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A Better Way to Push Your Buttons

by Victor Fleury

Introduction

Is there a better way to debounce the on/off push button of a handheld device? Some designers use discrete logic, flip-flops, resistors and capacitors. Others use an on-board microprocessor, which requires constant power—even after the handheld device has been turned off. Additionally, for multi-cell battery applications, a high voltage LDO is needed to drive the low voltage logic and microprocessor. All this extra circuitry not only increases board space, but also drains the battery when the handheld device has been turned off.

The LTC2950 family of parts eliminates all of these problems. The part incorporates the flexible timing needed to debounce the push button input during system power on and system power off. The LTC2950's wide input voltage range (2.7V to 26.4V) is designed to operate from single cell to multi-cell battery stacks, thus eliminating the need for an LDO. The part's set of features allows the system designer to turn off system power to all circuits except the LTC2950, whose very low quiescent

current (6 μ A) is an insignificant drain on the battery. The device is available in space saving 8-lead 3mm \times 2mm DFN and ThinSOT packages.

More than Just a De-Bouncer

The LTC2950 is not just a low power, high voltage push button de-bouncer. The debounced push button input toggles an open drain enable output. This low leakage output can be

used to control the shutdown pin of a DC/DC converter, and thus allow manual control of system power. The part also contains a simple microprocessor interface that provides intelligent power up and power down sequencing. During power up, an internal timer ensures that the system will not power into a short and

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drain the battery. During power down, the LTC2950 interrupts the microprocessor 1024ms before de-asserting the enable output. This gives the microprocessor time to perform housekeeping tasks (such as saving to memory) before power is turned off.

continued on page 3



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Issue Highlights

When a push button is pressed, the voltage on the pin does not seamlessly switch from the pull-up voltage to ground. The voltage fluctuates as the push button makes and breaks contact, potentially causing the microprocessor to see a series of on/off events. The LTC2950 solves this problem by ignoring all the noise and driving the enable pin high 32ms after the push button stops bouncing.

See our cover article for more about this breakthrough device.

Featured Devices

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

Power Solutions

The **LTC3454 is a synchronous buck-boost DC/DC converter**, designed for driving a single high power LED with

regulated currents up to 1A from a single Li-Ion battery. (Page 23)

The **LT3477** combines a traditional voltage feedback loop and two unique current feedback loops to operate as a constant-current, constant-voltage source. It is a current mode, **3A DC/DC converter with dual rail-to-rail 100mV current sense amplifiers** that can be configured as a buck mode or buck-boost mode LED driver. It is versatile enough to also be configured as an input-output current limited boost, SEPIC or inverting converter. (Page 25)

Hot Swap and 2-Wire Bus Solutions

The **LTC4253A** and **LTC4253A-ADJ** facilitate **safe board insertion and removal from a live backplane** by applying power in a controlled manner. Running off a simple, fast responding shunt regulated supply that allows

very high voltage operation, they are uniquely suited for applications on the -48V bus. (Page 16)

The **LTC4306 4-channel 2-wire bus multiplexer/switch with bus buffers** addresses a variety of capacitive buffering, addressing and Hot Swap issues. (Page 28)

Op Amps

The **LTC6241** dual and **LTC6242 quad CMOS op amps** compete head-on with bipolar op amps in noise, speed, offset voltage, and offset drift, while maintaining superior low input bias and noise current. (Page 5)


What sets the **LT1994** apart from other **fully differential amplifiers** are its low noise, low distortion, rail-to-rail output, and an input common mode range that extends to ground on power supplies as low as 2.5V. This eliminates the need for a negative power supply, and makes the **LT1994** uniquely able to interface to differential input ADCs while sharing the same power supply. (Page 13)

The **LT6555** and **LT6556 triple video multiplexers** offer up to 750MHz performance in compact packages, requiring no external gain-setting resistors to establish a gain of two or unity. A single integrated circuit, in a choice of either 24-lead SSOP or 24-contact QFN (4mm x 4mm), performs fast switching between a pair of three-channel video sources, such as RGB or component HDTV. (Page 20)

Photoflash Capacitor Chargers

The **LT3484** and **LT3485 photoflash capacitor chargers** squeeze high performance xenon flash technology into the small spaces afforded to cameras in cell phones and PDAs. (Page 9)

Design Ideas and Cameos

Design Ideas start on page 32, including a discussion of Li-Ion-based battery chargers and an op amp selection guide. A cameo about the exciting new **LTM4600 μ Module DC/DC converter** appears on page 45. 

Linear Technology in the News...


Linear Announces New Line of Power Modules

In October, Linear Technology announced a new line of high-density power modules. The new product line provides designers with simple, compact and reliable power supplies for a broad range of applications. Using these compact, board-ready products, designers can significantly accelerate time-to-market and reduce risk in implementing high performance power systems. The first product in the family, the **LTM[®]4600 μ Module[™]** is a 10A switchmode DC/DC step-down power supply in a small, surface mount package. This new product line leverages the company's core strengths in power management, resulting in a highly integrated module with record power density.

According to Don Paulus, Vice President and General Manager for Power Management Products, "Designers today are challenged to develop systems at ever-higher

power densities with relentless time to market pressure. With power supplies becoming more complex, designers increasingly require sophisticated power design expertise. Using the **LTM4600 μ Module**, the power design is virtually done, freeing system designers to focus on their core expertise."

Expanded Test Facilities

Last month, Linear Technology announced the opening of the company's second semiconductor test facility in Singapore. This expansion will allow the company to more than double its current production capacity, strengthening its ability to meet the growing global demand for Linear's high performance analog circuits. This growing facility, combined with Linear's two US wafer fabrication plants and its assembly facility in Malaysia, ensure that customers receive the highest quality ICs with fast delivery times. 

LTC2950, continued from page 1

Watch the Push Button Bounce

When a push button is pressed, the voltage on the pin does not seamlessly switch from the pull-up voltage to ground. The voltage fluctuates as the push button makes and breaks contact.

Figure 1 shows an application with significant bounce on the push button pin. The LTC2950 ignores all the noise and drives the enable pin high 32ms after the push button stops bouncing. The scope trace shows the turn on debounce time of 32ms—that is, no external capacitor at the ONT pin. This application requires only one external component (R1).

Need Longer Debounce Times?

It is no problem to extend the debounce time of the push button input. The power on and power off debounce times can be extended independently by placing an external capacitor on the ONT and OFFT pins, respectively. Figure 2 shows the turn on timing with an external 0.033µF capacitor on the ONT pin (~250ms). The following equations describe the relationship between total debounce time and external capacitors:

$$\text{Turn On Debounce Time} = 32\text{ms} + (6.7 \cdot 10^6) \cdot C_{\text{ONT}}$$

$$\text{Turn Off Debounce Time} = 32\text{ms} + (6.7 \cdot 10^6) \cdot C_{\text{OFFT}}$$

Typical Power On/Off Timing Sequence

Figure 3 shows a typical LTC2950-1 power on and power off sequence. A high to low transition on $\overline{\text{PB}}$ (t1) initiates the power on sequence. This diagram does not show any bounce on $\overline{\text{PB}}$. In order to assert the enable output, the $\overline{\text{PB}}$ pin must stay low continuously ($\overline{\text{PB}}$ high resets timers) for a time controlled by the default 32ms and the external ONT capacitor (t2 – t1). Once EN goes high (t2), an internal 512ms blanking timer is started. This blanking timer is designed to give sufficient time for the DC/DC converter to reach its final volt-

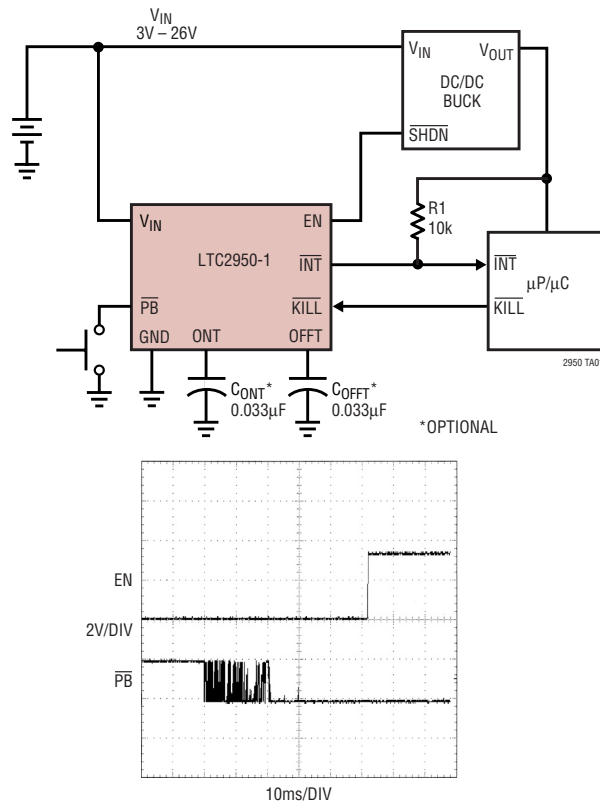


Figure 1. Typical circuit and de-bounce timing

age, and to allow the µP enough time to perform power on tasks. The $\overline{\text{KILL}}$ pin must be pulled high within 512ms after the EN pin went high. Failure to do so results in the EN pin going low 512ms after it went high (EN = low, see Figure 4). Note that the LTC2950 does not sample $\overline{\text{KILL}}$ and $\overline{\text{PB}}$ until after the 512ms internal timer has expired. The reason $\overline{\text{PB}}$ is ignored is to ensure that the system is not forced off while powering on.

Once the 512ms turn on blanking timer expires (t4), the release of the $\overline{\text{PB}}$ pin is then de-bounced with an internal 32ms timer. The system has now properly powered on and the LTC2950 monitors $\overline{\text{PB}}$ and $\overline{\text{KILL}}$ (for

a turnoff command) while consuming only 6µA of supply current.

A high to low transition on $\overline{\text{PB}}$ (t5) initiates the power off sequence. $\overline{\text{PB}}$ must stay low continuously ($\overline{\text{PB}}$ high resets de-bounce timer) for a period controlled by the default 32ms and the external OFFT capacitor (t6–t5). At the completion of the OFFT timing (t6), an interrupt (INT) is set, signifying that EN will be switched low in 1024ms. Once a system has finished performing its power down operations, it can set $\overline{\text{KILL}}$ low (t7) and thus immediately set EN low), terminating the internal 1024ms timer. The release of the $\overline{\text{PB}}$ pin is then de-bounced with an internal 32ms timer. The system is now in its reset state: where the LTC2950 is in low power mode (6µA). $\overline{\text{PB}}$ is monitored for a high to low transition.

What if the DC/DC Converter is Faulty at Power Up?

