 3).

Power-Up Sequence

A typical power-up timing sequence starts with the detection of an add-in card in the slot. Typically, this information is fed to the HPC, which instructs the LTC4242 to turn on the power switches via the ON/AUXON pins. Another alternative is to feed this information to the \overline{EN} pin. Figure 4 shows the power up waveforms in response to the \overline{EN} pin going low. The ON/AUXON pins, not shown in the figure, are high. An RC network can be added to this pin to provide debounce delay during card plug-in and removal. As shown in Figure 5, with $R_D = 47k\Omega$ and $C_D = 33nF$, the plug-in debounce delay is 1.4ms and the power to the slot is disabled 2.8ms after detection of card removal through the $\overline{BD_PRSNT}$ signal.

The power-up voltage rate of the $12V_{OUT}$ and $3.3V_{OUT}$ is approximately given by $dV/dt = 9\mu A/C_{G1,G2}$. For the internal power switch, the output rises at a slew rate of 1.2V/ms. Once the output voltages crosses the power good thresholds, the \overline{PGOOD} pin for the 12V/3.3V and the $\overline{AUXPGOOD}$ pin for the $3.3V_{AUX}$ (available on QFN38 only) are pulled low. The output voltages are continually monitored, and the \overline{PGOOD} and $\overline{AUXPGOOD}$ pins are pulled high when the output voltages drop below the power good thresholds.

Table 1. PCI Express power supply specifications.

Power Rail	Specification	10W Slot	25W Slot	75W Slot
3.3V	Voltage Tolerance Supply Current Capacitive Load	$\pm 9\%$ 3.0A(max) $1000\mu F$ (max)	$\pm 9\%$ 3.0A(max) $1000\mu F$ (max)	$\pm 9\%$ 3.0A(max) $1000\mu F$ (max)
12V	Voltage Tolerance Supply Current Capacitive Load	$\pm 8\%$ 0.5A $300\mu F$ (max)	$\pm 8\%$ 2.1A (max) $1000\mu F$ (max)	$\pm 8\%$ 5.5A (max) $2000\mu F$ (max)
$3.3V_{AUX}$	Voltage Tolerance Supply Current Capacitive Load	$\pm 9\%$ 375mA(max) $150\mu F$ (max)	$\pm 9\%$ 375mA(max) $150\mu F$ (max)	$\pm 9\%$ 375mA(max) $150\mu F$ (max)

The $\overline{\text{AUXPGOOD}}$ pin can also be used to sequence the output voltages. In the circuit shown in Figure 6, the main supplies turn on when the $3.3V_{\text{AUX}}$ output is above the power good threshold and not in fault. This circuit is useful if the $3.3V_{\text{AUX}}$ output is the supply to the board management and control functions.

Power-Down Sequence

During power-down, the gates of the external pass transistors are discharged with 1mA pull down current sources. The gate of the internal $3.3V_{\text{AUX}}$ FET is discharged by a weak current source. The power switches are turned off slowly to avoid glitching the power supplies. Internal pull down transistors discharge the output load capacitors. $\overline{\text{PGOOD}}$ and $\overline{\text{AUXPGOOD}}$ pull high immediately after $\overline{\text{EN}}$ goes high. Figure 7 shows the power down waveforms in response to $\overline{\text{EN}}$ going high, with load capacitors on the outputs.

LT3494, continued from page 20

than size, the addition of a small CMDSH-3 (or equivalent) Schottky diode connected between the SW and CAP pins can further improve the efficiency by a few percentage points. If an external Schottky diode is used, connect as shown in the Figure 2 schematic.

Dimming Control

The LT3494 also integrates a dimming control feature. This feature, available to the user via the CTRL pin, allows full control of the output. Applying an external voltage below 1.225V to the CTRL pin overrides the internal reference and lowers the output voltage for purposes of dimming or contrast adjustment.

Efficiency

This converter maintains good efficiency over the entire load range helped by its low quiescent current and adaptive switching frequency. At light load, the switching frequency is reduced which reduces switching losses. Efficiency can be even better for applications where the output discon-

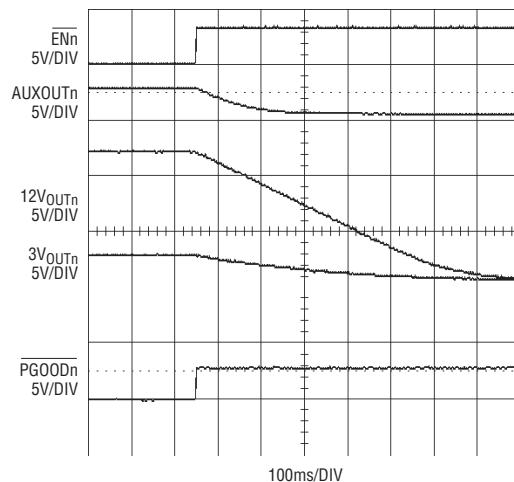


Figure 7. A typical power down sequence.

Conclusion

The LTC4242 provides a comprehensive solution to PCI Express Hot Swap applications. Fast current limiting and circuit breaker functions ensure that system disturbance is minimized during severe overloads and faults are quickly isolated. Integrated power

FETs reduce overall system complexity and cost. The LTC4242 is available in a 36-pin SSOP package and a 38-pin 5mm × 7mm QFN package. **L**

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nnect function is not needed. Figure 4 shows efficiency with the load at V_{OUT} or CAP for the LT3494 converter.

The LT3494A Supplies More Current

The converter in the schematic shown in Figure 2 shows a circuit with the LT3494A, a higher current version of the LT3494. The LT3494A has a typical switch current limit of 350mA, allowing approximately 50% more output current than the LT3494. With the increased current capability, a larger capacitor value of $0.47\mu\text{F}$ is recommended at the CAP node to maintain low ripple. Higher ripple levels have an adverse effect on efficiency. This larger capacitance produces an inrush current during start-up that stresses the internal diode in the LT3494A. To lower the stress in the internal diode, an external Schottky diode is necessary when the LT3494A is used. As with the LT3494, this diode also has the positive effect of increasing the efficiency by a few percentage points. Figure 5 shows the efficiency of this converter.

The LT3494 Uses All Ceramic Capacitors

Ceramic capacitors are well suited for most LT3494 applications because of their small size and low ESR. X5R or X7R types are recommended because they retain their capacitance over wider voltage and temperature ranges than other types such as Y5V or Z5U. The capacitors shown at the V_{OUT} nodes are rated at 25V and are of 1206 case to ensure enough capacitance is available for good stability margins and load transients.

Conclusion

The LT3494 simplifies the design of passive matrix OLED displays by integrating the Schottky diode, feedback resistor and output disconnect circuitry in a 3mm × 2mm package. Even though the LT3494A requires an external diode, both the LT3494 and the LT3494A maintain low output ripple at a non-audible frequency throughout the entire load range and provide full dimming control via the CTRL pin. **L**

Low Noise, High Current Regulated Charge Pump in 2mm × 2mm

Introduction

The LTC3204-5 and LTC3204-3.3 make it possible to produce low noise, high current, step-up or step-down power solutions in less than 0.04in². Both of these charge pumps are available in tiny 2mm×2mm DFN packages and include a patented technology

Both charge pumps are available in tiny 2mm × 2mm DFN packages, and only three tiny external ceramic capacitors are needed to complete a design.

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that reduces input noise generally associated with switch mode power supplies. Only three tiny external ceramic capacitors are needed to complete a design—no inductor is required. Table 1 shows the current ratings for a variety of applications.

Li-Ion to 5V or 2-AA to 3.3V

Figure 1 shows the LTC3204-5 in a circuit that produces a regulated 5V from a 2.7V to 4.2V Li-Ion battery. The available output current is as high as 150mA when the input is above 3.1V and 65mA for input voltages above 2.7V. A similar LTC3204-3.3 circuit generates 3.3V from two AA batteries. The 3.3V circuit supplies up to 50mA.

All of the capacitors are ceramic in 0603 size and produce low, predictable output ripple of 20mV with a 5V, 150mA output. The shutdown current is a mere 1µA and the no load input

by Hua (Walker) Bai

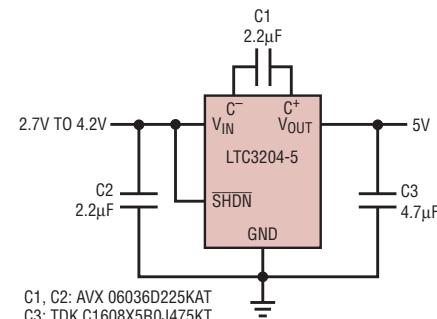


Figure 1. This power supply produces up to 150mA in less than 0.04 inch².

current is 60µA, thus improving battery life. Burst Mode operation reduces switching losses and greatly improves efficiency at light loads. The efficiency of the circuit is 81.3% when V_{IN} is 3V and I_{OUT} is 150mA. Features such as soft-start, short-circuit current limit and thermal shutdown enable robust operation for both LTC3204-5 and LTC3204-3.3.

