LINEAR TECHNOLOGY

IN THIS ISSUE...

Reduce Charge Time for High Capacity Li-Ion Batteries with 2A Continuous Charging

by Tom Hack

Introduction

The latest high capacity Li-Ion batteries meet the needs of power hungry portable devices, but they also increase the demands placed on battery chargers—demands that can be too much for a standard linear charger. For instance, a linear charger, operating at 1A charging current, charges a 1Ahr battery to 70% capacity within one hour, and fully charges it within three hours. Newer 2Ahr batteries need twice that current in order to be fully charged in the same amount of time. The problem is that a linear charger operating at 2A produces too much heat for continuous charging—it's just too inefficient. The LTC4001 solves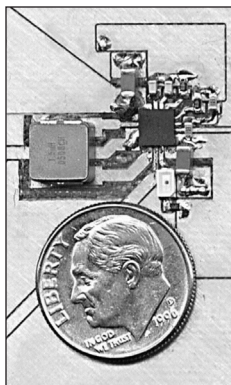 this problem by incorporating a high efficiency PWM charger to perform continuous 2A battery charging. It works with both standard and current-limited wall adapters—where the latter lowers battery charger dissipation and operating temperature.

Big Features; Small Footprint

A full-featured battery charger based on the LTC4001 requires an area not much larger than a dime (Figure 1). Fully programmable timer and charge rate terminations are included. Automatic battery "topping off" is also included. Filtering prevents accidental recharge from occurring in noisy environments (such as found in GPRS cellular phones). The LTC4001 works readily with NTC thermistors for battery temperature sensing. Remote battery sensing is included. Soft-start is fully programmable. The LTC4001 also drives charge status LEDs and provides logic signals for microprocessor-based designs.

The LTC4001 is tiny, fitting into a 4mm × 4mm package, but other factors also contribute to the charger's small footprint. High operating frequency (1.5MHz) reduces the size of the inductors and capacitors. Input short circuit blocking is built in so no external diode is required. Current

Figure 1. A typical LTC4001-based Li-Ion battery charger occupies minimal board real estate.

Issue Highlights

High capacity Li-Ion batteries meet the needs of power hungry portable devices, but they also increase the demands placed on battery chargers—demands that can be too much for a standard linear charger. The problem is that a linear chargers cannot efficiently produce enough current to quickly charge the battery. To solve this problem, the LTC4001 uses a high efficiency PWM charger, which makes continuous 2A battery charging practical. See our cover article for more about this breakthrough device.

Featured Devices

Below is a summary of the other devices featured in this issue.

The LT6411 **selectable-gain differential amplifier/ADC driver** is a good fit for power-critical high-speed signal chain applications. It can produce gains of 1, –1, and 2 with *no external components.* The dual amplifiers inside the LT6411 allow for easy single-ended-to-differential conversion for driving high-speed analog-to-digital converters (ADC). (Page 6)

The LT3486 **dual LED string driver** has two 1.3A channels with high PWM dimming capability in a small 5mm × 3mm DFN package. Since both channels' power switches are included in the IC, the circuit is simple and small. (Page 10)

The LTC4215 combines a robust **Hot Swap circuit with an I^{2}C interface and data converter** to allow power monitoring as well as hot-plug functionality and fault isolation. (Page 12)

The LTC3532 is a **300mA buck-boost converter**, which incorporates automatic Burst Mode® operation, adjustable switching frequency, and integrated soft-start. The LTC3532 is ideal for miniature disk-drive applications or any application that requires high efficiency over a wide range of output currents and input voltages. (Page 16)

The LTC2909 is a **highly customizable power supply monitoring solution** with adjustable input thresh-olds, input polarity selection, a multi-mode reset timer, and an open-drain $\overline{RST}$ output. Adjustable input thresholds allow the user to set any trip threshold for the comparator, subject only to the accuracy limitations of the part, instead of having to pick from a factory-set limited collection of thresholds. (Page 19)

The LTC2859 and LTC2861 combine a **logic-selectable integrated termination resistor with a rugged 20Mbps RS485/RS422 transceiver**, providing a single die impedance-matched network solution in a tiny package. (Page 27)

The LTC3026 is a **1.5A VLDO with input voltage capability down to 1.14V** and a low adjustable output voltage from 0.4V to 2.6V. The part also has a very low dropout voltage of only 100mV while delivering up to 1.5A of output current, enabling it to optimize battery run time from single cell applications with a low V_{IN} to V_{OUT} differential. (Page 30)

Design Ideas and Cameos

Design Ideas start on page 33, including a better way to combine battery packs using an ideal diode—improving on standard diode solutions. *LT*

Linear Technology in the News...

Linear's LTM4600 µModule Racks Up Awards

Linear Technology's LTM4600 10A DC/DC µModule™ has been steadily collecting major product awards. The LTM4600 was recently awarded *EDN* magazine's Innovation of the Year award in the Power Systems and Modules category, and named Product of the Year by both *Electronic Products* magazine and *AnalogZone.* To that, add an Innovation Award from China's *Electronic Engineering & Product World*, and a Product Award from *EDAW/ Nikkei Electronics* in China.

Linear Expands Design Capabilities

Linear Technology has further expanded its analog design resources in two new locations. In April, Linear announced the opening of the company's first European design center in Munich.

Lothar Maier, CEO of Linear Technology, stated, "We are excited to open our first European design center in the key technology hub of Munich. By centrally locating our latest design center on the European continent, this puts us in an even better position to develop the right products for our major European customers. Germany represents an ideal location for our first European design center, with its focus on the automotive, industrial and communications markets."

And in May, the company opened a new design center in Dallas. With this announcement, Linear now has 12 design centers worldwide.

Linear Announces Digital Power Alliance

Linear Technology has announced an alliance with Primarion, Inc. to produce digital point-of-load (POL) products for the networking, computing and telecommunications infrastructure markets. Linear's LTC7510 digital POL controller is the first in a family of PMBus-compliant digital DC/DC controllers. This family of digital power management and conversion products provides flexible, system-level control of sophisticated power management systems. These controller ICs ensure seamless power management in high availability systems, enabling real-time programming and monitoring of key parameters to optimize performance and maximize system uptime. *LT*

sensing is internal, so there is no need for an expensive milliohm-sized current sense resistor.

Inside the LTC4001

The LTC4001 is the basis for a complete 2A Li-Ion battery charger (Figure 2). A 50mA linear charger provides cell conditioning while a synchronous buck charger provides constant-current/constant-voltage high rate charging (up to 2A). Protection and lockouts guard against a variety of events including: shorts at the battery and wall adapter inputs; improper programming of the charge current; open battery and/or over-voltage battery; defective battery; insufficient wall adapter voltage; chip over-temperature; battery over- or under- temperature.

A Bare Bones Charger

Figure 3 shows a bare bones 2A battery charger. With only five additional components, this charger offers a high efficiency, high power solution. This implementation leaves out status indicator lights, battery temperature monitoring, and a timer (which may be provided by a microprocessor). In place of a timer, charge terminates when charge current drops below one-tenth the high rate charge current (an
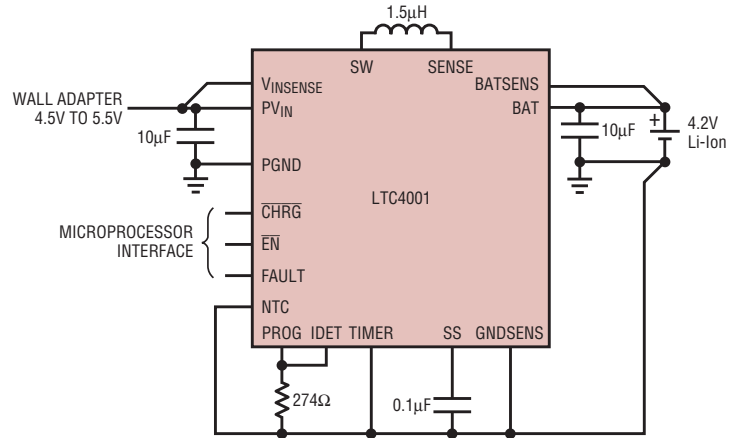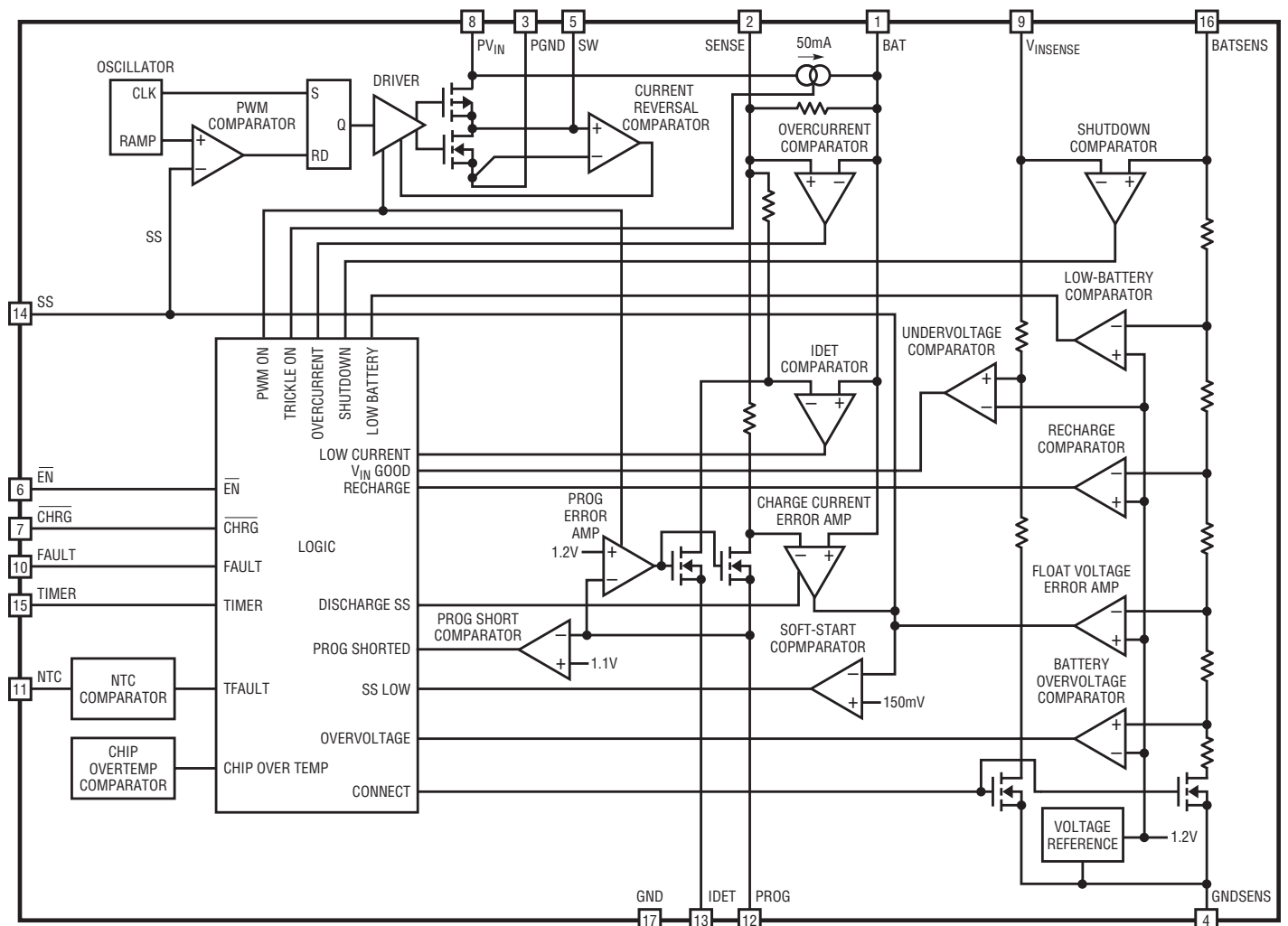

Figure 3. A bare bones battery charger

Figure 2. Simplified block diagram of the LTC4001 Li-Ion battery charger
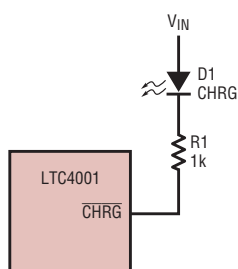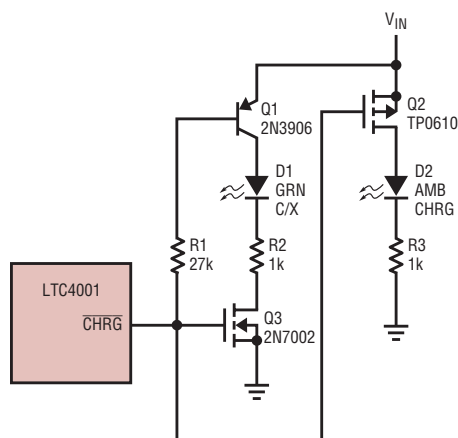
Figure 4. A simple status indicator

Figure 5. Full featured status indication

I_{DET} threshold equal to 200mA in this case). Internal charge termination may be completely defeated by connecting the timer pin to the I_{DET} pin instead of ground (allowing a microprocessor complete control of charge termination).

Adding Status Lights

The $\overline{CHRG}$ pin indicates a variety of charger states (Table 1). Adding a resistor and LED in series with this pin to V_{IN} (Figure 4) indicates charger off (LED off), high rate charging or cell conditioning (LED on continuously at high brightness), and battery temperature out of range/NTC fault (LED blinking).

The LED also indicates when the battery is nearly full charged. As the battery approaches the float voltage and charge current drops below the I_{DET} threshold the LED is dimly lit. This is difficult to see, so a better approach uses two LEDs to indicate all charger states (Figure 5).

Interfacing with Microprocessors

The interface in Figure 6 can distinguish between all states available on the $\overline{CHRG}$ pin. To detect cell conditioning or high rate charging, force the digital output pin, OUT, high and measure the voltage on the $\overline{CHRG}$ pin. The N-channel mosfet pulls $\overline{CHRG}$ low even with a 2k pull-up resistor. Near end of charge, the NMOS turns off, and $\overline{CHRG}$ sinks only 30µA. The IN pin is pulled high by the 2k resistor connected to OUT. If OUT is placed into a high impedance state, the 30µA sink current from the $\overline{CHRG}$ pin pulls IN low. When charging stops, $\overline{CHRG}$ opens and OUT stays high, even with a 390k pull up resistor.

If a battery temperature fault occurs during high rate charging, the $\overline{CHRG}$ pin blinks using a serrated pulse pattern. Nominal timing of this pattern is given in Figure 7. The extra edges provide rapid indication to a microprocessor and may be used to drive a microprocessor interrupt line for low processor overhead, but still provide for a visible fault indication when using LEDs.

Battery Temperature Sensing

By adding one resistor and one thermistor, battery temperature sensing may be included. The LTC4001 is designed for Vishay Dale's "R-T Curve 2" thermistors, but any thermistor with an R_{COLD}-to-R_{HOT} ratio of about 7 will also work. If battery sensing is not needed, the NTC pin is grounded.

Operation with Conventional and Current Limited Wall Adapters

Wall adapters with or without current limiting may be used with the LTC4001, but the lowest power dissipation battery charging occurs with a current limited wall adapter. To use this feature, program the LTC4001 above the wall adapter current limit. For example, if the wall adapter current limit is 2A, set the LTC4001 charge current slightly higher than 2A (allowing for tolerances).

To understand operation with a current limited wall adapter, assume battery voltage V_{BAT} is initially below V_{TRIKL}, the trickle charge threshold (Figure 8). Battery charging begins at approximately 50mA, well below the wall adapter current limit so the voltage into the LTC4001 (V_{IN}) is the wall adapter's rated output voltage ($V_{ADAPTER}$). Battery voltage rises eventually reaching V_{TRIKL}. The linear charger shuts off and the PWM (high rate) charger turns on using soft start. Battery charging current rises during the soft-start cycle causing a corresponding increase in wall adapter load current. When the wall adapter reaches current limit, the wall adapter output voltage collapses, and

Figure 6. A microprocessor interface

Table 1. $\overline{CHRG}$ Behavior

Charger State	$\overline{CHRG}$ Behavior
Not charging	Open
High rate Charging and $I_{BAT}>I_{DET}$ Or cell conditioning	NMOS turned on pulling pin low
High Rate Charging and $I_{BAT}<I_{DET}$	30µA pull down current
NTC temperature fault while charging at $I_{BAT}>I_{DET}$	Blink

Figure 7. $\overline{CHRG}$ temperature fault waveform

Figure 8. Idealized charging behavior

the LTC4001 PWM charger duty cycle ramps up to 100% (the top-side PMOS switch in the LTC4001 buck regulator stays on continuously.) As the battery voltage approaches V_{FLOAT}, the float voltage error amplifier commands the PWM charger to deliver less than I_{LIMIT}. The wall adapter exits current limit and V_{IN} jumps back up to $V_{ADAPTER}$. Battery charging current continues to drop as V_{BAT} rises, dropping to zero at V_{FLOAT}.
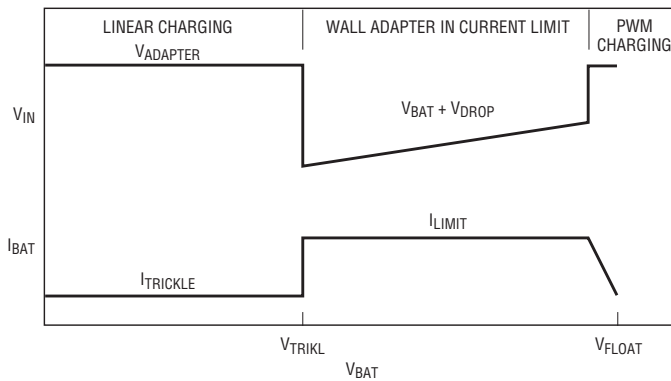
Because the voltage drop in the LTC4001 is very low when charge current is highest, power dissipation is also very low.

Low Dissipation

Trickle charging uses a linear charger but low charge current produces low power dissipation, typically 256mW ($V_{IN} = 5V$, $V_{BAT} = 0$). High rate charging uses a high efficiency buck switcher and total charger dissipation is approximately 1.2W at 2A (Figure 9). High rate charging with a current limited wall adapter produces even lower charger dissipation (537mW at $V_{BAT} = 4.2V$ with a 2A current limited wall adapter) because there is very little voltage drop for the battery charging path inside the LTC4001.
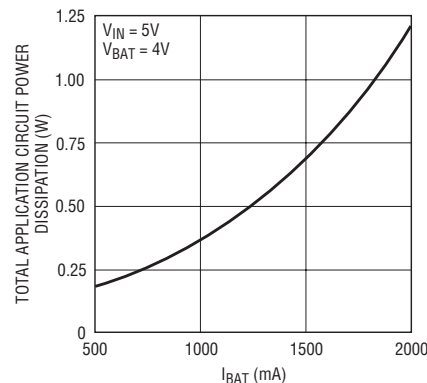

Figure 9. High rate charger power dissipation

So how does LTC4001 dissipation stack up against a 2A linear charger? Most of a linear charger's dissipation occurs in the series pass element so the dissipation is approximately equal to the voltage drop in the pass element times the charge current. Worst case dissipation occurs at the lowest battery voltage where high rate charging occurs (to make a valid comparison to the LTC4001 this would be 2.85V). For a 5.0V input, this translates into a dissipation of 4.3W! Higher input voltage makes the situation even worse.

A Charger with All the Bells and Whistles

A full featured battery charger is shown in Figure 10. It includes a three hour timer, battery temperature monitoring, programmable charge and I_{DET} currents, remote sensing, and status lights. A fault light has been included that indicates when a shorted battery is detected or when the battery is out of normal temperature range.

Conclusion

The LTC4001 sets a new standard for small, low parts count, full-featured, high efficiency Li-Ion battery chargers. Low power dissipation makes continuous 2A battery charging practical, cutting dissipation to approximately one fifth the dissipation of a straight linear charger. ☰
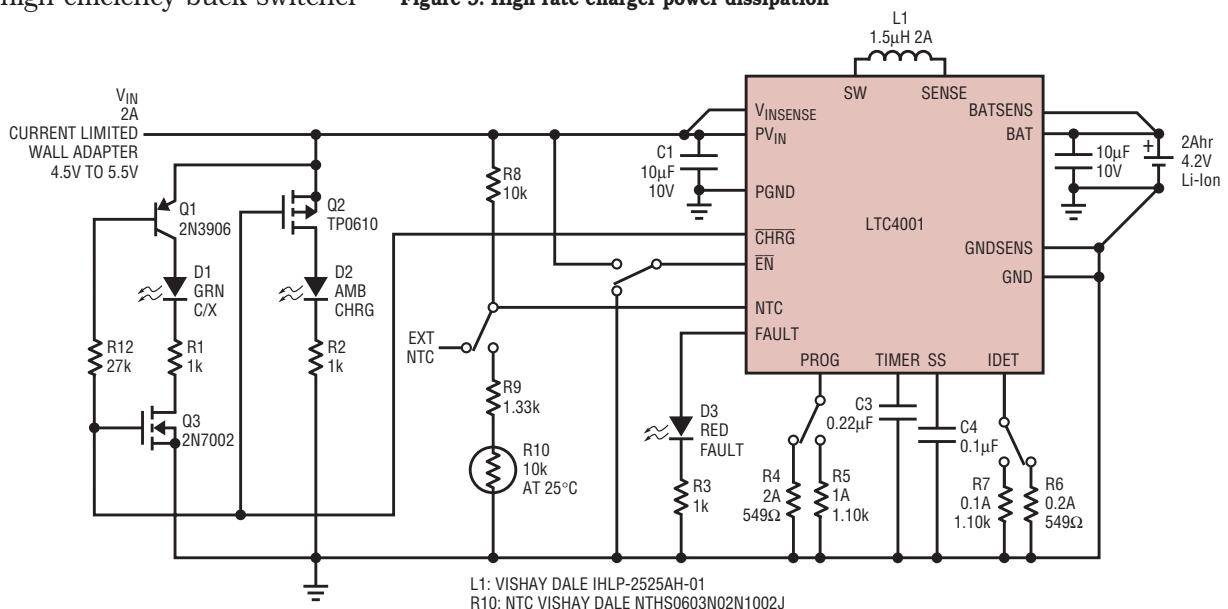

Figure 10. A full featured battery charger

650MHz Selectable-Gain Amplifier/Differential ADC Driver Has Small Form but Many Functions

by Cheng-Wei Pei

Introduction

Operational amplifiers have always been an important part of an analog designer's bag of tools. Even the most basic tools can be improved, and some recent advancements have increased the utility of the workhorse op amp. For instance, the advent of dual and quad op amp packages allowed engineers to produce a variety of applications from a single device. The new LT6411 selectable-gain differential amplifier/ADC driver makes a good thing better by adding internal gain and feedback resistors and an easy-to-use flow-through pin layout.

The LT6411 can produce gains of 1, –1, and 2 with *no external components*. The dual amplifiers inside the LT6411 allow for easy single-ended-to-differential conversion for driving high-speed analog-to-digital converters (ADC). The wide bandwidth (650MHz), low distortion (–77dBc harmonic distortion at 30MHz) and high slew rate (3300V/µs) preserve signal fidelity even at high frequencies, while the low supply current (16mA total) enables the LT6411 to be used in power-critical high-speed signal chain applications. Form factor is also not an issue—all of these features fit

> *When blazing fast speed, low power, or the flexibility of selectable gains is necessary, a circuit designer need only reach into his bag of tools and pull out the LT6411.*

into the tiny 3mm × 3mm 16-pin QFN package. When blazing fast speed, low power, or the flexibility of selectable gains is necessary, a circuit designer need only reach into his bag of tools and pull out the LT6411.

Internal Topology and Gain Selection

The LT6411 contains two internal current-feedback amplifiers with matched feedback and gain resistors. The integrated 370Ω resistors take the guesswork out of selecting the optimal feedback resistor for good AC response. Another common source of frustration with current-feedback amplifiers is maintaining a tight PC board layout to prevent excessive capacitance at the inverting input node. This node is

internal to the LT6411, so the layout of the circuit board is nearly optimal from the get-go.

Selecting between the basic gain configurations of the LT6411 is a simple matter of pin strapping. Figure 1 shows the standard non-inverting gain of 2 configuration, with differential input and output. Figure 2 shows the non-inverting gain of 1 configuration, and Figure 3 shows an inverting gain of 1 configuration. Note that in Figure 2, the non-inverting inputs are tied together with the gain resistor. In theory, the gain resistor pin could be left floating, but in practice, the parasitic pin and pad capacitances cause the gain to peak up to 6dB at high frequencies, which causes excessive ringing in the transient response. At 650MHz (the bandwidth of the LT6411), 2pF of parasitic capacitance has a reactance of just 122Ω.

Single-Supply Operation and Level-Shifting

The LT6411 operates on a wide supply voltage range, from 4.5V–12.6V. Figures 1–3 show the part with dual supplies. However, the LT6411 performance remains excellent with a
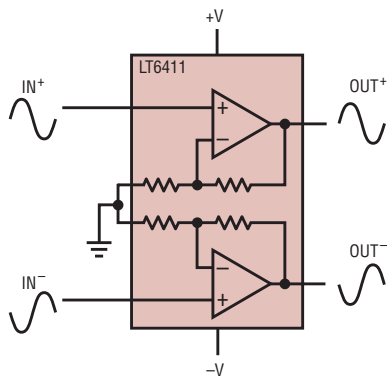

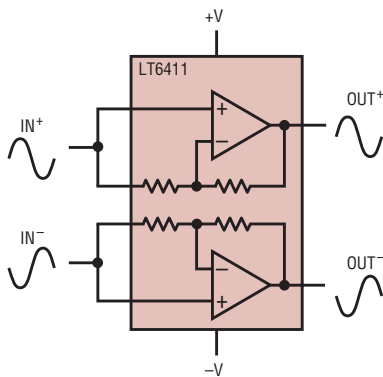Figure 1. Standard non-inverting gain of 2 configuration, shown with split supplies.

Figure 2. Standard non-inverting gain of 1 configuration, shown with dual supplies. The gain resistors, shown here tied to the inputs, should *not* be left floating (see text).
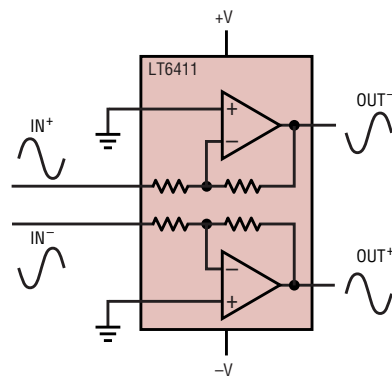

Figure 3. Standard inverting gain of 1 configuration, shown with dual supplies.

single supply, especially important for many common applications such as high speed ADC driving, where only a single supply is available. If the input signals into the LT6411 are already DC level-shifted above ground so that the input and output common-mode ranges are met (the input operates to within 1V of the supplies, the output with a 1k load swings to within 1.3V of the supplies), no additional work needs to be done. However, if the input signal is AC-coupled or centered around ground, then level-shifting is necessary. This section presents some of the methods to level-shift the input and output voltages of the LT6411.

Figure 4 shows a center-tapped transformer providing the DC voltages necessary for single-supply operation. The input signal (shown single-ended, but can also be differential by driving the other end of the primary) is provided through the transformer primary, and the transformer secondary presents a balanced signal to the LT6411. If the input signal is coming from a 50Ω signal source, the two 24.9Ω resistors provide the appropriate termination. Note that the DC voltage provided at the center tap of the transformer is $V_{O(DC)}/2$, due to the fact that the LT6411 has a non-inverting gain of 2. Alternatively, the two gain resistor pins could be AC-coupled to ground through a capacitor, and the DC gain of the LT6411 would be unity. The total differential gain in this configuration is 2; if a differential gain of 1 is desired, simply tie the gain resistor pins to the corresponding non-inverting input pins.