When a user turns on a handheld device, the LTC2950 EN output pin enables a DC/DC converter. The output of the converter can then power a µP, which in turn drives the $\overline{\text{KILL}}$ pin (see Figure 1). If there is a system fault

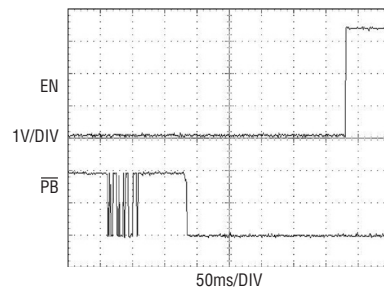


Figure 2. $\overline{\text{PB}}$ turn on de-bounce time increased with an external 0.033µF capacitor

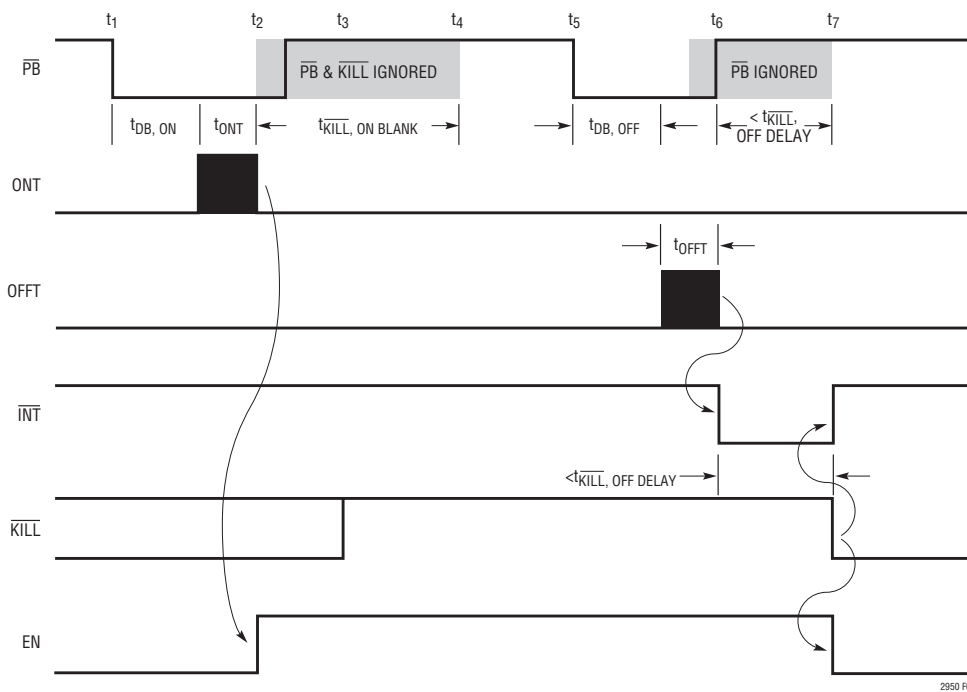


Figure 3. Typical Power On/Off Timing Sequence for LTC2950-1

(shorted DC/DC output, for example) that prevents the μP from driving the $\overline{\text{KILL}}$ input high within 512ms, the LTC2950 automatically releases its enable output. This in turn shuts off the converter and prevents the handheld device from turning on. Figure 4 depicts an aborted power on sequence.

Protect Against μP Hang Ups

Has this happened to you? Your PDA or laptop has frozen—not responding to any input. You try to restart the device by pressing the on/off button. Nothing happens. In frustration you resort to unplugging the device and removing any batteries to shut it down. The unresponsive push button is probably the result of an on/off push button that was de-bounced by an unresponsive μP —evidenced by the crash. The LTC2950 eliminates this common fault.

The LTC2950 *always* responds to the push button in some way. It does this by initiating a power down sequence (in response to the user pressing the push button) by asserting $\overline{\text{INT}}$ low and starting an internal 1024ms timer. This event alerts the μP of the impending power down. If the $\overline{\text{KILL}}$ pin remains high (μP not respond-

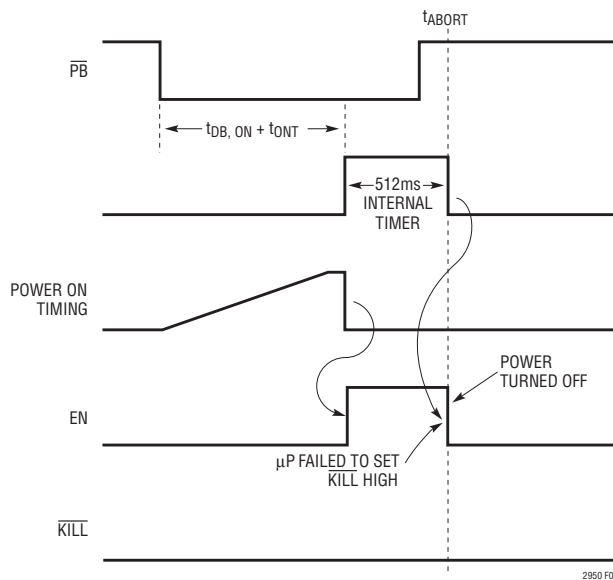


Figure 4. Aborted power on sequence for LTC2950-1

ing) at the end of the 1024ms timeout period, the LTC2950 automatically releases its enable pin, thus shutting

off system power. This fault protection feature makes sure that a user is always capable of turning off system power, even when the rest of the system is faulty or not responding.

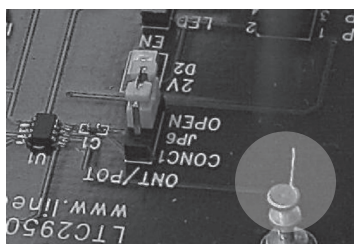


Figure 5. ESD Strikes PB Pin

$\overline{\text{PB}}$ Pin Survives Minor Lightning Strike

The $\overline{\text{PB}}$ and V_{IN} pins are both high voltage pins (33V absolute maximum). Additionally, high ESD strikes ($\pm 10\text{kV}$, HBM) will not damage the $\overline{\text{PB}}$ pin. Fig-

continued on page 46

Fast CMOS Op Amp Challenges Bipolar Amps on All Key Specs

by John Wright and Glen Brisebois

Introduction

The LTC6241 dual and LTC6242 quad CMOS op amps compete head-on with bipolar op amps in noise, speed, offset voltage, and offset drift, while maintaining superior low input bias and noise current. Crucial advances in these amplifiers' parameters translate to tighter system specs, lower complexity, and a wider supply voltage operating range than previous CMOS op amps. These extremely low input bias current op amps are optimized for high impedance transducer applications such as photodiode transimpedance amplifiers, TIAs, though they are also well suited to a variety of precision applications.

The LTC6241 and LTC6242 do not employ complicated post-package schemes to reduce offset voltage, yet their 125 μ V offset voltage and 2.5 μ V/ $^{\circ}$ C offset drift are among the best CMOS amplifiers available. The 18MHz gain bandwidth and very low noise further distinguishes them from a field of mediocre amplifiers. They are fully specified on 3V, and 5V, with an HV version that guarantees operation to \pm 5.5V. Supply current consumption is 2.2mA/amplifier maximum. Table 1 summarizes the conservative specs for these op amps.

The LTC6241 is available in the SO8, and for compact designs it is packaged in the tiny dual fine pitch leadless (DFN) package. The LTC6242 is available in a 16-Pin SSOP as well as a 5mm \times 3mm DFN package.

CMOS with Low 1/f Noise? What about Noise Current?

CMOS op amps have traditionally had much higher 1/f noise than bipolar amplifiers. It is common to find CMOS amplifiers with a 1/f corner above several kilohertz, but the LTC6241 rivals the best bipolar op amps with a 1/f noise corner of only 40Hz. This exceptionally low noise translates

to just 550nV_{P-P} in a 0.1Hz to 10Hz bandwidth, and represents the lowest 1/f noise available in a non-autozero CMOS op amp.

In I-to-V applications such as photodiode amplifiers, where the amplifier is operated inverting, noise current dominates at high frequency. CMOS op amp noise current has two sources. The first is the input device channel thermal noise coupling through the gate-to-source and gate-to-drain capacitances. The second noise current is derived from

the op amp's input capacitance, and capacitance associated with the input transducer. This input referred noise current (CV noise) is due to the amplifier's noise voltage, V_N , impressed across the total input capacitance, C_T , causing a current of magnitude $2\pi f C_T V_N$ to flow through the feedback resistor.

The way to make CMOS or bipolar low noise amplifiers is with large input transistors. The problem is that big input structures carry the burden of high input capacitance. High input capaci-

Table 1. LTC6241/LTC6242 Performance: $T_a = 25^{\circ}\text{C}$, $V_s = 5\text{V}/0\text{V}$ unless otherwise specified. The ● denotes specifications that apply over -40°C to 85°C .

Parameter	Conditions		Min	Typ	Max	Units
Offset Voltage	$V_{CM} = 0$			40	125	μV
	S8, LTC6241			50	150	μV
	GN16, LTC6242 DD, DHC, LTC6241/42			100	550	μV
$T_C V_{OS}$		●		0.6	2.5	$\mu\text{V}/^{\circ}\text{C}$
Input Bias Current		●		1		pA
				10	75	pA
Noise Voltage	$f = 1\text{kHz}$			7	10	$\text{nV}/\sqrt{\text{Hz}}$
	$f = 0.1\text{Hz to } 10\text{Hz}$			550		nV_{P-P}
Noise Current	$f = 100\text{kHz}$			110		$\text{fA}/\sqrt{\text{Hz}}$
Input Capacitance	$f = 100\text{kHz}$			0.5		pF
	C_{DM} C_{CM}			3		pF
Large Signal Gain	$R_L = 1\text{k}\Omega$ to $V_S/2$		90	215		V/mV
CMRR	$V_{CM} = -V$ to $+V - 1.5\text{V}$	●	80	105		dB
Operating Supply Range	LTC6241/42	●	2.8		6	V
	LTV6241HV/42HV	●	2.8		11	V
V_{OUT} Low	$I_{SINK} = 5\text{mA}$	●		190	325	mV
V_{OUT} High	$I_{SOURCE} = 5\text{mA}$	●		4.81	4.675	V
Supply Current	per amplifier	●		1.8	2.2	mA
Slew Rate	$A_V = -2$, $R_L = 1\text{k}\Omega$,	●	5	10		V/ μs
Gain Bandwidth Product	$R_L = 1\text{k}\Omega$	●	13	18		

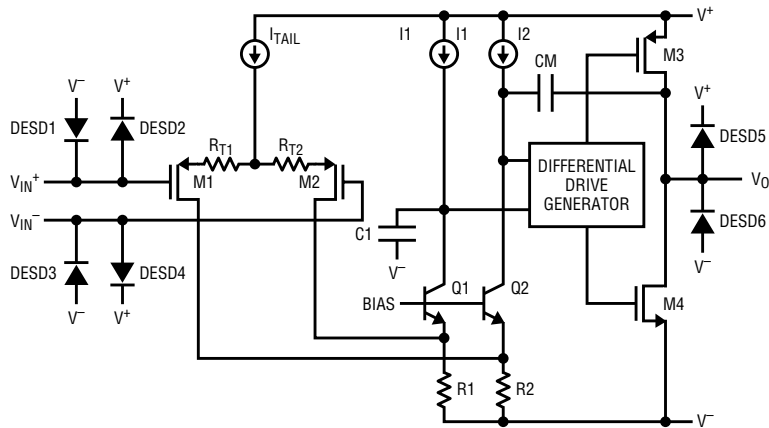


Figure 1. Simplified schematic

tance increases high frequency noise current, as well as reduces overall op amp speed. An uncommon feature of the LTC6241 is its low differential input capacitance of just 0.5pF, which is a major benefit in I-to-V amplifier designs. This input capacitance is 8 to 10 times lower than that of other CMOS amps.

Simple Architecture Yields Low Noise and DC Precision

Figure 1 is a simplified schematic of one half of the LTC6241, which has a pair of low noise input transistors M1 and M2. A simple folded cascode Q1, Q2, and R1, R2 allow the input stage to swing to the negative rail, while performing level shift to the differential drive generator. Transistors M1 and M2 along with current sources

I1 and I2 have been optimized for low noise and consume over 30% of the die area. Low offset is achieved by laser trimming resistors R_{T1} and R_{T2}. Stresses that occur during package assembly have minimal affect on this simple, stable architecture, and consequently, complicated post-package trim schemes that adjust offset voltage and drift are unnecessary.