Compact White LED Driver

Figure 2 shows a compact white LED driver for parallel LEDs, driving 3 LEDs at 15mA each from a single cell Li-Ion battery. Transistor Q1, which provides dimming, can be very small because it only needs to be rated for 45mA. This circuit can drive 10 LEDs at 15mA each if the input voltage remains above 3.1V. □

Table 1. Output Current Ratings

IC	V_{IN}	V_{OUT}/I_{OUT}
LTC3204-5	3.1V–5.5V	5V/150mA
	2.7V–5.5V	5V/65mA
LTC3204-3.3	1.9V–4.5V	3.3V/50mA
	1.8V–4.5V	3.3V/40mA

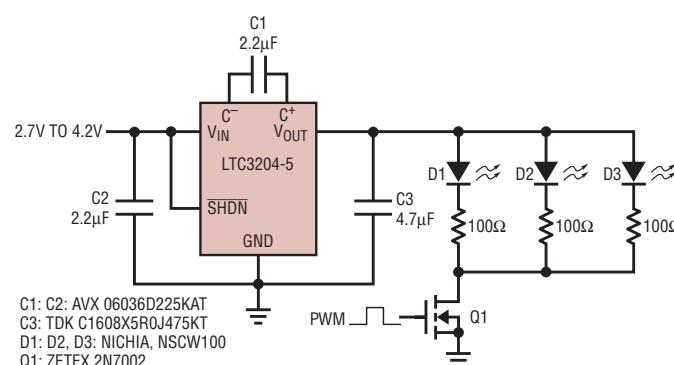


Figure 2. A compact LED driver with dimming control. When Q1 is on, the current in each LED is 15mA; when off, the current in each LED is zero. More levels of brightness can be achieved by applying a PWM signal to Q1, which effectively changes the average current at the LEDs.

Buck Controller with Low Offset Remote Sense Amplifier Allows Tight Regulation Despite Drops Due to Trace Resistance

by Narayan Raja

Introduction

The LTC3823 is a constant on-time, valley current mode synchronous buck controller with its on-time set by an external resistor. The on-time can be compensated for input and output voltage variations, minimizing frequency change with changing duty cycle requirements. The constant on-time architecture with a minimum programmable on-time of 50ns can accommodate very low duty cycle operation, without sacrificing switching frequency. Combined with a wide input supply range, this allows input voltages as high as 30V to be efficiently stepped down to output voltages as low as 0.6V. For output voltages lower than 3.3V, a low offset remote sense amplifier tightens regulation accuracy. The LTC3823 is available in a 32-lead 5mm × 5mm QFN, with a 28-lead SSOP narrow package option for customer preferring a leadless package.

Features

Remote Sense Amplifier

The recent trend in step-down power conversion has been towards lower output voltages, while the load cur-

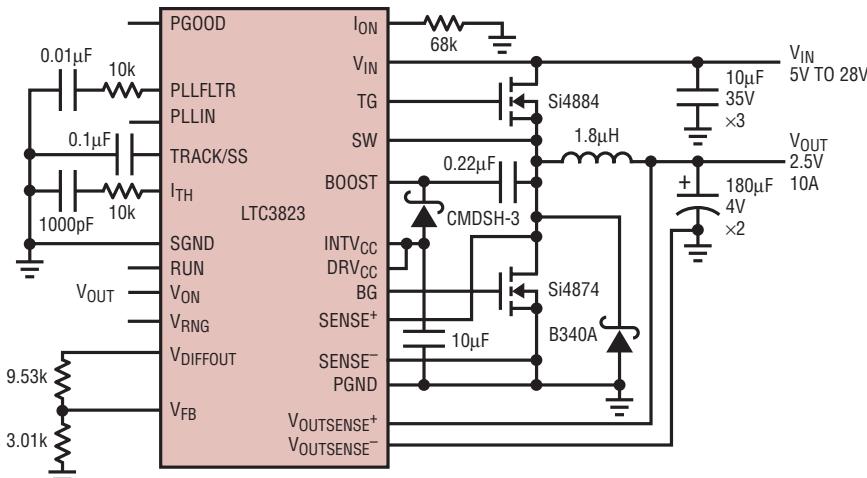


Figure 1. High efficiency step-down converter

rent requirements and regulation tolerances have been going up. Now more than ever, voltage drops due to parasitic trace resistance on the board contribute significantly towards power loss and regulation inaccuracies. Limited board space leads to compromised switching regulator layouts where the load may not be close to the MOSFETs and inductor, further aggravating the problem. A remote sense amplifier allows the direct sensing and regulation of the voltage across the load, allowing tighter output voltage control with

less than optimal placement of power components.

Post package trimming enables the remote sense amplifier of the LTC3823 to meet a 2mV offset specification. With a wide input common mode and output voltage range, the internal diffamp can be used with output voltages up to 3.3V. A greater than 2mA source capability allows the use of lower impedance resistor dividers off the amplifier output, while high input impedance minimizes IR drops along the sense lines. The op amp has an open loop gain of greater than 120dB with a unity gain bandwidth of over 3.5MHz.

Flexibility

The LTC3823 can be operated either in discontinuous conduction mode (DCM) or in forced continuous mode (FCM), set via the FCB pin. For improved performance in tracking mode, the part is designed to start up in DCM and switch over to the selected mode after the output reaches 80% of the desired voltage. A RUN pin allows the user to enter a low I_Q sleep mode

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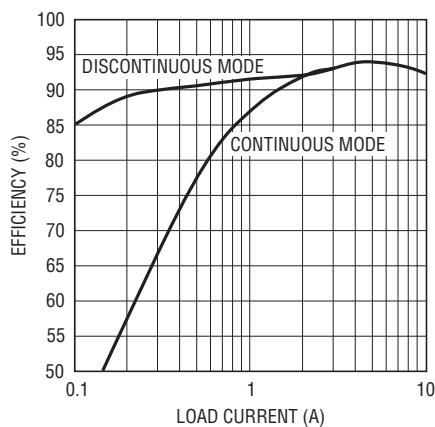


Figure 2. Efficiency and power loss for the circuit in Figure 1

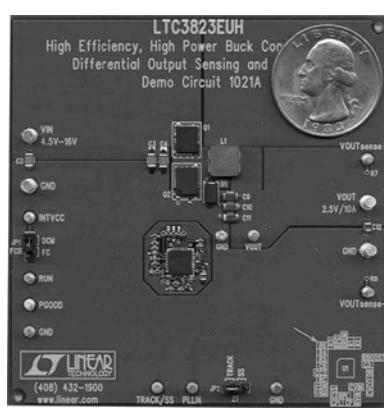


Figure 3. Layout of the circuit in Figure 1

PowerPath Controllers Improve Efficiency and Reduce Heating in Power Supply ORing and Undervoltage/Ovvoltage Protection Applications

by Luke A. Perkins

Introduction

Power supply switchover and load protection from overvoltage and undervoltage conditions are difficult circuits to design using off-the-shelf components. Standard components, when used in these applications, are power hungry and do not provide precise thresholds and timing often demanded in today's power circuits. The LTC4416 and the LTC4416-1 ICs provide solutions to these problems.

The LTC4416 and LTC4416-1 are dual, interconnected PowerPath controllers designed specifically to drive large or small Q_G PFETs. Unlike traditional NFET hot swap controllers, PFET controllers permit highly efficient ORing of multiple power sources for extended battery life and low self-heating. The primary advantage of a PFET controllers is they do not require a noise generating, power consuming charge pump.

The LTC4416 targets power supply switchover applications. When the primary power supply has sufficient voltage, it provides the power to the load. When the primary power supply drops below a user-defined threshold, the LTC4416 transitions the load from the primary to the secondary power source. The transition is performed in such a way as to minimize the voltage droop on the load. The transition thresholds are defined by a simple, three-resistor network.

The LTC4416-1 will work for power switchover conditions where the input supplies move rapidly; however, the LTC4416-1 will also provide overvoltage and undervoltage protection. This protection is accomplished with three external PFETs.

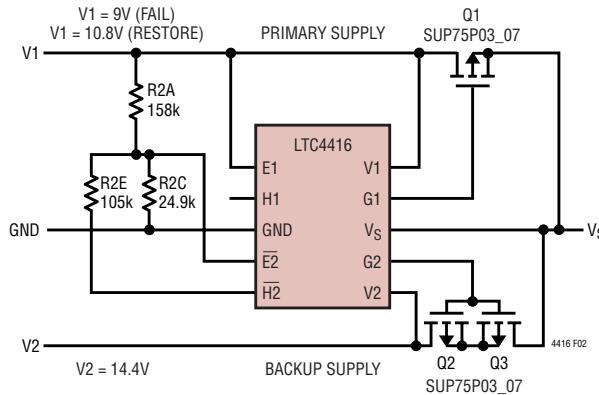


Figure 1. Power supply switchover application where voltage of the primary (V1) is nominally lower than that of the secondary (V2).

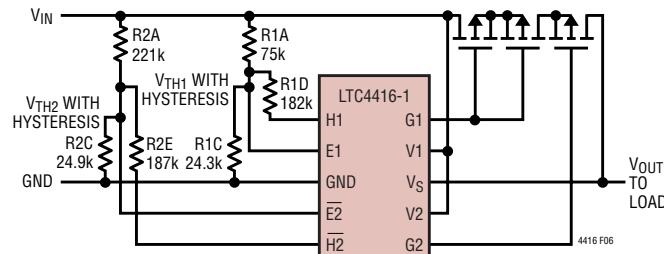


Figure 2. Overvoltage and undervoltage application

The overvoltage and the undervoltage thresholds are established with external resistor networks. The gate drivers in the LTC4416-1 are designed specifically for quick-transition to maximize the protection to the load.