For a single-ended input, Figure 5 shows a simple AC-coupled method of providing the correct input and output DC levels for the LT6411. The input is AC-coupled through a large capacitor (typically 0.1µF or larger), and the total differential gain at the output is +2. The DC voltage is provided through the second non-inverting input.

In the case of a differential inputs, Figure 6 shows a similar configuration, where the two inputs are AC-coupled to the LT6411, and the DC level is provided through two resistors. The choice of size for the resistance R is
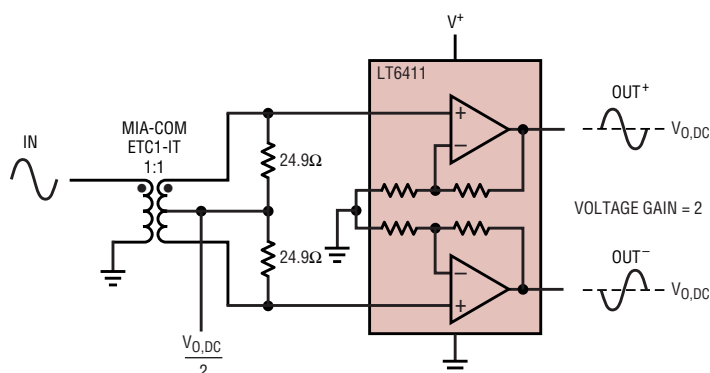

Figure 4. DC level-shifting through the center tap of a transformer for single-supply operation. The DC voltage source at the center tap should serve as a low-impedance AC ground. The two 24.9Ω resistors provide a 50Ω termination, if necessary.

$$f_{CO} \approx \frac{1}{2\pi \cdot 370\Omega \cdot C_{LARGE}}$$

Figure 5. A simple AC-coupled method of providing DC level shifting. The DC voltage source, which does not need to have a low impedance, is provided at the non-inverting input of the second amplifier. The AC input impedance of this circuit is 370Ω.

$$f_{CO} \approx \frac{1}{2\pi C_{LARGE} \cdot R}$$

Figure 6. A DC voltage source and two resistors sets the DC input and output level of the LT6411. The two resistors must be large enough not to overload the inputs.

a trade-off between loading the differential inputs (for smaller values of R) and increasing voltage offset and noise (for larger values of R, due to the input bias currents and current noise in the non-inverting inputs). Values of R up to 10k work well in practice. The DC and AC inputs both have a gain of 2 in this configuration. For a gain of 1, simply tie the gain resistor pins to the corresponding non-inverting input pins.

Figures 4–6 all have a lower frequency limitation, defined by either the transformer's magnetizing inductance or the size of the AC-coupling capacitor. What if voltage level-shifting and response down to zero Hertz

Figure 7. Two identical resistive dividers (R_A and R_B) shift the DC level of the input (and output) to within the limits of the LT6411. $V_{O,DC}$ must have a low impedance at the frequencies of interest, and must be capable of sourcing and sinking currents through the internal resistors.

$$GAIN_{DIFF} = 2 \cdot \frac{R_B}{R_A + R_B}$$

$$V_{O,DC} = \frac{(V^+ \cdot R_A) + (V_{I,DC} \cdot R_B)}{R_A + R_B}$$
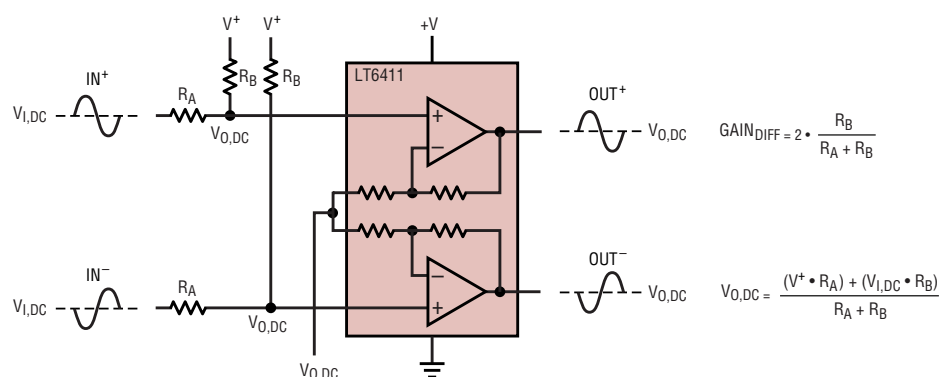
(DC) is necessary? Figure 7 shows one method of level-shifting and maintaining signal response down to DC. The R_A-R_B resistive dividers set the input common-mode voltage for the LT6411. Choose R_A and R_B so that the common-mode voltage at the input of the LT6411 is the same as $V_{O(DC)}$, the desired output common-mode voltage. The gain of the LT6411 with the resistive divider is $2 \cdot R_B/(R_A + R_B)$, and this circuit has a response to DC. If the input DC common-mode level ($V_{I, DC}$) is greater than the $V_{O,DC}$, then R_B should be attached to ground instead of the positive supply.

High Speed ADC Driving

Modern high resolution, high speed ADCs typically feature differential inputs with capacitive sample-and-hold circuits. Driving these inputs requires an amplifier with high bandwidth, fast settling times, good transient response and good distortion performance at the frequencies of interest. The figures in this section show various configurations designed to get the maximum performance out of the LT6411.

The LT6411 is optimized for driving 12-bit and 14-bit high-speed converters such as the 14-bit, 80Msps LTC2249. For many applications, the only additional recommended interfacing components would be small series resistors, to help isolate the LT6411 from the ADC's capacitive input. Figure 8 shows the simplest configuration for the LT6411 driving the ADC. The single supply operation means that either the inputs must have an appropriate DC common-mode level

(that is, V_{DC} in the Figure 11 must be appropriate for both the amplifier and the ADC), or that one of the techniques from the previous section must be used to level-shift the input.

If dual supplies are used for the LT6411, then interfacing with a single-supply ADC might require AC-coupling at the output, as shown in Figure 9. The DC level of the ADC input is established by V_{DC} and two 499Ω resistors.

Figure 8's circuit, though simple, has an important drawback: all of the wideband noise of the amplifier couples into the ADC input, and thus degrades the signal-to-noise ratio (SNR) of the signal. In most cases, the input signal is limited to some frequency band less than the DC–650MHz bandwidth of the LT6411; thus, any extra bandwidth beyond that introduces unnecessary noise. The simplest approach to fixing this problem aptpears in Figure 10. A single shunt capacitor creates an RC lowpass filter that limits the noise bandwidth of the amplifier output, improving the SNR.

For an even sharper cutoff lowpass filter, Figure 11 shows a more involved approach. The inductors and capacitor create a second-order lowpass filter, with R1 ensuring that the frequency peaking is not excessive. R2's primary function is to ensure that the ADC

Figure 8. LT6411 shown interfacing to an ADC. A small series resistance is recommended to isolate the ADC's capacitive input.

$$f_{CO} \approx \frac{1}{2\pi \cdot 499\Omega \cdot C_{LARGE}}$$

Figure 9. Level-shifting the LT6411 output to within the input common-mode range of the single-supply ADC. The resistors used must be large enough not to excessively load the outputs of the LT6411. Some ADCs have a V_{CM} output that can be used for V_{DC}.

inputs do not see too high of a source impedance. Figure 12 shows some sample values, configured for a cutoff of around 50MHz and almost no gain peaking near the cutoff frequency. Simulating the filter is a good way to determine optimal component values, especially when taking into account the series resistance of the inductor and component tolerances.

Some high frequency applications contain very narrow-band signals, where a bandpass filter would provide the best noise limiting, and thus the highest SNR. Figure 13 shows 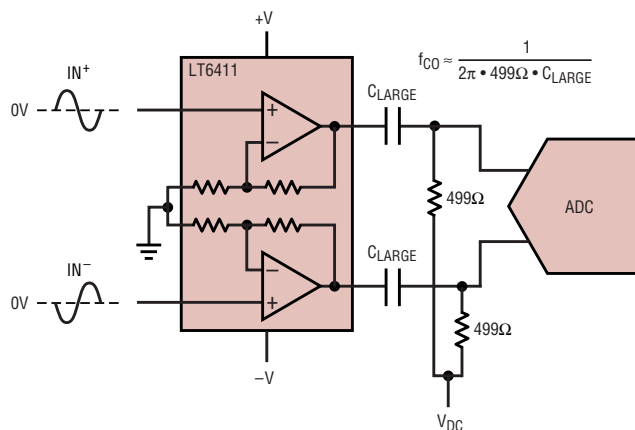a simple RLC bandpass filter. The value of R determines the quality factor (Q) of the filter—the larger the resistor value, the more narrow-band the filter. This comes at the cost of more pass-band loss (depending on the parasitic components of the inductor and capacitor) and higher sensitivity to component tolerances and variations. As the bandpass filter gets narrower, a small shift in center frequency can significantly attenuate the desired output signal.

Conclusion

The LT6411 selectable-gain amplifier/ADC driver is extremely flexible, featuring a multitude of possible configurations with a minimal number of external components. The LT6411 is a dual op amp, a differential ADC driver, and a selectable-gain amplifier all in one. In most basic dual op amp functions, all that is required is a few power supply bypass capacitors to get excellent AC performance to over 600MHz. In addition, the LT6411 comes in the tiny 3mm × 3mm 16-lead QFN package and consumes only 16mA total. The LT6411 also has a shutdown feature that reduces the power supply current to 700µA total. With this unique set of features, the LT6411 can provide myriad functions without breaking the budget of size or power. LT

for the latest information on LTC products, visit www.linear.com

Figure 10. An RC lowpass filter is shown at the output of the LT6411. Reducing the bandwidth at the output also reduces the amount of wideband noise seen by the ADC.

$$f_{CUTOFF} = (4\pi RC)^{-1}$$

Figure 11. A second-order LC lowpass filter, offering a flatter passband and sharper stopband rolloff than an RC filter. The series resistor R1 controls peaking near the cutoff frequency, and the parallel resistor R2 ensures that the ADC sees a low source impedance at very high frequencies.

$$f_{CUTOFF} > \left(2\pi\sqrt{2LC}\right)^{-1}$$
$$R1 + R2 \leq 150\Omega$$

Figure 12. The circuit of Figure 11, configured for a cutoff frequency of around 50MHz.

Figure 13. For narrow-band applications, an LC bandpass filter does an excellent job limiting the noise at the ADC input while maintaining a low component count. The series resistor affects the width of the pass-band and the pass-band attenuation.

$$f_{CENTER} = (2\pi\sqrt{LC})^{-1}$$

Dual Step-Up Converter Drives White LEDs with 1000:1 PWM Dimming

by Keith Szolusha

Introduction

Notebook computers, large-screen handheld PDAs, dashboard displays, and automotive and avionic in-cabin entertainment LCD panels are illuminated with strings of high power white LEDs. White LEDs are the preferred over other backlight technologies because they provide true white light at a high enough intensity 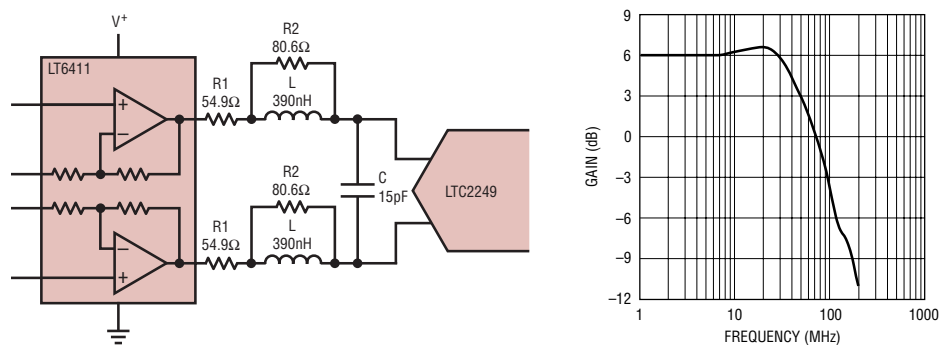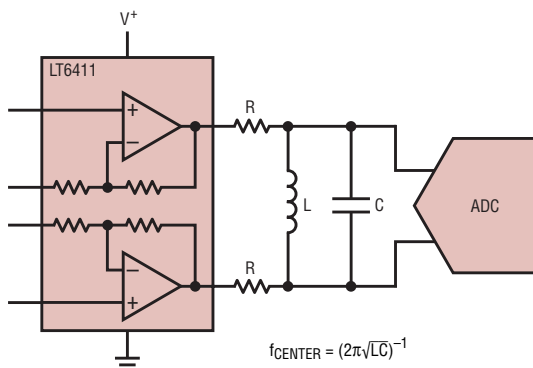for daylight viewing, and enough dimming capability for nighttime use. LEDs also offer relatively long life spans and a lack of hazardous materials. LED strings lining the edges of these LCD panels provide uniform brightness when driven with a constant current.

The maximum switch voltage of an LED driver limits the number of LEDs that it can drive in series. It may seem that paralleling LEDs is a good way to increase the capacity of a driver IC, but parallel LEDs must be well-matched in forward voltage; otherwise un-matched LED strings cause uneven currents and thus uneven brightness. LEDs can be specially sorted (binned) for matching characteristics, but this increases cost.

A better solution is a dual channel LED driver to drive two strings of LEDs. This saves the space and cost of duplicating components, such as the driver IC and input capacitors. Each string is driven with the same regulated constant current, thus providing uniform brightness.

One IC that has these features is the LT3486 dual LED string driver, which has two 1.3A channels with high PWM dimming capability in a small 5mm × 3mm DFN package. Since both channels' power switches are included in the IC, the circuit is simple and small.

Dual LED String Step-Up Driver

The LT3486 is a dual step-up LED driver. Each channel has an efficient, low side 1.3A npn power switch with low V_{CE}(sat) of 300mV (at 750mA switch current). The IC is designed to drive a string of LEDs from a wide input voltage range. Each LED string total voltage can be as high as 38V in a typical application, but may be limited to 34V if the overvoltage protection (OVP) pin is used to protect the switch when the LED string is open.

Figure 1 demonstrates the LT3486 as a dual LED string step-up converter driving a total of 16–20 white LEDs at 100mA from a 4V–16V input voltage range source. The total voltage of the LEDs cannot exceed 34V. The circuit is kept small and simple with the single ceramic input capacitor and two small ceramic output capacitors. With a high 800kHz switching frequency, the inductors and capacitors can be small in size while the efficiency of the circuit remains high, as shown in Figure 2. As PWM duty cycle is decreased from 100%, the circuit efficiency drops slightly, but remains high during the PWM on-time. Not only is the operating efficiency high, but the converter shutdown current

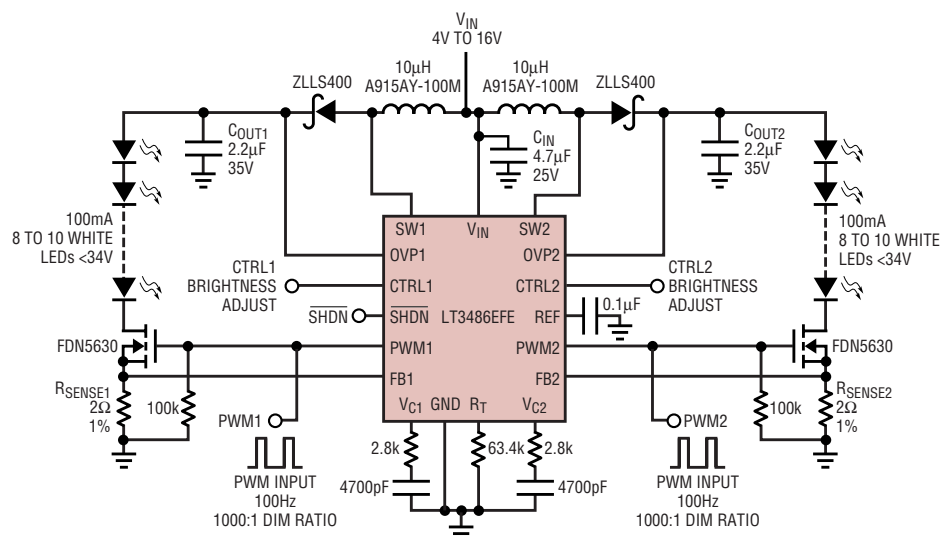**Figure 1. LED driver uses 4V–16V input to drive two strings of eight-to-ten 100mA LEDs (less than 34V total in string) with 1000:1 PWM dimming**
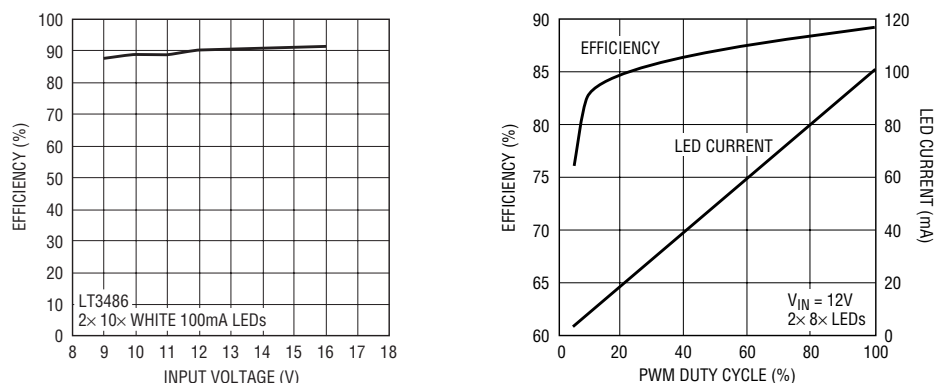

Figure 2. Efficiency of the circuit in Figure 1 and efficiency as a function of PWM duty cycle
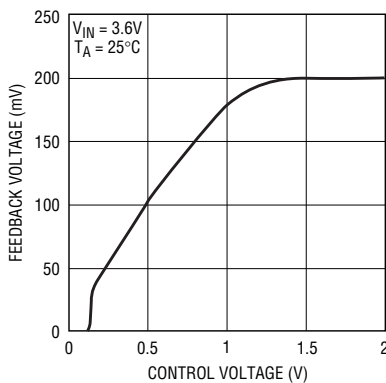
Figure 3. FB pin voltage vs CTRL pin voltage

$V_{IN} = 12V$ 0.2ms/DIV
8/8 LEDs
PWM FREQ = 1kHz

Figure 4. PWM dimming waveforms

consumption is less than 1µA (typically 100nA), merely sipping from the battery when the IC is off.

1000:1 PWM Dimming and 10:1 Brightness Control

As shown in Figure 1, LED brightness can be controlled on the LT3486 with an analog voltage input to the CTRL pin or a digital PWM signal to the gates of the PWM dimming MOSFET and the PWM pin. Analog brightness control reduces the LED current from 100mA to a lower value by reducing the internal sense resistor voltage. Although this is a simple way to decrease the brightness of the LED, the accuracy of the LED current control is reduced and the chromaticity of the LED changes at lower currents. The graph in Figure 3 displays the LT3486 typ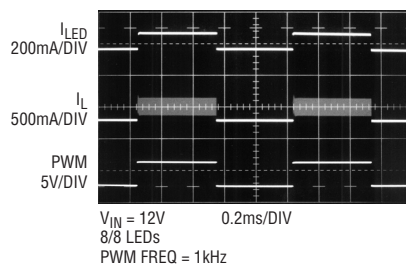ical FB pin voltage dropping as a function of CTRL pin voltage. The low 200mV FB pin (and current sense voltage) accuracy is typically 3% at full current with the CTRL pin pulled high (above 1.5V) but as the CTRL pin voltage is lowered to 150mV, the FB pin voltage is also reduced to about 40mV. Below this 5:1 dimming ratio, the LEDs are turned off as the CTRL pin voltage is pulled below 75mV.

Another method of reducing the brightness of the LEDs is digital PWM dimming. The PWM MOSFET in series with the LEDs creates the waveform shown in Figure 4 when the string of LEDs is PWM'd at 100mA constant current. During PWM on-time, the current is a well-regulated 100mA. During PWM off-time, the current is zero. Because the current is either 100mA or zero, the LED color is preserved as if the LED were driven by a constant 100mA current. Dimming is simply a function of the average, instead of instantaneous, current. The advanced PWM function in the LT3486 is particularly fast in returning the LED to its programmed LED current. Its short minimum dimming on-time (10µs on-time) allows a 1000:1 digital PWM dimming ratio with 100Hz PWM frequency—fast enough to avoid visible flicker. For instance, a combination of two LT3486s driving four LED strings (R-G-G-B) in a top-end display provides 1000:1 dimming while maintaining the true-color of the display even during very dim nighttime operation.

When a PWM signal is used for brightness control, but less than a 5:1 dimming range is needed and the chromaticity of the LEDs is not especially important, the PWM signal can be fed into an RC filter such as the one in Figure 6. This turns the PWM input into an analog CTRL pin voltage controlling the LED current directly, eliminating the need for the PWM dimming MOSFETs. The 5V, 16-LED converter in Figure 5 can deliver up to a 5:1 analog dimming range at the CTRL pins with such a filter without the need for the two additional PWM dimming MOSFETs. In this case, the LT3486 PWM pins are tied high to the 1.25V REF pin.

Doubler Delivers Greater than 34V to LED Strings

GPS navigation and in-cabin entertainment displays are increasingly popular in mainstream consumer vehicles. The advantage of using two LED drivers each with 8-LED strings, instead of a single 16-LED string, is that the maximum switch voltage remains that of a single 8-LED string (less than 34V total string voltage at 100mA). Even so, LCD panel screen sizes are pushing beyond the standard 6" and 7", requiring more LEDs and string voltages above 34V.

The circuit in Figure 7 uses a charge pump voltage doubler to drive two strings of LEDs to voltages as high
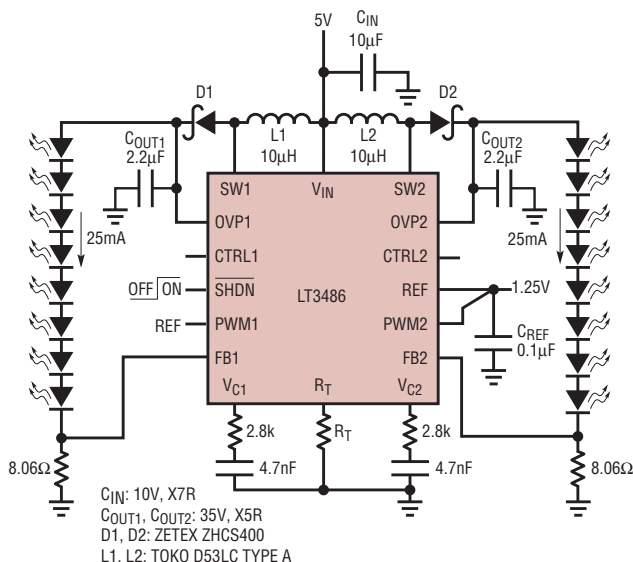
Figure 5. LED driver uses 5V input to drive two strings of eight 25mA LEDs (less than 34V total in either string) with 5:1 brightness control
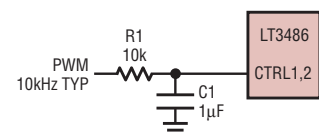

Figure 6. Achieving 5:1 brightness control with a filtered PWM signal

Hot Swap Controller Monitors and Reports Power Supply Status   by Josh Simonson

Introduction

High availability systems are designed to achieve an ideal of zero down time. To approach this goal, the system needs to be able to operate during routine maintenance and upgrades, which often involves cards being inserted and removed from a live backplane. These systems must also be designed for failsafe operation by isolating faulty boards before they cause backplane disturbances.

Hot swapping requires a power switch to initially isolate the board, and a controller to turn on the switch slowly to minimize backplane disturbances. Since the Hot Swap controller monitors card voltage and current, it is an obvious place to integrate higher level monitoring with a data converter. This provides detailed information about the health of the power path and the power consumption of downstream circuits. Such information
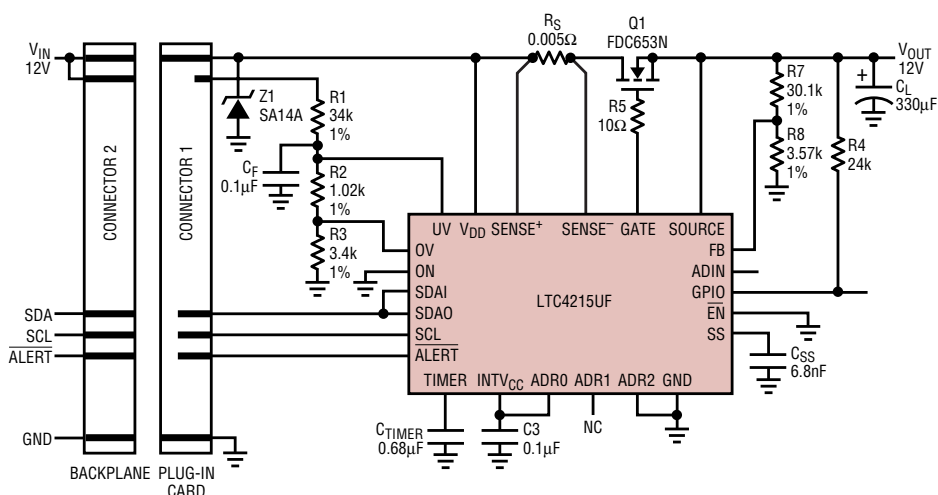

Figure 1. In a typical application the LTC4215 uses an external N-channel pass transistor to isolate the hot swapped board from the backplane when it is first inserted. After a debounce time the controller can begin to apply power to the board or wait for a turn-on command from a host processor. Power is ramped gradually to minimize any backplane disturbance. After the power-up process is complete, the LTC4215 continues to monitor for faults in the power path.

can be used to monitor performance over time and identify boards that are drifting towards failure or marginal performance.

The LTC4215 combines a robust Hot Swap circuit with an I^2C interface and data converter to allow power monitoring as well as hot-plug functionality

Table 1. A few of the LTC4215's many features

Feature	Benefits
Wide Input Voltage Range: Operates from inputs of 2.9V to 15V, with 24V absolute maximum	❏ Suitable for 3.3V, 5V and 12V systems ❏ Simplifies design because part functions on a semi-regulated supply ❏ Large overvoltage transient range eases design tolerances for transient protection
8-bit ADC: ADC monitors current, output voltage and external pin voltage and measures off-state current in the FET to determine FET failures	❏ Increases reliability ❏ Board power information provides an early warning of board failure ❏ Verify board is staying within its allotted power ❏ Allows integrity check of redundant supply paths ❏ Allows active power management to safely maximize power utilization within the chassis cooling constraints
I^2C/SMBus: Communicates as a read-write slave device using a 2-wire serial interface	❏ Improves integration with the host system. Interface allows the host to configure the part, determine which faults are present or have occurred, and read back ADC measurements
Fast Short Circuit Response: Fast (<1µs) current limit response to shorts	❏ Protects connector from overcurrent ❏ Limits the disturbance to the input supply from a short circuit
Alerts Host after Faults: When configured (using I^2C), faults activate an active pull-down on the $\overline{ALERT}$ pin	❏ Interrupting the host for immediate fault servicing limits system damage ❏ Reduces the bus traffic for polling

and fault isolation (see Table 1). In a typical application the LTC4215 uses an external N-channel pass transistor to isolate the hot swapped board from the backplane when it is first inserted (Figure 1). After a debounce time the controller can begin to apply power to the board or wait for a turn-on command from a host processor. Power is ramped gradually to minimize any backplane disturbance. After the power-up process is complete, the LTC4215 continues to monitor for faults in the power path.