The LTC6241 and LTC6242 were intentionally designed without a rail-to-rail input stage as to not compromise their noise specs. Many CMOS rail-to-rail input amplifiers show large offset shift and higher noise when the common mode voltage is operating in this top side transition region, limiting their usefulness.

The LTC6241 and LTC6242 have reverse-biased ESD protection diodes on all inputs and outputs as shown in Figure 1. These diodes protect the amplifiers from ESD strikes up to 1.7kV. No current flows into the gate on a DC basis, but these ESD protection diodes are the source of input bias current specified on the data sheet. These diodes have leakage current that doubles approximately every 7°C, but input current typically remains below 10pA up to 85°C ambient.

Capacitor C1 reduces the unity cross frequency and improves the frequency stability without degrading the gain bandwidth of the amplifier. Capacitor C_M sets the overall amplifier gain bandwidth. The differential drive generator supplies signal to transistors M3 and M4 that swing the output from rail-to-rail.

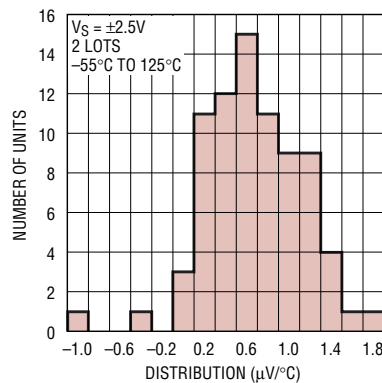
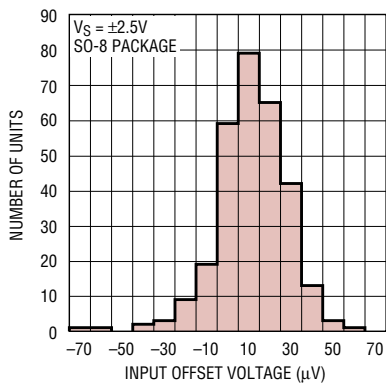


Figure 2. Vos distribution and Vos temperature coefficient distribution

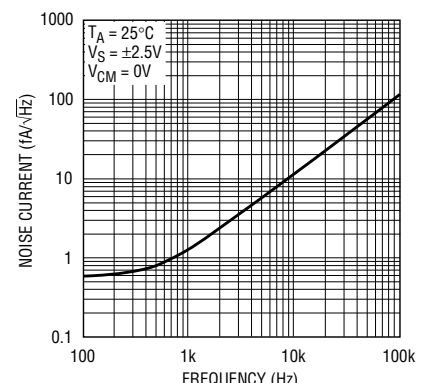
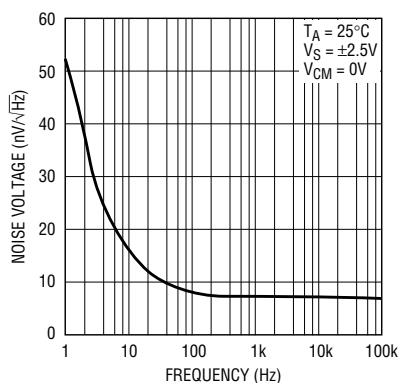
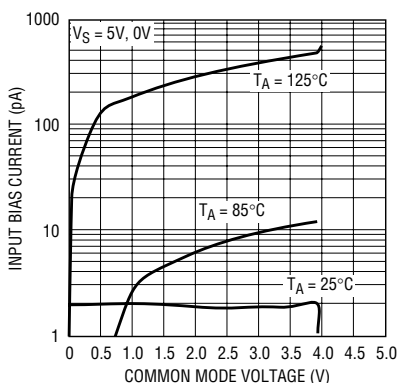


Figure 3. Input bias current vs common mode voltage and voltage and current noise vs frequency

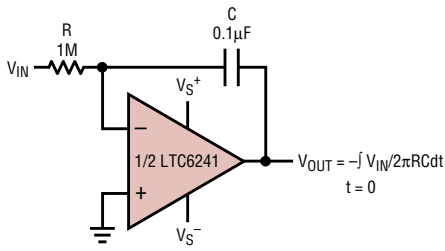


Figure 4. A textbook integrator is inverting

Figure 2 shows the distribution of offset voltage and offset voltage drift. Figure 3 shows the input bias current vs common mode voltage as well as the noise voltage and current spectrum.

Applications

Non-Inverting Integrator

Integrators are used widely in feedback control systems and filters. CMOS input amplifiers like the LTC6241 are preferred for this function because the low input bias current allows the use of large value resistors and small capacitors for a given integration time constant. The most common form of integrator is the inverting form, shown in Figure 4. It has a transfer function of:

$$V_{OUT} = - \int_{t=0} \frac{V_{IN}}{2\pi RC} dt$$

If inversion is not desired in the feedback control loop using the circuit in Figure 4, a designer must add another op amp to invert again. A simpler overall solution produces a non-inverting integrator using just one op amp. Figure 5 shows the circuit.

At low frequencies, $R1 \cdot C1$ does not attenuate, and the non-inverting integration function is provided by the op amp gain and its feedback components $C2$ and $R2$. At higher frequencies, $C2$ becomes a short circuit so the op amp goes to a gain of one, and the integration function is provided by $R1$ and $C1$. If the time constants are matched, the integrator conformance is excellent. Matching is not easy. In most loops, to guarantee that the phase of the integrator does not exceed 90 degrees, the time constants can be intentionally skewed so that $R1 \cdot C1 < R2 \cdot C2$. For an example

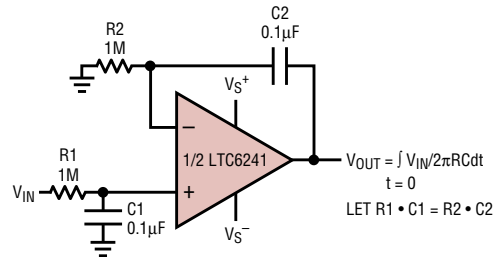


Figure 5. A non-inverting integrator can be very simple. Ideally, $R1 \cdot C1 = R2 \cdot C2$, but mismatch is inevitable. To avoid any phase buildup from a mismatch, the time constants may be skewed so that $R1 \cdot C1 < R2 \cdot C2$.

of a specific closed loop utilization of a non-inverting integrator, see LTC Design Note DN254.

Piezoelectric Accelerometers: Inverting vs Non-Inverting

Figures 6 and 7 show two different approaches to amplifying signals from a capacitive sensor using the LTC6241. The sensor in both cases is a 770pF piezoelectric shock sensor accelerometer, which generates charge

under physical acceleration. Figure 6 shows the classical “charge amplifier” approach. The op amp is in the inverting configuration so the sensor looks into a virtual ground. All of the charge generated by the sensor is transferred across the feedback capacitor by the op amp action. Because the feedback capacitor is 100 times smaller than the sensor, the output is forced to a voltage 100 times what would have

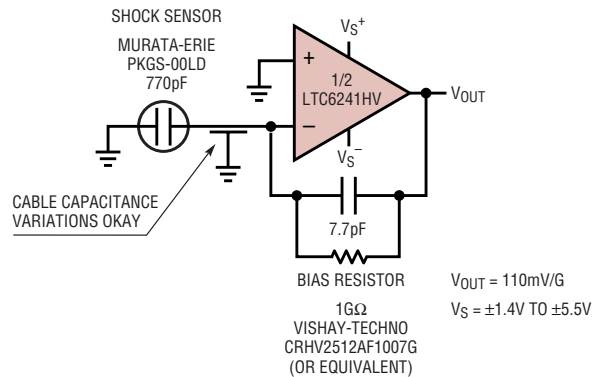


Figure 6. Classical inverting charge amplifier. Variations in cable capacitance (i.e. length) do not affect the signal gain. Use this circuit when the accelerometer is remote from the amplifier and the cable length is unspecified. Drawbacks are that gain is set by the low valued feedback capacitor and low frequency performance is set by the bias resistor working into the same.

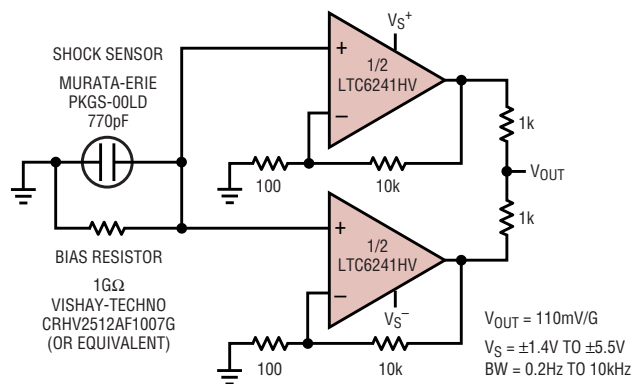


Figure 7. Non-inverting charge amplifier offers several advantages. Stages can be paralleled for lower voltage noise. Bias resistor works into higher capacitance for better low frequency response.

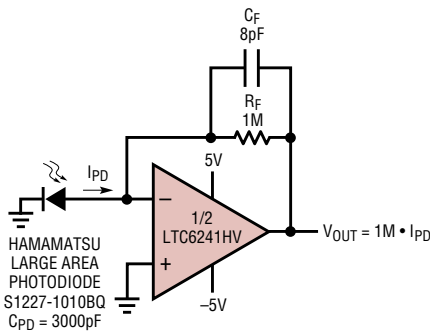


Figure 8. Large area photodiode amplifier provides about 25kHz bandwidth. DCs are good but output is noisy.

been the sensor’s open circuit voltage. Thus, the circuit gain is 100.

The benefit of this approach is that the signal gain of the circuit is independent of any cable capacitance introduced between the sensor and the amplifier, making this a good solution for remote accelerometers where the cable length may vary. Difficulties with the circuit are inaccuracy of the gain setting with the small capacitor, and low frequency cutoff due to the bias resistor working into the small feedback capacitor.

Figure 7 shows a non-inverting amplifier approach. This approach has many advantages. First, the gain is set accurately with resistors rather than with a small capacitor. Second, the low frequency cutoff is dictated by the bias resistor working into the large 770pF sensor, rather than into a small feedback capacitor, for lower frequency response. Third, the non-inverting topology can be paralleled and summed (as shown) for scalable reductions in voltage noise. The only

drawback to this circuit is that the parasitic capacitance at the input reduces the gain slightly. This circuit is favored in cases where parasitic input capacitances such as traces and cables are relatively small and invariant.

Consider making the bias resistor larger than bandwidth calculations would suggest. This actually reduces the noise floor at low frequency. For example, to support frequencies down to 10Hz at -3dB, the bias resistor would calculate to:

$$\frac{1}{2\pi \cdot 10\text{Hz} \cdot 770\text{pF}} = 20\text{M}\Omega$$

At 10Hz, the 20M resistor would contribute 580nV/√Hz of noise, and be 3dB down just like the signal. Making the resistor 1GΩ as shown, its 4000nV/√Hz voltage noise would be attenuated down to effectively 80nV/√Hz by the accelerometer capacitance, while the signal would barely be attenuated at all. That’s an easy seven-fold improvement in the signal-to-noise ratio.

Large Area Photodiode Amplifiers

Figure 8 shows the LTC6241 used as a transimpedance amplifier for a high capacitance large area photodiode. The circuit has unity noise gain at DC, so resolution is entirely noise limited. The bandwidth rolls due to the fact that the photodiode impedance drops with frequency raising the effective gain (the noise gain), which the op amp looks into. This severely limits the bandwidth and increases the output noise. The -3dB bandwidth for this

circuit was measured at 25kHz, and the output noise density at 10kHz was measured at 1.6μV/√Hz. That may be good enough for many applications. If it’s not good enough, keep reading.