Switchover Application

Figure 1 shows a power supply switchover circuit for an application where the secondary power source has a nominally higher voltage than the primary. For example, where the primary is a 12V power supply and the secondary is a 4-cell Li-Ion battery pack.

When the primary (V1) is 12V, E2 disconnects V2 and Vs through Q2 and Q3 by forcing G2 to V2—H2 is open circuit. E1 is connected to a voltage greater than the V_{REF} to keep the V1 to Vs path active. The Vs output can be shut completely off by grounding the E1 input. The LTC4416 takes its power from the higher of V1, V2 and Vs. This configuration provides power from V1 to Vs until the V1 supply drops below 9V.

When V1 drops below 9V, the H2 pin closes to GND, G2 drops to a V_{CLAMP} below V2 and G1 rises to the Vs voltage level. V2 supplies current to Vs until

V1 rises above 10.8V. The transition from the V2 to V1 is accomplished by slowly (10ms) turning off Q2 and Q3 allowing the Q1 to turn on rapidly when V_S matches V1. The H1 output is open until the E1 input drops below the V_{REF} voltage level. The V1 V_{FAIL} is determined by:

$$\begin{aligned} V_{FAIL} &= V_{ETH} \cdot \frac{R2A + R2C}{R2C} \\ &= 1.222V \cdot \frac{158k + 24.9k}{24.9k} \\ &= 8.98V \end{aligned}$$

Determine V1 $V_{RESTORE}$ by:

$$\begin{aligned} V_{RESTORE} &= V_{ETH} \cdot \frac{(R2A + (R2C \parallel R2E))}{R2C \parallel R2E} \\ &= 1.222V \cdot \frac{158k + (24.9k \parallel 105k)}{24.9k \parallel 105k} \\ &= 10.81V \end{aligned}$$

Undervoltage and Overvoltage Shutdown

Figure 2 shows an application that disables the power to the load when the input voltage gets too low or too high. When V_{IN} starts from zero volts, the load to the output is disabled until V_{IN} reaches 5.5V. The V1 path is enabled and the load remains on the input until the supply exceeds 13.5V. At that voltage, the V2 path is disabled. As the input falls, the voltage source is reconnected to the load when the

input drops to 12V and the V2 path is enabled. Finally, the load will be removed from the input supply when the voltage drops below 5V.

Undervoltage

$$\begin{aligned} V_{FAIL} &= V_{ETH} \cdot \frac{R1A + R1C}{R1C} \\ &= 1.222V \cdot \frac{75k + 24.3k}{24.3k} \\ &= 4.99V \end{aligned}$$

$$\begin{aligned} V_{RESTORE} &= V_{ETH} \cdot \frac{(R1A + (R1C \parallel R1D))}{R1C \parallel R1D} \\ &= 1.222V \cdot \frac{75k + (24.3k \parallel 182k)}{24.3k \parallel 182k} \\ &= 5.497V \end{aligned}$$

Overvoltage

$$\begin{aligned} V_{FAIL} &= V_{ETH} \cdot \frac{R2A + R2C \parallel R2E}{R2C \parallel R2E} \\ &= 1.222V \cdot \frac{221k + 24.9k \parallel 187k}{24.9k \parallel 187k} \\ &= 13.51V \end{aligned}$$

$$\begin{aligned} V_{RESTORE} &= V_{ETH} \cdot \frac{R2A + R2C}{R2C} \\ &= 1.222V \cdot \frac{221k + 24.9k}{24.9k} \\ &= 12.07V \end{aligned}$$

The over and undervoltage lockout circuits are shown here working in tandem. It is possible to configure the circuit for either over or undervoltage

lockout by using only one of the voltage paths and eliminating the components from the other. Only one PFET is required in this case. The LTC4416-1 should be used in this configuration rather than the LTC4416 because the LTC4416-1 turns off rapidly if an over or undervoltage condition is detected.

Conclusion

The LTC4416 provides power supply switchover solutions that cannot be easily generated using off-the-shelf components. The LTC4416 also provides power efficiencies not available with traditional NFET Hot Swap controllers. These efficiencies reduce the IDD of the solution by not having active switching gate drivers. The power losses are also reduced by decreasing the voltage drop across the PFETs to 25mV. The LTC4416 provides a smoother transition between the backup and the secondary power supplies.

The LTC4416-1 dual gate drivers provide a single controller solution to not only protect loads from overvoltage conditions, but also undervoltage conditions. The user can externally program the overvoltage and undervoltage thresholds using simple external resistor networks. These resistor networks also provide hysteresis to prevent chattering between the power source and the load. □

LTC6103/LTC6104, continued from page 8

and below this point. Make sure that the lowest expected output level is higher than pin 4 (V) by at least 0.3V to ensure that negative going output swings remain linear.

H-Bridge Load Current Monitor

The H-bridge power-transistor topology remains popular as a means of driving motors and other loads bi-directionally from a single supply potential. In most cases, monitoring the current delivered to the load allows for real-time operational feedback to a control system.

Figure 7 shows the LTC6104 used in monitoring the load current in an H-bridge. In this case, the LTC6104 operates with dual supplies. The output resistance is connected directly to ground instead of connected to a voltage reference. The output ranges from 0V to 2.5V for $V_{SENSE_A} = 0mV$ to 100mV, and from 0V to -2.5V for $V_{SENSE_B} = 0mV$ to 100mV.

Conclusion

The LTC6103 and LTC6104 are precise high side current sensing solutions. The parts can operate to 60V, making

them ideal for high voltage applications such as those found in automotive, industrial and telecom systems. Low DC offset allows the use of a small shunt resistor and large gain-setting resistors. The fast response time makes them suitable for overcurrent-protection circuits. Configurable gain means design flexibility. In addition, the open-drain output architecture provides an advantage for remote-sensing applications. □

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A Simple Integrated Solution to Drive Avalanche Photo Diodes

by Jesus Rosales

Introduction

Avalanche Photo Diodes (APDs) are expensive and electrically delicate modules that must be protected under many different adverse conditions. They require monitored bias voltage levels as high as 80V that are generated from a 3.3V or 5V supply. A high side current monitor is necessary since the APD anode is committed to the receiver amplifier's summing point. Traditionally, the challenge of driving and monitoring APDs has been addressed by the use of separate circuits. The main circuits are the step-up converter, the voltage monitor and the current monitor. Implementing separate circuits presents high side biasing problems and board space challenges. The LT3482 addresses these challenges with an integrated solution.

APD Driver Provides 10V to 80V at 2mA

The circuit in Figure 1 shows the LT3482 configured to produce an output voltage ranging from 10V to 80V from a 3V to 12V source—capable of delivering up to 2mA of load current. Its operation is straightforward. The LT3482 contains a 48V, 260mA internal switch, which boosts V_{OUT1} to one half the APD output voltage level. This voltage is doubled through an internal charge pump to generate V_{OUT2} . All boost and charge pump diodes are

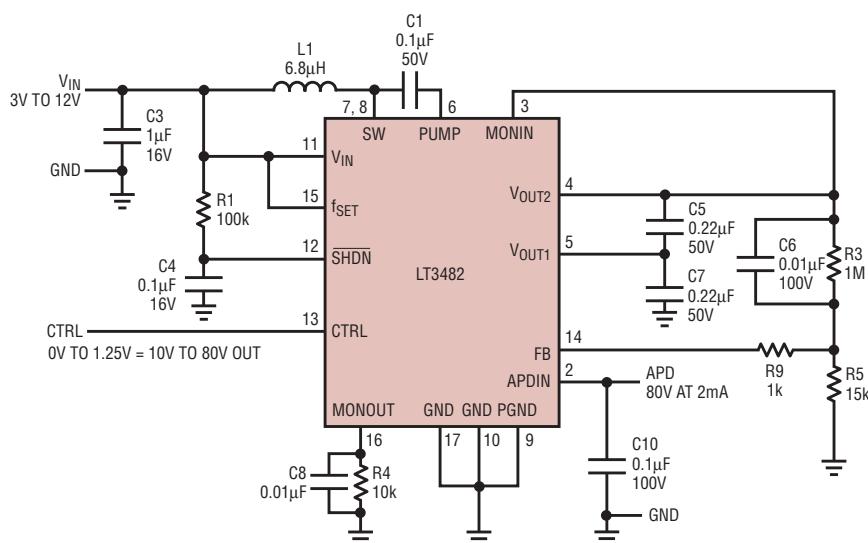


Figure 1. An APD converter operating at 1.1MHz; 3V input to 12V-80V output at 2mA

integrated. V_{OUT2} is regulated by the internal voltage reference and the resistor divider made up of R3 and R5. At this point, V_{OUT2} goes through the integrated high side current monitor (MON IN), which produces a current proportional to the APD current at the MON OUT pin, and produces a voltage across R4 which can be used to digitally program the output voltage via the CTRL pin.