The LTC4215 provides the means for quantitatively measuring the board current and voltages with an onboard ADC and multiplexer. It reports this information using the I^{2}C serial communication bus when polled by a host processor. The device interrupts the host for specific fault conditions, if configured to do so.

The LTC4215 works in applications from 12V (with transients to 24V) down to 3.3V where the operating input voltage could drop to 2.9V. Functionally, the LTC4215 is very similar to the LTC4260 (*Linear Technology*, Nov. 2004) operating in a lower voltage range. Table 2 compares the major features of the LTC4215 and LTC4260. Special attention should be paid to the power-up sequence because of the added soft start pin and some changes in the function of the TIMER pin relative to the LTC4260. Since both parts can be used for 12V systems, Table 2 may be used to select the part with optimal set of features for a specific 12V application.

An N-channel pass transistor, Q1, controls the application of power to the board as in Figure 1. A series sense resistor, RS, allows the LTC4215 to measure the current in the power-path. Resistor R5 suppresses self oscillations in Q1. Resistors R1–R3 select the undervoltage (UV) and overvoltage (OV) thresholds. Capacitor C$_F$ allows these thresholds to be filtered as needed. R7 and R8 select the power-good threshold and set the foldback current limit level. Capacitor C$_{SS}$ sets a maximum slew rate to control the inrush current and C$_{TIMER}$ is used to set the startup time. C3 is used to bypass the internal core voltage.

Typically, the pins on the connector are staggered so that bulk power is applied first with the longest pins, followed by communication lines on medium length pins, and last, Hot Swap control lines such as the supply for the $\overline{\text{UV}}$, OV, or $\overline{\text{EN}}$ pins. The UV, OV and $\overline{\text{EN}}$ pins must be in the correct state for a programmable debounce period of 100ms before Q1 is allowed to turn on. At this point the ON pin turns the part on immediately if it is high, or holds the part off if it is low. When the ON pin is held low, Q1 is turned on through the I^{2}C bus by writing to the ON bit in the control register.

Measure Real-Time Board Power with Integrated ADC

Monitoring the supply voltage and current in real-time is a useful way of tracking the health of the power path. New data can be compared with historical data for the same card to detect changes in power consumption that could indicate that the card is behaving abnormally. An abnormal card can be shut down and flagged for service, perhaps before a more severe fault or system malfunction occurs. The LTC4215 includes an 8-bit data converter that continuously monitors three voltages: the ADIN pin, the SOURCE pin and the current sense voltage between the SENSE+ (V$_{DD}$) and SENSE– pins. The ADIN pin is an uncommitted ADC input which allows the user to monitor any available voltage.

The ADIN pin is monitored with a 1.235V full scale. The ADIN pin is connected directly to a data converter input without any signal scaling. The SOURCE pin uses a 1:12.5 divider at the input which gives a 15.4V full

Table 2. Comparison of the LTC4260 and LTC4215		
Feature	**LTC4215**	**LTC4260**
V$_{DD}$ Abs Max	24V	100V
V$_{DD}$ Min	2.9V	8.5V
Recommend TransZorb for 12V	Yes	No
Current Limit/Circuit Breaker	25mV	50mV
Circuit Breaker Precision	10%	10%
Current Limit Precision (FB = 0)	35%	25%
Current Limit vs Circuit Breaker V$_{TH}$	75mV, 25mV	Both 50mV
Current Limit Foldback	Only during startup	Always
Gate RC Network	Optional	Required
Soft Start	Yes, Required	No
Timer Pin	Optional	Required
OC Timer	20µs	External/Adjustable
ΔV$_{GATE}$ at 12V (0µA)	6V	8.5V
Built In Overvoltage Threshold	15.6V	None
ADC Source LSB	60mV	400mV
ADC V$_{SENSE}$ LSB	151µV	300µV
ADC ADIN LSB	4.85mV	10mV
Internally Generated V$_{CC}$	3.1V	5.5V
Package	4mm × 5mm QFN	5mm × 5mm QFN

scale. The SENSE voltage amplifier has a voltage gain of 32, which results in a 38.4mV full scale. The converter uses a sophisticated oversampling and offset cancellation method that preserves the full 8-bit dynamic range on the SENSE channel.

If the data converter reads more than 1mV on the V_{DD}-SENSE channel while the external switch is turned off, the LTC4215 generates a FET-SHORT fault to indicate that the switch may be damaged. The presence of this condition is indicated in STATUS register bit C5 and logged to FAULT register bit D5. The LTC4215 takes no action in this condition other than logging the fault and generating an alert if configured to do so.

The results from each conversion are stored in three ADC registers (see Table 3) and updated 10 times a second. Setting the test mode control register bit halts the data converter so that the registers can be written to and read from for software testing.

Versatile Inrush Current Control

Once the inputs to the LTC4215 reach the correct values for the part to turn the external switch on and an internal 100ms debounce timer has expired, the LTC4215 turns on. The startup time is determined by the capacitor on the TIMER pin, or 100ms if the TIMER pin is tied to V_{CC}. During this time the circuit breaker is disabled to prevent an overcurrent fault from occurring, the power-good signal from the GPIO pin is also disabled to prevent turning on a load before the current limit has reached the full value via the Soft Start and Foldback pins. The inrush current slew rate (dI/dt) is limited via the SS pin. The inrush current is also folded back from 25mV to 10mV via the FB pin. An optional RC network on the external MOSFET gate can be used to set the inrush current below the foldback level by setting the maximum slope of the output voltage. The various inrush current profiles obtainable by these three methods are detailed in Figures 2 thru 6, which show a 12V system with a 25mΩ sense resistor,

Table 3. LTC4215 registers

Register	Description
CONTROL	Register turns-on or turns-off the pass transistor and controls whether the part will auto-retry or latchoff after a fault. It also configures the behavior of the GPIO pin.
ALERT	Alert register enables which faults interrupt the host using the $\overline{ALERT}$ pin. At power-up the default is to not alert on faults.
STATUS	Status register provides pass transistor (on/off), $\overline{EN}$ (high/low) and GPIO (high/low) conditions. It also lists five fault present conditions.
FAULT	Fault register logs overcurrent, overvoltage, undervoltage, power-bad, FET short and $\overline{EN}$ changed state faults.
SENSE	ADC data for the V_{DD}-SENSE voltage measurement.
SOURCE	ADC data for the SOURCE pin voltage measurement.
ADIN	ADC data for the ADIN pin voltage measurement.

or 1A current limit, starting up into a 470µF capacitive load.

At the end of the startup period the current limit circuit is checked. If the current limit is still regulating the current, the LTC4215 determines that the output failed to come up and generates an overcurrent fault. If the current limit circuit is not active then the current limit threshold is moved to 75mV, the power-good signal to the GPIO pin is enabled and the 25mV circuit breaker is armed.

The SS pin sets the current slew rate limit at startup. It starts at ground, which corresponds to a negative voltage on the sense resistor and results in the MOSFET being turned off. A current into the soft-start capacitor produces a ramp that corresponds to increasing V_{DD}-SENSE voltage. When the current limit circuit releases the gate (when the commanded V_{DD}-SENSE voltage becomes positive) the current from the SS pin is stopped to wait for the GATE pin to rise and start to turn on the MOSFET. Once the current limit circuit begins to regulate the V_{DD}-SENSE voltage, the current from the SS pin is resumed and the ramp continues until it reaches the foldback level. It is important that the SS pin stop the ramp while the GATE pin slews because the ramp would otherwise continue and result in an uncontrolled step in current once the MOSFET threshold is reached.

An uncontrolled step may violate inrush specifications and cause supply glitches on the backplane.

If the soft-start ramp reaches the foldback level, the foldback circuit stops the ramp, as shown in Figure 5. The ramp is allowed to continue as the voltage at the FB pin rises and increases the foldback current limit, still limited in slope and limited in magnitude by foldback as well.

If an RC network is placed on the GATE pin to manually set the inrush current to a value below the foldback level (Figure 4), the current limit circuit will leave regulation when it is unable to achieve the V_{DD}-SENSE voltage commanded by the SS and FB pins. If the startup timer expires during this inrush an overcurrent fault is not generated because the current limit is not active. The power-good output for GPIO is allowed to relay the state of the FB pin, and the circuit breaker is armed. Either the output voltage finishes rising and a power good is asserted when the FB pin crosses it's 1.235V threshold, or the current rises to the circuit breaker threshold and the part generates an overcurrent fault.

In the event there is an overcurrent condition after startup, the current limit circuit limits the V_{DD}-SENSE voltage to 75mV while the circuit breaker waits for a 20µs timeout before producing an overcurrent fault. After any overcurrent fault, the part waits
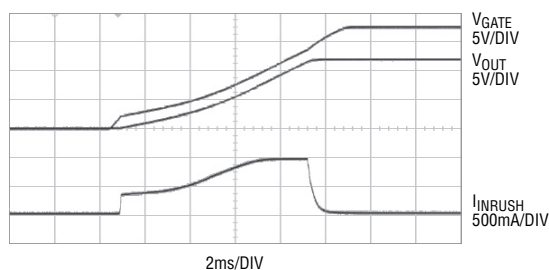
Figure 2. Inrush current is limited by foldback. This allows the fastest startup of the load, with the inrush only lasting 9ms with a peak current of 1A.
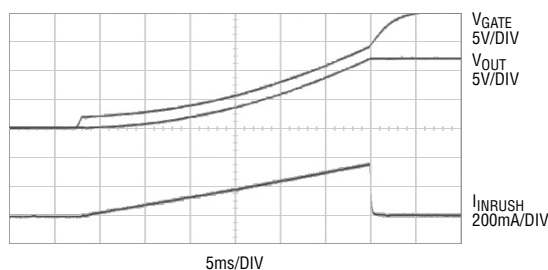

Figure 3. Inrush current is limited by a 1µF SS capacitor. This provides the fastest startup in a system with demanding inrush current slew rate requirements. Inrush dI/dt is reduced to 12mA/ms.
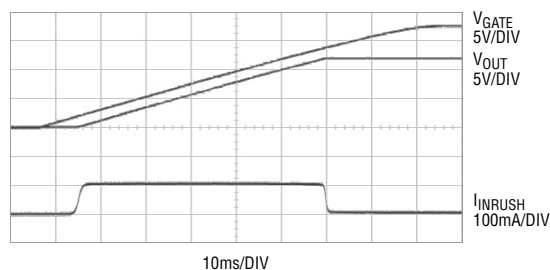

Figure 4. Inrush current is limited by a 0.1µF GATE capacitor. This minimizes the power in the switch, allowing the use of smaller components at the cost of speed. The inrush current is only 100mA, but startup takes 55ms.
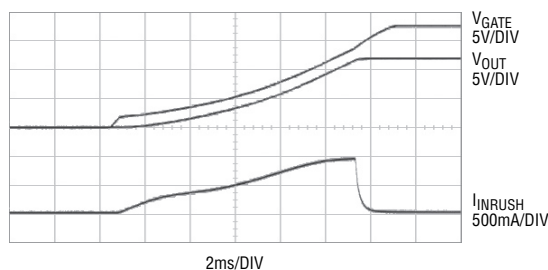

Figure 5. Inrush current is limited by a 68nF SS capacitor and FB. Soft start controls the inrush current slew rate while the current limit is modulated by foldback. This allows the fastest startup while also protecting the backplane from current surges.
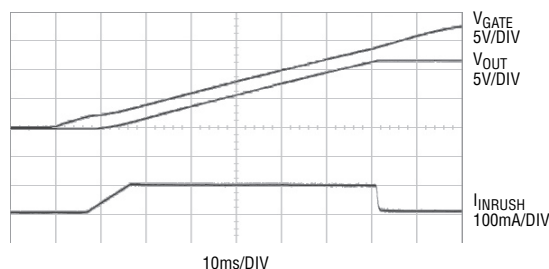

Figure 6. Inrush current with 1µF SS capacitor and 0.1µF GATE capacitor. Soft start limits the inrush current slew rate until the GATE capacitor limits the inrush current by limiting the dV/dt at the output. This minimizes the power in the switch while protecting the backplane from inrush current surges.

for a cool-down period of 50 times the startup time before allowing the part to restart by any means, including auto-retry, I^2C, or cycling the $\overline{EN}$, UV or ON pins.

Controlled Turn-Off

When the LTC4215 is turned off by a fault or I^2C transaction, the GATE pin is pulled down with a 1mA current source. Once the GATE pin is below the SOURCE pin, a diode from SOURCE to GATE turns on and the voltage at the SOURCE pin is discharged by the same 1mA current.

If there is a short that causes the sense voltage to exceed 75mV, a 400mA pull-down from GATE to SOURCE removes the gate charge of the switch. Once the sense voltage falls to 75mV, the current limit regulates there for 20µs before turning the gate off with the 1mA current source.

If there is significant inductance between V_{DD} (SENSE+) and bulk capacitance, across a connector for instance, it is possible that a short circuit at the output with a very fast rise time could cause the input voltage to collapse while the current through this inductance slews. In this case, after 2µs, the V_{DD} undervoltage lockout circuit turns on and discharges the GATE pin with the 400mA pull-down to the SOURCE pin and quickly turns the switch off.

Save Power with Precise 25mV Circuit Breaker

For supplies with lower voltages and higher currents, a 50mV circuit breaker threshold may result in too much power dissipation in the sense resistor, or cut excessively into the input supply voltage tolerance of downstream circuits. To reduce this problem the LTC4215 has a precision circuit breaker at 25mV with a low 10% tolerance. This allows the use of smaller and less expensive sense resistors with lower power ratings.

In systems where the circuit breaker has only 20% accuracy the designer must be able to safely provide 40% more power than the card actually consumes to ensure that the slot doesn't suffer from heat and supply limitations on the high side or produce a fault in normal operation on the low side. The precise 10% accuracy of the LTC4215 cuts this guard-band in half and safely allows the use of 20% more

Efficient Buck-Boost Converter Ideal for Power Saving Modes and Wide Input Voltage Ranges

by Kevin Ohlson

Introduction

Portable handheld electronic tools, gadgets, and toys are approaching the multi-function equivalent of a Swiss army knife. Previously separate functions are now miniaturized and combined into single pocket sized package. Unless owned by a teenager or a Hollywood agent, most gadgets spend most of their time in a low power state, waiting to come to life to make a phone call, pulse a photoflash or spin music and video to a tiny hard disk drive. These devices are powered from a variety of power sources, which means at some point the input voltage is higher than or lower than the 3.3V or 3.6V voltage needed to power the internal electronics. The DC/DC converters in the latest portables must be able to step up and step down voltage, maintain very high efficiency during idle and standby modes, and respond quickly and efficiently during peak power demands.
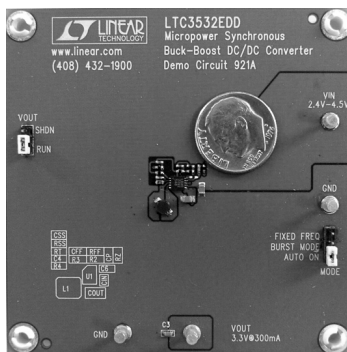

Figure 1. Lithium ion battery to 2.5V–5.25V converter in 1.4cm^2

Linear Technology offers a family of buck-boost converters capable of supplying from 200mA to 2A with excellent efficiency. The latest addition to this lineup is the LTC3532, a 300mA buck-boost converter, which incorporates automatic Burst Mode operation, adjustable switching frequency, and integrated soft-start. The LTC3532 is ideal for miniature disk-drive applications or any application that requires high efficiency over a wide range of output currents and input voltages.

Features

The input voltage range of a LTC3532-based converter is 2.4V to 5.5V and its output range can be programmed from 2.4V to 5.25V, making it ideal for devices operating from multiple sources such as battery, USB, and wall adapters. The LTC3532 is available in either an MS10 package, which is pin compatible with the LTC3440 converter, or in an exposed pad 3mm × 3mm DFN. With these tiny packages, an entire converter can be squeezed into the smallest spaces, as shown in Figure 1.

Using a fixed frequency four-switch architecture and a patented control method, the converter needs only a single inductor to regulate a constant output voltage with input voltages greater than or less than the output. The four switch topology of the LTC3532 (see the output stage schematic in Figure 2) allows the regulator to smoothly transition from buck mode to buck-boost mode and boost mode by correctly phasing the four output switches (A, B, C, and D) in response to the error amp output voltage, V_C. During buck mode, switch D is on while switches A and B act like a buck converter. At the other extreme, boost mode, switch A is always on while switches C and D implement a synchronous boost converter. When V_{IN} and V_{OUT} approach the same voltage, all four switches commutate with the on time for each pair controlled by the voltage at V_C. The four switch architecture inherently provides output disconnect, which prevents current flow between V_{IN} and V_{OUT} in shutdown mode.
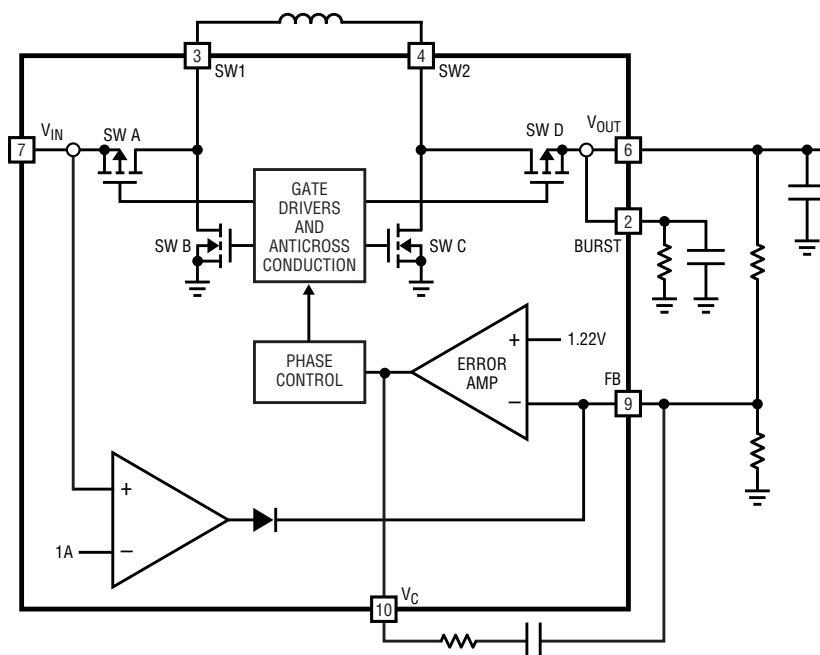

Figure 2. The LTC3532 four switch buck-boost converter uses a single inductor and features peak current clamp and automatic Burst Mode control.
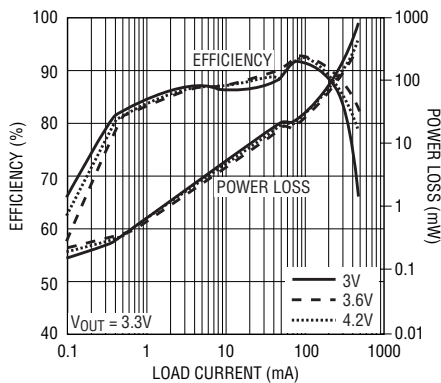
Figure 3. High efficiency is possible over a wide range of load currents using automatic Burst Mode control.

Figure 4. A soft clamp peak current control keeps the converter in continuous control mode when peak inductor current is reached.

The capabilities of the four switch architecture is exploited in the LTC3532 when in Burst Mode operation as well. An innovative (patent pending) Burst Mode control circuit optimizes the firing of the four switches depending on whether in buck mode, boost mode, or buck-boost mode. Optimal switching control and a low 35µA Burst Mode quiescent current means the converter increases the battery life of a system by keeping efficiency above 80% in Burst Mode operation at loads as low as 300µA. Measured efficiency over a load current ranging from 0.1mA to 500mA is shown in Figure 3. A resistor and filter capacitor connected to the burst pin set the level of load current at which the converter automatically switches between continuous and Burst Mode operation.

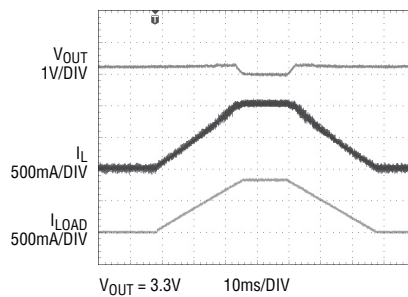Peak inductor current is limited two ways. The first method monitors current in switch A and sources a fraction of that current into the FB pin when the peak current exceeds 1.1A. This effectively lowers the V_{OUT} set point providing a closed loop method of controlling the peak current. In higher load and transient situations a comparator opens switches A and B, thus providing a hard peak current limit of 1.3A. Figure 4 shows inductor current and V_{OUT} responding to increasing load. As the inductor current reaches current limit, V_{OUT} drops and the control loop stays in continuous operation. When V_{OUT} is low either at start-up or recovering from a short circuit the current limit clamp level is reduced by half providing a foldback function.

Switching frequency may be programmed with an external resistor to a frequency between 300kHz and 2MHz, which allows a trade off between component size and efficiency. Soft start is performed by controlling the slew on the $\overline{SHDN}$/SS pin. Once the converter is enabled, when the $\overline{SHDN}$/SS pin reaches 1V, the internal control voltage is clamped until the pin rises to 2V.

Li-Ion to 3.3V Converter Ideal for Miniature Hard Disk Drives

A miniature, 1" or smaller disk drive in standby may draw 40mA. In idle or track seeking modes the drives current increases to 150mA and when reading or writing data the load might peak at 200mA to 300mA. Even during these transitions, the supply should be well regulated with very low ripple throughout the discharge cycle of the system battery. Compared with a step-down regulator in a 3.3V system a buck-boost converter such as the LTC3532, which maintains regulation even as the battery drops below 3.3V, allows all the energy in the battery to be used. The converter keeps accurate regulation during load transitions when battery ESR may cause the input voltage to drop below V_{OUT}. A Li-ion to 3.3V application, which uses a tiny multi-layer chip inductor, is shown in Figure 5. Capable of load steps up to 400mA with battery voltage as low as 3.0V, the converter delivers efficiency greater than 90% in continuous mode with loads between 30mA and 200mA as shown in Figure 6. Lower power burst mode efficiency is greater than 80% at sub 1mA loads. This circuit uses a soft start capacitor connected to $\overline{SHDN}$/SS to limit inrush current. Transient responses to load steps are shown in
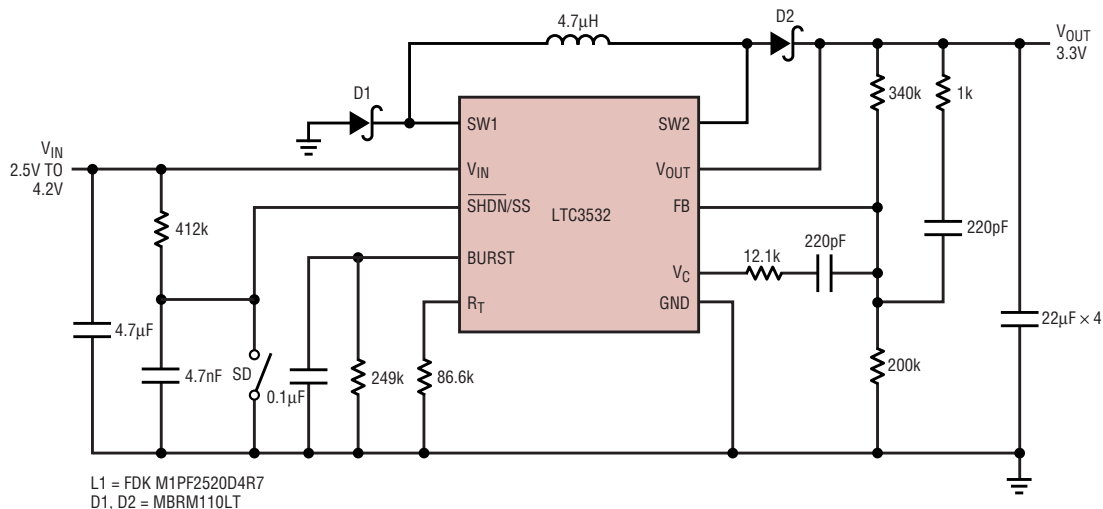

Figure 5. A high efficiency converter using a tiny multi-layer inductor ideal for miniature HDD applications
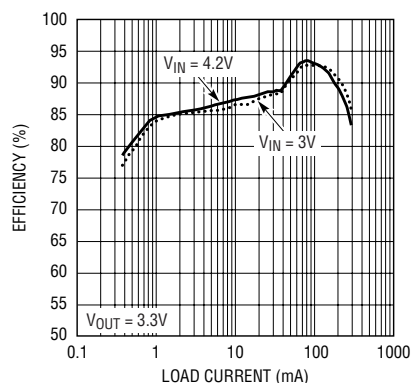
Figure 6. Efficiency is over 85% for all power saving modes of a typical 1" hard disk drive

Figure 7. V_{OUT} response to load transitions is well controlled.

Figure 7. As the load is increased from 10mA to 50mA the transition from burst mode to continuous operation occurs. Steps from 150mA to 300mA and back show V_{OUT} transients have peak amplitude of only 20mV.

An RC network on the BURST pin enables automatic Burst Mode operation to maintain high efficiency at light loads without external control. Burst operation is controlled by the voltage on the burst pin. During operation a small fraction of the output current passing through switch D is mirrored out of the BURST pin. The mirrored current produces a voltage across the burst pin resistor that is proportional to the average load current. Figure 7 shows the burst pin responding with a voltage proportional to the load current. When the load increases and forces the burst pin voltage above
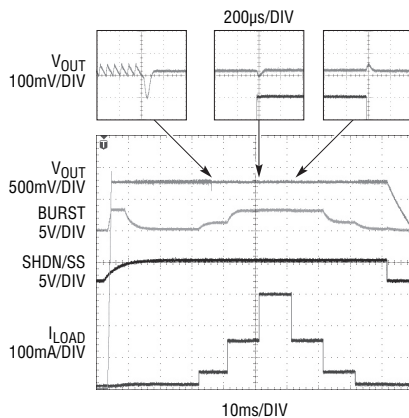
1.12V the converter switches to continuous mode. As the load decreases and the voltage on the burst pin drops below 0.88V, the converter switches back to Burst Mode operation. Raising
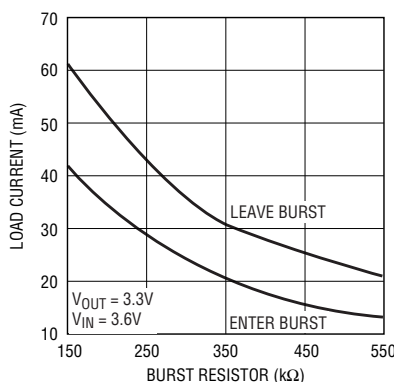

Figure 8. The load at which the converter transitions from Burst Mode to continuous mode is programmable with a single resistor.

the value of the resistor on the BURST pin lowers the current at which burst mode is entered. Figure 8 shows the relationship between the burst pin resistor and the output current value at which the transition between continuous and burst operation takes place. If desired, the operating mode may be forced by driving the BURST pin above or below the thresholds.