The main problem with the previous circuit is the large capacitance of the photodiode. The perfect thing to do is to bootstrap that capacitance with a low noise JFET. Figure 9 shows the circuit. The low noise JFET source follower runs about 1mA down through the 4.99k resistor, with the source sitting about 0.6V above ground. Now the effective input voltage noise placed across the photodiode capacitance is the 1nV/√Hz of the JFET rather than the 8nV/√Hz of the op amp. The op amp is looking into its own 3pF of input capacitance plus the 2pF of gate-drain capacitance, plus parasitics. That’s a much better situation than looking into 3000pF!

The effects of this simple modification are drastic. The compensation capacitor C_F can be reduced, and bandwidth is improved to 220kHz (1.58μs rise time). Output noise density at 10kHz is reduced to 221nV/√Hz, as shown in Figure 10. DC performance remains excellent because the JFET is not involved; it simply provides a slight reverse bias to the photodiode.

Conclusion

The LTC6241 and LTC6242 combine the low noise, offset, and drift of the best bipolar op amps with low input bias and noise current of CMOS op amps. These amplifiers operate from 2.7V to ±5.5V and represent all-in-one solutions for fast, low noise signal processing.

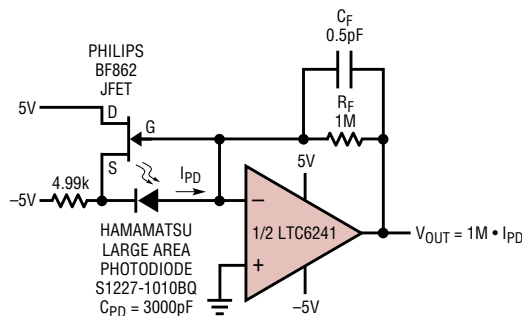


Figure 9. A simple bootstrap circuit drastically improves the ACs while leaving the DCs excellent. Output noise is now 221nV/√Hz at 10kHz, and bandwidth is 220kHz. Rise time is 1.58μs from a 3000pF photodiode at 1MΩ of gain!

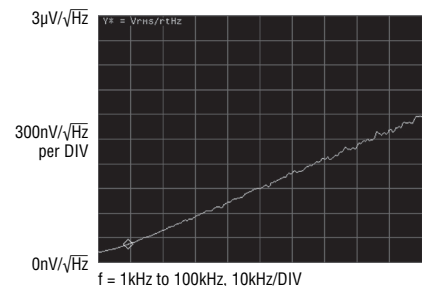


Figure 10. Output noise spectral density of the bootstrap circuit of Figure 9

Photoflash Capacitor Chargers Keep Up with Shrinking Cameras

by Mike Negrete

Introduction

Camera-phones have come a long way since the first generation of integrated cameras offered low-resolution CMOS images through the eye of a plastic lens. Now PDAs and high-end cell phones include high quality cameras with 2 megapixel resolutions and glass optics. Since these devices are carried by most users at all times, size is of the utmost importance. LED flashes were introduced in early model cell phone cameras, but they cannot produce enough light and lack the spectral quality required for higher-end cameras. Although xenon flashes are an optimal source of light for photography, they required substantially more board space than LED flashes until

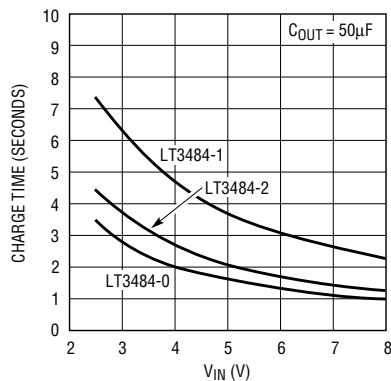


Figure 2. Charge time for the LT3484

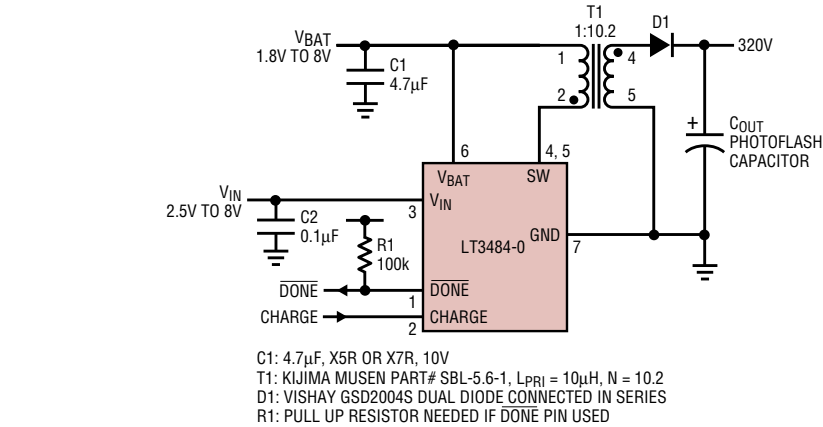


Figure 1. Compact, 320V photoflash capacitor charging circuit needs no external Schottky diode

the LT3468 allowed xenon flashes to fit into the spaces of cell phones and PDAs. The LT3484 and LT3485 photoflash capacitor chargers improve upon the LT3468.

The LT3484 and LT3485 are based on the LT3468's patented control scheme, providing well controlled battery current, fast charge times and high efficiency. Both series of parts use the same tiny, low-profile transformers as the LT3468. Available in a 6-Lead 2mm × 3mm DFN, the LT3484 reduces the board space significantly with its smaller package and total solution size compared to the

LT3468. The LT3484 has also added an additional pin, V_{BAT}, to allow it to operate from two alkaline cells. For xenon photoflash applications with an IGBT, the LT3485 decreases the solution size further with the same photoflash functionality as the LT3484 and an integrated IGBT driver in its 10-Lead 3mm × 3mm DFN package. The LT3485 also features an output voltage monitor pin.

Overview

A typical application circuit for the LT3484 is shown in Figure 1. With a high level of integration inside the part,

Table 1. Photoflash capacitor charger features

	LT3484-0	LT3484-1	LT3484-2	LT3485-0	LT3485-1	LT3485-2	LT3485-3
Peak SW Current (A)	1.4	0.7	1.0	1.4	0.7	1.0	2.0
Average Input Current (mA) (V _{IN} = 3.6V, V _{OUT} = 225V)	500	250	400	500	250	400	750
Charge Time Coefficient Kijima (τ)	0.65	0.30	0.50	0.75	0.34	0.51	NA
Charge Time Coefficient TDK (τ)	0.62	0.32	0.51	0.73	0.37	0.51	1.10
Minimum Battery Voltage(V)		1.8			1.8		
Integrated IGBT Drive + V _{OUT} Monitor		No			Yes		
External Schottky Diode Required		No			No		
Package		2mm × 3mm DFN 6L			3mm × 3mm DFN 10L		

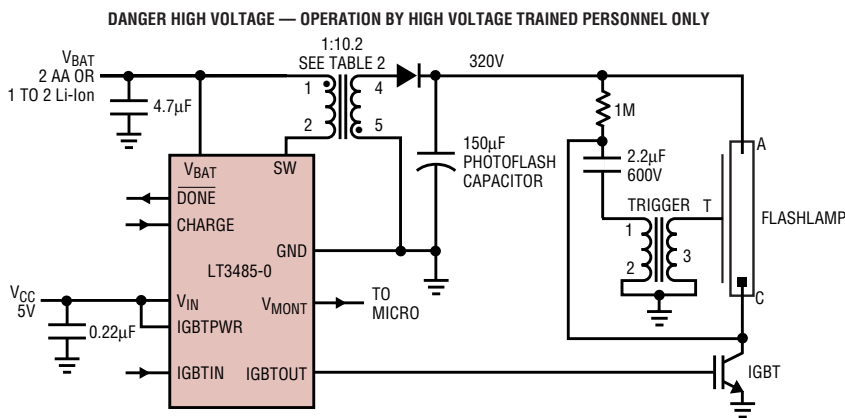


Figure 3. Compact, 320V photoflash capacitor charging circuit with integrated IGBT drive

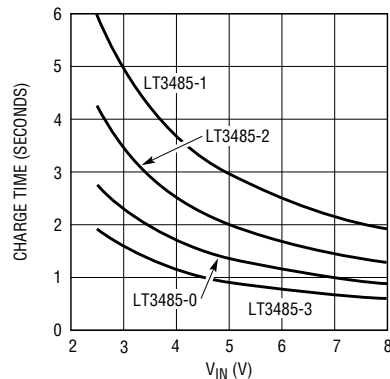


Figure 4. Charge time for the LT3485

the application circuit only requires a tiny, low-profile transformer, a high voltage diode, and an input bypass capacitor to charge any size output capacitor to 320V. Despite requiring only 70mm² of valuable board space, the patented control scheme with its high power, integrated low resistance NPN power switch produces fast charge times shown in Figure 2. There are three versions of the LT3484 depending on charge time and input current requirements. The LT3484-0 has the highest input current at 500mA, while the LT3484-1 has the lowest average input current at 225mA. The LT3484-2 has an input current at 375mA.

A typical application circuit for the LT3485 is shown in Figure 3. In addition to the photoflash capacitor charging circuitry, the LT3485 integrates an IGBT drive and a voltage output monitor. The integrated IGBT

drive saves valuable board space and cost by eliminating several external components. The voltage output monitor provides a solution to monitor the output voltage without resorting to a resistor divider on the output, which would drain the output capacitor. Along with identical current level versions of the LT3484, the LT3485 series features a high input current part, the LT3485-3, at 750mA. Typical charge times are shown in Figure 4.

Operation

Figure 5 shows a block diagram for the LT3484 and LT3485, which have identical operation except for the IGBT drive and voltage output monitor in the LT3485—highlighted in the diagram. A low-to-high transition on the CHARGE pin initiates the part. An edge triggered one-shot triggered by the CHARGE pin

puts the various latches inside the part into the proper state.

The part begins charging by turning on the power NPN transistor Q1. With Q1 on, the current in the primary of the flyback transformer increases. When it reaches the current limit, Q1 is turned off and the secondary of the transformer delivers current to the photoflash capacitor via diode D1. During this time, the voltage on the SW pin is proportional to the output voltage. Since the SW pin is higher than V_{BAT} by an amount roughly equal to $(V_{OUT} + 2 \cdot V_D)/N$, the output of the discontinuous conduction (DCM) mode comparator is high. In this equation, V_{OUT} is the photoflash capacitor voltage, V_D is the rectifying diode forward drop, and N is the turns ratio of the transformer.