The output voltage is available for the APD at the APD IN pin. C6 minimizes low frequency output noise due to internal reference and error amplifier noise (See *Linear Technology Design Note 273*). The CTRL pin serves to override the internal refer-

ence. By tying this pin above 1.25V, the output voltage is regulated with the feedback at 1.25V. By externally setting the CTRL pin to a lower voltage, the feedback and the output voltage follow accordingly.

The SHDN pin not only enables the converter when 1.5V or higher is applied, but also provides a soft-start function to control the slew rate of the switch current, thereby minimizing inrush current. The switching frequency can be set to 650kHz, or 1.1MHz by tying the FSET pin to ground or to V_{IN} , respectively. Fixed frequency operation allows for an output ripple that is predictable and easier to filter.

Figure 2 shows the output ripple for the Figure 1 application, and Figure 3 shows a typical layout.

Conclusion

The LT3482 provides a complete APD biasing solution with its integrated 48V, 260mA internal switch, boost and charge pump diodes and current monitor. Its fixed frequency, soft-start function, internal compensation and small footprint make the LT3482 a very simple and small solution not only for APDs but also for Optical Receivers, Fiber Optic Network Equipment and other applications.

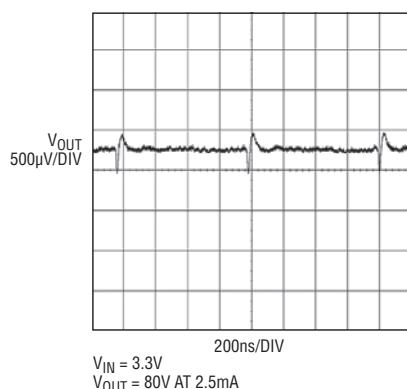


Figure 2. Output voltage ripple for the application in Figure 1.

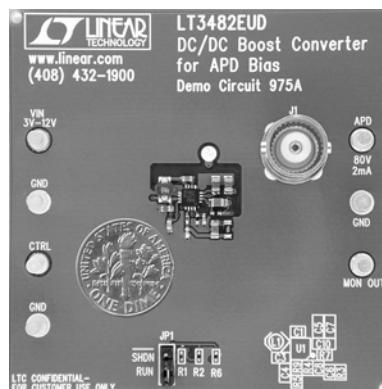


Figure 3. Typical layout for the Figure 1 converter.

Tiny 2.25MHz Monolithic Step-Down Regulator Delivers Low Ripple and Fast Transient Response

by Theo Phillips and Stephanie Dai

Introduction

A truism of contemporary power supply design is that portable electronics require high efficiency regulators to prolong battery life. This requirement rules out switching regulators that slosh inductor current back and forth even at the lightest loads. The most efficient power saving schemes use Burst Mode operation at light loads, but many of today's radio-equipped devices cannot tolerate the output voltage ripple and resulting system noise. The LTC3410 is designed to reconcile these demands, minimizing wasted current at light loads while keeping ripple to a tolerable level.

The LTC3410, in a 1mm high SC70 package, provides up to 300mA from an input of 2.5V to 5.5V. Its 2.25MHz switching frequency allows the use of tiny, surface mount components and keeps the switching noise well above the passband of most communication systems. Its low-ripple implementation of Burst Mode operation increases efficiency at light loads, consuming just 26µA of supply current at no load. In shutdown mode, less than 1µA is consumed. Current mode operation provides excellent line and load transient response at output voltages down to 0.8V. The internal 0.7Ω synchronous switch increases efficiency while eliminating the need for an external Schottky diode. The P-channel top MOSFET allows noise-free dropout operation at 100% duty

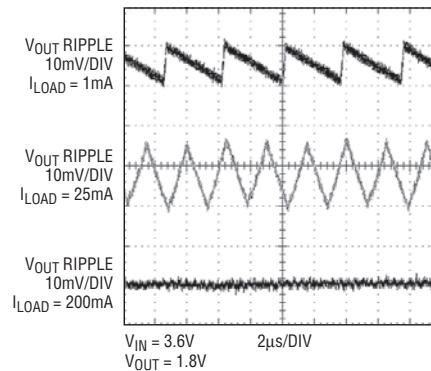


Figure 1. The LTC3410 produces low output voltage ripple throughout its load range.

cycle, further extending battery life in portable applications. Compensation and soft-start are internal, reducing the need for external components.

Low Ripple

Figure 1 shows the minimal ripple produced by a typical LTC3410 application at light load. Burst Mode operation causes the internal power MOSFETs to operate intermittently based on the required load, with intervening sleep intervals in which the output capacitor supplies the load current. But with just a few pulses with very short on-times in each burst, the peak-to-peak value holds to around 10mV at light load and 20mV at moderate load, thus reducing possible interference with audio circuitry. Other power saving methods, with longer pulse trains and on times, can

produce about ten times the ripple. At maximum load, the regulator operates continuously at 2.25MHz and ripple is just 5mV.

Fault Protection

The LTC3410 protects against output short-circuit and power overdissipation conditions. When the output is shorted to ground, the frequency of the oscillator slows to 1/7 the nominal switching frequency, or around 310kHz, to prevent inductor current runaway. The frequency returns to 2.25MHz when the feedback node is allowed to rise to 0.8V. At very high temperatures, a thermal protection circuit shuts off the power switches until the overtemperature condition clears.

Tiny 1.8V/300mA Step-Down Regulator Using All Ceramic Capacitors

Figure 2 shows a schematic of an LTC3410 application using all ceramic capacitors. It supplies 1.8V/300mA from a lithium-ion battery input range (2.7V–4.2V) with a nominal value of 3.3V. Its ceramic capacitors are very small and have low equivalent series

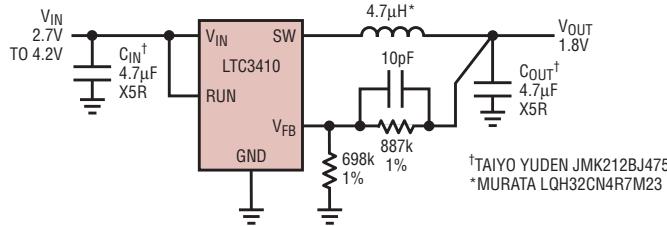


Figure 2. A typical LTC3410 supply, delivering 1.8V at up to 300mA, requires few external components.

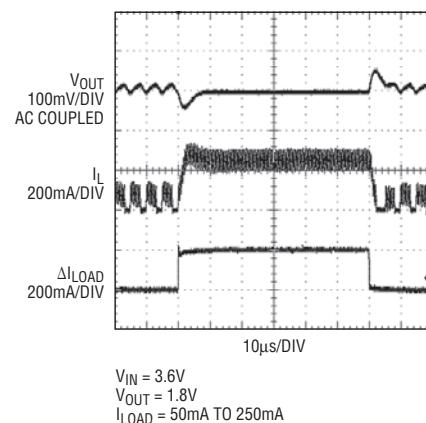


Figure 3. Quick transient response of the LTC3410.

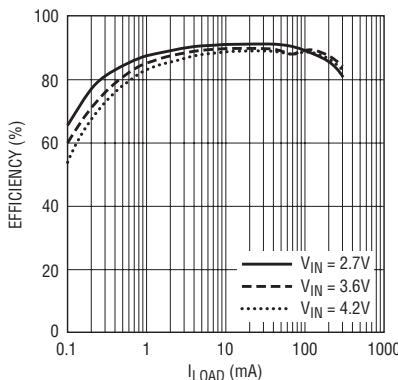


Figure 4. Efficiency for the regulator of Figure 1.

resistance, permitting extremely small ripple voltage at the input and output. The LTC3410's control loop does not require the higher ESR of tantalum or electrolytic capacitors for stable operation. Figure 3 shows the converter's fast response to a load transient from 50mA to 250mA.

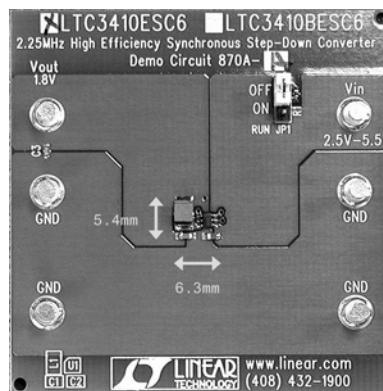


Figure 5. An LTC3410 application can fit into the tiniest of spaces. A typical layout occupies just one-third of a square centimeter.