Control Input Current for USB Applications

Many devices now are powered and recharged from USB ports which have the restriction of a maximum current draw of 500mA. A converter that typically supplies 500mA would not nominally exceed the USB current limits. However, tolerances of host regulators, USB bus powered hubs, and cable drops result in a rather poorly regulated USB voltage which may vary from 5.25V down to 4.35V or lower during a transient. For example, if a 5V to 3.6V converter circuit is responding to a peak load of 500mA and the USB voltage drops to 4.35V, at 80% efficiency the input current exceeds 500mA.

Figure 9 shows an input current monitor, which controls V_{OUT} to clamp the input current to 500mA. Current injected into the FB node changes the effective set point of the output voltage within the voltage mode control loop. In fact, the output voltage of a
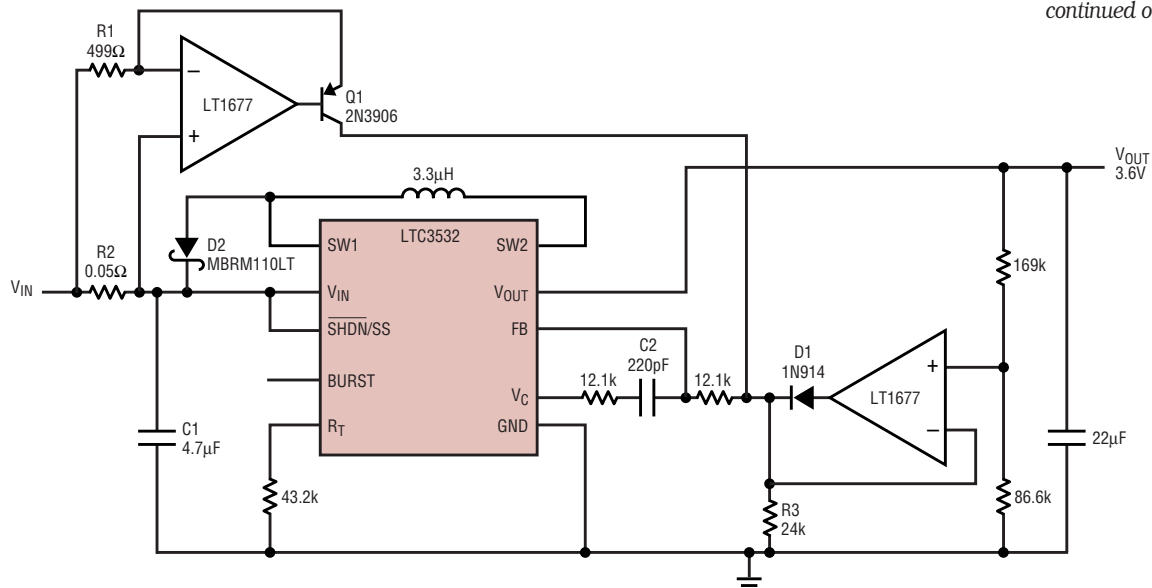
Figure 9. A few components may be used to limit input current for USB and other applications

Dual/Triple Power Supply Monitor for Undervoltage and Overvoltage on Positive *and* Negative Supplies

by Andrew Thomas

Introduction

An accurate power supply monitor can signal when a supply overvoltage or undervoltage condition threatens to cause system failures, allowing the system to deal with the situation gracefully.

The LTC2909 is a highly customizable monitoring solution with adjustable input thresholds, input polarity selection, a multimode reset timer, and an open-drain $\overline{RST}$ output. Adjustable input thresholds allow the user to set any trip threshold for the comparator, subject only to the accuracy limitations of the part, instead of having to pick from a factory-set limited collection of thresholds.

Each adjustable input can be configured in either polarity, allowing it to monitor negative or positive supply voltages for undervoltage or overvoltage. Polarity selection is controlled by simple connection of the SEL pin—no external components required.

The multimode timer pin can be configured a number of ways to suit a large variety of applications, allowing full control over the reset timeout, elimination of the external timing capacitor, or removal of the timeout altogether. The $\overline{RST}$ pin is an open-drain output—it can be pulled up to an appropriate voltage for the device receiving the $\overline{RST}$ signal, independent of the supply for the LTC2909. The output can be wired-OR connected with other supervisors or other open-drain logic, allowing any of a number of conditions to issue a reset.

Minimal Space Required

Figure 1 shows how the LTC2909, with just a few components, can monitor a 24V supply for both undervoltage and overvoltage. Almost any two re-
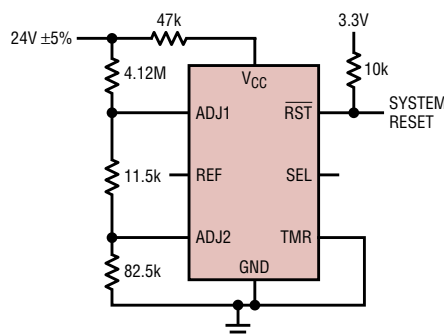

Figure 1. A 24V undervoltage and overvoltage monitor

set conditions in any system can be monitored by appropriate connection of the LTC2909. The small size of the LTC2909 (available in 8-pin 3mm × 2mm DFN and TSOT-23 packages) keeps the monitoring solution small, and the high accuracy of the part keeps

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The two adjustable inputs can be configured in either polarity, allowing the LTC2909 to monitor negative or positive supply voltages for undervoltage or overvoltage. Polarity selection is controlled by simple connection of the SEL pin—no external components required.

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system uptime high without sacrificing reliability. The separate V_{CC} pin of the LTC2909 incorporates a shunt regulator, which allows the part to be powered from any high availability supply, even a high voltage rail. Furthermore, the low quiescent current consumed by the LTC2909 makes it suitable for low power applications like battery-powered handheld devices.

Any Polarity, Undervoltage or Overvoltage

The most common application of a supply monitor is determining when a positive supply is below some critical threshold required for the proper operation of powered devices. Less common, but no more difficult for the LTC2909, are scenarios that require negative supply monitoring, or determining when the voltage exceeds some value beyond which functionality might be impaired or powered devices damaged.

The connection of the SEL three-state input pin determines whether each of the ADJ input comparators is configured as positive-polarity (input must be above the threshold or $\overline{RST}$ is asserted low) or as negative-polarity (input must be below the threshold or $\overline{RST}$ is asserted low). Inputs that are configured as negative-polarity are useful for resetting when the monitored voltage is more positive (or less negative) than it should be. In other words, a negative-polarity input can monitor a positive supply for overvoltage (OV) or a negative supply for undervoltage (UV). Similarly, a positive-polarity input is useful for issuing a reset when the monitored voltage is more negative (or less positive) than it should be, so it may monitor a positive voltage for undervoltage or a negative voltage for overvoltage. Conventionally, the terms overvoltage and undervoltage refer to the absolute value of the monitored voltage, so a –5V supply at –4.3V is undervoltage.

Connecting SEL to ground configures both adjustable inputs as negative polarity. In this mode, the part may be used as a dual negative undervoltage monitor, or a dual positive overvoltage monitor. If desired, it also functions as

a single negative undervoltage monitor with a single positive overvoltage monitor. Connecting SEL to V_{CC} configures both inputs as positive polarity, useful for dual positive undervoltage or dual negative overvoltage monitors, as well as a single positive undervoltage monitor with a single negative overvoltage monitor. Finally, leaving the SEL pin open configures ADJ1 as positive polarity, and ADJ2 as negative polarity. In this configuration, the part can monitor one positive and one negative supply both for undervoltage, or both for overvoltage. It can also function as a window (undervoltage and overvoltage) monitor for one positive or negative supply. These polarity selections and the corresponding applications are summarized in Table 1.

Adjustable Inputs

The LTC2909 inputs are fully adjustable for ultimate monitoring flexibility. Each ADJ pin connects directly to the high-impedance input of a comparator whose other input is tied to an internal 500mV (nominal) reference. Setting the threshold voltage is as simple as connecting a resistor divider from the supply so that the ADJ input is at 500mV when the monitored supply is at the desired threshold. By choosing the correct external resistors, the nominal trip point can be set to *any* desired value.

The typical configuration of resistors for a positive supply is as shown in Figure 2. For a negative supply, some offset is needed to allow the resistor tap point to lie at 500mV. This offset is provided by the REF pin on the LTC2909, which provides a buffered 1V reference (with 1.5% accuracy over the operating temperature and supply voltage range). Thus, the typical divider connection for a negative supply is as shown in Figure 3. Note that positive supplies with nominal trip points below 500mV should be considered "negative" for monitoring purposes (since they require an upwards shift to reach 0.5V). Monitoring a single supply for UV and OV can be accomplished with three resistors, as shown in Figure 4 for a positive supply and Figure 5 for a negative supply.

Selection of resistor values is driven by two factors: nominal trip point and current consumption. In particular, the selection of R1 is driven by current consumption, and the ratio of the other resistors to R1 determines the trip point. If the monitored voltage is typically close to its nominal trip threshold, the voltage across R1 is approximately 0.5V, so the current consumed by the resistor divider is about 0.5V/R1. Supplies that operate substantially away from their threshold cause the current consumption to deviate from the estimate above by about the same percentage by which they deviate from the threshold.

In most applications, the current consumption should be minimized. However, as the current is reduced, the impact of leakage at the tap point on the monitoring accuracy becomes more severe. The leakage current is drawn from the driving-point impedance at the ADJ input, so the fractional error is approximately:

$$\frac{I_{LEAK} \bullet R1 \bullet R2/(R1+R2)}{500mV}$$

or for UV/OV circuits:

$$\frac{I_{LEAK} \bullet (R1+R2) \bullet R3 / (R1+R2+R3)}{500mV}$$

and

$$\frac{I_{LEAK} \bullet R1 \bullet (R2+R3) / (R1+R2+R3)}{500mV}$$

As a rule of thumb, the current in the divider should be at least 100 times the expected leakage, including the 15nA maximum internal to the part and any external leakage sources.

The rest of the resistor values are determined by the choice of trip point. Since the accuracy of the LTC2909 thresholds is guaranteed to 1.5% over the operating temperature and supply range, the trip points should usually be set 1.5% beyond the specified operating range of the monitored supply. For example, a 5V±10% supply should have a 4.425V undervoltage trip point, not 4.5V. See the sidebar on threshold accuracy for an explanation.

Given a desired trip point, and the value of R1 chosen as above, it is then possible to calculate the appropriate
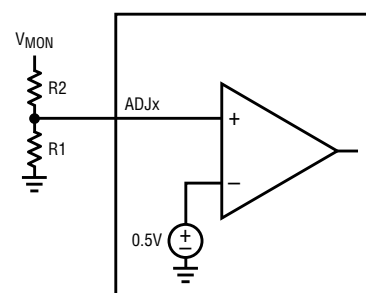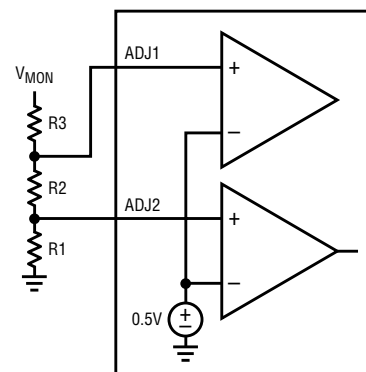

Figure 2. Monitoring a positive supply

Figure 4. Monitoring a positive supply for UV and OV
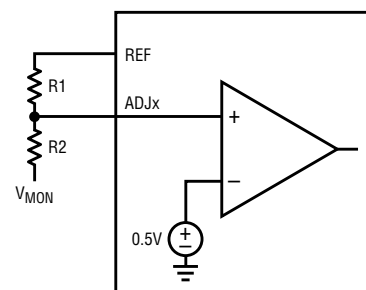

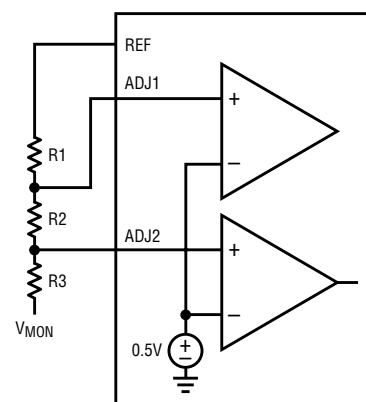**Figure 3. Monitoring a negative supply**

Figure 5. Monitoring a negative supply for UV and OV

values of the rest of the resistors. When monitoring a positive supply for a single fault condition, the user should choose

$$R2 = R1 \cdot \frac{V_{TRIP} - 500mV}{500mV}$$

Similarly, for a negative supply (or positive supply with trip voltage below 0.5 V),

$$R2 = R1 \cdot \frac{500mV - V_{TRIP}}{500mV}$$

Note that if the desired trip voltage is below ground, the value V_{TRIP} should be negative. The situation is slightly more complicated when only three resistors are used to monitor a single supply for UV and OV. For a positive supply with desired trip thresholds $V_{TRIP(UV)}$ and $V_{TRIP(OV)}$, the appropriate values are

$$R2 = R1 \cdot \frac{V_{TRIP(OV)} - V_{TRIP(UV)}}{V_{TRIP(UV)}}$$

and

$$R3 = R1 \cdot \frac{V_{TRIP(UV)} - 500mV}{500mV} \cdot \frac{V_{TRIP(OV)}}{V_{TRIP(UV)}}$$

Table 1. SEL connection for various input polarities

ADJ1	ADJ2	SEL Pin
Positive polarity: Positive UV or Negative OV	Positive polarity: Positive UV or Negative OV	V_{CC}
Positive polarity: Positive UV or Negative OV	Negative polarity: Negative UV or Positive OV	Open
Negative polarity: Negative UV or Positive OV	Negative polarity: Negative UV or Positive OV	GND

Finally, for a negative supply with desired trip thresholds $V_{TRIP(UV)}$ and $V_{TRIP(OV)}$, the appropriate values are:

$$R2 = R1 \cdot \frac{V_{TRIP(UV)} - V_{TRIP(OV)}}{1V - V_{TRIP(UV)}}$$

and

$$R3 = R1 \cdot \frac{500mV - V_{TRIP(UV)}}{500mV} \cdot \frac{1V - V_{TRIP(OV)}}{1V - V_{TRIP(UV)}}$$

Tables 2 and 3 show suggested values of resistors for monitoring a number of standard supply voltages for UV, OV or UV and OV. Table 2 gives values for nominal supply accuracy of 5% (6.5% trip points), and Table 3 gives values for 10% supplies (11.5%

trip points). In the tables, the values of R1 have been chosen to minimize the threshold error using standard 1% resistor values, while maintaining the divider current consumption near 5µA.

UVLO

The LTC2909 features a third high-accuracy comparator on the V_{CC} pin, which allows the part to function in some applications as a triple supply monitor. The polarity of the V_{CC} comparator is fixed to be positive, so the comparator creates an accurate UVLO. The threshold of the UVLO is also fixed, and is set at 11.5% below the nominal threshold voltage specified in the part number. Versions are available for standard logic supplies: LTC2909-2.5 for 2.5V supplies (2.175V nominal threshold), LTC2909-3.3 for 3.3V supplies (2.921V nominal threshold), and LTC2909-5 for 5.0V supplies (4.425V nominal threshold). The LTC2909-2.5 is recommended for designs that do not want monitoring of the V_{CC} pin. The UVLO then functions merely to ensure that $\overline{RST}$ is not allowed to go high while the V_{CC} voltage is too low to guarantee proper accuracy of the ADJ input thresholds. The accuracy of the UVLO threshold is the same as the ADJ thresholds: ±1.5% guaranteed over the operating temperature range.

Glitch Immunity

A monitored supply generally has high-frequency components riding on its DC value. These may be caused by load transients acting on non-zero output impedance (whether due to supply line impedance or regulation bandwidth), output ripple of the supply, coupling

Table 2. Suggested resistor values for 5% monitoring

Nominal Voltage	5% UV		5% OV		5% UV and OV		
	R1	R2	R1	R2	R1	R2	R3
24	232k	10.2M	102k	5.11M	82.5k	11.5k	4.12M
15	115k	3.09M	200k	6.19M	76.8k	10.7k	2.37M
12	49.9k	1.07M	102k	2.49M	76.8k	10.7k	1.87M
9	115k	1.82M	78.7k	1.43M	162k	22.6k	2.94M
5	137k	1.15M	137k	1.33M	76.8k	10.7k	732k
3.3	221k	1.15M	340k	2.05M	76.8k	10.7k	453k
2.5	115k	422k	51.1k	221k	137k	19.1k	576k
1.8	63.4k	150k	115k	324k	82.5k	11.5k	221k
1.5	59.0k	107k	137k	301k	76.8k	10.7k	158k
1.2	127k	158k	102k	158k	187k	26.1k	267k
1.0	200k	174k	100k	113k	107k	15.0k	105k
−5	133k	1.37M	118k	1.37M	174k	20.0K	2.00M
−9	97.6k	1.74M	115k	2.32M	182k	22.6k	3.65M
−12	107k	2.49M	40.2k	1.07M	40.2k	5.11k	1.07M
−15	107k	3.09M	309k	10.2M	309k	40.2k	10.2M

from nearby high-frequency signals, or noise. Ideally, the supply monitor should decide whether the supply voltage transient threatens the functionality of any of the devices which are powered by that voltage rail, and issue a reset if (and only if) it does. Unfortunately, a real supervisor cannot use an omniscient algorithm to know what exactly is connected to the bus or how those devices respond to supply transients. Given this, a number of possible approaches exist, addressing some of the concerns related to supply transients. These techniques focus on eliminating two undesirable situations that result from using a simple comparator.

One undesirable effect that must be prevented is rapid toggling of the reset output ("chattering"), caused by ripple, coupling, or noise on a supply voltage that is near the threshold. A common solution is to add hysteresis to the monitor threshold, which prevents chattering as long as the transient amplitude is less than the amount of hysteresis. Adding hysteresis effectively worsens the threshold accuracy, thereby unnecessarily reducing system uptime, or tightening the system requirements on supply voltage. For this reason, the LTC2909 uses other methods to prevent chattering, and does not have threshold hysteresis, unless the part is configured in comparator mode, where it would otherwise be more susceptible to chattering than usual (as explained below).

The primary defense against chattering is the programmed reset timeout period. If at any time during the reset timeout the supplies become invalid, the timer is immediately zeroed, and starts timing again from the beginning of the period when the supplies become valid again. Thus, any time the supply voltage is close enough to the threshold that the amplitude of the supply transients take the supply into the invalid region, $\overline{RST}$ remains low as long as the time between transients is less than the reset timeout. That is to say, the reset timeout prevents transients with frequency greater than $1/t_{RST}$ from causing undesired

Nominal Voltage	10% UV		10% OV		10% UV and OV		
	R1	R2	R1	R2	R1	R2	R3
24	102k	4.22M	115k	6.04M	39.2k	10.2k	2.05M
15	200k	5.11M	200k	6.49M	41.2k	10.7k	1.33M
12	115k	2.32M	107k	2.74M	41.2k	10.7k	1.05M
9	113k	1.69M	140k	2.67M	73.2k	19.1k	1.37M
5	113k	887k	113k	1.15M	115k	30.1k	1.13M
3.3	221k	1.07M	294k	1.87M	226k	59.0k	1.37M
2.5	102k	348k	301k	1.37M	41.2k	10.7k	178k
1.8	137k	301k	86.6k	261k	63.4k	16.5k	174k
1.5	48.7k	80.6k	43.2k	102k	51.1k	13.3k	107k
1.2	137k	154k	63.4k	107k	80.6k	21.0k	115k
1.0	200k	154k	137k	169k	174k	45.3k	169k
−5	115k	1.13M	200k	2.43M	115k	24.3k	1.37M
−9	127k	2.15M	215k	4.53M	51.1k	11.8k	1.07M
−12	115k	2.55M	41.2k	1.15M	130k	30.9k	3.57M
−15	115k	3.16M	309k	10.7M	47.5k	11.5k	1.62M

Table 3. Suggested resistor values for 10% monitoring

toggling at the reset output. Because the timeout is defeated in comparator mode, the LTC2909 is free to chatter in that mode, so a small amount of one-sided hysteresis is added to the comparator thresholds. See "Timeout Control" below for a description of the hysteresis behavior.

The other concern that must be addressed is identifying which transients cause a problem for the devices on the supply bus. It can generally be assumed that those devices can continue to operate through short

Figure 6. Allowable glitch duration as a function of magnitude

duration excursions outside the valid supply region, particularly because local decoupling capacitors help prevent such transients from appearing at the devices. If possible, the supervisor should not issue a reset during these conditions.

Consider, for example, what happens when a system spins up a hard drive connected to a monitored supply bus. The bus voltage briefly dips, possibly falling outside the valid region, and then returns, approximately, to its previous value. This is normal, expected behavior, and a microprocessor that is also connected to that bus should function normally through the transient (otherwise there is no way the system can ever safely use the hard drive). The supply monitor should not issue a reset to the microcontroller during such a transient.

To solve this problem, the LTC2909 has low-pass filtering on the comparator outputs, so that short duration glitches on the monitored supply are not passed through to the control logic. For most systems, the response of the system to a glitch depends on the
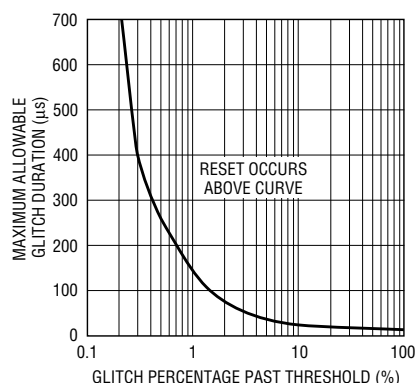
energy contained in the glitch, rather than just the voltage amplitude of the glitch. The duration of the glitch also factors into that energy, so the probability of a failure increases as the duration of the glitch increases (e.g. a 20% glitch on the supply may only be tolerable for 100µs, whereas a 5% glitch is tolerable for 1ms). The filtering on the LTC2909 comparators reflects this tendency. Figure 6 shows a typical curve of the maximum glitch duration that does not result in the LTC2909 issuing a reset, versus the percentage amount the glitch goes into the invalid region.

Some of these concerns can be exacerbated by circuit board layout, so it is also important that some care be taken in the layout near the LTC2909. In applications which use negative polarity comparators, capacitive coupling from the $\overline{RST}$ output to the negative-polarity input can cause the part to oscillate at approximately $1/t_{\overline{RST}}$ if the negative-polarity input is sufficiently close to threshold: the capacitive coupling creates AC negative feedback around the part. To prevent this oscillation, the $\overline{RST}$ line should be kept away from the relevant ADJ inputs, and, where

possible, from the corresponding supply. Negative-polarity applications may also oscillate when the $\overline{RST}$ is driving a large load, which causes a voltage difference between the ground of the 0.5V internal reference, and the ground of the monitored voltage. Several factors can help eliminate this source of oscillation. First and foremost, the current sunk by $\overline{RST}$ should be kept below 1mA if possible. Good grounding practice is also important. Input resistor dividers which connect to ground should have a Kelvin-sense trace directly to the GND pin, and the path from the monitored supply ground to the GND pin should be low impedance (preferably through a good ground plane).

Timeout Control

As described above, the LTC2909 has a reset timeout delay which helps reduce the sensitivity of the monitor to supply glitches. For convenience, this reset timeout can be controlled in three different ways. If a 200ms timeout is appropriate for the application (based on expected noise distributions and system timing specifications), no external components are needed to set the timeout—simply tie the TMR pin
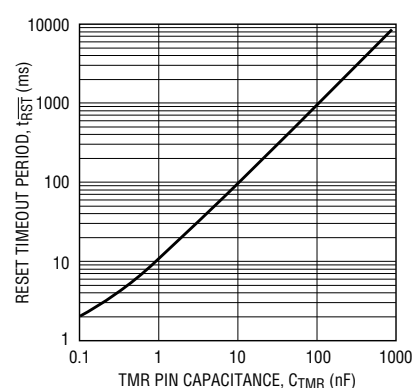

Figure 7. Reset timeout period as a function of capacitance

to ground, and the LTC2909 uses an internal 200ms delay generator.

For applications that require timeout periods other than 200ms, the delay can be set by connecting the TMR pin to a grounded capacitor, where the delay is set at approximately 9ms per nF of capacitance. To ensure timer accuracy, the timing capacitor should be a low leakage ceramic type. Leakage currents over 500nA substantially impair timer function. As an example, for a 50ms delay, the timer capacitor should be 50/9 = 5.6nF.

Figure 7 shows the typical timeout period as a function of the capacitor on the TMR pin. Due to inherent capacitance on the TMR pin, the minimum attainable timeout period in external mode is about 400µs, with no external capacitor connected to the pin. The maximum timeout is limited to nine seconds (1µF capacitor) by startup concerns. Assuming that the timer capacitor is initially discharged during the power-up sequence, the LTC2909 initially sees that the TMR voltage is near ground, and thus operates in internal timeout mode. As soon as the part is powered, a 2µA current source begins pulling up on the TMR pin, charging the timer capacitor towards the ground sense threshold (approximately 250mV). If all three supply inputs (V_{CC} and both ADJ inputs) become valid, and the 200ms internal timeout period completes before the TMR voltage reaches the ground sense threshold, $\overline{RST}$ goes high after a much shorter delay than was intended. If this startup behavior is not a problem in a given system, the

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Why Is Threshold Accuracy Important?

In monitored systems, there is some voltage level beyond which the proper function of the devices connected to a supply bus cannot be guaranteed. Ideally, that is the voltage at which the supervisor should issue a reset, since this guarantees the proper function of the system while permitting the maximum allowable variation in supply voltage. Thus, in the ideal case, the power supply tolerance is as loose as the devices on the bus will tolerate.

Of course, any real supervisor has limited accuracy, which tightens the system constraints. Typically, monitor accuracy is specified as a percentage band around the nominal trip point in which the threshold is guaranteed to lie, such as ±1.5%. To prevent nuisance resets when the supply is operating normally, the supply tolerance and monitor accuracy bands should not overlap.

As an example, a supply with a specified tolerance of ±5%, monitored by a 1.5% accurate monitor must have its nominal threshold set at 6.5% to prevent nuisance resets. With that accuracy band, the supervisor is not guaranteed to issues a reset until the supply has reached the other end of the monitor accuracy band, at 8%. Therefore, the devices attached to the supply must function properly to at least an 8% deviation in supply voltage. If this is not possible, a supply with tighter tolerance must be provided. For comparison, if the 1.5% accurate supply monitor is replaced by a less accurate 2.5% device, the power supply tolerance must be tightened to ±3% to ensure the same 8% operation band, thus complicating the power supply design. *LT*

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maximum timeout is limited only by the availability of large capacitors with leakage currents below 500nA.

Finally, there are some systems where the reset timeout delay is undesirable. For example, this may be the case in applications where the user is not using the LTC2909 $\overline{RST}$ pin as a system reset line. If the user ties the TMR pin to V_{CC}, the LTC2909 is put into comparator mode. In comparator mode, the timeout delay is bypassed, and the comparator outputs are connected directly to the $\overline{RST}$ drive circuitry. Due to the glitch-rejecting low-pass filter in the comparators, there will still be some delay from the inputs to the $\overline{RST}$ output, based on the amount of overdrive on the input. As shown by Figure 6, the propagation delay for large overdrives is about 25µs.

In comparator mode, because the reset timeout has been removed, the glitch and oscillation immunity of the part have been decreased. To prevent undesired "chattering" of the $\overline{RST}$ output when the input voltages are very close to threshold, a small amount of one-sided hysteresis is added to all three comparators. The hysteresis is "one-sided" in the sense that the valid-to-invalid transition is unaffected, but the invalid-to-valid threshold is moved about 0.7% into the valid region. Thus, for the ADJ inputs, the threshold voltages in comparator mode are a function of the SEL pin state. Nominal values are shown in Table 4.