Once the current in the secondary of the transformer decays to zero,

Table 2. Pre-designed transformers — typical specifications unless otherwise noted

For Use With	Transformer Name	Size (W × L × H) mm	L _{PRI} (µH)	L _{PRI} -Leakage (nH)	N	R _{PRI} (MΩ)	R _{SEC} (Ω)	Vendor
LT3484/5-0	SBL-5.6-1	5.6 × 8.5 × 4.0	10	200 Max	10.2	103	26	Kijima Musen Hong Kong Office 852-2489-8266 (ph) kijimahk@netvigator.com (email)
LT3484/5-2	SBL-5.6-1	5.6 × 8.5 × 4.0	10	200 Max	10.2	103	26	
LT3484/5-1	SBL-5.6S-1	5.6 × 8.5 × 3.0	24	400 Max	10.2	305	55	
LT3484/5-0	LDT565630T-001	5.8 × 5.8 × 3.0	6	200 Max	10.4	100 Max	10 Max	TDK Chicago Sales Office (847) 803-6100 (ph) www.components.tdk.com
LT3484/5-1	LDT565630T-002	5.8 × 5.8 × 3.0	14.5	500 Max	10.2	240 Max	16.5 Max	
LT3484/5-2	LDT565630T-003	5.8 × 5.8 × 3.0	10.5	550 Max	10.2	210 Max	14 Max	
LT3485-3	LDT565630T-041	5.8 × 5.8 × 3.0	4.7	150 Max	10.4	90 Max	16.4 Max	
LT3485-0	T-15-089	6.4 × 7.7 × 4.0	12	400 Max	10.2	211 Max	27 Max	Tokyo Coil Engineering Japan Office 0426-56-6262 (ph) www.tokyo-coil.co.jp
LT3485-1	T-15-089	6.4 × 7.7 × 4.0	12	400 Max	10.2	211 Max	27 Max	
LT3485-1	T-15-083	8.0 × 8.9 × 2.0	20	500 Max	10.2	675 Max	35 Max	
LT3485-3	T-17-109A	6.5 × 7.9 × 4.0	5.9	300 Max	10.2	78 Max	18.61 Max	

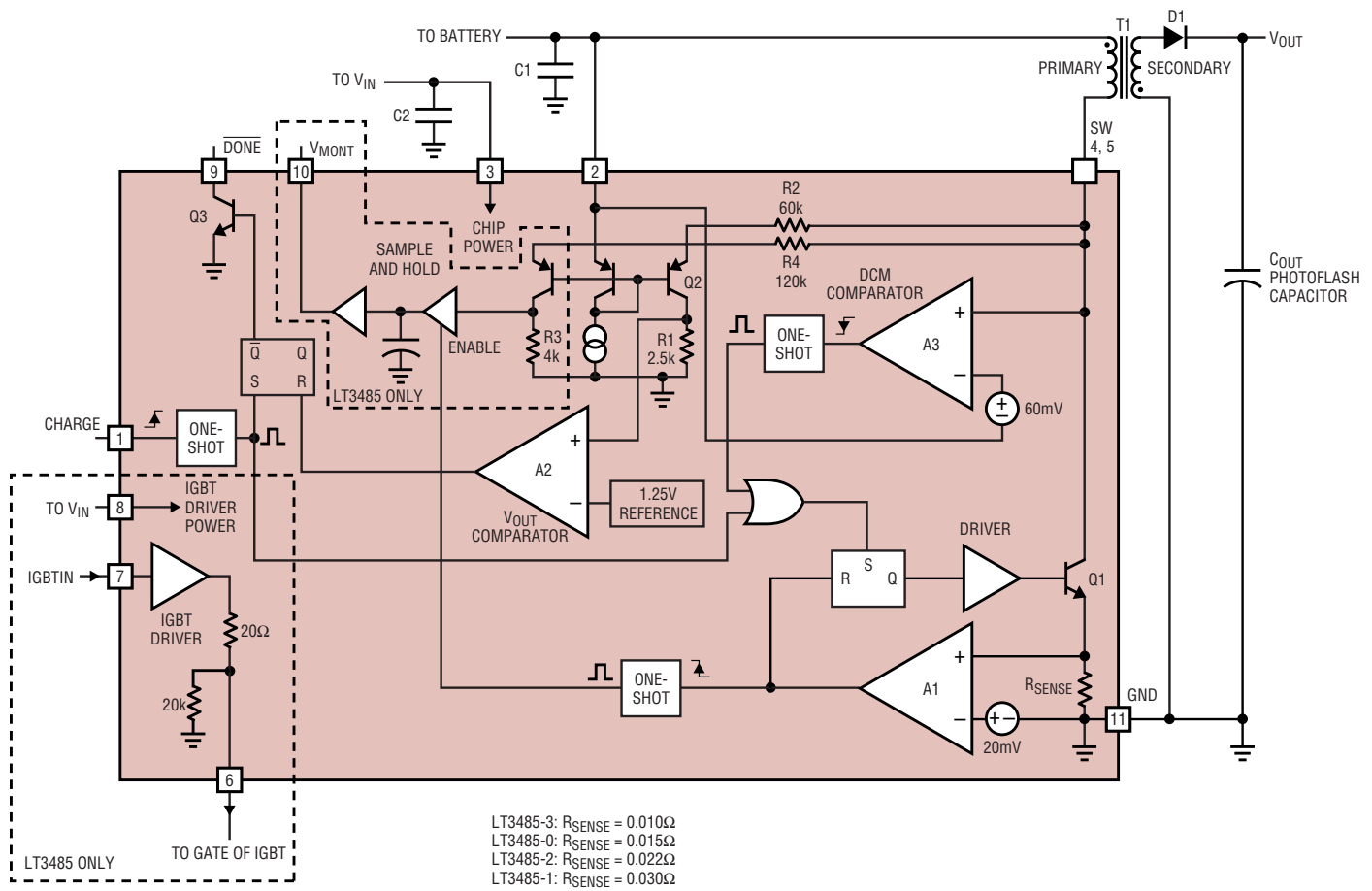


Figure 5. Block diagram for the LT3484 and the LT3485

the voltage on the SW pin collapses to V_{BAT} , or lower. As a result, the output of the DCM comparator goes low, which triggers the one-shot. This leads to Q1 turning on again and the cycle repeats.

Output voltage detection is accomplished via comparator A2. When the SW pin is 31.5V higher than V_{BAT} on any cycle, the output of A2 goes high. This resets the master latch and the part stops delivering power to the photoflash capacitor. Power delivery can only restart by taking the CHARGE pin low and then high.

Note that the flux in the flyback transformer is brought to zero on each switching cycle. This is generally referred to as boundary mode operation since the transformer is operated in between continuous conduction mode and discontinuous conduction mode (CCM and DCM respectively). When the CHARGE pin is forced low at any time, the LT3484/LT3485 ceases power delivery and goes into shutdown mode, thus reducing quiescent current to less than 1μA. Figure 6 shows some typical waveforms for the LT3484 and LT3485.

output, V_{MONT} , proportional to the capacitor voltage. This output can easily be read by a microcontroller with an ADC.

Figure 7 shows the measured output of V_{MONT} . Because of the high speed nature of the circuit and the high dV/dt of the switch pin, there is a small amount of ripple on the V_{MONT} output, which can be reduced by adding a 0.1μF capacitor to the output or by using the ADC to sample the V_{MONT} output multiple times and taking the average.

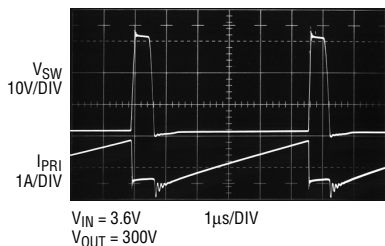


Figure 6. A LT3485 switching waveform at 300V output

Voltage Output Monitor

Camera manufacturers continue to try to differentiate their product with novel features such as strobe shots and sequential shots. These new features rely on fast capacitor charging to be done in the time between shots. If the capacitor is not fully charged, is the voltage high enough to produce a flash? The LT3485 addresses this problem by including a 1V full-scale

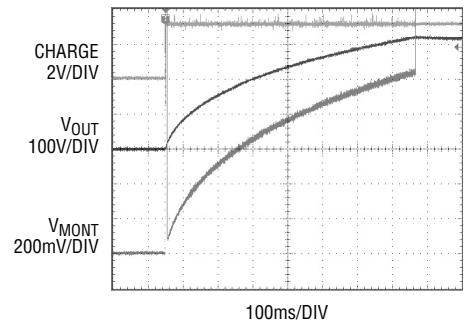


Figure 7. Voltage output monitor waveform during charging

IGBT Drive

Most camera flashes are capable of reeye reduction and light-feedback flashing. These features quench, or stop, the flash before the capacitor drains completely. This added level of control requires a high current, high voltage Insulated Gate Bipolar Transistor (IGBT). An IGBT has the advantage of a BJT's high voltage and high current capabilities but does not need base current since it has a MOSFET gate as the input. The tradeoff for these two advantages is speed. Since a flash is on the order of milliseconds, speed is not an issue in this application and an IGBT fits perfectly for the role.

Like a MOSFET, the gate acts like a capacitor. The IGBT driver's job is to charge and discharge the gate. The IGBT driver does not need to be fast, and actually a fast driver can potentially destroy the device. The IGBT turns on when the IGBTIN pin is above 1.5V and turns off when the IGBTIN pin is below 0.3V. When the input is high, the driver draws a small amount of current to hold the gate high with a PNP. When the input is low, the driver has zero quiescent current. During transitions the driver is capable of delivering 150mA of current.

The speed of the driver needs to be carefully controlled or the IGBT may be destroyed. The IGBT driver does not need to pull up the gate fast because of the inherently slow nature of the IGBT. A rise time of 2 μ s is sufficient to charge the gate of the IGBT and create a trigger pulse. With slower rise times, the trigger circuitry does not have a fast enough edge to create the required 4kV pulse. The fall time of the IGBT drive is critical to the safe operation of the IGBT. The IGBT gate

is a network of resistors and capacitors. When the gate terminal is pulled low too quickly, the capacitance closest to the terminal goes low but the capacitance further from the terminal remains high, causing a small portion of the IGBT device to handle the full 100A of current which quickly destroys the device. The pull down circuitry therefore needs to be slower than the internal RC time constant in the gate of the IGBT. To slow down the driver, a 20 Ω series resistor is integrated into the LT3485.

Which Part to Use

The LT3484 and LT3485 families of photoflash capacitor chargers suit about any photoflash need. The basic photoflash functionality in each part is identical and both parts are capable of operating from 2AA cells. The integrated IGBT drive and voltage output monitor differentiate the LT3485 from the LT3484, along with its higher current capabilities. The LT3484 is the smallest solution available if quenching the bulb is not needed. When using an IGBT to trigger the flash, the LT3485 offers valuable board space savings over the LT3484 by eliminating several external components. Table 1 shows the major functional differences between these seven parts.

Once the decision is made on the integrated IGBT driver, choosing a current option is a matter of balancing the inherent trade-off between input current and charge time. For a given photoflash capacitor size, the device which results in the highest input current offers the fastest charge time. The limit on how much current the photoflash charger can draw is usually set by the battery technology used, and how much load they

can handle. The LT3485-3 offers the fastest charge times of the chargers discussed here.


The following equation predicts the charge times (T) in seconds for the seven parts:

$$T = \frac{C_{OUT} \cdot (V_{OUT(FINAL)}^2 - V_{OUT(INIT)}^2)}{\tau \cdot V_{IN}}$$

where C_{OUT} is the value of the photoflash capacitor in Farads, $V_{OUT(FINAL)}$ is the target output voltage, $V_{OUT(INIT)}$ is the initial output voltage, V_{IN} is the battery voltage to which the flyback transformer is connected, and τ is the charge time coefficient listed in Table 1.

The charge time coefficients for each part are different depending on the transformer due to differences in efficiency and average input current. The charge time coefficients are given for Kijima Musen and TDK transformers, with part numbers and typical specifications for these transformers listed in Table 2.

Conclusion

The LT3484 and LT3485 provide simple, efficient capacitor charging solutions for digital still cameras and integrated digital cameras in cell phones. The high level of integration reduces the amount of external components while also producing tightly controlled output voltage and average input current distributions. The three current limits in the LT3484 family and the four current limits in the LT3485 family allow for flexibility in the trade-off between input current and charge time. The LT3485 saves even more space for some applications by integrating an IGBT driver and voltage output monitor. 

LT6555/56, continued from page 22

Demonstration Circuits Available

The LT6555 and LT6556 have Demo Boards available that make evaluation of these parts a simple plug-and-play operation. To evaluate the LT6555 ask for DC858A (SSOP-24 package) or DC892A-A (QFN package). To evalu-

ate the LT6556 ask for DC892A-B (in QFN package). All three of these demo circuits have high-quality 75 Ω BNC connections for best performance and illustrate high-frequency layout practices that are important to obtaining the best performance from these super-fast amplifiers. 