Figure 4 shows efficiency curves for the circuit working from typical Li-ion single cell voltages. The low ripple implementation of Burst Mode operation comes at a small sacrifice in efficiency compared with other small monolithics

LTC2912/13/14, continued from page 13

$$\begin{aligned} R_{B1} &= \left| \frac{0.5V}{I_{n1}} \cdot \frac{V_{n1}}{V_{UV1}} \right| - R_{A1} \\ &= \left| \frac{0.5V}{10\mu A} \cdot \frac{5V}{4.75V} \right| - 45.3k\Omega \\ &\approx 7.15k\Omega \end{aligned}$$

$$\begin{aligned} R_{B3} &= \left| \frac{0.5V}{I_{n3}} \cdot \frac{V_{n3} - 1V}{V_{UV3} - 1V} \right| - R_{A3} \\ &= \left| \frac{0.5V}{10\mu A} \cdot \frac{-5V - 1V}{-4.75V - 1V} \right| - 46.4k\Omega \\ &\approx 5.76k\Omega \end{aligned}$$

R_C is obtained from Equation (3) of the "3-Step Design Procedure."

$$\begin{aligned} R_{C1} &= \left| \frac{V_{n1}}{I_{n1}} \right| - R_{A1} - R_{B1} \\ &= \left| \frac{5V}{10\mu A} \right| - 45.3k\Omega - 7.15k\Omega \\ &\approx 442k\Omega \end{aligned}$$

$$\begin{aligned} R_{C3} &= \left| \frac{V_{n3}}{I_{n3}} \right| - R_{A3} - R_{B3} \\ &= \left| \frac{5V - 1V}{10\mu A} \right| - 46.4k\Omega - 5.76k\Omega \\ &\approx 549k\Omega \end{aligned}$$

±3.3V Supply Monitoring

The resistor values required to monitor the ±3.3V supplies with a ±5% tolerance are found by repeating the "3-Step Design Procedure" for each supply:

For the +3.3V supply,

$$R_{A2} \approx 47.5k$$

$$R_{B2} \approx 5.11k$$

$$R_{C2} \approx 274k$$

For the -3.3V supply,

$$R_{A4} \approx 48.7k$$

$$R_{B4} \approx 3.83k$$

$$R_{C4} \approx 374k$$

48V Supply Monitor Example

Consider a system with a single 48V supply. System requirements specify a single LED to indicate a powergood condition. The powergood LED indicates if the supply is within ±10% of the nominal 48V. To provide some time for the outputs to settle into the powergood range, a timeout of 85ms is chosen for the outputs. Figure 5 shows the monitoring system.

The LTC2912-2 monitors a single supply without output latching. The

such as the LTC3405, but the extra energy devoted to the aggressive burst comparator reduces maximum ripple by a factor of at least five.

Conclusion

The LTC3410 is a high performance monolithic synchronous step-down DC/DC converter, which provides up to 300mA from a 2.5V–5.5V input while optimizing efficiency and low output voltage ripple. A typical LTC3410 application occupies just 34mm² with a maximum height under 1mm (Figure 5). With its high switching frequency, low $R_{DS(ON)}$ switches and small number of ancillary components, the LTC3410 is an excellent choice for space-constrained environments such as cellular phones, MP3 players, wireless modems and digital cameras. □

\overline{OV} and \overline{UV} outputs are tied together to create a single fault output. These open-drain outputs sink current away from the powergood LED causing it to turn off during a fault condition. The 200k R_Z keeps quiescent current low. This results in approximately 200μA of quiescent current in the LTC2912. Choosing a 1.25μA nominal current through the resistive divider and following the 3-Step Design Procedure results in the following resistor values.

$$R_A \approx 365k$$

$$R_B \approx 78.7k$$

$$R_C \approx 38.3M$$

To generate the 85ms timeout period, 10nF is chosen for C_{TMR} .

Conclusion

The LTC2914, LTC2913, and LTC2912 supervisor family provides accurate overvoltage and undervoltage supply monitoring. With separate overvoltage and undervoltage detection inputs and an internal shunt regulator, each supervisor can monitor any supply level while offering adjustable fault reporting options. □

Buck-Boost Converters Increase Handheld Battery Runtimes by 20%

by David Canny

Introduction

A common power supply problem in today's portable devices is generating a regulated voltage that falls somewhere in the middle of the full voltage range of the battery—for instance, producing 2.5V from two AA cells (1.8V to 3.2V), or 3.3V from a single Li-Ion cell (2.7V to 4.2V). A typical solution is to use a SEPIC converter. However a SEPIC has some inherent drawbacks, including the requirement of both a coupled inductor and a high current flyback capacitor. Another solution is a circuit that cascades a boost converter with either an LDO or a buck converter, but this is a costly and inefficient solution that reduces battery runtimes. Linear Technology's dedicated buck-boost converters offer a number of advantages—typically increasing battery runtimes by around 20% over the LDO solutions, reducing cost, and saving precious PC board real estate. Figures 1 and 2 show compact and high efficiency buck-boost solutions using the LTC3530 converter.

Figure 3 shows typical efficiencies versus input voltages for the different topologies mentioned above. The LTC3530 solution is the only one of the group that can reach 95% efficiency and it maintains significantly better efficiency than the other solutions across the entire operating voltage

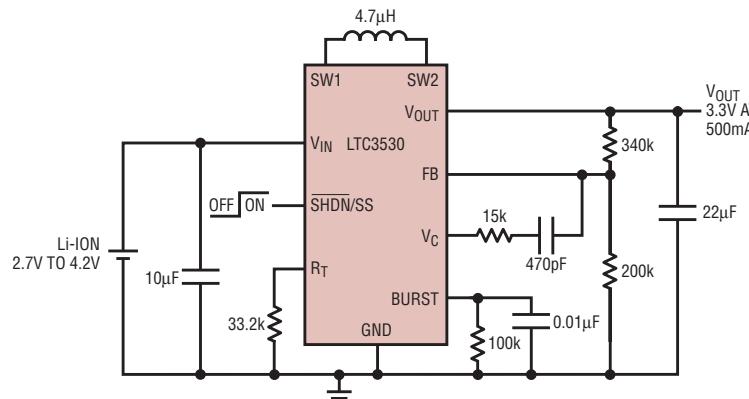


Figure 1. Li-Ion cell input to 3.3V output at 500mA

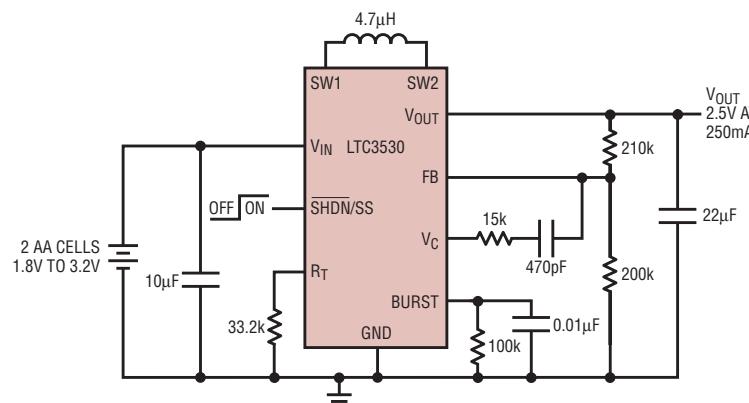


Figure 2. Two AA cell input to 2.5V output at 250mA.

range. Shutdown current is <1µA and automatic Burst Mode operation further improves battery runtime in light load current conditions. For added

flexibility, the LTC3530 allows the user to program the load current threshold for Burst Mode operation.

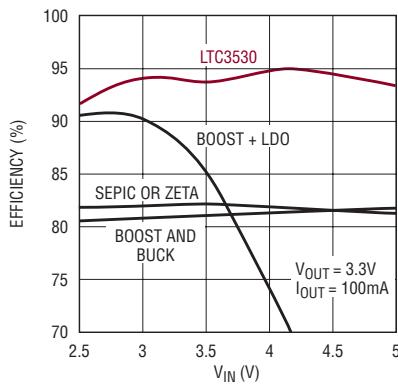


Figure 3. Efficiency of a Linear Technology buck-boost converter is far better than other, less-compact options, translating to possible battery runtime improvements of 20%.

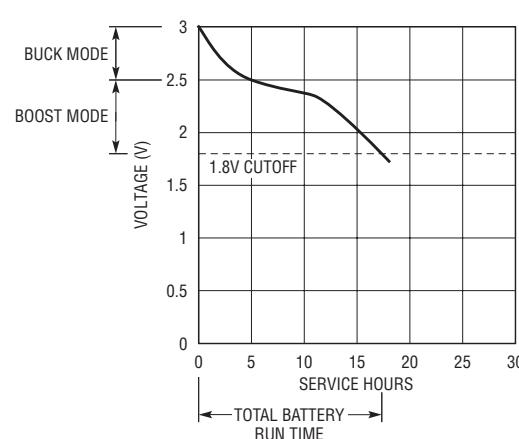


Figure 4. Discharge profile of two AA alkaline cells (constant current drawn = 125mA).

Better than Buck-Only or Boost-Only Solutions

To avoid the cost or real estate requirements of traditional SEPIC or cascaded boost-buck topologies, some designers opt for buck-only or boost-only solutions. For example, in two AA alkaline cell applications such as MP3 players, 2.5V often serves as the main rail since it drives both the flash memory and the main processor I/O. In such applications, some designers use a synchronous boost converter to save cost and space. The problem is that the boost converter is very inefficient while the battery voltage is above 2.5V because a boost converter

incurs both the losses inherent in an LDO and the switching losses an LDO doesn't have. Figure 4 shows that the boost converter operates inefficiently for 28% of the battery runtime (the portion of the battery life when the battery's voltage declines from a fully charged 3V to 1.8V). An LTC3530 solution results in significantly longer battery runtimes compared to these solo boost or buck solutions.