Shunt Regulator

In most systems, it is possible to identify one supply as the one with highest availability—that is to say the supply which is most likely to be on, first to power up, last to shut down, and so on. There are a number of advantages to powering a supply supervisor from this highest-availability supply. First, the $\overline{RST}$ pull-down circuits are powered by the part supply. Thus, having the part supply come up first helps guarantee that $\overline{RST}$ never floats high due to insufficient pull-down strength. Conversely, powering the part from a high-availability supply helps maximize the uptime of the system because the LTC2909 will not release the $\overline{RST}$ output unless the part is properly powered.

The problem in many systems is that the high-availability supply is also a relatively high-voltage supply. For example, the highest availability supply in an automotive system is the 12V (nominal) battery voltage, and in a telecom system it is likely to be a 48V supply. Most supply supervisors require an external voltage regulator to operate from these supplies, but the LTC2909 saves components by integrating a 6.5V shunt regulator into the V_{CC} pin. All that is required is a series-dropping resistor between the high-voltage supply and the V_{CC} pin. This scheme allows the LTC2909 to be powered from an arbitrarily high voltage, subject only to constraint by the power dissipation in the shunt resistor. Furthermore, the V_{CC} pin can be used to power other low voltage parts, as long as their supply current (which should be less than 5mA) is factored into the selection of the resistor.

The shunt regulation voltage is nominally 6.5V, and is guaranteed to lie between 6.0V and 6.9V across the entire operating temperature range and across a wide range of shunt current. Selection of the series resistor is driven by the shunt regulator bias current. The shunt regulator bias will be set by the amount of current flowing through the resistor (based on its value and the voltage drop across it), minus the supply current of the part, including any load drawn from the REF pin, and the load currents of any other devices that take advantage of the 6.5V supply at the V_{CC} pin. The series resistor should be chosen to bias the shunt regulator somewhere between 50µA and 10mA, ideally around 1mA.

These design constraints impose the following limits on the series resistor. The maximum load drawn from the reference, plus the maximum load drawn by other devices connected to the V_{CC} pin, plus 150µA for the LTC2909 must be less than the minimum current through the resistor by at least 50µA:

$$(I_{REF} + I_{DEVICES})_{MAX} + 150µA + 50µA$$
$$\leq \frac{V_{SUPPLY(MIN)} - 6.9V}{R_{SERIES}}$$

This ensures that the shunt regulator is biased with at least 50µA of current. On the other side, the minimum load on the reference, plus the minimum load drawn by other devices on V_{CC} must be less than the maximum current through the resistor by at most 10mA:

$$(I_{REF} + I_{DEVICES})_{MIN} + 10mA$$
$$\geq \frac{V_{SUPPLY(MAX)} - 6V}{R_{SERIES}}$$

This ensures that the regulator is never shunting more than 10mA of current. In summary, the series resistor is required to satisfy:

$$\frac{V_{SUPPLY(MAX)} - 6V}{(I_{REF} + I_{DEVICES})_{MIN} + 10mA} \leq R$$
$$\leq \frac{V_{SUPPLY(MIN)} - 6.9V}{(I_{REF} + I_{DEVICES})_{MAX} + 200µA}$$

As an example, consider operation from an automobile battery. For purposes of this example, the operating range of the battery supply is approximately 10V to 60V, and we can suppose that the user's loading of REF and external current use can each range

Table 4. Nominal ADJ thresholds in comparator mode				
Input		**SEL = GND**	**SEL Open**	**SEL = V_{CC}**
ADJ1	Rising	500.0mV	503.5mV	503.5mV
	Falling	496.5mV	500.0mV	500.0mV
ADJ2	Rising	500.0mV	500.0mV	503.5mV
	Falling	496.5mV	496.5mV	500.0mV

from 0μA to 100μA. The minimum value of R is then 54V/10mA = 5.4k, and the maximum is 3.1V/400μA = 7.75k. Given these constraints, a value of 6.8k is probably optimal.

The above equation is actually overly restrictive. In cases where the supply voltage is very close to the shunt regulation voltage, it may be impossible to satisfy the above equation because the maximum allowable value is less than the minimum. In these cases, it may be assumed that the maximum allowable value is 1k instead of the value predicted by the formula above, as long as the V_{CC} pin is not used to power other devices. There are scenarios where the shunt regulator cannot satisfy the needs for V_{CC} (e.g. those with a very large possible supply range). These applications must use an external voltage regulator of some sort, which, of course, should have a regulation voltage below 6V.

A final consideration is the power dissipation in the series resistor, which may be quite high for high voltage supplies. The series resistor must be rated to handle a power of at least

$$\frac{\left(V_{SUPPLY(MAX)} - 6V\right)^2}{R_{SERIES}}$$

A rough rule of thumb suitable for many applications (those that have fairly constant REF current draw, and have minimum supply voltages well above 6V) is that the resistor rating should be at least 0.1 Watt per 100

Figure 8. ±12V undervoltage monitor with pushbutton reset

volts of maximum supply, multiplied by the ratio of maximum to minimum supply voltage.

Returning to the automobile battery example from above, the power dissipated in the 6.8k resistor could be as large as $54^2/6800 = 0.43W$ (the rule of thumb would give 0.36W), so a 0.5W resistor is best. In reality, of course, the battery is unlikely to stay at 60V for long enough to heat up the resistor substantially. If we were to take a more reasonable DC maximum of 16V, the resistor only needs to handle about 15mW.

Applications

±12V UV Monitor with Manual Reset
Figure 8 shows a LTC2909 configured as an undervoltage monitor for a system with ±12V supplies, and a 1.8V logic bus. The part is powered from the high-availability 12V supply
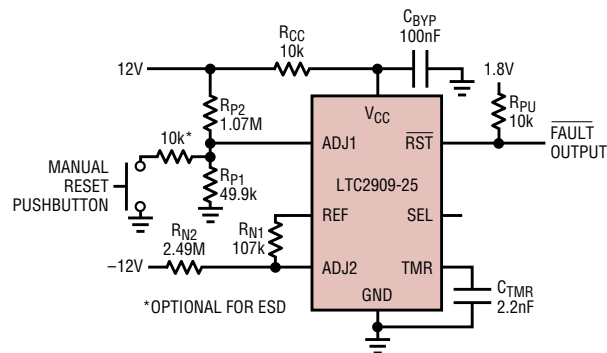
through the series dropping resistor R_{CC}. The floating condition of SEL sets the polarity for one positive and one negative UV. The reset timeout is set to 20ms nominal by C_{TMR}, which allows faster recovery from faults. Finally, the pushbutton allows the user to drive ADJ1 to ground, manually forcing a reset condition. The release of the pushbutton is debounced by the LTC2909's reset timeout. If ESD from people touching the pushbutton is a concern, a 10k resistor in series with the pushbutton limits the current flow into the LTC2909 to prevent damage.

48V Telecom UV/OV Monitor with Hysteresis
Telecom supply specifications usually require some amount of hysteresis in the acceptable voltage range. Since the LTC2909 does not generally have hysteresis in its thresholds, the hysteresis must be externally added. Figure 9 shows the LTC2909 configured to monitor a 48V nominal supply bus for UV and OV. The NMOS devices lower the UV threshold (by reducing R2 for ADJ1) and raise the OV threshold (by reducing R1 for ADJ2) while the RST is high. This has the effect of widening the acceptable supply window once the supply becomes good. The resistors are chosen so that the window is 43.3V–70.2V when the supply is outside the window, and 38.7V–71.6V once the supply is good. Since the part is powered from the 48V bus, the series-dropping resistor is required to be a 0.25W device to handle the power dissipated when the bus is overvoltage.
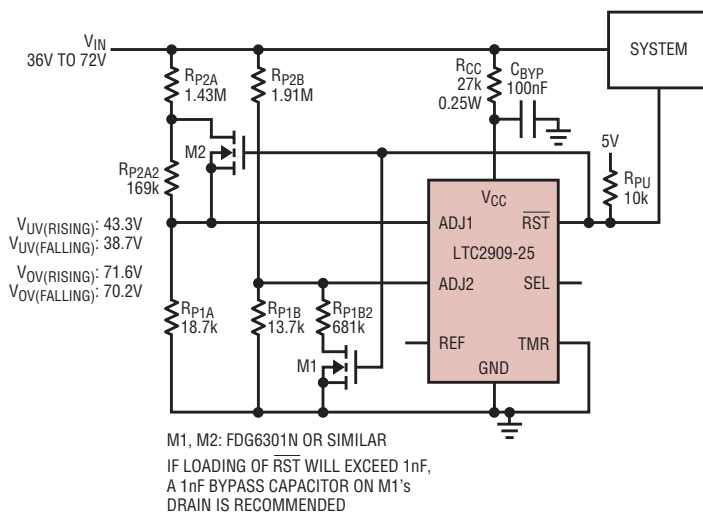

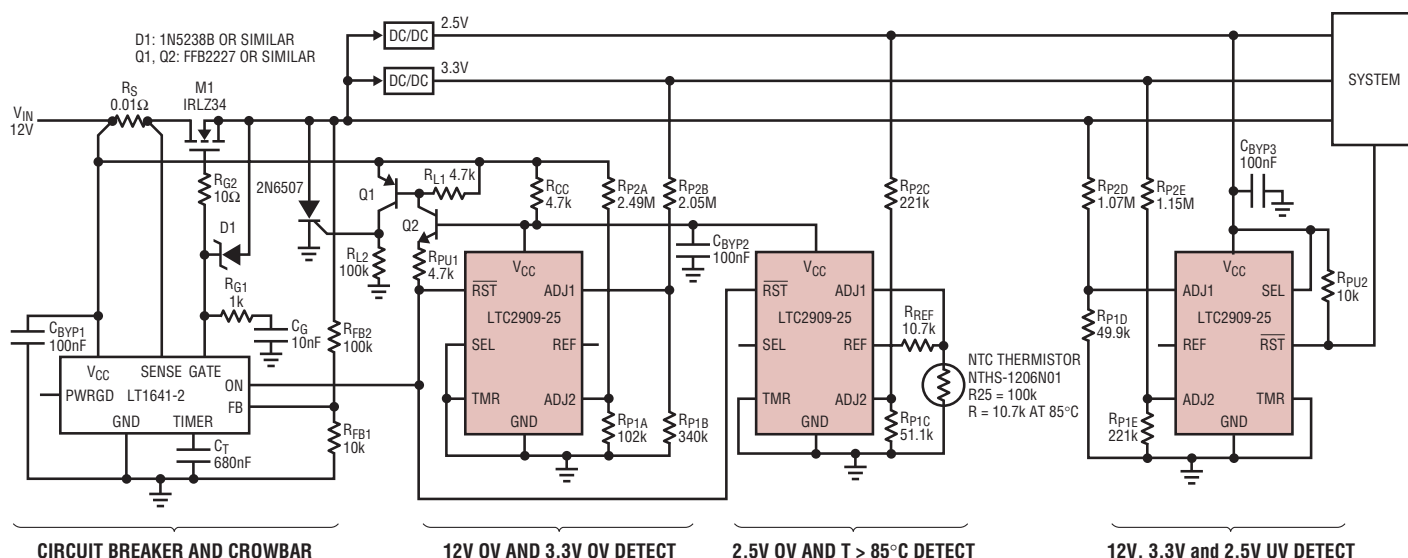Figure 9. A 48V telecom UV/OV monitor with hysteresis

Figure 10. Automotive supply system with overvoltage, overcurrent and overtemperature protection

The recommended NMOS device is FDG6301N, which combines both NMOS devices in one SC70-6 package. Other devices may be used as long as the threshold voltage is guaranteed to be much less than 5V, and the drain-source breakdown is greater than 10V. Note that if the $\overline{RST}$ output is loaded with a large capacitance, the feedback through the gate-drain capacitance of M1 can cause the circuit to oscillate unless a bypass capacitor is placed on M1's drain.

Automotive Supply System

Figure 10 shows three LTC2909s in a full-featured automotive supply system, providing overvoltage, overcurrent, and over-temperature protection in addition to an undervoltage system reset. The system uses an LT1641-2 Hot Swap controller as a controlled electronic circuit breaker. The IRLZ34 logic-NFET serves as the disconnect switch, and the 10mΩ sense resistor sets a current limit of 4.7A. After an over-current fault, the LT1641-2 reconnects after a delay of 160ms (set by C_T).

The two LTC2909s on the left are responsible for detecting overvoltage and over-temperature conditions. To guarantee that they function properly, they must be powered from the 12V input. The V_{CC} pins are tied together, and the supply current flows through just one dropping resistor, so the voltage tends to regulate at whichever of the shunt regulation voltages is the lower of the two.

When any of the supply voltages goes overvoltage, or the temperature sensor is heated above 85°C, the shared $\overline{RST}$ line is pulled low by one of the two LTC2909s. This takes the LT1641-2 ON input low, disconnecting the power switch. At the same time, current is pulled through Q2, turning on Q1, which triggers the 2N6507 SCR and thereby crowbars the 12V supply to the system, removing the overvoltage condition. After the fault condition disappears, the LTC2909s apply a 200ms timeout before reconnecting to the 12V input.

The third LTC2909 serves to provide a master reset to the system when any of the three supplies are undervoltage, whether because insufficient input voltage is present, or because one of the protection faults has tripped. The third monitor function is provided by the UVLO.

A Dale NTHS-1206N01 NTC thermistor with room temperature resistance of 100k is used to detect the temperature, and may be physically located wherever temperature monitoring is needed. The thermistor forms part of a resistor divider from the buffered reference output to ground. As long as the temperature is below 85°C, the thermistor resistance is greater than R_{REF}, so ADJ1 is above its threshold, and $\overline{RST}$ is allowed to go high. If the temperature rises, the thermistor resistance decreases, pulling down on ADJ1, and causing a reset when its resistance is equal to or lower than R_{REF}.

Conclusion

The LTC2909 is a true one-size-fits-all power supply monitor—a way to simplify design and parts stock. It provides a compact solution to monitoring any two supplies for almost any fault condition, where input polarity selection and a buffered reference output allow monitoring of OV conditions and negative supplies. Precision comparators, including a third input on the part's V_{CC}, increase system reliability. To simplify design further, no regulated voltage is required—a built-in shunt regulator on V_{CC} allows operation from a high-voltage high-availability supply. An accurate model of the LTC2909 is included with SwitcherCAD (available at www.linear.com), as an aid to rapid development. *LT*

For more information on parts featured in this issue, go to http://www.linear.com

High Speed Low Power RS485 Transceivers with Integrated Switchable Termination

by Ray Schuler and Steven Tanghe

Introduction

The LTC2859 and LTC2861 combine a logic-selectable integrated termination resistor with a rugged 20Mbps RS485/RS422 transceiver, providing a single die impedance-matched network solution in a tiny package. The low power driver features logic-selectable reduced-slew rate mode for operation below 250kbps with low EMI emissions. The 1/8-unit load receiver provides a failsafe output over the full RS485 common mode range for up to 256 nodes. Both receiver inputs and driver outputs feature robust ESD protection exceeding ±15kV. The half duplex LTC2859 is available in a 3mm × 3mm DFN, while the full duplex LTC2861 is available in both 4mm × 3mm DFN and 16-pin SSOP packages (see photo in Figure 1). Block diagrams for the LTC2859 and LTC2861 are shown in Figure 2.

Switchable Termination

RS485 transceivers typically communicate over twisted-pair cables with characteristic impedance ranging from 100Ω to 120Ω. Proper termination of the cable is important to minimize reflections that can otherwise corrupt

Figure 1. Photograph of the LTC2861 SSOP, LTC2861 4mm × 3mm DFN, and LTC2959 3mm × 3mm DFN packages

the data. The termination usually consists of discrete resistors that have the same resistance as the characteristic impedance of the cable, connected differentially across the cable at both ends. When using LTC2859/LTC2861 transceivers, however, no external resistors are necessary. These devices have an integrated 120Ω resistor on the receiver inputs that can be enabled logically to terminate the cable where needed. Figure 3 shows an example of a properly connected network using LTC2859 transceivers with integrated termination resistors enabled on the two end devices. Short connection to
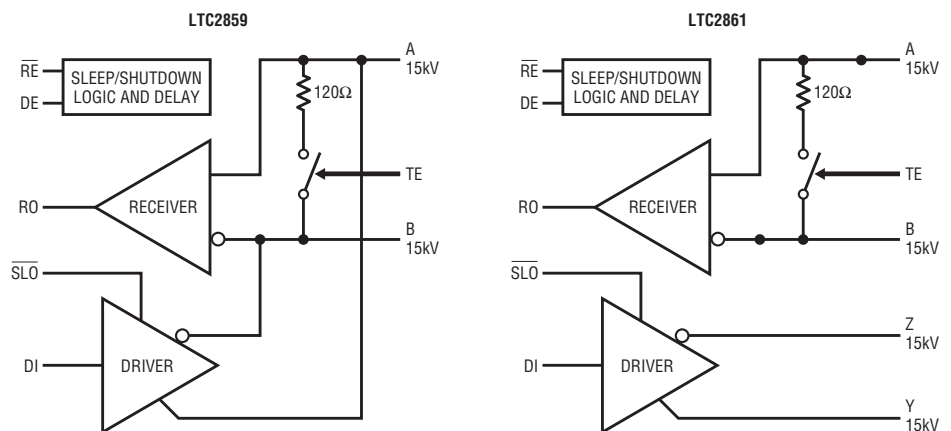
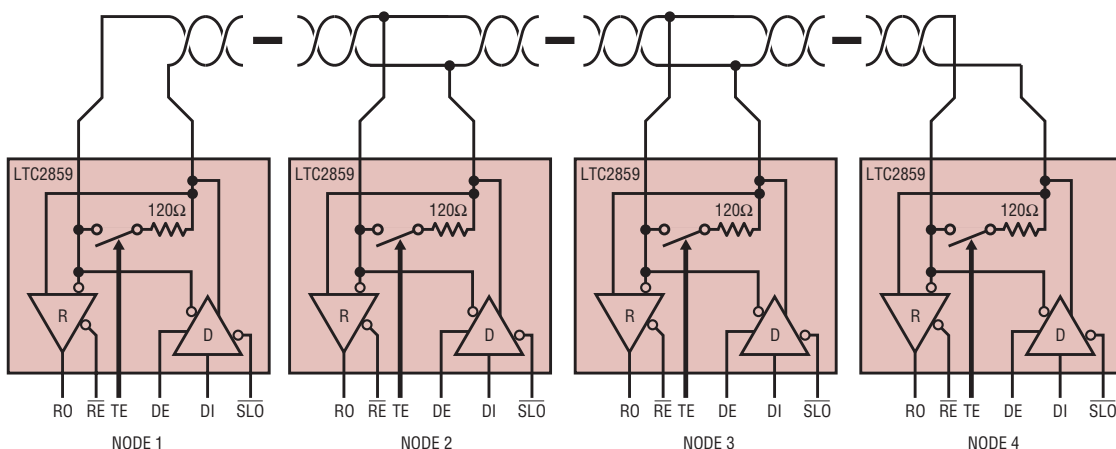

Figure 2. Block diagrams of the LTC2859 and LTC2861.

Figure 3. A properly connected RS485 network using LTC2859 transceivers with selectable terminators.

TE1 = 5V   400ns/DIV
TE2 = 0V
TE3 = 0V
TE4 = 5V

(a)

TE1 = 5V   400ns/DIV
TE2 = 5V
TE3 = 0V
TE4 = 0V

(b)

TE1 = 5V   400ns/DIV
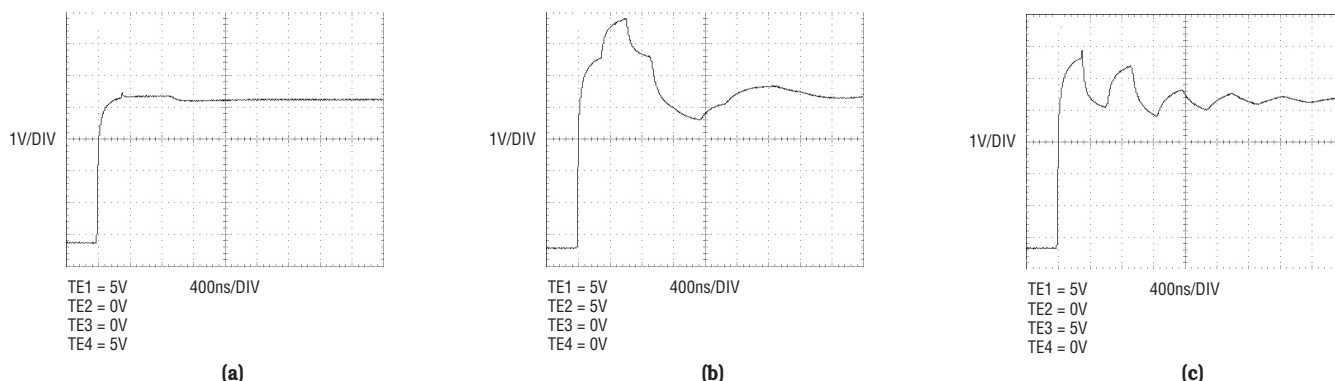TE2 = 0V
TE3 = 5V
TE4 = 0V

(c)

Figure 4. Differential received signals at node 4 from network in Figure 3. properly terminated with termination at nodes 1 and 4 (a). Improperly terminated with termination at nodes 1 and 2 (b), and nodes 1 and 3 (c)
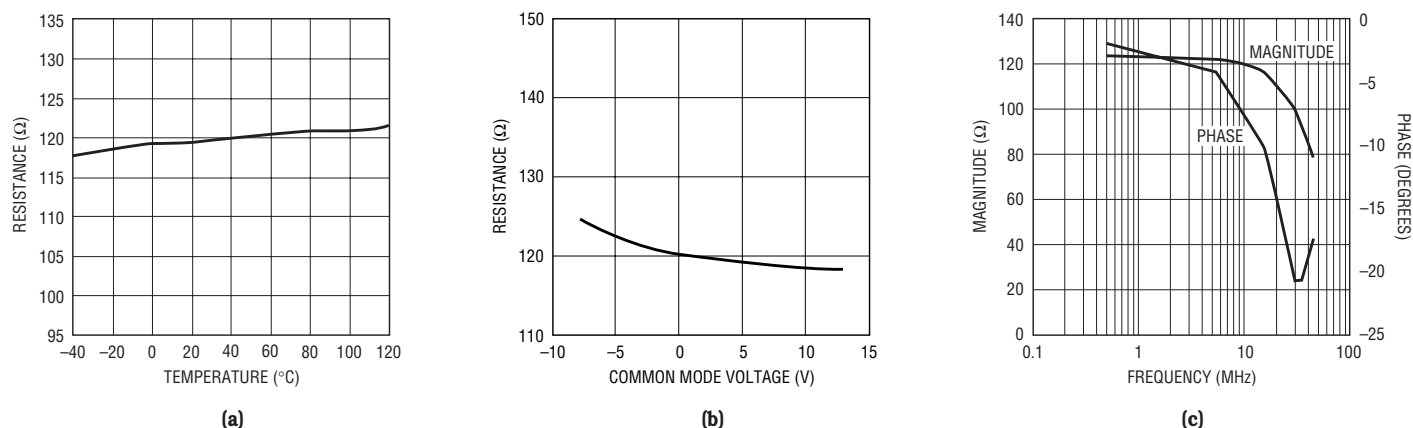

(a)

(b)

(c)

Figure 5. LTC2859 termination resistance vs temperature (a), common mode voltage (b), and frequency (c)

the intermediate stages along the cable should be maintained as these stubs will produce unwanted reflections.

To illustrate the importance of proper termination placement in an RS485 system, consider the network of Figure 3 where four LTC2859 transceivers are spaced equidistantly along three hundred feet of cat5e cable. Signals are driven from node 1 and received at node 4. The integrated termination resistors in the LTC2859 devices are switched in or out at various locations along the cable to illustrate the effect of termination placement on the received waveforms. No external resistors are used.

With proper resistive termination applied at the ends of the line (nodes 1 and 4) the received waveform has clean transitions, as shown in Figure 4a. If the end resistive termination is moved from node 4 to nodes 2 or 3, the waveforms of figures 4b and 4c result, respectively. It is clear that placement of termination resistors

can have a large impact on the signal integrity.

The termination resistor in the LTC2859 and LTC2861 is enabled by pulling the termination enable (TE) pin high. The resistor is disconnected when the termination enable is pulled to a logic low or the device is unpowered. Figures 5 (a), (b), and (c) show the resistance is maintained well over temperature, common mode voltage, and frequency.

The inclusion of a selectable 120Ω resistive termination on the LTC2859 and LTC2861 is a significant advantage over other RS485 transceivers. When modifications or additions are made to an RS485 network, the required termination changes can be made by logical control of the termination enable pin on the desired transceiver. This can be done through the use of a simple jumper or through higher level system control where manual intervention is prohibitive. A valuable benefit is that *every* node in the network is capable of

providing termination without the use of external resistors, making network re-configuration more manageable.

Controller based configuration of a network with the aid of switchable termination can be extended beyond simple network additions and reductions to include fault protection as well. RS485 networks supporting bus lengths up to 4000 feet are at risk of breaks or disconnects in the cabling that can interrupt service. Figure 6 shows a dual-master controlled ring network making use of the LTC2859 logic-controlled termination to protect against such open cable faults.

When a break in a network is detected, through loss of response from one or more nodes, the master controller enters a low data rate mode. Even though the break imposes a severe impedance mismatch, low data rate communication is still possible up to the point of the break (as a rule of thumb, communication without termination is possible if the two-way
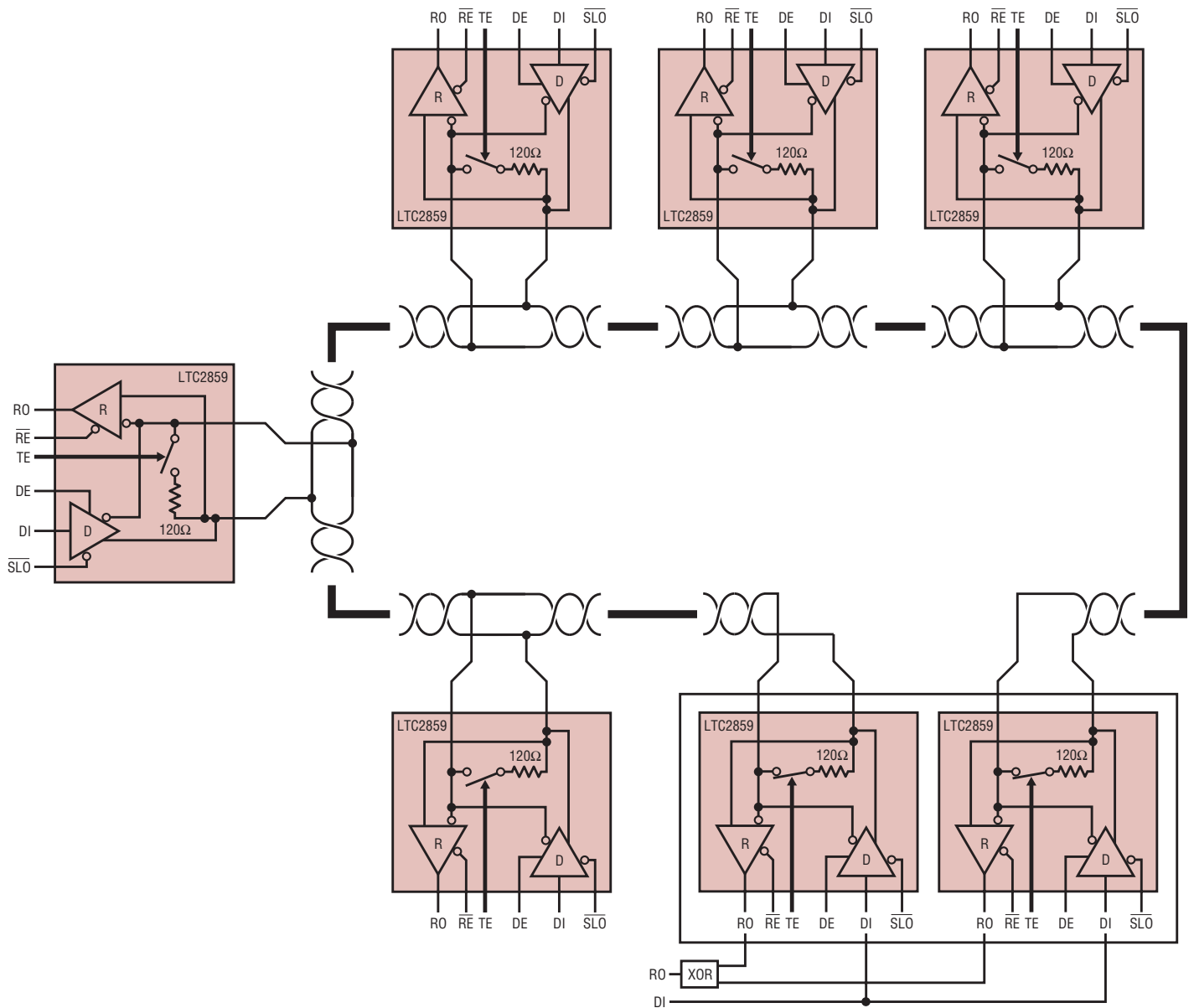
Figure 6. Line break-tolerant RS485 topology

cable propagation delay is less than 10% of the bit time). Each sequentially addressed transceiver on the network is polled by the masters to determine where the break has occurred. If the network is constructed of LTC2859 and LTC2861 transceivers with selectable termination, the masters can instruct the two nodes on either side of the line break to enable termination resistance. The two master controllers may now access each node of the bus at high data rates until physical repairs have been made. An XOR of the RO pin from each transceiver on the master controller can optionally

reduce the I/O pin count to the micro controller.