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

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Fully Differential Amplifier with Rail-to-Rail Outputs Offers 16-Bit Performance at 1MHz on a Single 2.5V Supply

by Arnold Nordeng

Introduction

With increasing levels of IC integration, and shrinking transistor geometries, A/D converter supply voltages have decreased and their inputs have been designed to process signals differentially to maintain good dynamic range. These ADCs typically run from a single low voltage supply with an optimal common mode input somewhere near mid-supply. The LT1994 facilitates interfacing to these ADCs by providing differential conversion and amplification, common mode translation of wide band, ground referenced, single-ended or differential input signals. It comes in an 8-pin MSOP or DFN package, which is pin-for-pin compatible with other commercially available fully-differential amplifiers.

What sets the LT1994 apart from other fully-differential amplifiers are its low noise, low distortion, rail-to-rail output, and an input common mode range that extends to ground on power supplies as low as 2.5V. This eliminates the need for a negative power supply, and makes the LT1994 uniquely able to interface to differential input ADCs while sharing the same power supply. This saves the user system cost, and power.

Performance of LT1994

The first advantage of the LT1994 is that it can convert and level-shift ground referenced, single-ended or differential signals to V_{OCM} pin referenced, differential output signals. Figure 1 shows how. A single-ended $5V_{P-P}$ ground referenced signal (which swings 2.5V below the supply of both the ADC and the LT1994) is translated by the LT1994 from being a ground referenced signal to a differential mid-supply referenced signal. This is accomplished within the LT1994 by two feedback loops: a differential feedback loop, and a common mode

feedback loop. Both loops have high open loop gain, around 100dB. The common mode feedback loop forces the instantaneous average of the two outputs to be equal to the voltage on the V_{OCM} pin. Its feedback loop is internal to the LT1994. The differential feedback loop works similarly to traditional op amps forcing the difference in the summing node voltages to zero. As a result, the differential output is simply governed by the equation:

$$V_{OUT} = V_{OUT+} - V_{OUT-} \approx \frac{R_F}{R_I} \cdot V_{IN}$$

By eliminating the need for a negative supply, the LT1994 gives the user maximum dynamic range at minimal power. Since each output of the LT1994 is capable of swinging rail-to-rail, and with the LT1994's $3nV/\sqrt{Hz}$ input referred voltage noise (see Figure 2 for the LT1994's noise spectral density plot), applications such as the one shown in Figure 1 have a signal-to-noise ratio approaching 96dB in a 10MHz noise bandwidth. This represents a 6dB increase in dynamic range compared to single ended output rail-to-rail amplifiers

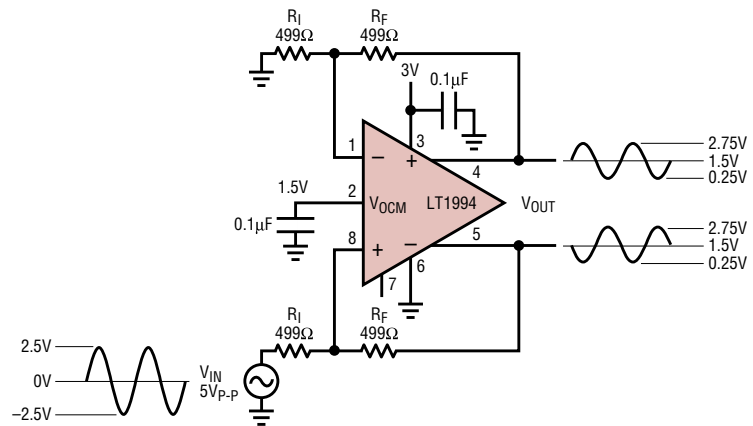


Figure 1. Common mode translation of V_{IN} using the LT1994

Table 1. LT1994 key specifications

Parameter	Typical Specification
Supply Current at 3V	13.3mA
e_n – Input referred Voltage Noise	$3nV/\sqrt{Hz}$
HD2 at $V_{IN} = 2V_{P-P}$ 1MHz	-94dBc
HD3 at $V_{IN} = 2V_{P-P}$ 1MHz	-108dBc
Gain-Bandwidth	70MHz
Slew Rate	65V/ μ s
0.01% Settling on a 2V step	120ns

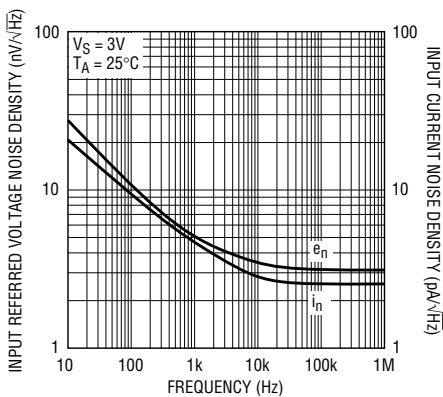


Figure 2. LT1994 input referred noise spectral density

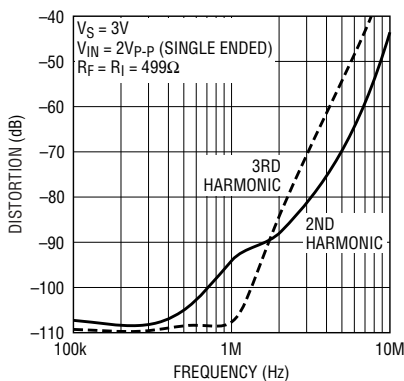


Figure 3. LT1994 distortion vs frequency

with the similar noise floors. Some of the LT1994's key specifications are tabulated in Table 1.

Another benefit of fully-differential signal processing is that interference such as ground noise or power supply noise appear as common mode signals and are rejected by the internal matching and balance of the amplifier. Power supply rejection and common-mode rejection becomes limited primarily by internal transistor matching and are typically around 100dB.

Linearity is enhanced using a fully differential architecture allowing the cancellation of even order harmonics. To see how this works, a pure single tone sine wave input is applied to the LT1994 as shown in Figure 1. The outputs of the LT1994 can be represented by a Taylor series expansion:

$$V_{OUT+} = K_1 \left(\frac{V_{IN}}{2} \right) + K_2 \left(\frac{V_{IN}}{2} \right)^2 + K_3 \left(\frac{V_{IN}}{2} \right)^3 + K_4 \left(\frac{V_{IN}}{2} \right)^4 + \dots$$

$$V_{OUT-} = K_1 \left(-\frac{V_{IN}}{2} \right) + K_2 \left(-\frac{V_{IN}}{2} \right)^2 + K_3 \left(-\frac{V_{IN}}{2} \right)^3 + K_4 \left(-\frac{V_{IN}}{2} \right)^4 + \dots$$

V_{OUT} is the difference:

$$V_{OUT} = V_{OUT+} - V_{OUT-} = 2K_1 \left(\frac{V_{IN}}{2} \right) + 2K_3 \left(\frac{V_{IN}}{2} \right)^3 + \dots$$

leaving just the odd harmonic terms.

Figure 3 shows a plot of distortion vs frequency with the LT1994 configured in the closed-loop unity gain configuration shown in Figure 1. With a 2V_{P-P}, 1MHz, single-ended input, the 2nd harmonic measures -94dBc, and the 3rd harmonic measures -108dBc.

Getting the best distortion out of the LT1994 requires careful layout, paying close attention to symmetry and balance. In single supply applications, it is recommended that high quality, low ESR, surface mount 1μF and 0.1μF caps be paralleled and tied directly

across the power supplies with short traces with the V⁻ tied directly to a low-impedance ground plane. On split supplies, additional 0.1μF high quality, low ESR, surface-mount bypass caps should be used to bypass each supply separately to a low-impedance ground plane.

Interfacing to ADCs

The sampling process of ADCs create a sampling glitch caused by switching in the sampling capacitor on the ADC front end which momentarily “shorts” the output of the amplifier as charge is transferred between the amplifier and the sampling cap. The amplifier must recover and settle from this load transient before this acquisition period ends for a valid representation of the input signal.

In general, the LT1994 settles faster from these periodic load impulses than from a 2V input step, but it is a good idea to place a small RC filter network between the output of the LT1994 and the input of the ADC to help absorb the charge injection that comes out of the ADC from the sampling process (Figure 4 shows an example of this). The capacitance of this decoupling network serves as a charge reservoir to provide high frequency charging during the sampling process, while the two resistors of the decoupling network are used to dampen and attenuate any charge kickback from the ADC.

The selection of the RC time constant is trial and error for a given ADC, but the following general guidelines are recommended: Too large a resistor in the decoupling network leaving

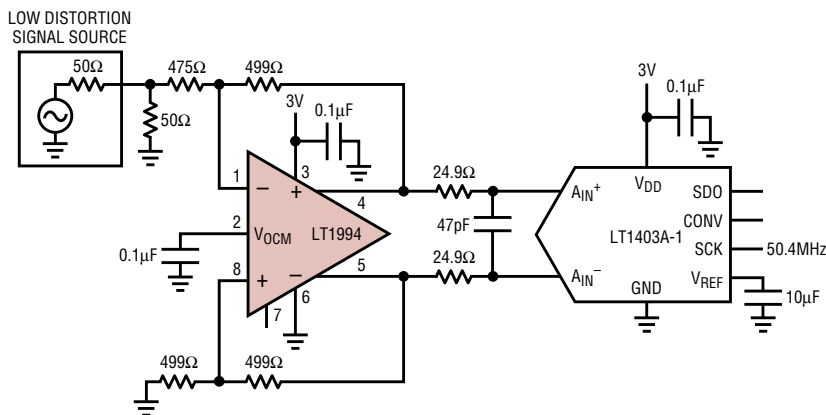


Figure 4. ADC buffering with common mode translation and differential conversion

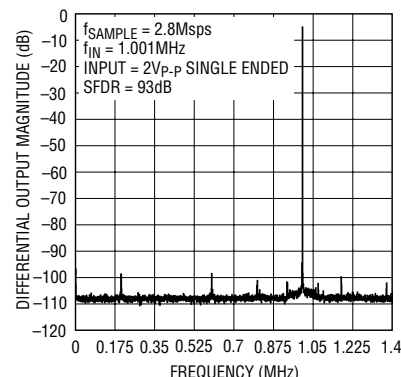


Figure 5. 4096 sample FFT of the LT1994 driving a 14-bit ADC

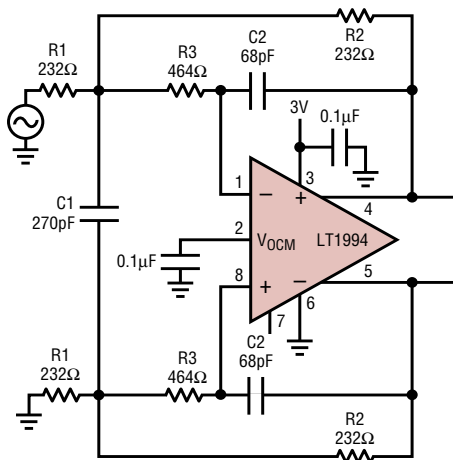


Figure 6. Low noise differential active RC filter

insufficient settling time creates a voltage divider between the dynamic input impedance of the ADC and the decoupling resistors. Too small of a resistor possibly prevents the resistor from properly dampening the load transient caused by the sampling process, prolonging the time required for settling.