Conclusion

Linear Technology's synchronous buck-boost converter simplifies the design of lithium-ion or 2-AA-cell powered handheld devices that require up

to 600mA output. Programmable soft-start and switching frequency, as well as external compensation, make the LTC3530 a flexible and compact solution. The buck-boost topology helps a designer extend battery runtime while the automatic Burst Mode operation further maximizes the runtime in applications with widely varying load requirements. 



LT3476, continued from page 5

with a LED string. With this PFET disconnect circuit, the switch off time is less than 2 μ s.

Boost Circuit for Automobile Lighting

It is straightforward to use the LT3476 for a boost application given the fact the main power switch is tied to the ground. Figure 5 shows a boost circuit for applications such as automotive exterior and interior lighting. This circuit provides 350mA to eight Luxeon LEDs per channel from a car battery. The efficiency is over 92% with a 16V input.

Triple the Number of LED Strings with the LT3003

Each LT3476 channel can be configured to drive three parallel LED strings by adding the LT3003. In such a configuration, each LED string uses one third of the output current of the channel. The LT3003 easily operates in boost mode, or in buck mode with an architecture that allows the power ground (V_{EE}) to move with the output capacitor voltage. Figure 6 shows LT3476 channel 1 plus a LT3003 circuit in buck mode. The string-to-string current matching is 5%, important to maintaining uniform LED brightness between the strings. Figure 7 shows a LT3477 and a LT3003 circuit in boost mode. The V_{MAX} of the

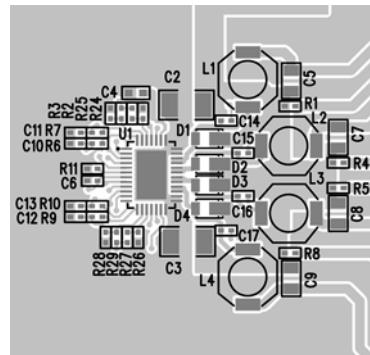


Figure 8. Recommended parts placement and layout

LT3003 should be tied to the highest voltage in a circuit. In the buck mode, it is PVIN. In the boost mode, it is the cathode of D1.

Layout Considerations

For proper operation and minimum EMI, care must be taken during the PCB layout. Figure 8 shows the recommended components placement for LT3476 in buck mode for a 4-layer board. The schematic is shown in Figure 1. In a buck circuit, the loop formed by the input capacitors (C2 and C3), the SW pins and the catch diodes (D1, D2, D3 and D4) should be as small as possible because of the present of high di/dt pulsing current in this loop. The second layer should be an unbroken ground plane. The SW nodes should be as small as possible. From each sense resistor, the traces

to the CAP pin and to the LED pin should be a Kelvin trace pair. Those traces should be in the third layer for best shielding. The fourth layer should be another ground plane.

If long wires are used to connect a power supply to PVIN of the LT3476, an aluminum-electrolytic capacitor should be used to reduce input ringing which could break down the LT3476 internal switch. See *Linear Technology Application Note 88* for more information.

To ensure reliable operation, good thermal designs for both the LT3476 and the LT3003 are essential. The exposed pads on the bottom of the packages must be evenly soldered to the ground plane on the PCB so that the exposed pads act as heat sinks. Unevenly soldered IC package degrades the heat sinking capability dramatically. To keep the thermal resistance low, the ground plane should be extended as much as possible. For the LT3476, on the top layer, ground can be extended out from the pins 19, 20, 21, 30, 31 and 32. This also allows tight loop components placement mentioned above. The second and fourth layers should be reserved for the ground plane. Thermal vias under and near the IC package helps transfer the heat from the IC to the ground plane and from inner layers to outer layers. 

Ultra-Low Quiescent Current Regulated DC/DC Converter for Light Load Applications

by Vui Min

In lightly loaded battery applications that require regulated power supplies, the quiescent current drawn by the DC/DC converter can be a substantial portion of the average battery current drain. In such applications, minimizing the quiescent current of the DC/DC converter becomes a primary objective because this results in longer battery life and/or an increased power budget for the rest of the circuitry.

The LTC3221 is a micropower charge pump designed to produce up to 60mA of output current while drawing only 8 μ A of quiescent current at no load. The part uses the Burst Mode architecture to provide a regulated output voltage. The low quiescent current of the LTC3221 may render shutdown of the output unnecessary because the 8 μ A quiescent current is less than the self-discharge rate of many batteries. However, the part is also equipped with a 1 μ A shutdown mode for additional power saving.

Figure 2 shows a plot of the LTC3221-3.3 excess input supply current vs load. The input current for an ideal regulating voltage doubler is always twice the output current. The

excess input current is used to power the LTC3221 internal circuitry and stray capacitance. At light load of up to 100 μ A, the quiescent current remains low at 8 μ A. As the load current increases further, the LTC3221 switches more frequently and the supply current starts increasing.

In a conventional μ power charge pump, the charge pump switches are controlled by a hysteretic comparator and reference to provide output regulation. The switches are either delivering maximum current to the output or are turned off completely. A low frequency ripple appears at the output, which is required for regulation. The amplitude of this ripple is heavily dependent on the load current, the input voltage and the output capacitor size. At high input voltage and light load, the output ripple can become substantial because the increased strength of the charge pump causes fast edges that may

outpace the regulation circuitry. This high amplitude ripple can also result in poor line and load regulation. One solution to reduce the amplitude of the output ripple is to use a higher value output capacitor, greater than 10 μ F, but this of course takes more board space and increases expense.

The LTC3221 overcomes this problem by using a constant current to charge the output if the output is low, thus keeping the output ripple fairly constant over the full input voltage operating range. The part requires only a 4.7 μ F capacitor, 0603 size, at the output to achieve an output ripple of <30mV, which is <1% of a 3.3V output. The part can work with a 2.2 μ F capacitor at an output ripple of 50mV, which is about 1.5% of a 3.3V output. Figure 3 shows the output ripple of the LTC3221-3.3 at 2.5V V_{IN} and with 60mA load.

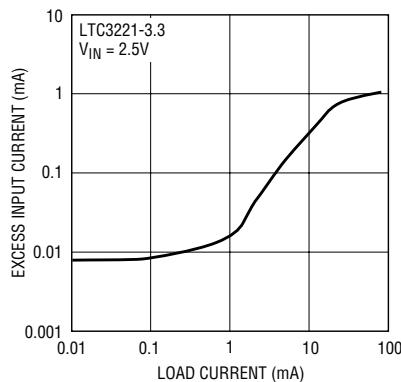


Figure 2. Excess input supply current vs load.
The input current for an ideal regulating voltage doubler is always twice the output current. The excess input current is used to power the LTC3221 internal circuitry and stray capacitance.

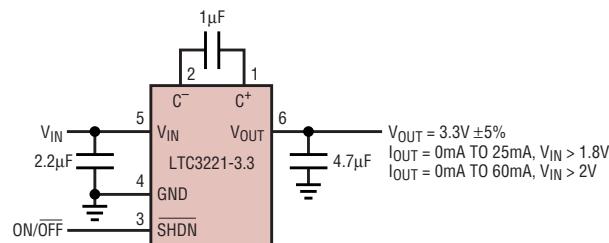


Figure 1. Regulated 3.3V output from 1.8V to 4.4V input

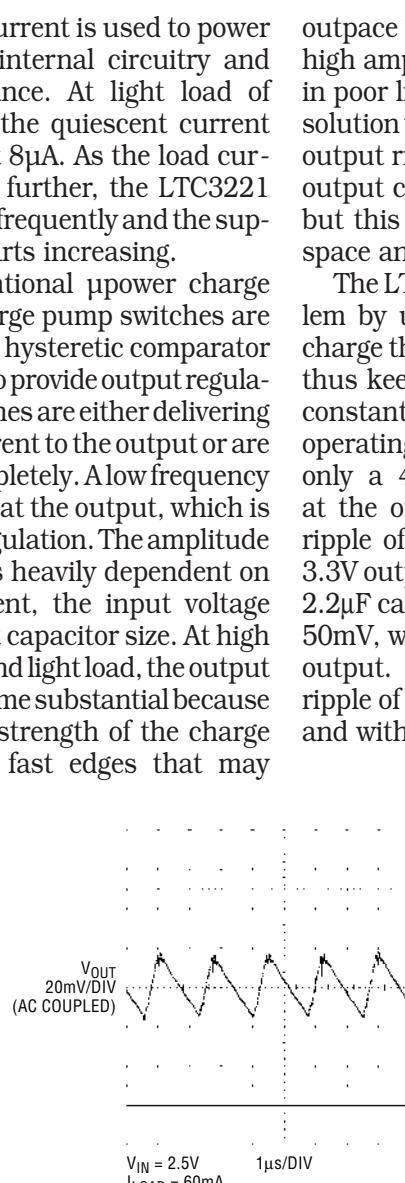


Figure 3. The LTC3221 uses a constant current to charge the output if the output is low, thus keeping the output ripple fairly constant over the full input voltage operating range.

The LTC3221 can operate with the input supply voltage as low as 1V with limited output drive. This feature allows the output to drop gracefully when the battery terminal voltage starts decreasing, further prolonging the battery life.

The LTC3221 family is available in the 2mm × 2mm 6-pin DFN package and it requires only three external capacitors to operate, achieving a very small total component area. The LTC3221 family comes in three output versions: fixed 3.3V, fixed 5V and adjustable.