Driver

The LTC2859/LTC2861 drivers can deliver RS485/RS422 signals up to 20Mbps. Figure 7 shows waveforms of the part operating at the maximum data rate. The LTC2859/LTC2861 also feature a reduced-slew rate mode ($\overline{SLO}$ mode), which is entered by setting the $\overline{SLO}$ pin to a logic low level.

$\overline{SLO}$ mode increases the driver transition time to reduce high frequency EMI emissions from equipment and cables. In this mode the driver data

rate is limited to about 250kbps. Slew limiting also mitigates the adverse affects of improper line termination and long stubs.

Figure 7. Driver outputs toggling at the maximum data rate of 20Mbps

1.5A VLDO Operates Down to 0.4V Output and Maintains 100mV Dropout

by Bill Walter

Introduction

As the power supply voltage levels for digital ICs continue to fall, supply noise and efficiency become increasing concerns. Although many efficient switching DC/DC converters are available, switching noise at the output can be unacceptable at low supply levels, or the size can be prohibitive where multiple supply voltages are required. One commonly used solution is to follow a switching DC/DC converter with a low dropout (LDO) regulator to provide an efficient low noise supply. Nevertheless, LDOs can introduce other problems when used as a low voltage post regulator, such as requiring large and expensive tantalum capacitors, dropping out at a voltage of 300mV or more, requiring supply voltages greater than 1.8V, no ability to produce output voltages below 1.2V, poor supply rejection, poor load regulation, etc.

The LTC3026 is a 1.5A VLDO with input voltage capability down to 1.14V and a low adjustable output voltage from 0.4V to 2.6V. The part also has a very low dropout voltage of only 100mV while delivering up to 1.5A of output current, enabling it to optimize battery run time from single cell applications

with a low V_{IN} to V_{OUT} differential. The low V_{IN} capability is important as many emerging handheld applications are using a 1.5V main rail, and require output voltages from 1.25V to 0.5V to drive low voltage microprocessor and microcontroller cores. Additionally, the LTC3026 offers low input quiescent current (<1mA) at 1.5V input and less than 1µA in shutdown, maximizing run time in battery-powered applications. Internal protection circuitry includes current limiting, thermal limiting, and reverse-current
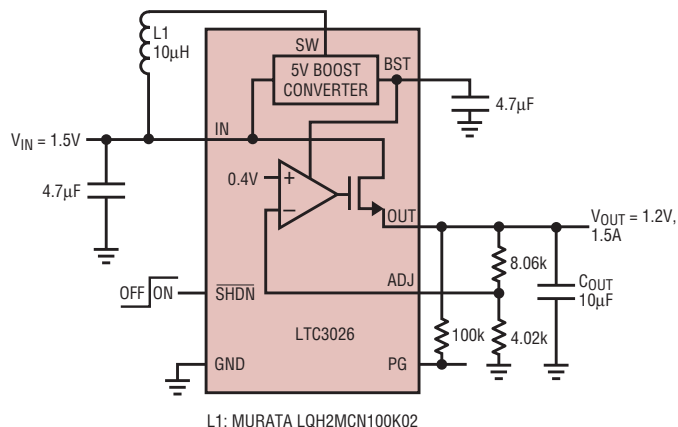

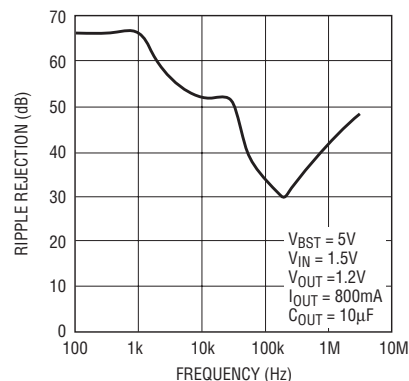**Figure 1. Block diagram of the LTC3026 and typical application circuit**

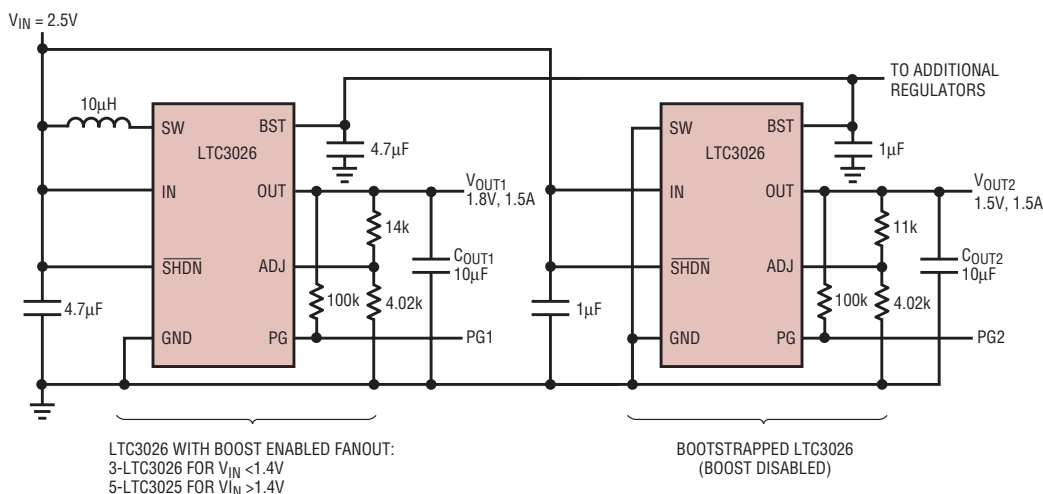**Figure 2. Ripple rejection of the LTC3026**

Figure 3. Multiple LTC3026s operating from single boost converter.

protection. The LTC3026 regulator is available in a low profile 10-lead DFN (3mm × 3mm × 0.75mm) and 10-pin MSOP packages with exposed pad, offering a very compact and thermally efficient solution.

Low Voltage Operation and Excellent Ripple Rejection

To allow operation at low input voltages, the LTC3026 includes a boost converter that provides the necessary headroom for the internal LDO circuitry as shown in Figure 1. This feature offers the flexibility of lower input voltage capability if an external 5V rail is not available in the system. Figure 2 shows the input ripple rejection for the application circuit of Figure 1. From Figure 2 we see that the ripple rejection is greater than 30dB all the way up to 3MHz. The exceptional ripple rejection makes the LTC3026 an excellent choice for post DC/DC switching supply regulation.

Multiple VLDO Outputs from Single Boost

The boost converter of the LTC3026 can be disabled by connecting the SW pin to ground (GND). Disabling the boost converter allows the Boost pin to be driven by an external 5V supply or from the Boost pin of a second LTC3026 with boost converter

Figure 4. Dropout voltage vs output current for various output voltages

enabled. Figure 3 shows an application with the boost converter of one LTC3026 driving the boost pin of a second LTC3026 with boost converter disabled. In this application up to five LTC3026 LDOs can be bootstrapped from one LTC3026 with its boost converter enabled. Figure 4 shows that the dropout voltage of the LTC3026 is very low and well controlled over a wide range of output voltages.

Excellent Transient Response and Efficiency Post DC/DC Buck Converter

Figure 5 shows a typical application for the LTC3026 as post regulator for a buck DC/DC switching converter. The LTC1773 an efficient 1.8V output while the LTC3026 provides over 80%

efficiency dropping the 1.8V input down to 1.5V. In this application the LTC3026 boost converter is disabled and the Boost pin is driven by the external 5V supply.

The fast transient response of the LTC3026's output stage overcomes the traditional tradeoffs between dropout voltage, quiescent current and load transient response inherent in most LDO regulator architectures. Figure 6 shows that the LTC3026 only undershoots about 10mV for a full load transient step (0A to 1.5A), and overshoots about 30mV for a full load transient step (1.5A to 0mA). Additionally, the LTC3026 is designed to detect an overshoot condition and automatically loads the output to bring it back into regulation, as can
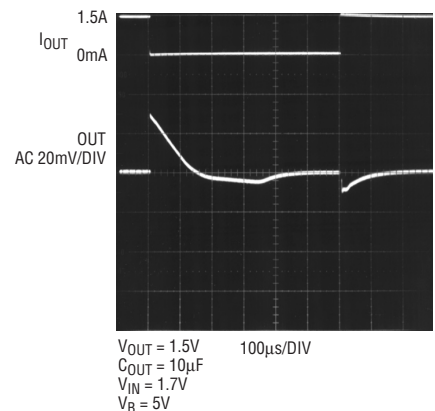
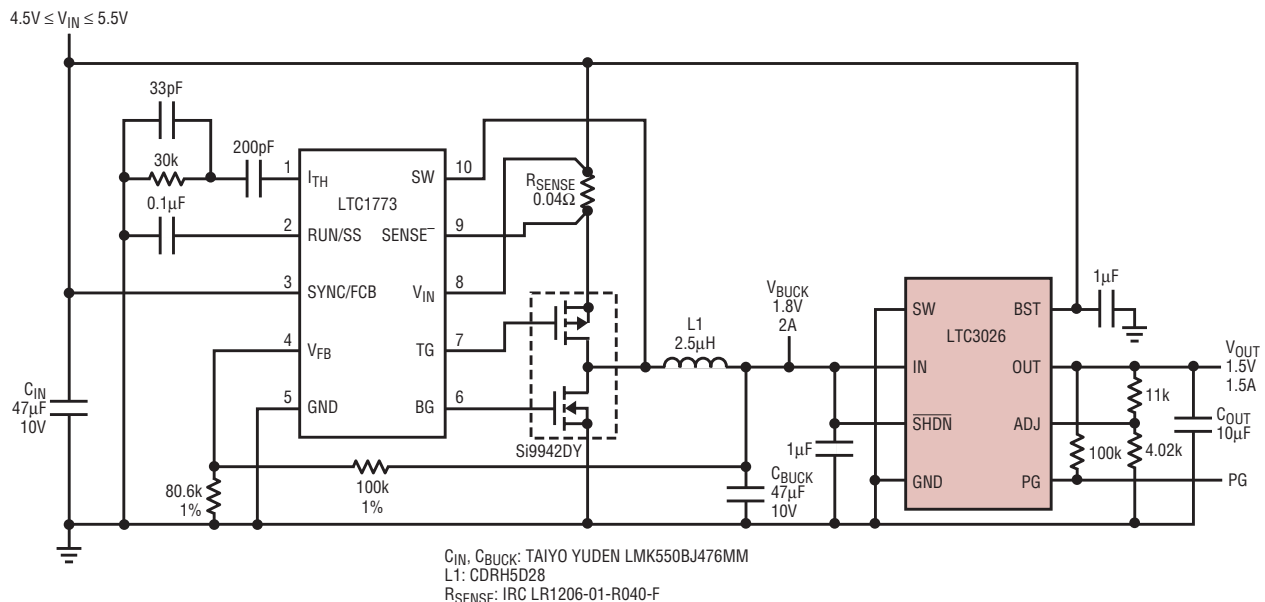**Figure 6. Output load transient response**

Figure 5. Efficient low noise 1.5V output from 1.8V DC/DC buck converter

be seen in Figure 6. Once regulation is achieved the part disables the output load thus keeping supply current low. The transient scope photo also shows that the LTC3026 has excellent load regulation. The LTC3026 is designed to be stable with a wide range of ceramic output capacitors as small as 10µF.

Conclusion

The LTC3026 is an excellent choice for low voltage applications where efficient, low noise supplies are required. The exceptional ripple rejection and very low dropout of the LTC3026 makes it especially well suited as a post regulator for switching supplies. *LT*

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Figure 8 shows single-ended and differential driver outputs in normal and $\overline{SLO}$ mode with corresponding frequency spectrums operating at 250kbps. $\overline{SLO}$ mode significantly reduces the high frequency harmonics.

The LTC2859 and LTC2861 drivers feature current limiting that protects them from faults such as shorting the outputs to the power supply or ground. Short circuit current is limited to below the ±250mA RS485 standard, with typical clamp currents of ±150mA. If fault voltages are greater than approximately ±10V, currents are reduced further to limit power dissipation. The LTC2859 and LTC2861 also feature thermal shutdown protection, disabling the part if a fault condition causes it to overheat. Figure 9 shows the driver output I-V characteristic when driven by a curve tracer. Overcurrent protection engages on the positive and negative sweeps limiting the driver output current.

Receiver

The LTC2859 features a low power receiver using just 540µA of current (typical). The LTC2859/LTC2861 fail-safe feature guarantees the receiver output to be a logic HIGH state when the inputs are shorted, open, or terminated, but not driven for more than about 3µs. The delay prevents signal zero crossings from being interpreted as a shorted input and causing RO to go inadvertently high. This failsafe feature is guaranteed to work for inputs spanning the entire common mode range of –7V to +12V.

The receiver output is internally driven high (to V_{CC}) or low (to ground) with no external pull-up needed. The RO pin of the disabled receiver becomes high impedance with leakage
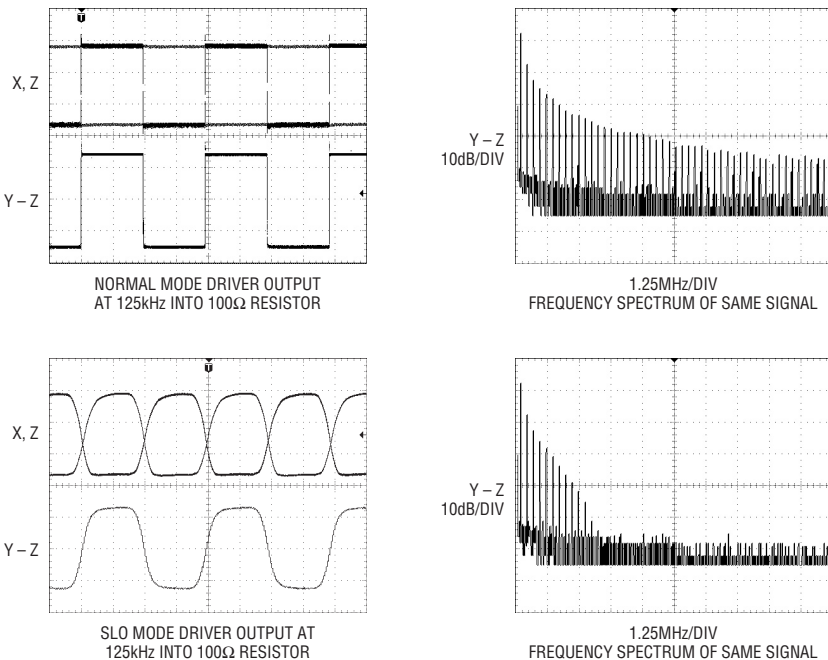

X, Z

Y – Z

NORMAL MODE DRIVER OUTPUT AT 125kHz INTO 100Ω RESISTOR

Y – Z
10dB/DIV

1.25MHz/DIV
FREQUENCY SPECTRUM OF SAME SIGNAL

X, Z

Y – Z

SLO MODE DRIVER OUTPUT AT 125kHz INTO 100Ω RESISTOR

Y – Z
10dB/DIV

1.25MHz/DIV
FREQUENCY SPECTRUM OF SAME SIGNAL

Figure 8. Time and frequency domain waveforms of the LTC2859 driver output in normal and reduced-EMI $\overline{SLO}$ mode

50mA/DIV

DE = 5V
DI = 0V

5V/DIV
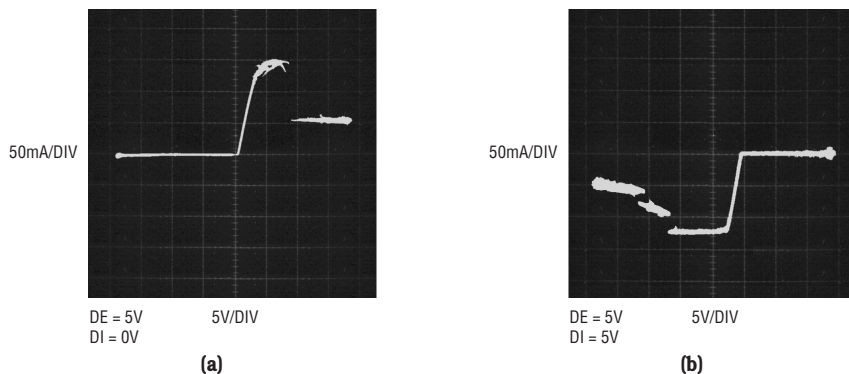
(a)

50mA/DIV

DE = 5V
DI = 5V

5V/DIV

(b)

Figure 9. Curve traced I-V characteristic of the LTC2861 driver output showing current limiting. (a) Pin is driven low by the LTC2861 driver. (b) Pin is driven high by the LTC2861 driver

of less than ±1µA for voltages within the supply range.

Conclusion

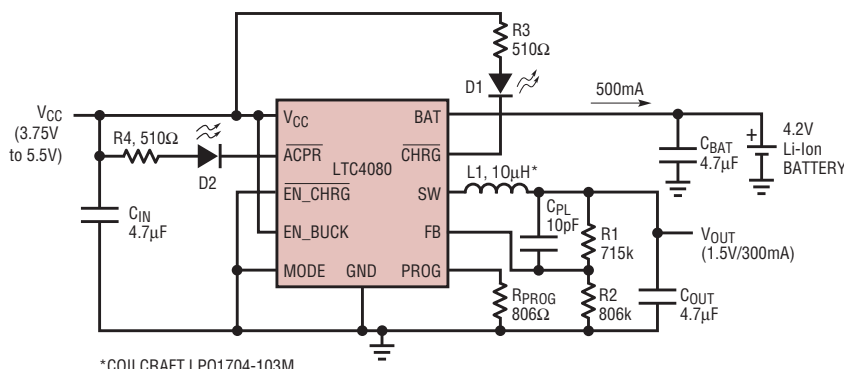Improperly terminated RS485 cabling can severely distort signals leading to losses in data integrity. Correcting network termination without logic-selectable termination often requires physical inspection of expansive networks. The inclusion of selectable termination resistance on the rugged LTC2859 and LTC2861 provide complete solutions to RS485 networking with next-generation remote network tuning capability. *LT*

A Complete 500mA Linear Charger and 300mA Synchronous Buck Converter in a Tiny 3mm × 3mm DFN Package

by Ashish Kirtania

Introduction

The LTC4080 is a full-featured, single-cell 4.2V Li-Ion battery charger with an integrated synchronous buck DC/DC converter designed primarily for handheld applications. Its tiny 3mm × 3mm DFN package and low external component count provide space-savings in today's crowded circuit boards. The high operating frequency (2.25MHz) of the switching regulator minimizes overall solution footprint further by allowing the use of tiny, low profile inductors and ceramic capacitors. To extend battery life, the buck regulator offers high efficiency burst mode operation in which the

Figure 1. Full featured Li-Ion charger with thermal management and efficient buck regulator in a compact, single IC solution

regulator typically consumes only 20µA at no load.

Battery Charger Features

The LTC4080 battery charger uses a unique constant-current, constant-voltage, constant-temperature algorithm with programmable charge current up to 500mA and a final float voltage of 4.2V±0.5%. The maximum charge current is programmed using a single external resistor (R_{PROG}) from the PROG pin to ground. The charge current (I_{CHRG}) out of the BAT pin can be determined at any time by monitoring the PROG pin voltage (V_{PROG}) and applying the following equation:

$$I_{CHRG} = \frac{V_{PROG}}{R_{PROG}} \bullet 400$$

In typical operation, the charge cycle begins in constant current mode. When the battery approaches the final float voltage of 4.2V, the charge current starts to decrease as the battery charger switches to constant-voltage mode. When the charge current drops to 10% of the full-scale charge current, commonly referred to as the C/10 point, the open-drain charge status pin, $\overline{CHRG}$, assumes a high impedance state.

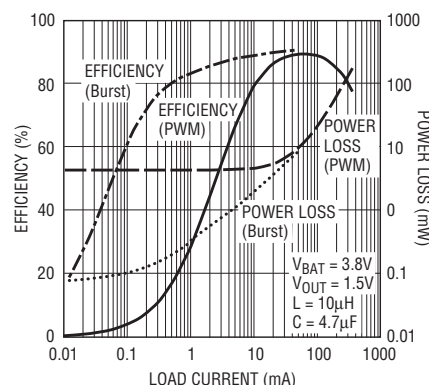An internal thermal regulator reduces the programmed charge current

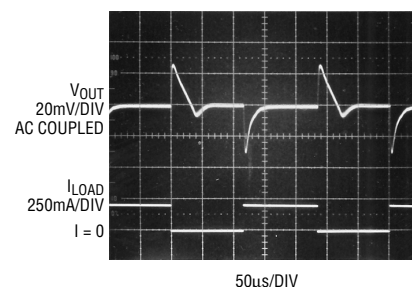**Figure 2. Efficiency of the buck regulator in Figure 1**

Figure 3. Transient response of the buck regulator to a 0.5mA–200mA load step

if the die temperature attempts to rise above a preset value of approximately 115°C. This feature not only protects the LTC4080 and external components from excessive temperature, it can also reduce the charge time by allowing the user to set a higher maximum charge current—essentially taking into account typical, instead of worst case, ambient temperatures for a given application.

An internal safety timer sets the maximum time for a charge cycle, typically 4.5 hours. When this time elapses, the charge cycle terminates and the $\overline{\text{CHRG}}$ pin assumes a high impedance state even if C/10 has not been reached yet. A new charge cycle of 2.25hr automatically starts if the battery voltage falls below the recharge threshold (typically 4.1V).

Trickle Charge and Defective Battery Detection

At the beginning of a charge cycle, if the battery voltage is below 2.9V, the charger goes into trickle charge mode, reducing the charge current to 10% of the programmed value. If the low voltage condition persists for one quarter of the charge cycle (1.125hr), the battery is assumed to be defective, the charge cycle terminates and the $\overline{\text{CHRG}}$ output blinks at a frequency of 2Hz with a 75% duty cycle. If, for any reason, the battery voltage rises above 2.9V, the charge cycle restarts.

Undervoltage Lockout

An internal undervoltage lockout circuit monitors the input voltage and keeps the battery charger in shutdown until the input rises above 3.6V and approximately 80mV above the battery voltage. The undervoltage condition is indicated by a high-impedance state of the open-drain status output pin $\overline{\text{ACPR}}$.

Undervoltage Charge Current Limiting

The battery charger in the LTC4080 includes undervoltage charge current limiting that prevents full charge current until the input supply voltage reaches approximately 300mV above the battery voltage. This feature is particularly useful if the LTC4080 is powered from a supply with long leads or any relatively high output impedance.

Buck Converter Features

The buck converter in LTC4080 is powered from the BAT pin and has a programmable output voltage (0.8V to V_{BAT}) providing a maximum load current of 300mA. It has two modes of operation, constant frequency mode and Burst Mode operation, selectable via the MODE pin. In constant frequency mode, also referred to as PWM mode, the switching regulator uses current mode control scheme with internal compensation and provides efficiencies up to 91% with very low ripple. The operating frequency of the switching regulator is set at 2.25MHz to minimize possible interference with the AM band. The switching regulator and the battery charger can run simultaneously or independently of each other.

Burst Mode Operation

Burst Mode operation offers higher efficiency at light loads at the cost of higher ripple at the output voltage. In this mode, the inductor current swings between a maximum value (I_{PEAK}) and a minimum value (I_{ZERO}) irrespective of the load as long as the FB pin voltage (V_{FB}) is less than the reference voltage of 0.8V. Once V_{FB} exceeds 0.8V, the control logic turns off both switches along with most of the circuitry and the regulator draws only about 20μA from the battery. When the output voltage drops about 2% from its nominal value, the switching regulator wakes up and the inductor current starts ramping again. To minimize the output voltage ripple, the regulator is limited to a maximum load current of 55mA in Burst Mode operation.

Short-Circuit Protection

In the event of a short circuit at the output or during start-up, the shallow negative slope (~V_{OUT}/L) of the inductor current may prevent the inductor from discharging enough to avoid a cumulative runaway situation over a number of switching cycles. Even the hard current limit on the main PMOS switch is no guarantee against inductor current runaway because of current sense blanking. The switching regulator in the LTC4080 prevents inductor current runaway by imposing a current limit on the synchronous NMOS switch. If the inductor current through the NMOS switch at the end of a discharge cycle is not below this limit, the regulator skips the next inductor charging cycle.

Buck Undervoltage Lockout

To prevent unreliable operation, when V_{BAT} is less than 2.7V, an undervoltage lockout circuit prevents the switching regulator from turning on. However, if the regulator is already running and the battery voltage is dropping, the undervoltage comparator does not turn it off until V_{BAT} becomes less than 2.5V.

Global Thermal Shutdown

The LTC4080 includes a global thermal shutdown which turns off the entire part (both battery charger and switching regulator) if the die temperature exceeds 160°C. The part resumes normal operation once the temperature drops approximately 14°C.

Conclusion

The LTC4080, with its complete Li-Ion battery charger and a moderately high current buck converter in a small 3mm × 3mm package, offers a very compact solution with minimum number of external components. Thermal regulation of the battery charger and the high efficiency of the converter reduce charge times and simplify thermal management. *LT*

for the latest information on LTC products, visit www.linear.com

Triple Amps Handle High-Res Workstation Video from Single Supplies

by Jon Munson

Introduction

Cutting edge, high resolution workstation displays demand cutting edge bandwidth and slew rate specs of their video amplifiers. Displays supporting 1920 × 1200 pixels, for example, must handle over 200Mpixels/s, less than five nanoseconds per pixel! System supply voltages have also been dropping in order to accommodate the new, faster digital processor technologies. Historically the fast amplifiers used for these formats required a total power supply of at least 6V, particularly in cable driver applications. Enter the LT6557 and LT6558, ultra-fast triple video amplifiers with internally fixed gains of two and unity, respectively. These devices have been specifically engineered to operate on single supply voltages down to 3.3V and yet maintain high bandwidth. Now it is practical to use a low voltage digital supply to directly power the analog video circuitry within high resolution products.