Start with 25Ω on each output to decouple the ADC input capacitance. Then choose a capacitance (taking account of the sampling capacitance), which gives the amplifier time to settle to desired accuracy during the acquisition period. In 16-bit applications, this typically requires a minimum of 11 RC time constants. The capacitor chosen should have a high quality dielectric (for example, COG multi-layer ceramic). Figure 4 shows the LT1994 driving the LTC1403A-1, a 14-bit ADC, sampling at 2.8MHz on a single 3V supply. Figure 5 shows its 4096-

point FFT. The spurious free dynamic range is about 93dB and is limited by the non-linearities of the ADC rather than the LT1994 (The SFDR of the LTC1403A-1 is specified around 86dB at 1.4MHz). This shows that the LT1994 has no problem settling and accommodating the LTC1403A's 39ns acquisition times.

Single 3V Supply, 2.5MHz, 2nd Order Fully Differential Butterworth Filter

Figure 6 shows a low noise, single supply, butterworth active filter with a 2.5MHz bandwidth suitable for anti-aliasing applications. The differential output spot noise at 50kHz is about 7nV/√Hz, and the amplifier provides about 40dB of stopband rejection at 25MHz. The filter's frequency response is shown in Figure 7. The filter's low frequency gain is set by the ratio of R2 to R1. If a different cutoff frequency is desired, the capacitors C1, and C2 can easily be scaled inversely with cutoff frequency.

Gain-of-2 Amplifier (No resistors required)

Figure 8 shows the LT1994 configured in circuit configuration in which the output consists of an in-phase and an out-of-phase representation of the input signal. The circuit has the benefit of having high input impedance. The input-to-output transfer function is governed by the equation: $V_{OUT} = 2 \cdot V_{IN}$

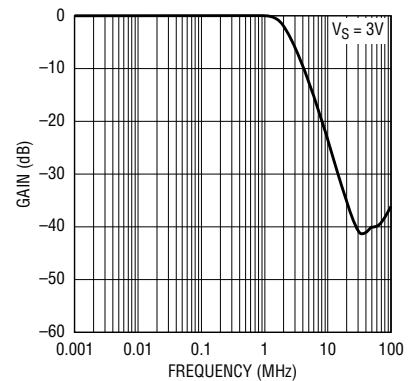



Figure 7. Differential filter response

The circuit works well enough, but the consequence of such a configuration is that it reflects the performance of the common mode path, rather than the differential path. Because of this, the output does not have the benefit of the differential noise (3nV/√Hz), but rather is swamped by the common mode noise of 15nV/√Hz gained up by a factor of two (30nV/√Hz). This is a consequence of mismatch in feedback factors from the LT1994 outputs to their respective inputs. In fact, whenever the two feedback paths from the output to the input mismatch, and to the degree they mismatch, common mode noise is converted to differential noise at the output.

$$e_{NO(DIFF)} = 2e_{N(VOCM)} \frac{(\beta_{F1} - \beta_{F2})}{(\beta_{F1} + \beta_{F2})}$$

where β_{F1} , and β_{F2} are the two feedback factors from each output to their respective input.

Conclusion

The LT1994's low noise, low distortion, and high performance make it an ideal amplifier for interfacing with single supply ADCs. Its rail-to-rail outputs, low distortion, and 3nV/√Hz input referred voltage noise maximize dynamic range, and its ability to common mode to ground eliminates the need for a negative supply in single supply systems, saving cost and power. 

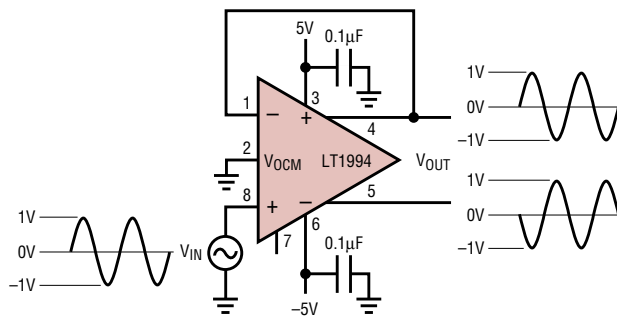


Figure 8. Gain of two (no resistors required)

For more information on parts featured in this issue, see <http://www.linear.com/designtools>

Negative High Voltage Hot Swap Controllers Incorporate an Accurate Supply Monitor and Power Module Sequencing

by Kevin Wong

Introduction

When a circuit board is inserted into a live backplane slot, discharged supply bypass capacitors on the board can draw large transient currents from the system supplies. In high-voltage systems like the -48V backplanes prevalent in high reliability telecom systems, such transients can reach hundreds of amps and damage connector pins, PCB traces and board components. In addition, current spikes can cause voltage glitches on the power bus, causing other boards in the system to reset. This is particularly unacceptable in telecom systems where the ability to safely Hot Swap modules is a primary system requirement.

The LTC4253A and LTC4253A-ADJ facilitate safe board insertion and removal from a live backplane by applying power in a controlled manner. Running off a simple, fast responding

shunt regulated supply that allows very high voltage operation, they are uniquely suited for applications on the -48V bus.

User programmable, high accuracy undervoltage and overvoltage detectors act as supply monitors and ensure the supply is stable and within tolerance before applying power to the load. An inrush current control loop then takes over, resulting in a controlled startup current profile. When the external pass transistor is fully enhanced, Power Good status outputs then allow time adjustable or load feedback enabled sequencing of up to four load modules. Short circuits or excessive supply current events trigger protective circuits which quickly isolate the fault to prevent glitching of the backplane supply. With all these features, these devices offer a comprehensive solution for -48V Hot Swap applications.

Typical Application

Figure 1 shows a typical -48V Hot Swap application using the LTC4253A. The LTC4253A floats on the negative rail and uses an internal shunt regulator that together with R_{IN} and C_{IN} , regulates V_{IN} to about 13V above the negative rail. The MOSFET N-channel transistor Q1 is placed in the power path to control turn on and turn off with input from resistor R_S which senses the load current. R_C and C_C provides compensation for the current limit loop. R_1 and R_2 form a resistive divider that allows MOSFET turn on only when the -48V supply is between the user programmed undervoltage and overvoltage thresholds. The resistive divider R_1/R_2 connects to the -48V RTN rail via a short pin so that during plug in, the MOSFET is held off by the undervoltage condition until the longer power pins are properly seated. The five opto-couplers form an

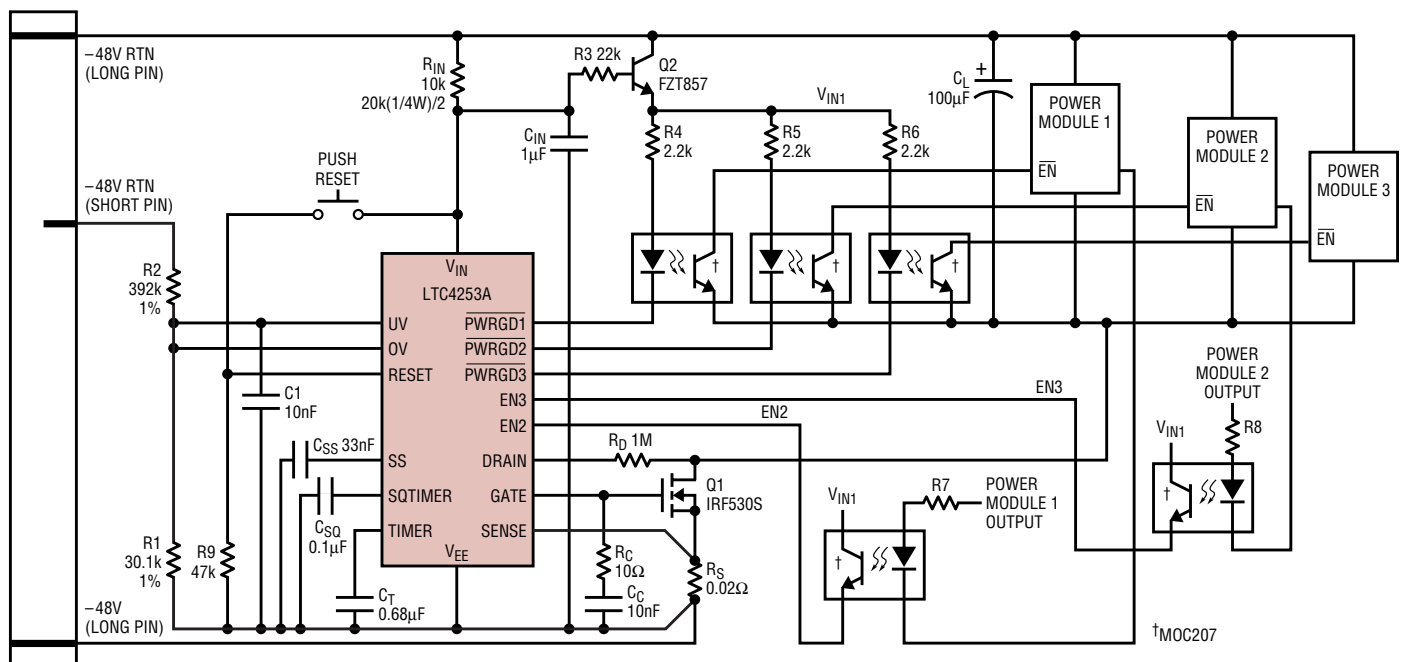


Figure 1. $-48\text{V}/2.5\text{A}$ Hot Swap controller with opto-isolated Power Good sequencing

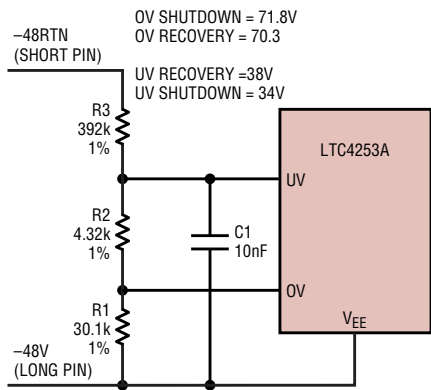


Figure 2. Undervoltage and overvoltage resistive divider connection to the LTC4253A

electrically isolated interface between LTC4253A and the load modules for power sequencing.

Undervoltage and Overvoltage Detection

The LTC4253A and LTC4253A-ADJ have 1% accurate undervoltage and overvoltage threshold detectors that can be set to any desired power supply range. This level of accuracy and flexibility allow these parts to be easily designed to conform to any operating

ranges specified by the various prevailing -48V standards.

In the LTC4253A, an UV hysteric comparator detects undervoltage conditions at the UV pin, with the following thresholds (with respect to V_{EE}):

- UV low-to-high (V_{UV}) = 3.08V
- UV low-to-high hysteresis (V_{UVHST}) = 0.324V

An OV hysteric comparator detects overvoltage conditions at the OV pin, with the following thresholds (with respect to V_{EE}):

- OV low-to-high (V_{OV}) = 5.09V
- OV low-to-high hysteresis (V_{OVHST}) = 0.102V

The undervoltage recovery and overvoltage shutdown thresholds are designed to match the standard telecom operating range of 43V to 71V with the UV and OV pins shorted as in Figure 1. The undervoltage shutdown and overvoltage recovery thresholds are then 38.5V and 69.6V respectively. The UV and OV pins can also be separated for implementing different operating ranges as shown in Figure 2.