LTC3410, continued from page 10

500mA of Output Current from a 2mm × 2mm Package

For more output current in a slightly bigger package than the LTC3410, Linear Technology offers the LTC3542, a 500mA monolithic step-down converter available in both a 2mm × 2mm DFN package and 6-lead SOT-23. Burst Mode operation or pulse skipping mode can be easily selected through the MODE/SYNC input. Quiescent current is only 26µA and the output voltage ripple in Burst Mode operation is only 20mV_{P-P}. The device also offers soft-start to prevent excessive current draw on the input supply during start-up.

The LTC3542 also offers external frequency synchronization which can be used to avoid potential problems with radiated electromagnetic interference (EMI) from the switching currents in the inductor. Although the LTC3542 (as well as LTC3410 and LTC3547) mitigates EMI problems by carefully controlling the turn-on and turn-off of the integrated switches to reduce the EMI magnitude, setting the switching frequency to be outside of

the frequency band of interest is effective as well.

To enable frequency synchronization, connect an external clock to the MODE/SYNC input. The frequency can be synchronized anywhere between 1MHz and 3MHz. Pulse skipping mode is automatically selected when using this input to sync the switching frequency.

Figures 7 and 8 show a typical application circuit and efficiency graph. The circuit takes up about 55mm² of board space, as shown in Figure 9.

A Dual 400mA/800mA Synchronous Buck Converter

For applications needing more power, the LTC3548 can supply 400mA and 800mA respectively from two output channels. The LTC3548 is a dual synchronous buck regulator in a 10-lead MSOP/DFN package. With no load, both converters draw only 40µA. Burst Mode operation and pulse skipping mode can be selected via the MODE/SYNC pin, and external frequency synchronization is also supported. Figure 10 shows a typical application.

With the tiny 6-pin 2mm × 2mm DFN package and low external parts, the LTC3221 family of charge pumps is perfect for space-constrained applications. The low operating current of these parts make them ideal for low power DC/DC conversion. □

A power-on reset output can be monitored by a microprocessor to ensure proper start-up. Internal undervoltage and overvoltage comparators on each output pull the POR output low if either output is not within ±8.5% of its set voltage. The POR output is delayed by 262,144 clock cycles (about 175ms if the switching frequency is left at 2.25MHz) after achieving regulation, but is pulled low immediately once either output falls out of regulation.

Even though the LTC3548 has more features and can handle twice the power of the LTC3547, it still only takes up 140mm² of board space. Figure 11 shows a photo of the demo circuit.

Conclusion

The LTC3410 and LTC3542 provide extremely compact solutions for high efficiency, single channel step-down outputs. The LTC3547 and LTC3548 generate dual channel step-down outputs. All of these converters require a minimal number of external components and are available in small packages to reduce the required board real estate. □

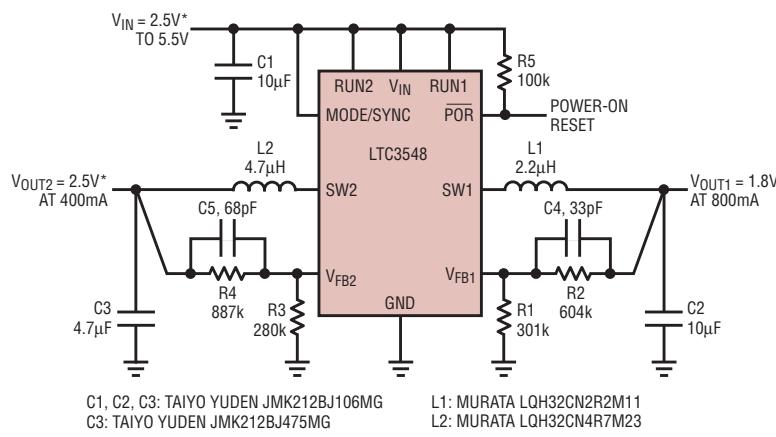


Figure 10. Dual output step-down application yields 1.8V at 800mA and 2.5V at 400mA.



Figure 11. Less than 150mm² is needed for two DC/DC converters (LTC3548).

Digitally Control the Operating Frequency of Switching Regulators that Have No Sync Function

by Tom Gross

Introduction

Certain applications require on-the-fly adjustment of a switching regulator's operating frequency to avoid interference or to match a system clock. Programming the operating frequency is easy if the switching regulator has a synchronization function (a SYNC pin), but what if there is no sync function? This article shows how to use a DAC to adjust the switching frequency for regulators that have a resistor-set operating frequency.

Figure 1 shows such a circuit. A 10-bit, micropower voltage output DAC, the LTC1669, controls the operating frequency of a 1A output current synchronous step-down switching regulator, the LTC3561. The DAC adjusts the LTC3561's switching frequency over its 500kHz to 4MHz frequency range by driving the LTC3561's SHDN/R_T pin. The DAC output voltage is scaled via R_{DAC} and R_T to match the adjustment range of the LTC3561. In a typical application of the LTC3561 (without DAC control), R_T sets the current out of the SHDN/R_T pin, I_{FREQ} (See Figure 1). If R_T is 324k, I_{FREQ} is about 2.5μA and the nominal operating frequency is 1MHz. I_{FREQ} determines the switching frequency

of the LTC3561 (f_{SW}) according to the relation:

$$\begin{aligned} f_{SW} &= \left(\frac{9.78 \cdot 10^{11}}{R_T} \right)^{1.08} \text{ MHz} \\ &= \left(\frac{9.78 \cdot 10^{11}}{V_{R(T)}/I_{FREQ}} \right)^{1.08} \text{ MHz} \end{aligned}$$

where V_{R(T)} equals 0.8V, the voltage present at the SHDN/R_T pin. Rearranging the equation for I_{FREQ}:

$$\begin{aligned} I_{FREQ} &= \left(\frac{V_{R(T)}}{9.78 \cdot 10^{11}/f_{SW}^{1.08}} \right) A \\ &= \left(\frac{0.8V}{9.78 \cdot 10^{11}/f_{SW}^{1.08}} \right) A \end{aligned}$$

The DAC increases or decreases the current that sets the operating frequency from the nominal value set by R_T. I_{DAC} varies I_{FREQ} from its nominal value by the following relation:

$$\begin{aligned} I_{FREQ} &= I_{DAC} + I_{RT} \\ &= \left[\left(\frac{V_{R(T)} - V_{DAC}}{R_{DAC}} \right) + \left(\frac{V_{R(T)}}{324k} \right) \right] A \end{aligned}$$

Setting the upper frequency limit of 4MHz to correspond to a DAC output of zero, and using the equation above, the value of R_{DAC} calculates to approximately 93.1k. Equating the two I_{FREQ} equations results in:

$$\left(\frac{V_{R(T)} - V_{DAC}}{93.1k} \right) + \left(\frac{V_{R(T)}}{324k} \right) = \left(\frac{0.8V}{9.78 \cdot 10^{11}/f_{SW}^{1.08}} \right)$$

and rearranging for V_{DAC}:

$$V_{DAC} = 0.8V \left\{ 1 + \left[\frac{1}{324k} - \frac{1}{9.78 \cdot 10^{11}/f_{SW}^{1.08}} \right] 93.1k \right\}$$

A 10-bit hexadecimal input code adjusts the DAC output from 0V to 2.5V, in approximately 2.4mV steps. Thus, for a given output voltage, the code in its equivalent decimal form is:

$$\begin{aligned} \text{CODE(DECIMAL)} &= \left(\frac{2^{10}}{2.5V} \right) V_{DAC} \\ &= \left(\frac{1024}{2.5V} \right) V_{DAC} \end{aligned}$$

continued on page 38

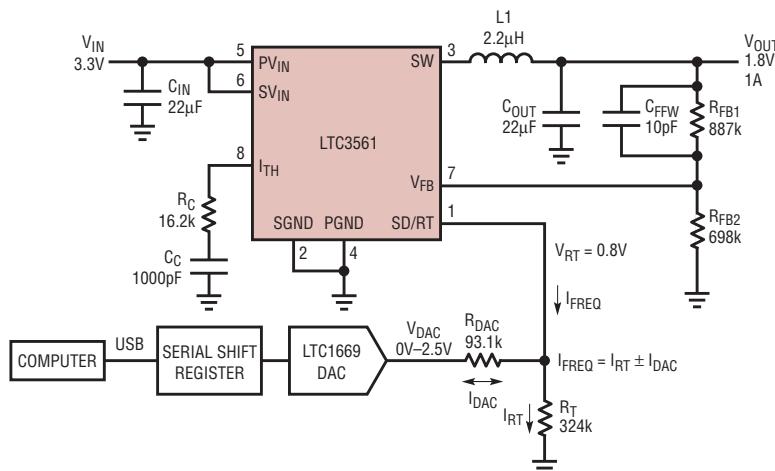


Figure 1. The operating frequency of the LTC3561 switching regulator is digitally controlled

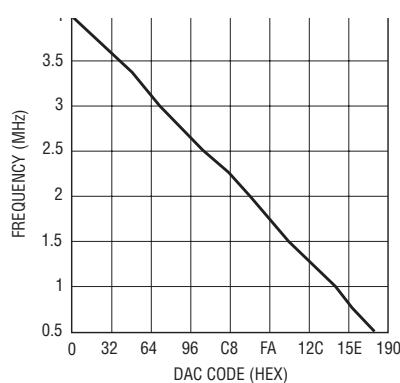


Figure 2. Regulator operating frequency vs DAC code for the circuit in Figure 1

New Device Cameos

SOT-23 Spread Spectrum Clock for Switching Regulators

The LTC6908 is a tiny spread spectrum silicon oscillator optimized for switching regulators. Using a single resistor, the LTC6908 is programmable to any frequency from 50kHz to 10MHz. The LTC6908 comes in two configurations, each with dual outputs. The LT6908-1's outputs are 180° out of phase and the LTC6908-2's outputs are 90° out of phase.