Features for Performance and Ease-of-Use

The LT6557 and LT6558 are bipolar voltage-feedback topology parts that are designed for exceptionally high slew-rate and large output swing capabilities for their operating voltage. A blazing slew rate of 2200V/µs is responsible for assuring that 400MHz of bandwidth is available, regardless of signal amplitude. These parts also include a single-resistor-programmable biasing feature that eliminates having to place resistor networks in the signal path to establish the correct DC operating point in single-supply operation. In the typical application for the LT6557 as a cable-driver or the LT6558 as an input buffer, as shown in Figure 1 and Figure 2, the only external components in the signal path are the coupling capacitors and termination resistors, simplifying layout and preventing frequency response anomalies.

The output voltage of these amplifiers can swing to within 800mV of either rail, thus there is 1.7V of swing available on 3.3V, and 3.4V of swing on 5.0V. This means that there is plenty of swing available for RGB or HD video waveforms, plus allowance for offset variations due to AC-coupled picture content, for the unity gain LT6558 to operate on 3.3V or the LT6557 with gain of two to operate on 5V. Figure 3 shows the time response of the LT6558 to 700mV$_{P-P}$ pulses 6ns wide while operating from 3.3V (as in Figure 2). Note that for high fidelity waveform capture, a coupling circuit like that in Figure 1 is used, with a blocking capacitor and the signal measurement taken after a 75Ω double-termination. The same circuit operating at 5V exhibits even less overshoot. The typical current consumption is about 22mA per amplifier, and an enable feature is provided to permit a less than 1mA total current draw when the part is not in use. Both parts are available in a leaded SSOP-16 package or the leadless 5mm × 3mm DFN-16. The DFN model includes a bottom-side ground pad for enhanced thermal performance.
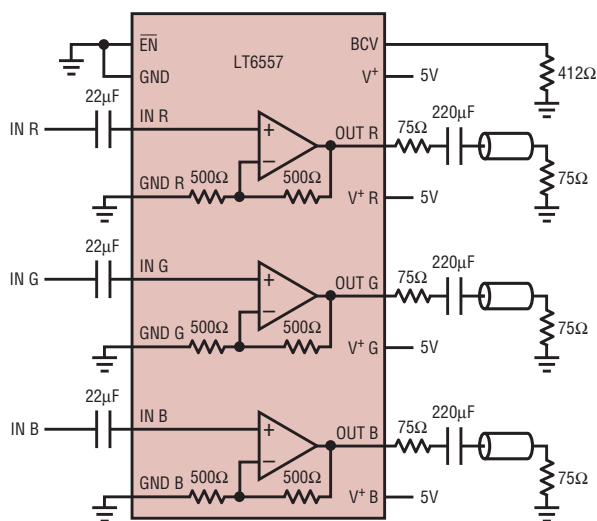

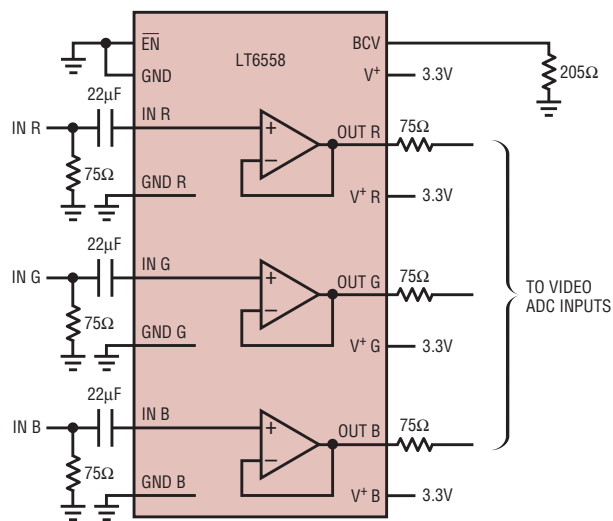
Figure 1. An LT6557 single-supply RGB cable driver

Figure 2. An LT6558 single-supply RGB buffer/ADC driver

500mA Output Current Low Noise Dual Mode Charge Pump
by Yang Wen

Introduction

Charge pump (inductorless) DC/DC converters are popular in space-constrained applications with low to moderate load current (10mA–500mA) requirements. The devices in the LTC3203 family are low noise, high efficiency regulating charge pumps that can supply up to 500mA of output current from a single 2.7V to 5.5V supply. The LTC3203-1 and LTC3203B-1 produce a selectable fixed 4.5V or 5V output. The LTC3203B produces an adjustable output voltage. The LTC3203-1 features automatic Burst Mode operation at light load to achieve low supply current whereas the LTC3203B and LTC3203B-1 operate at constant frequency to minimize both input and output noise. High switching frequency (1MHz) makes it possible to use only four tiny low cost ceramic capacitors and two resistors for operation. The device also has two user selectable conversion modes for optimizing the efficiency of the charge pump. Additional features include low shutdown current (<1µA), soft-start at power-on and short circuit protection. The LTC3203 family is available in a 10-lead thermally enhanced DFN package, making it possible to build a complete converter in less than 0.04in^2. A typical application circuit is shown in Figure 1.

Low Noise Operation

The constant frequency architecture achieves regulation by sensing the output voltage and regulating the amount of charge transferred per cycle. This method of regulation provides much lower input and output voltage ripple than that of burst mode regulated switched capacitor charge pumps. The LTC3203B and LTC3203B-1 make filtering input and output noise less demanding than burst mode switched capacitor charge pumps where switching frequencies depend on load current and can range over sev-
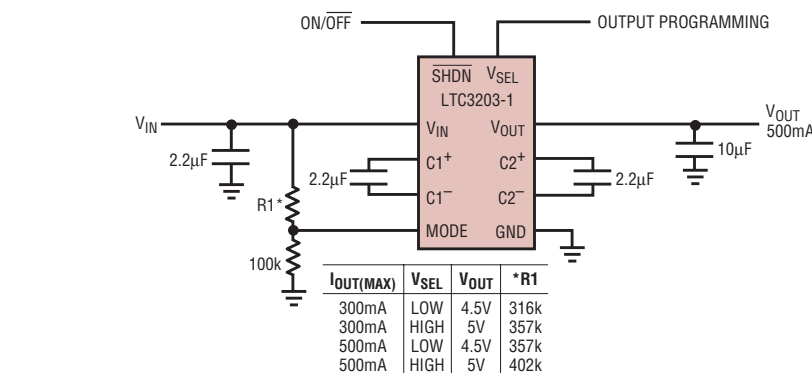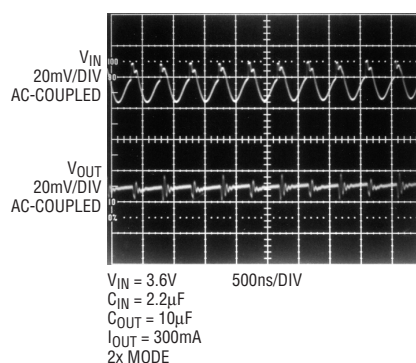

$I_{OUT(MAX)}$	V_{SEL}	V_{OUT}	*R1
300mA	LOW	4.5V	316k
300mA	HIGH	5V	357k
500mA	LOW	4.5V	357k
500mA	HIGH	5V	402k

Figure 1. Typical application

V_{IN} = 3.6V
C_{IN} = 2.2µF
C_{OUT} = 10µF
I_{OUT} = 300mA
2x MODE

Figure 3. Input and output noise in 2× mode

V_{IN} = 4V
C_{IN} = 2.2µF
C_{OUT} = 10µF
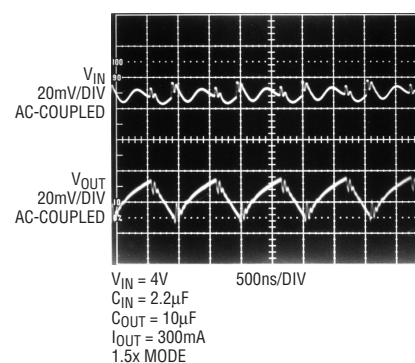I_{OUT} = 300mA
1.5x MODE

Figure 2. Input and output noise in 1.5× mode

eral orders of magnitude. The charge pump operates on two phases, where a break-before-make circuit prevents switch cross-conduction. The higher frequency noise due to the non-overlap "notches" is easily filtered by a small input capacitor and PCB parasitic inductance. Figures 2 and 3 shows the low input and output ripple with a 300mA load. The device is powered from a 3.6V input and produces a regulated 4.5V output. The input voltage source has 0.1Ω impedance.

Dual Mode Conversion

The LTC3203 family offers both 1.5× and 2× boost modes—selected by the mode pin. In the 2× mode, the chip works as a dual-phase regulated voltage doubler. The flying capacitors are charged on alternate clock phases from V_{IN}. While one capacitor is being charged from V_{IN}, the other is stacked on top of V_{IN} and connected to the output. The two flying capacitors operate out of phase to minimize both input and output ripple. Alternatively, in 1.5× mode, it uses a split-capacitor technique rather than doubling. The flying capacitors are charged in series during the first clock phase, and stacked in parallel on top of V_{IN} on the second clock phase. With this technique, the input current is reduced from more than twice the load current to just over 1.5 times the load current, resulting in approximately 25% less input current than what would be required for operating in 2× mode charge pump to drive the same load. Therefore, the efficiency at higher V_{IN} is increased to approximately 90% with V_{IN} at somewhere between 3V and 4V. Figure 4 shows the conversion efficiency at 300mA load current for 4.5V V_{OUT} and 5V V_{OUT}, respectively.

The conversion mode should be chosen based on considerations of efficiency, available output current and V_{OUT} ripple. With a given V_{IN}, the 1.5× mode gives a higher efficiency at lower available output current. The 2× mode gives a higher available output current at lower efficiency. Moreover, the output voltage ripple in the 2× mode is lower due to the out-of-phase operation of the two flying capacitors. Typically, at low V_{IN}, the 2× mode should be selected, and at higher V_{IN}, the 1.5× mode should be selected.

The MODE pin has a precision comparator. By connecting a resistive divider from V_{IN} to the MODE input pin, the user can accurately program

Figure 4. Efficiency vs V_{IN} at 300mA load current

the V_{IN} threshold at which the charge pump will switch from 1.5× mode to 2× mode as V_{IN} falls and vice versa.

The 10% hysteresis on the MODE pin prevents the chip from hunting between the two modes.

Conclusion

With low operating current, low external parts count and robust protection features, the LTC3203 family is well suited for low power step-up/step-down DC/DC conversion. The shutdown, dual mode conversion, selectable output voltage and low noise operation features provide additional value and functionality. The simple and versatile LTC3203 family is ideal for moderate power DC/DC conversion applications. ⟋⟋

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power than a slot with a 20% accurate circuit breaker.

Detect Insertion Events via the ENABLE Pin

The $\overline{EN}$ pin can be used to sense the insertion of a board when the LTC4215 is used in backplane resident application. A short pin on the connector pulls $\overline{EN}$ to ground once the other, longer pins have already been connected. Once the $\overline{EN}$ pin crosses its falling 1.107V threshold the LTC4215 turns on the external switch after a 100ms debounce delay. Because a falling edge on the $\overline{EN}$ pin corresponds to the

insertion of a new board, the LTC4215 clears the fault register (except for the $\overline{EN}$ Changed State bit) so that a previously recorded fault does not prevent the new board from starting up. Whenever the $\overline{EN}$ pin rises or falls, the $\overline{EN}$ Changed State bit in the FAULT register is set to indicate that a board has either been inserted or removed. A STATUS register bit contains the complement of the state of the $\overline{EN}$ pin to indicate if a board is present. When the board is unplugged, the short $\overline{EN}$ pin is the first to disconnect. The $\overline{EN}$ pin pulls up with an internal 10µA current source until the voltage reaches the rising 1.235V threshold,

at which point the external switch is turned off with a 1mA current.

Conclusion

The LTC4215 is a smart power gateway for hot swappable circuits. It provides fault isolation, closely monitors the health of the power path and provides an unprecedented level of control over the inrush current profile. It logs faults, provides real-time status information, and can interrupt the host if necessary. Meanwhile an internal 8-bit ADC continuously monitors board current and voltages. These features make the LTC4215 an ideal power gateway for high availability systems. ⟋⟋

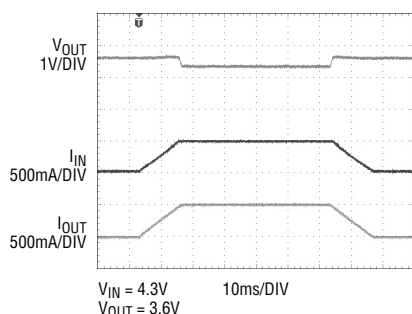DC/DC converter may be dynamically programmed by sourcing or sinking

Figure 10. As load increases, the input current is clamped to 500mA using the circuit of Figure 9

current at the FB node. Referring to Figure 9, the equation for the input current clamp level is:

$$I_{CLAMP} = \frac{1.22V}{R3} \cdot \frac{R1}{R2}$$

Figure 10 shows V_{OUT} dropping when input current reaches 500mA as the load increases. In USB applications where the input voltage is nominally 5V, a Schottky diode is used to limit peak voltages on the SW1 pin.

Conclusion

Linear Technology's new LTC3532 synchronous buck-boost converter

simplifies the design of Lithium-Ion or multi-cell powered handheld electronics. With a highly efficient automatic Burst Mode operation, the converter maximizes battery life in portable devices with widely varying load requirements. Soft start, programmable switching frequency and external compensation make the LTC3532 suitable to a wide variety of applications. Two package options, an MS10 leaded package and a 3mm × 3mm DFN, plus the ability to operate efficiently at high frequency, enable the designer to minimize board area and component height. ⟋⟋

Tiny Buck-Boost Converter for Low Current Applications

by Eddy Wells

Introduction

One common challenge for many battery powered portable applications is creating a regulated output voltage above or below the input source. Traditional buck-boost approaches, such as a dual inductor SEPIC converter or cascaded regulators, are unacceptable in most portable devices because of their large solution size and low efficiency. Smaller footprint, integrated charge-pump solutions can switch between buck and boost operation, but charge-pumps achieve good efficiency at only a few operating voltages, while efficiency dips below 50% at others. Another compact and simple approach forgoes a portion of the battery capacity and uses a buck (step-down) only solution, but the advantages are hard to justify when much of the battery capacity is not used, as with certain Li-Ion chemistries and a 3.3V output, or with two alkaline cells and a 3.0V or 2.5V output.
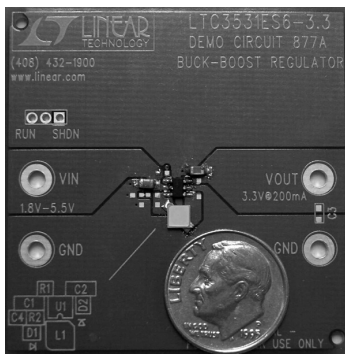

Figure 1. Compact 3.3V buck-boost application

The LTC3531 is a single inductor 200mA buck-boost converter that generates a regulated output voltage from a wide input voltage between 1.8V and 5.5V while maintaining high efficiency. It is an excellent fit for low power applications where a tiny total solution size is required. The LTC3531 is available with fixed outputs [3.0V or 3.3V] or with an adjustable output that can be set between 2.0V and 5.0V. The fixed output versions require only two small ceramic capacitors, a miniature inductor, and the ThinSOT IC. All versions of the part are available in a thermally enhanced 3mm × 3mm DFN packages. A photo of the LTC3531 demo-board is shown in Figure 1.

Generating a Clean 5V from a Noisy USB Cable

Generating a clean 5V output from a USB cable or wall adapter can be a challenge when the combination of source impedance and load transients cause noise and voltage droops. USB cable voltage can vary between 5.25V and 4.35V, while the maximum allowed decoupling capacitance on the input is 10µF. The trace labeled "V_{IN}" in Figure 2 shows what can occur with a 4.7µF input capacitor and a 100mA load step from a powered device. The LTC3531 produces a clean 5V output (V_{OUT}) with less than 100mV of peak-to-peak ripple using a 22µF V_{OUT} capacitor. V_{IN} and V_{OUT} are DC aligned at 500mV per division in Figure 2, showing significant improvement in noise and voltage droop. Inductor current is also shown with operation in both boost ($V_{IN} < V_{OUT}$) and buck-boost ($V_{IN} \cong V_{OUT}$) modes.

A complete schematic of the USB to 5V application is shown in Figure 3 along with efficiency and power loss curves versus load current. The
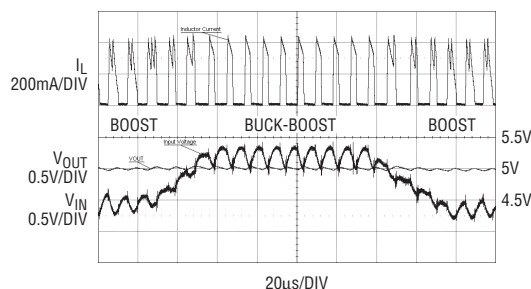

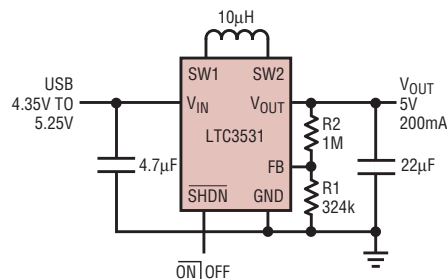**Figure 2. Noisy USB cable input to clean 5V output**

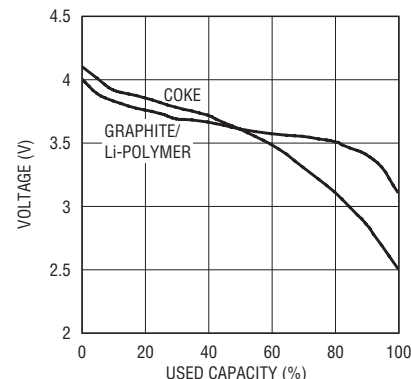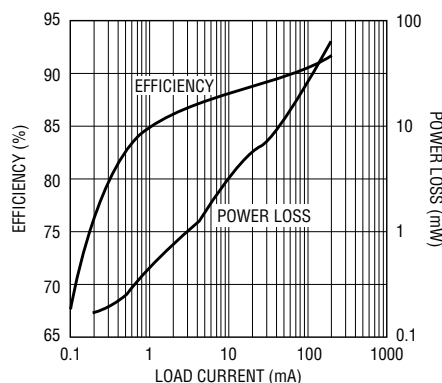**Figure 3. USB to 5V application**

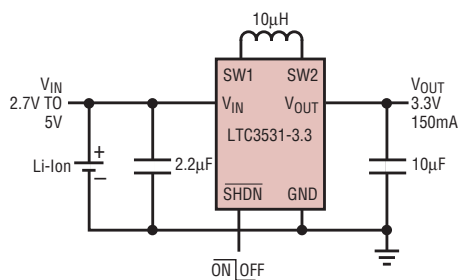**Figure 4. Typical 1C lithium-ion/polymer capacity curves**

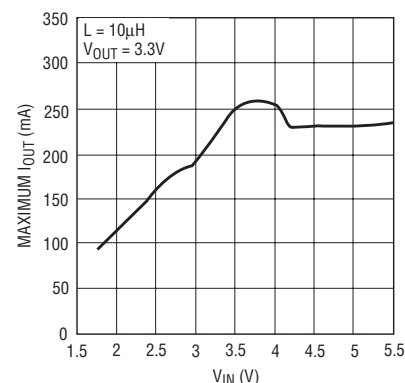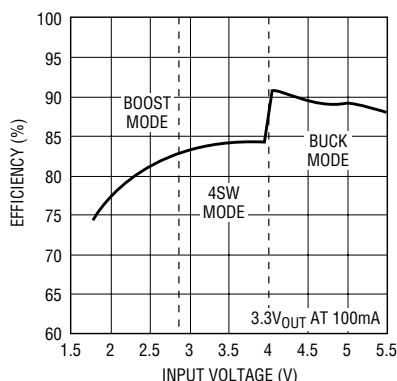Figure 5. Lithium Ion to 3.3V schematic and 100mA efficiency curve.

Figure 6. Output current capability vs input voltage (V_{OUT} = 3.3V)

LTC3531 operates in Burst Mode operation, with just 20μA of quiescent current, providing high efficiency over several decades of load current. All four switches have an $R_{DS(ON)}$ of about 0.5Ω when operating at 5V, providing >90% efficiency at higher load currents.

Maximizing Li-Ion Capacity for 3.3V

When compared to a straight buck converter, the LTC3531 allows lower input voltage operation when providing a 3.3V output from a Li-Ion input source. Typical capacity curves for coke and graphite anode Li-Ion batteries are shown in Figure 4. Coke types have a lower cut-off voltage at 2.5V, where graphite types have a flatter discharge curve and a 3.0V cut-off. Solid lithium polymer batteries have discharge curves similar to graphite.

The equivalent series resistance (ESR) of the Li-Ion battery causes additional voltage drops at the terminal at higher load currents. To make matters worse, the battery protector circuit adds additional series resistance and the effects of ESR lower system efficiency as the battery is discharged.
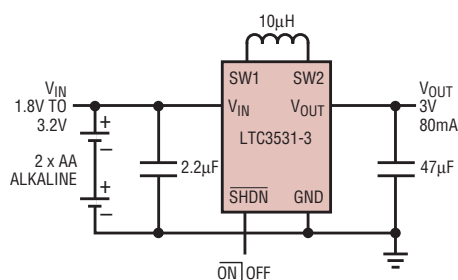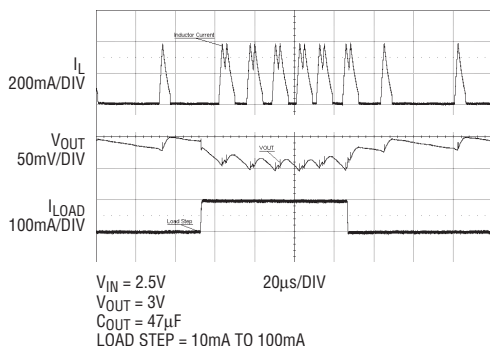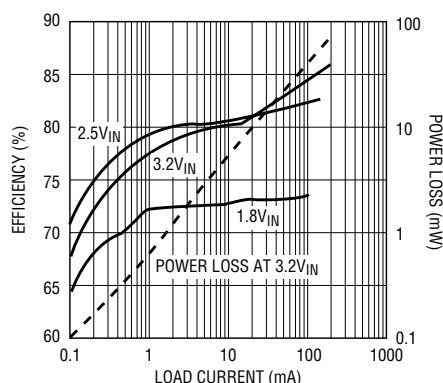
To guarantee a 3.3V output, a buck only design may need to use a cut-off voltage of 3.5V or 3.6V. This translates to a capacity loss of approximately 45% for the coke cell and 20% for the graphite—both significant reductions in run time. Furthermore, while graphite or polymer cells are more popular because of their flat discharge curve, new chemistries with greater capacity per volume are on the horizon with expected discharge curves resembling the coke anode.

The wide input voltage range of the LTC3531 allows a regulated 3.3V to be produced from all Li-Ion chemistries, two or three alkaline cells, or a 5V source such as USB. The LTC3531 automatically transitions between buck, 4-switch (buck-boost), and boost modes based on the voltage difference between V_{IN} and V_{OUT}. Figure 5 shows a 3.3V application circuit, along with efficiency vs input voltage for a 100mA load. Maximum load current capability vs input voltage (V_{OUT} = 3.3V) is shown in Figure 6. As expected, efficiency and load current capability are reduced with input voltage.

Figure 7. Two AA or two AAA to 3.0V application

Figure 8. Transient response of the circuit in Figure 7

3.0V Flash Memory Application from Two Alkaline Cells

Inexpensive MP3 players and other relatively low capacity, low cost portable devices often replace a hard disk drive (HDD) with flash memory and Li-Ion batteries with disposable alkaline—a good fit for the LTC3531. A complete schematic for a two cell alkaline to 3V flash memory supply is shown in Figure 7. Efficiency is better

Ideal Diodes Combine Battery Stacks (Minimize Heat and Voltage Loss)

by Mitchell Lee

Introduction

Combining multiple battery stacks to serve a common load is an easy task for a diode. Each stack delivers whatever current it can muster to the load, but back-feeding from a fresh battery to one mostly discharged is precluded by the presence of the blocking diode. If you're concerned with heat dissipation in the diode and voltage drop at end of discharge, diodes may leave you pining for a better solution.

Active Diode

The circuit in Figure 1 implements an active, ideal diode using a low resistance MOSFET and a micropower op amp. The MOSFET functionally replaces the diode. Placing it in the negative lead of the battery stack permits use of an N-channel device in a simple arrangement, driven on by an op amp if there is a slight forward voltage and off if the voltage reverses.

The forward voltage drop is regulated at 26.4mV, giving freedom from oscillations and preventing reverse current flow. Static current drain for the entire circuit is less than 4µA.

Modifications

Operation over a range of 10V to 36V is practical, and lower voltage operation is feasible by converting to a logic level MOSFET. Because the forward regulation point is a function of the battery voltage, the 10M–10kΩ divider should be adjusted to keep the drop across the 10kΩ resistor in the range of 10mV to 50mV.

In Figure 1 the forward drop exceeds 26.4mV when the product of the load current and MOSFET $R_{DS(ON)}$ thus dictate. For the 55mΩ IRF540, this point is reached at load currents of 500mA.

Given some finite $R_{DS(ON)}$ there is a practical limit for the load current in any MOSFET, where $R_{DS(ON)} \bullet I_{LOAD}$ ceases to provide any advantage over a diode. In the case of the IRF540 this point arrives in the 5A-to-10A range. For higher current applications, substitute a lower $R_{DS(ON)}$ MOSFET. *LT*
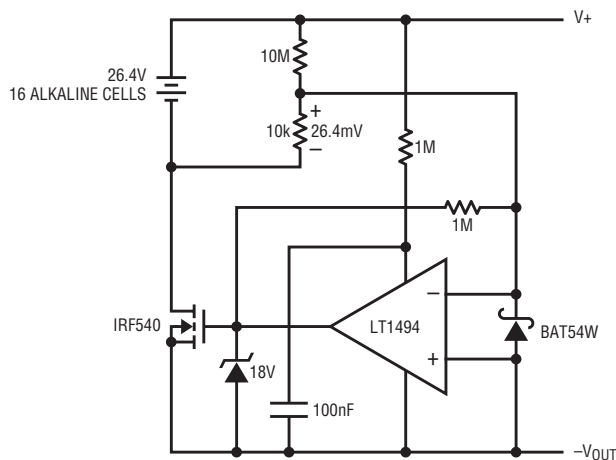

Figure 1. Battery pack ideal diode

than 80% with fresh alkaline cells and better than 70% with depleted batteries. Note that overall efficiency is lower in this application (relative to the Li-Ion application) because of the lower drive voltage for the switches—switch $R_{DS(ON)}$ increases with decreasing drive voltage. The adjustable version of the part can be used to power lower voltage flash memories (i.e. 2.5V) with similar performance results.

Peak current requirements for flash memory are typically lower than a HDD, since there is no disk spin-up,

but load transients still occur when memory is accessed. The LTC3531's response to a 10mA–100mA load step is shown in Figure 8. Burst Mode® operation provides a rapid transient response since there is no compensation loop to slew. Peak-to-peak voltage ripple plus load step is under 50mV with a 47µF output capacitor. The output voltage ripple stays fairly constant over input voltage. Half the output capacitance (22µF) results in approximately twice the load step plus voltage ripple (100mV).

Summary

The LTC3531 provides a simple, compact buck-boost solution for lower current, portable applications. A complete solution, 1mm in height, can fit in a 35mm² footprint. The part maintains high efficiency over a wide range of input voltages and load currents, extending battery run time, while providing the flexibility to address many designs such as 2-cell alkaline, USB, and present day or emerging Li-Ion chemistries. *LT*

High Efficiency, Low Input Voltage, Synchronous Buck Controller Drives up to 15A Load Current

by Joseph Duncan

Introduction

The LTC3822 is a synchronous stepdown DC/DC converter that drives external N-channel power MOSFETs to maximize average current drive for the lowest cost. Its No R_{SENSE} constant frequency architecture minimizes the number of external components, and a programmable frequency of up to 750kHz allows the use of small surface-mount inductors and capacitors. This DC/DC controller is optimized for $3.3V_{IN}$ and Lithium-Ion applications allowing V_{OUT} as low as 0.6V while maintaining 1% precision. The all N-channel MOSFET drive simplifies component selection as well as drastically increasing the current capabilities of a typical circuit. Even with 3.3V gate drive, the LTC3822 is capable of controlling more than 15A load current while maintaining high efficiency.

Compact, 1.8V, 8A Application

Figure 1 shows a 1.8V, 8A application that operates over input voltages between 2.75V and 4.5V, perfect for 3.3V or Li-Ion inputs. This application occupies much less space than would be expected for its current capabilities, as shown in Figure 2.

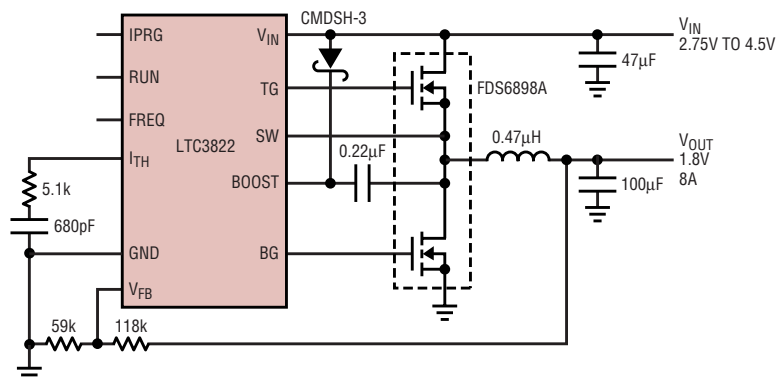During startup, the internal soft-start circuitry smoothly

Figure 1. Typical application delivering 1.8V at 8A.

Figure 2. Sample footprint for application circuit in Figure 1
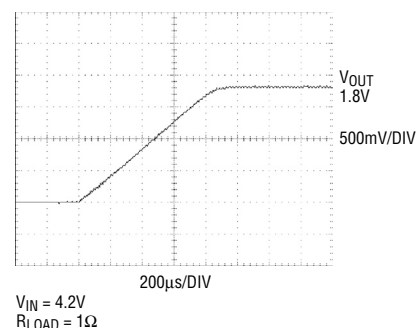

V_{IN} = 4.2V
R_{LOAD} = 1Ω

Figure 3. Internal soft-start ramps the output voltage smoothly without requiring an external capacitor.

ramps the output voltage from 0V to its final value in 800µs (Figure 3). This is done without the need for an external capacitor. The LTC3822 incorporates No R_{SENSE} technology to sense the inductor current from the drain to source voltage (V_{DS}) of the top-side

power MOSFET. The maximum load current that the controller is capable of driving is determined by the $R_{DS(ON)}$ of this MOSFET. Since the LTC3822 incorporates all N-channel MOSFET drive, lower $R_{DS(ON)}$ (and cheaper) devices are available for the top-side MOSFET, when compared to traditional complementary MOSFET drive.

Increasing the Current to 20A

Figures 4 and 5 show two ways to raise the current capability of the regulator by lowering the $R_{DS(ON)}$ of the MOSFETs. In Figure 4, MOSFETs with a much lower $R_{DS(ON)}$ than those of Figure 1 are used. Because they are in individual SO-8 packages, their thermal capabilities are also higher. This application is designed for a 15A continuous current load. Figure 5 in-
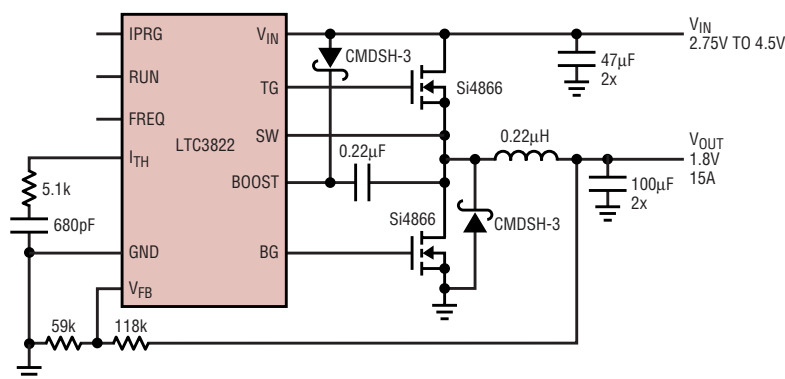

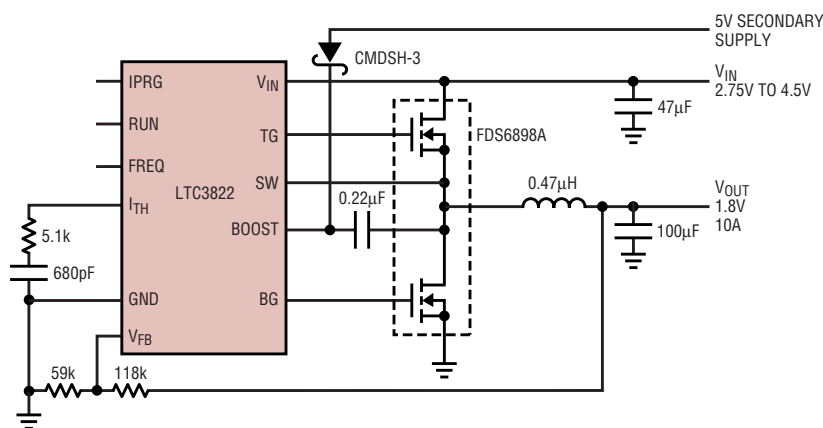**Figure 4. High current application delivering 1.8V at 15A.**

Figure 5. High efficiency application deriving gate drive voltage from a secondary 5V supply.

Figure 6. Transient performance of the converter in Figure 1

stead utilizes a secondary 5V supply to provide a higher gate drive voltage to the MOSFETs. Higher gate drive voltages lower $R_{DS(ON)}$ while simultaneously allowing the use of cheaper logic-level MOSFETs. The maximum load current can also be tailored using the current limit programming pin, IPRG. This three-state pin sets the peak current sense voltage across the top-side MOSFET. Combining all three high current approaches (utilizing low $R_{DS(ON)}$ MOSFETs, powering

the gate drive from a secondary 5V supply, and setting current limit to its highest value) enables applications in excess of 20A.

OPTI-LOOP Compensation

The LTC3822 incorporates OPTI-LOOP® compensation to enable the user to choose optimal component values to compensate the loop over a wide range of operating conditions with the minimum number of output capacitors. Figure 6 shows the tran-
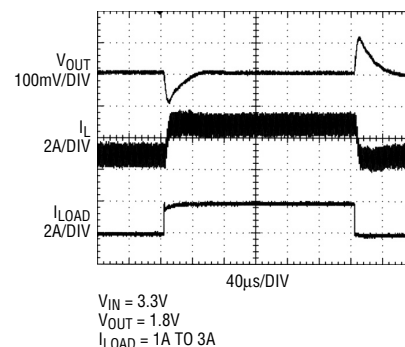
sient response for the circuit in Figure 1 with a load step of 1A to 3A. The output overshoots by approximately 100mV on a 1.8V output and then settles in about 50µs.

Conclusion

The LTC3822 delivers currents as high as 20A for single-output applications using a minimum number of components in a tiny complete solution footprint. ![LT]

Automatic Biasing Feature

The LT6557 and LT6558 are designed specifically with single-supply AC-coupled operation in mind. Each input includes an internal current-controlled bias voltage source like that shown in Figure 3. A single external resistor R_{BCV} programs the input bias voltages as shown in Figure 4 for the LT6557. The LT6558 R_{BCV} function is similar to Figure 4, but is optimized for producing higher biasing levels to account for the lower gain and auto-

matically tracks downward with the supply if below 4V. The selection of input bias point may depend on the application, but the values shown for the programming resistors in Figures 1 and 2 are representative of most designs.

Conclusion

The LT6557 and LT6558 triple video amplifiers are optimized specifically for operation on low voltage single supplies. With preset gain and pro-

grammable biasing, these devices offer minimal parts-count AC-coupled amplifier solutions for very high-resolution applications. The LT6557, with its gain of two, is designed for RGB output ports such as in video routers and KVM switch products. The LT6558, with unity gain, is designed as an RGB input port buffer and/or ADC driver, such as in computer or home-theater display products. ![LT]

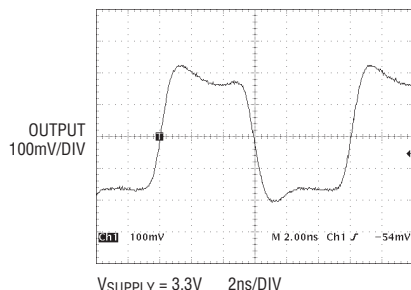

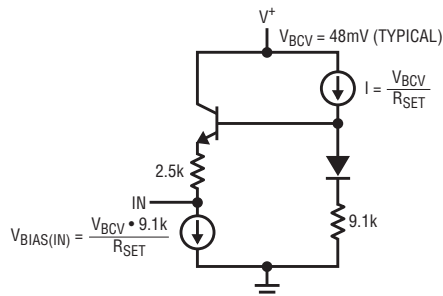**Figure 3. Fast pulse response of LT6558 on 3.3V single supply**

Figure 4. Simplified schematic of LT6557 input biasing circuit (LT6558 similar)
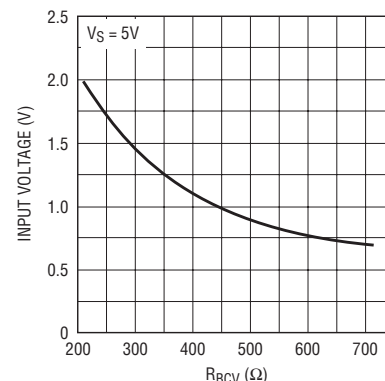

Figure 5. Relationship of LT6557 input bias voltage to programming resistor R_{BCV}

Step-Up/Step-Down Charge Pump DC/DC Converter Provides up to 150mA in a Tiny 2mm × 2mm DFN Package

by Julian Zhu

Introduction

A wide variety of handheld and portable applications are powered by Li-Ion batteries or AA cells. The wide input voltage range of a single Li-Ion battery (2.7V–4.2V) or 2 AA cells (1.8V–3.0V) requires a DC/DC converter that can step-up or step-down the input voltage to provide a fixed output voltage such as 3.3V or 2.5V. The new LTC3240 step-up/step-down DC/DC converter is ideally suited for such applications and can provide up to 150mA in a tiny 6-lead 2mm × 2mm DFN package.

For input voltages greater than the regulated output voltage the LTC3240 operates as a low dropout regulator. When the input voltage decreases to within about 150mV of the regulated output voltage, the LTC3240 automatically switches to step-up

Figure 1. LTC3240-3.3 step-up/step-down converter capable of delivering current up to 150mA

mode. In step-up mode, the LTC3240 operates as a constant frequency (1.2MHz) voltage doubling charge pump. The LTC3240 requires only three tiny external ceramic capacitors for an ultra small application footprint as shown in Figure 1.
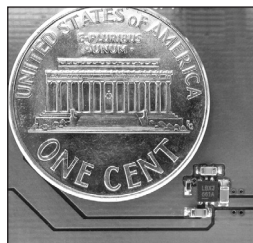
The LTC3240 features low no load operating current (65µA typical) and ultra low shutdown current (<1µA). Built-in soft-start circuitry prevents excessive inrush current during start-up. The thermal-shutdown and current-limit circuitry allow the parts to survive a continuous output short-circuit.

Application for Li-Ion or Three AA Battery Input to 3.3V Out

A typical application circuit for LTC3240-3.3 is shown in Figure 2. The input can be a single Li-Ion battery or three AA cells. Figure 3 shows the output voltage variation for the entire input voltage range at a load current of 30mA.

A new or recharged battery starts out at its highest terminal voltage. As the battery discharges, its terminal voltage continues to drop until the next recharge. The LTC3240 optimizes the output efficiency by continuing to operate in the step-down (LDO mode) for most of the battery life. When the battery voltage gets low enough it automatically switches into charge-pump mode to squeeze out maximum energy from the battery before its next
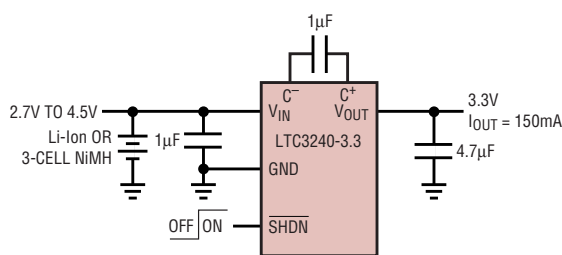

Figure 2. The regulated 3.3V from battery voltage

Figure 3. Output voltage vs input voltage (full range)

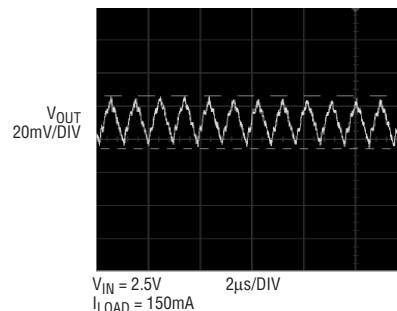**Figure 4. Efficiency vs input voltage (LTC3240-3.3)**

Figure 5. Output voltage ripple (LTC3240-3.3)

recharge cycle. Figure 4 shows the efficiency of LTC3240-3.3 as a function of input voltage.

In step-up (charge pump) mode, the LTC3240 uses a unique architecture to optimize the charge transferred to the output in each clock cycle, thus minimizing the output ripple. The part only needs a 4.7µF, 0603 size ceramic capacitor to obtain a 32mV maximum

ripple voltage (<1% for 3.3V output) at 150mA (See Figure 5).

To extend battery life at light loads, in charge pump mode, the part operates in high efficiency Burst Mode operation. In this case, the LTC3240 delivers a minimum amount of charge for a few cycles, and then enters a low current state until the output drops low enough to require another burst of charge.

Conclusion

The LTC3240 step-up/step-down charge pump DC/DC converter provides fixed regulated output voltage with currents up to 150mA from a wide input voltage range in a small 6-lead 2mm × 2mm DFN package. It is ideally suited for efficient DC/DC conversion in space-constrained applications such as battery-powered handheld electronics. ![LT]

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as 70V while both providing both overvoltage protection *and* remaining below the 42V maximum switch voltage. The charge pump Schottky diodes and capacitors double the effective output voltage for a given duty cycle while the LT3486 LED driver continues to regulate the 100mA constant LED current. The LEDs in Figure 7 have higher forward voltage than those in Figure 1 at 100mA,

resulting in a total string voltage as high as 40V. If more LEDs are needed, the string voltage can be stacked up to 70V before hitting the overvoltage protection level, but the peak switch current limit cannot be exceeded. As the string voltage and LED current goes up, the minimum input voltage also rises. Figure 8 shows the typical peak switch current limit dropping as duty cycle increases. In addition to

the peak inductor current, the voltage doubler also adds additional charge pump capacitor current.

Conclusion

The LT3486 is a dual 1.3A LED string driver with 1000:1 PWM dimming capability. Its 3% LED current accuracy, low sense voltage, low shutdown current, overvoltage protection and wide input voltage range make it ideal for high power LCD panels in a variety of applications including automotive displays and notebook computers. The simple 5:1 analog dimming ratio and more precise 1000:1 PWM dimming ratio provide the displays with enough brightness control for daylight and nighttime use while retaining their color characteristics across brightness levels. ![LT]
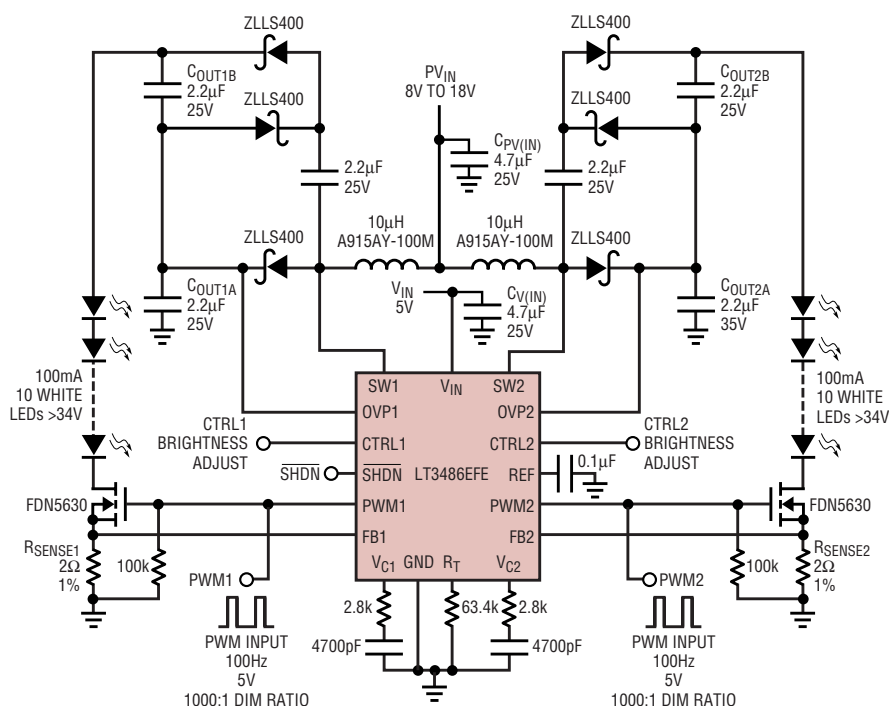

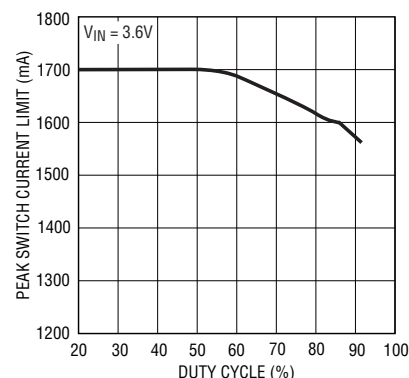**Figure 7. LED driver uses 8V–18V input to drive two strings of ten 100mA LEDs (40V max per string) with 1000:1 PWM dimming**

Figure 8. Typical peak switch current limit drops as duty cycle increases above 50%

New Dual Input USB/AC Linear Li-Ion Battery Chargers

by Alfonso Centuori

Introduction

Digital Cameras, PDAs, mobile phones and MP3 players all use batteries that are commonly charged via either a wall adapter or through the USB. The LTC4076 and LTC4077 lithium-ion battery chargers are specifically designed to detect power at the inputs and automatically select the appropriate source for charging. Using a constant current/constant voltage algorithm, the chargers can be programmed to deliver up to 950mA of charge current with a final float voltage accuracy of ±0.6%. The LTC4076 and LTC4077 include an internal P-channel power MOSFET and thermal regulation circuitry with no blocking diode or external sense resistor required. Thus, the basic dual-source charger requires only three external components.
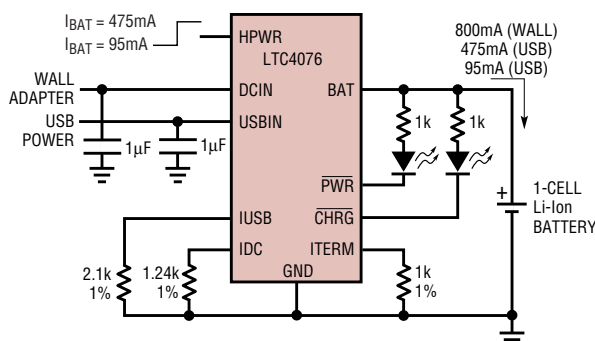
The LTC4076 offers a programmable current based termination scheme. The $\overline{CHRG}$ open-drain status pin can be programmed to indicate the battery charge state according to the needs of the application. The $\overline{PWR}$ open-drain status pin indicates that enough voltage is present at one of the inputs to charge a battery. With power applied on both inputs, LTC4076 and LTC4077 can be put into shutdown mode, reducing the DCIN supply current to 20µA, the USBIN supply current to 10µA and the battery drain current to less than 2µA.

Internal thermal feedback regulates the charge current to maintain

Figure 1. An LTC4076 USB/wall adapter Li-Ion charger

a constant die temperature during high power operation or high ambient temperature conditions.

USB Compatibility

The HPWR pins of the LTC4076 and LTC4077 provide an easy method to choose between two different USB power modes: high power (usually ≤ 500mA) and low power (usually ≤ 100mA).

With the LTC4076, a logic high on the HPWR pin sets the charge current to 100% of the current programmed by the IUSB pin resistor, while a logic low on the HPWR pin sets the charge current to 20% of the current programmed by the IUSB pin resistor.

With the LTC4077, a logic high on the HPWR pin sets the charge current to the value programmed by the IUSB pin resistor, while a logic low on the HPWR pin sets the current to the value programmed by the IUSBL pin resistor.

A weak pull down on the HPWR pin sets the part to default to the low power state. The HPWR pin provides a simple control for managing charge current as shown in Figure 1 and Figure 2. The presence of a wall adapter takes priority over the USB input and the IDC 1.24kΩ resistor sets the total maximum charge current to 800mA. When USB power is present, and a wall adapter is not, the IUSB 2.1kΩ resistor sets the charge current to 476mA with HPWR in its high state, and 95mA with HPWR in its low state for the LTC4076 (or 100mA with IUSBL 2kΩ resistor for the LTC4077).

Programmability

The LTC4076 and LTC4077 provide a great deal of design flexibility including programmable charge current and programmable current termination. The charge currents are programmed using a resistor from the IDC, IUSB and IUSBL (LTC4077 only) pins to ground as indicated in the following equations:

$$I_{CHG} = \frac{1000V}{R_{IDC}} \quad \text{(Wall Adapter Present)}$$

$$I_{CHG} = \frac{1000V}{R_{IUSB}} \quad \text{(USB HPWR = High)}$$

$$I_{CHG} = \frac{200V}{R_{IUSB}} \quad \text{(USBHPWR = Low, LTC4076 Only)}$$

$$I_{CHG} = \frac{200V}{R_{IUSBL}} \quad \text{(USBHPWR = Low, LTC4077 Only)}$$

Figure 2. An LTC4077 USB/wall adapter Li-Ion charger

Figure 3. LTC4076 or LTC4077 dual input charger circuit. The wall adapter charge current and the USB charge current (HPWR = High) are both programmed to be 500mA with just one resistor.

Both the LTC4076 and LTC4077 terminate the charge cycle based on the battery current. For the LTC4076 the current threshold is programmable and for the LTC4077 the current threshold is fixed, as described below.

For LTC4076, the programmable current detection threshold, I_{TERM}, is set by connecting a resistor, R_{ITERM}, from ITERM to ground. The following formula programs the termination current:

$$I_{TERM} = \frac{100V}{R_{ITERM}}$$

For the LTC4077 the termination current is fixed at 10% of the programmed charge current as set by IDC or IUSB (HPWR = High). When HPWR in its low state, the termination current is 50% of the current programmed by R_{IUSBL}.

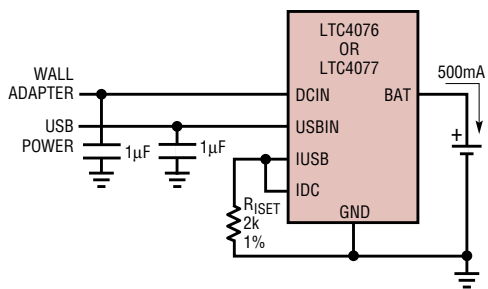The condition of the $\overline{CHRG}$ pin indicates the charge state. A strong pull-down on the $\overline{CHRG}$ pin indicates that the battery is charging. When the current termination threshold is reached the $\overline{CHRG}$ pin assumes a high impedance state.

Avoiding Unnecessary Charge Cycles

LTC4076 and LTC4077 are designed to avoid unnecessary charge cycles to extend the life of Li-Ion batteries. When power is first applied or when exiting shutdown, the LTC4076 and LTC4077 check the voltage on the BAT pin to determine its initial state. If the BAT pin voltage is below the recharge threshold of 4.1V (which corresponds

to approximately 80%–90% battery capacity), LTC4076 and LTC4077 enter charge mode and begin a charge cycle. If the BAT pin is above 4.1V, the battery is nearly full and the charger does not initiate a charge cycle and instead enters standby mode. When in standby mode, the chargers continuously monitor the BAT pin voltage. When the BAT pin voltage drops below 4.1V, the charge cycle is automatically restarted. This feature eliminates the need for periodic charge cycle initiations, ensures that the battery is always fully charged, and prolongs battery life by reducing the number of unnecessary charge cycles.

Conclusion

LTC4076 and LTC4077 are complete Linear Li-Ion battery chargers compatible with portable USB applications. They are designed to accommodate charging from both a wall adapter and a USB input. The versatility, low quiescent current, simplicity, high level of integration and small size of the LTC4076 and LTC4077 provide an ideal choice for many portable USB applications. LTC4076 and LTC4077 are available in a small 10-lead low profile 3mm × 3mm DFN package. *LT*

New Device Cameos

New Timing and High Voltage Output Enhance Push Button On/Off Controller

The LTC2954 is an upgrade to the LTC2950 family of push button controllers. Two features have been added: new push button controlled interrupt timing, and a high voltage open drain enable output capable of 33V operation.

The LTC2954 now provides interrupt and power down modes to allow more flexible and reliable system shut down. Once in the powered on state, momentary presses on the push button provide interrupts to system logic. This can be useful for system housekeeping and power down under

firmware control. Should the system require a forced power down, a long duration press and hold of the push button automatically releases the enable pin, thus turning system power off. The timing of this latter power down mode is adjustable with external capacitance on the power down timing pin.

The new high voltage enable output of the LTC2954-2 is ideally suited to drive the gate of a high voltage power PFET. This allows a user to connect/disconnect input power from its load simply by toggling the de-bounced push button input.

The LTC2954 operates over a wide 2.7V to 26V input voltage range while

consuming only 6µA of supply current and is available in 8-Pin 3mm × 2mm DFN and ThinSOT™ Packages. Two versions of the part are available to accommodate either positive or negative enable polarities. *LT*

For further information on any of the devices mentioned in this issue of *Linear Technology*, use the reader service card or call the LTC literature service number:

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Power Management for Portable Products—The solutions in this product selection guide solve real-life problems for cell phones, digital cameras, PDAs and other portable devices, maximizing battery run time and saving space. Circuits are shown for Li-Ion battery chargers, battery managers, USB support, system power regulation, display drivers, white LED drivers, photoflash chargers, DC/DC converters and RF PA power supply and control.

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