Enabling the pseudo-random spread-spectrum provides a simple, effective way to reduce EMI. In the event that the switcher bandwidth is limited, the LTC6908 modulation rate

can be adjusted to one of three settings. Implementing spread-spectrum for switchers is now trivial—the user sets the frequency with one resistor and selects a modulation rate.

Fully specified over the temperature range of -40°C to 125°C, the LTC6908 offers the same outstanding features available with Linear Technology's silicon oscillator family; rugged and reliable operation under extreme conditions, fast start-up and low power consumption. These parts are available in a compact 6-lead ThinSOT™ package and a 2mm × 3mm DFN.

The gate driver maximum output voltage is clamped to ground with a 12V zener.

The LTC4210-3/LTC4210-4 allows safe board insertion and removal with inrush current control. The LTC4210-3/LTC4210-4 also can be utilized as high side gate driver to control a small footprint logic level MOSFET.

14-Bit 125Msps Low Power Dual ADC Enhances High Efficiency Basestation Transceivers

The LTC2285 is a 14-bit 125Msps dual high-speed analog to digital converter (ADC) with low power dissipation of just 395mW per channel. This high speed device is optimized for use in power efficient, multi-carrier wireless basestation transceiver applications including WiBro and WiMAX standards with performance of 71.3dB SNR and 78dB SFDR at 140MHz. The high sampling rate allows designers to capture wider channel bandwidths, doubling the capacity of existing systems that are typically sampling at 65Msps.

In addition to the 14-bit LTC2285, Linear Technology offers the pin compatible 12-bit LTC2283 and 10-bit LTC2281 125Msps dual ADCs. These three dual ADCs complete a 3V family of 10-, 12- and 14-bit parts ranging from 10Msps up to 125Msps. The pin compatibility offers designers more flexibility during product development, providing a fast and cost-effective

upgrade path for existing designs. The ADCs provide very low crosstalk between channels of -110dB.

The LTC2285 low power family is packaged in a small 9mm × 9mm QFN package. The parts include integrated bypass capacitance and 50Ω series output matching for a small total solution size. They provide the flexibility to choose between two input spans of 1V_{P-P} or 2V_{P-P}. The 125Msps dual ADCs also offer a data-ready clock-out pin for latching the output data buses. The ADCs are optimized for undersampling signals up to 140MHz, and have a wide analog input bandwidth of 640MHz. For downconversion signal chains, Linear Technology recommends the LT5516 direct conversion quadrature demodulator and LT6402 300MHz low distortion/low noise ADC driver.

All three devices are supported with demo boards for quick evaluation and can be purchased online at www.linear.com.

Table 1 provides an overview of the LTC2285 dual ADC product family. All parts are available in optional lead-free packages for RoHS compliance. A table of Linear Technology's entire low power high speed ADC family can be found at <http://www.linear.com/design-tools/hsadcs.jsp>.

36V, 2A, 2.8MHz Step-Down DC/DC Converter Offers 50µA Quiescent Current

The LT3481 is a 2A, 36V step-down switching regulator with Burst Mode operation to keep quiescent current under 50µA. The LT3481 operates within a V_{IN} range of 3.6V to 34V, making it ideal for load dump and cold-crank conditions found in automotive applications. Its 3.2A, 0.18Ω internal switch can deliver up to 2A of continuous output current to voltages as low as 1.26V. Switching frequency is user programmable from 300kHz to 2.8MHz, enabling the designer to optimize efficiency while avoiding critical noise-sensitive frequency bands. The combination of its 3mm × 3mm DFN-10 package (or thermally enhanced

Table 1. Overview of the LTC2285 dual ADC product family

Part	Resolution	Speed (Msps)	Power/Ch. (mW)
LTC2285	14-bit	125	395
LTC2284	14-bit	105	270
LTC2299	14-bit	80	222
LTC2298	14-bit	65	205
LTC2297	14-bit	40	120
LTC2296	14-bit	25	75
LTC2295	14-bit	10	60
LTC2283	12-bit	105	395
LTC2282	14-bit	105	270
LTC2294	12-bit	80	211
LTC2293	12-bit	65	205
LTC2292	12-bit	40	120
LTC2291	12-bit	25	75
LTC2290	12-bit	10	60
LTC2281	10-bit	125	395
LTC2280	10-bit	105	270
LTC2289	10-bit	80	211
LTC2288	10-bit	65	205
LTC2287	10-bit	40	120
LTC2286	10-bit	25	75

MSOP-10E) and high switching frequency keeps external inductors and capacitors small, providing a compact, thermally efficient footprint.

The necessary boost diode, oscillator, control and logic circuitry are also integrated. Output ripple in Burst Mode operation is below 15mV_{P-P} and current mode topology enables fast transient response and excellent loop stability.

LTC3823, continued from page 26

if needed. This pin can also be used as an enable pin. The TRACK/SS pin can either be used to soft start the part (by placing a capacitor to ground) or as a track pin (by connecting it to another regulated supply). If used in tracking mode, the soft start charge current can be turned off at the user's discretion.

A separate V_{IN} sense pin enables the user to run the chip off a different supply than the drain of the top external MOSFET. This configuration allows sensing of the top MOSFET's drain voltage to adjust the on-time, maintaining constant frequency operation. The LTC3823 has an internal 50mA, 5V LDO brought out to the INTCC pin. Connecting this pin to the DRVCC pin supplies the bottom MOSFET driver. If a different voltage is desired, an external LDO up to 7V can be used to supply the DRVCC and/or INTVCC pins.

The LTC3823 can be synchronized to a fixed frequency that is about 50% higher or lower than the set frequency for those applications that require

Low Voltage Current Limiting Hot Swap Controller

The LTC4210-3 and LTC4210-4 are new members to the LTC4210 family of tiny SOT-23 Hot Swap controllers. These two parts are ideal for low voltage applications from 2.7V to 7V where superior current limit response is essential to high performance systems. The LTC4210 rides through short duration of overload transients.

Severe load faults are isolated after a programmable circuit breaker timeout to prevent system and MOSFET damages. The LTC4210-3 retries after circuit breaker timeout, whereas the LTC4210-4 latches off until system reset. 

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tighter frequency control. Separate SENSE⁺ and SENSE⁻ pins make it possible to sense the inductor current through the R_{DS(ON)} of the bottom FET, eliminating the sense resistor for applications demanding the highest possible efficiency. However, for applications where maximum accuracy is desired, the user can choose to use an external sense resistor. The current limit can be set externally via the VRNG pin. The top FET turn-off to bottom FET turn-on dead time can be adjusted via the Z0 pin, improving efficiency by minimizing bottom FET body diode conduction losses.

Protection

The LTC3823 features an accurate UVLO circuit with over 750mV of hysteresis, enabling the use of less bulk capacitance on V_{IN}. The UVLO monitors all internal supply rails, including INTV_{CC}, for added protection. A thermal shutdown circuit protects the regulator in situations when adequate heat-sinking or airflow has not been provided.

The Power Good monitor sets the PGOOD flag when the output is out of regulation. A timer prevents the occurrence of false power good signals. The OV portion of this circuit turns on the bottom MOSFETs to protect the load from high voltage damage. The current limit has a foldback feature that reduces the limit as the feedback signal falls. This is primarily designed to protect the switcher from hard shorts. During start-up, the current limit foldback is disabled, as it may interfere with tracking. The soft start circuit limits in-rush currents at start-up.

The constant on-time architecture allows the controller to respond quickly to load current changes, thus limiting the output voltage swing seen by the load.

Conclusion

The LTC3823 packs many critical power management functions into a single IC, offering a high level of flexibility and protection in low output voltage applications. 

DAC Control, continued from page 36

Substituting the previous solution for V_{DAC} results in an equation for the corresponding DAC decimal code of a given switching frequency:

$$\text{CODE(DECIMAL)} = \left(\frac{1024}{2.5V} \right) \cdot 0.8V \left\{ 1 + \left[\frac{1}{324k} - \frac{1}{9.78 \cdot 10^{11}} \frac{1}{f_{SW}^{1.08}} \right] 93.1k \right\}$$

Which simplifies to:

$$\text{CODE(DECIMAL)} = 421.84 - \frac{3.12 \cdot f_{SW}^{1.08}}{105}$$

For example, suppose the desired operating point is 2MHz. Input 2MHz into the above code equation to produce a code of 223, or hex code 0DF for the input to the DAC.

Figure 2 shows a graph of the DAC code vs switching frequencies. 

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