The LTC4253A-ADJ offers additional flexibility in allowing the user to implement any required undervoltage recovery, undervoltage shutdown, overvoltage recovery and overvoltage shutdown thresholds. It achieves this by having two extra pins UVL and OVL connected to the internal comparators as shown in Figure 3. The undervoltage comparator has multiplexed inputs from UVL and UV, which is tapped off a resistive string across the power supply as in Figure 4. When comparator output UVD is high, UV is multiplexed to the comparator input UVIN. When UVD is low, UVL is multiplexed to UVIN. The overvoltage comparator similarly implements the overvoltage function. The various thresholds to note are (with respect to V_{EE}):

- UV low-to-high (V_{UVHI}) = 3.08V
- UVL high-to-low (V_{UVLO}) = 3.08V
- OV low-to-high (V_{OVHI}) = 5.09V
- OVL high-to-low (V_{OVLO}) = 5.08V

By tapping UVL, UV, OVL and OV off a resistive string across the power supply, undervoltage recov-

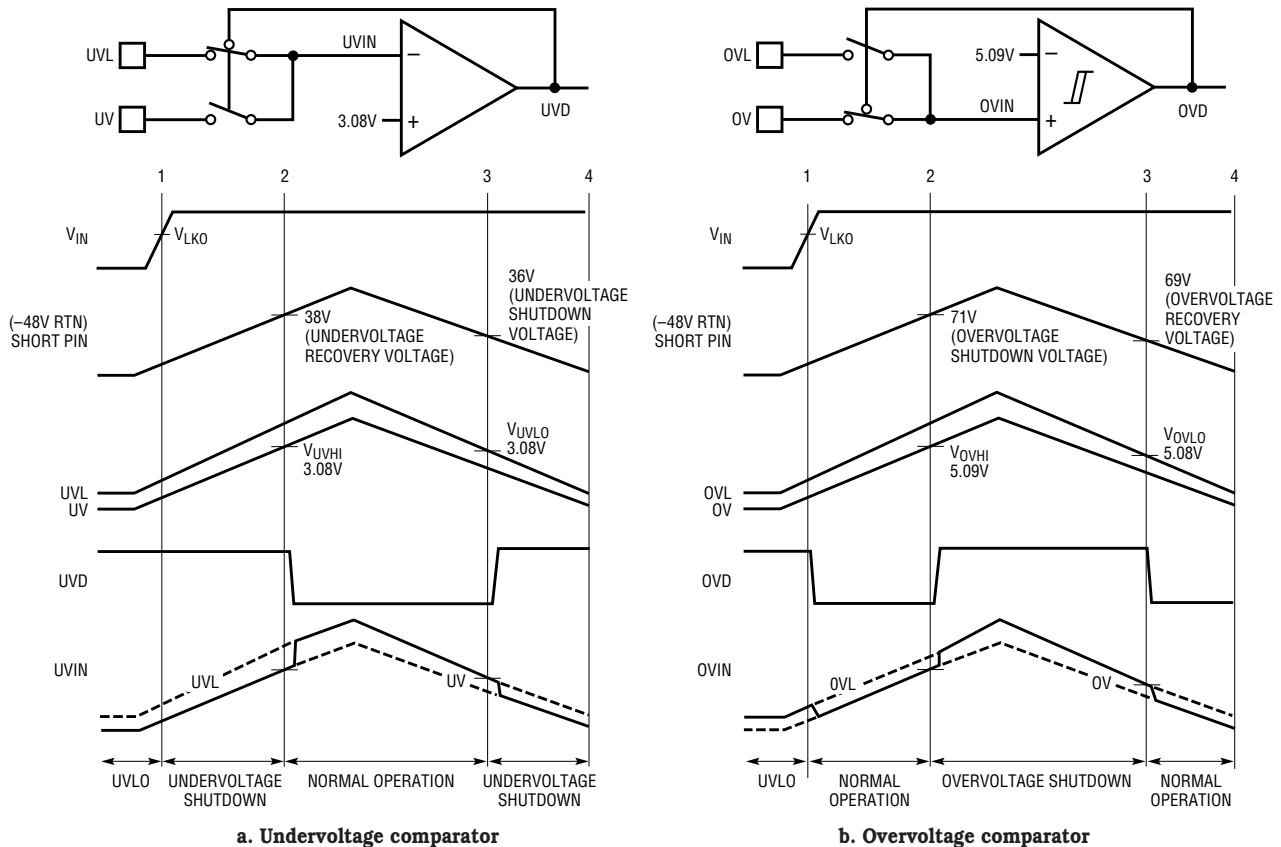


Figure 3. The LTC4253A-ADJ UV/OV detector block

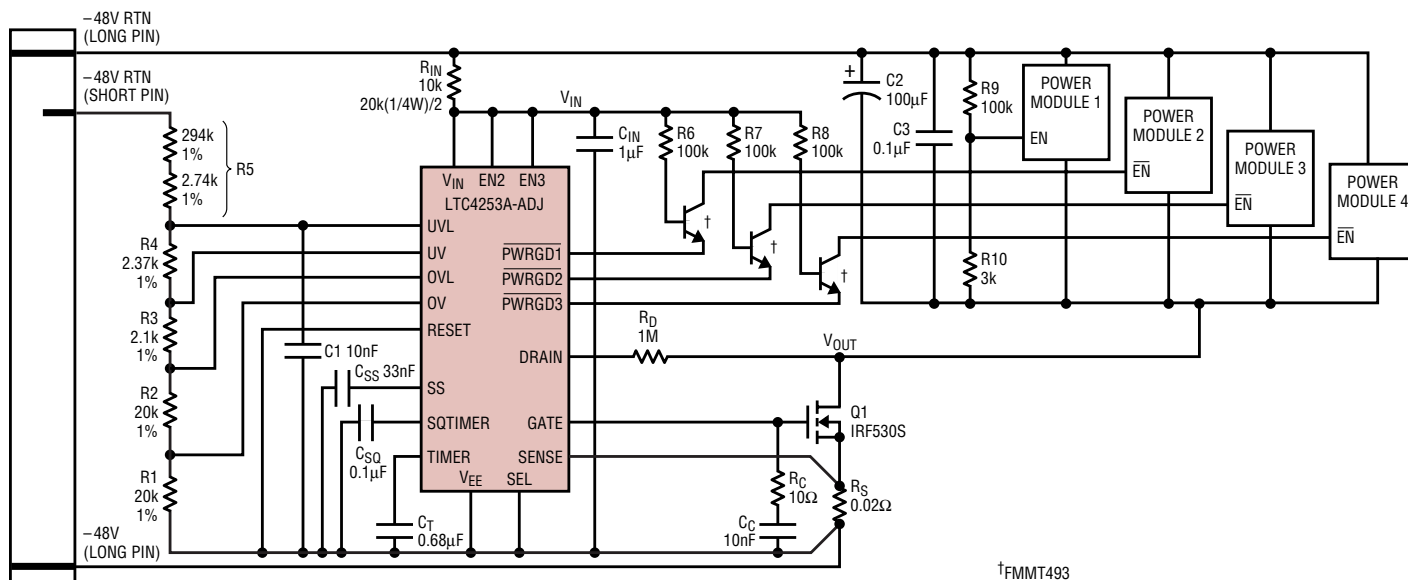


Figure 4. A -48V/2.5A Hot Swap controller with transistor enabled Power Good sequencing

ery = 43V, undervoltage shutdown = 39V, overvoltage recovery = 78V and overvoltage shutdown = 82V are implemented in Figure 4. Any required supply operating range can thus be implemented with great accuracy.

di/dt Soft Start

The LTC4253A offers a current soft start pin (SS) that acts as the reference for the analog current limit amplifier ($V_{ACL} = V_{SS}/20$). By attaching a capacitor at the SS pin, the analog current limit threshold ramps up in an exponential profile with an RC time constant equal to $50k\Omega \cdot C_{SS}$. The analog current limit amplifier forces the inrush current to follow this profile when the GATE pin rises above the

external MOSFET threshold and turns on the MOSFET. In this way, inrush current ramps up with a controlled slew rate (di/dt) that is approximately fixed and adjustable by C_{SS} (Figure 5a). Controlling the load current slew rate reduces system EMI and disturbances to the supply rail during startup.

The LTC4253A-ADJ offers an additional mode when the SEL pin is held low (it has an internal pullup to V_{IN} of $20\mu A$). In this mode, the SS pin is servoed from the time the GATE pin is released until it clears the external MOSFET threshold and turns the MOSFET on. The result is that the LTC4253A-ADJ enters analog current limit with V_{ACL} ramping up from close to zero. The resultant inrush current

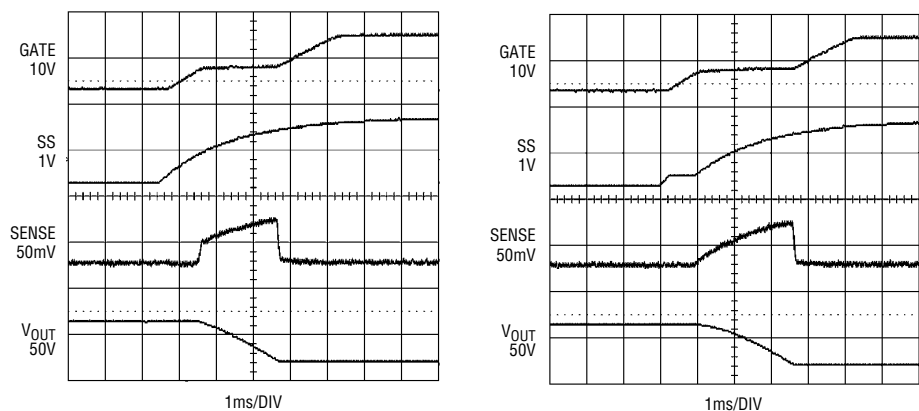
profile presents a smooth ramp up from zero and the load current slew rate is able to maintain an approximately fixed di/dt gradient right from turn on (Figure 5b). This di/dt gradient is similarly adjustable by C_{SS} .

Power Good Sequencing

The LTC4253A has three sequenced \overline{PWRGD} outputs and two enable (EN) inputs. This allows three load modules to be enabled sequentially, minimizing any sudden load power demand on the backplane supply.

The three load modules can be timer sequenced as in Figure 4 where the EN pins are enabled by tying them to V_{IN} . The three \overline{PWRGD} signals assert sequentially with a fixed time delay adjustable by capacitor C_{SQ} (approximate $TD = 600ms \cdot (C_{SQ}/1\mu F)$). The load modules can also be load feedback sequenced as in Figure 1 where the load modules control the EN pin inputs. In this way when Load Module 1 is enabled by $\overline{PWRGD1}$ and fully started up, it can signal back via the EN2 input to enable Load Module 2, which is enabled after one SQTIMER delay ramp. Load Module 2 can similarly enable Load Module 3 when it is ready. This mode of sequencing is shown in Figure 6.

The interface between the Hot Swap controller and the load modules is implemented with opto-couplers as in Figure 1 to take care of the differing



a. di/dt inrush current control using the LTC4253A or LTC4253A-ADJ with SEL = 1 (Figure 1 circuit)

b. Enhanced di/dt inrush current control using the LTC4253A-ADJ with SEL = 0 (Figure 4 circuit)

Figure 5. di/dt soft-start waveforms

signal common. If the load modules' \overline{EN} inputs have sufficient protection against negative bias current, a simpler NPN interface can be implemented as in Figure 4.

Figure 7 highlights an additional feature of the LTC4253A-ADJ. The $\overline{PWRGD1}$ signal only activates after one $SQTIMER$ ramp delay from the time $GATE$ goes high and $DRAIN$ goes low. This feature can be exploited to provide an additional $EN1$ signal so up to four load modules can be sequenced as in Figure 4.

Short Circuit Operation

Current faults are controlled in three stages using three thresholds: 50mV for a timed circuit breaker function, 60mV for an analog current limit loop and 200mV for a fast comparator that limits peak current in the event of a catastrophic short-circuit. This three-stage fault current minimizes backplane supply disturbances due to current faults.

A voltage across the $SENSE$ resistor (R_S) of greater than 50mV triggers $TIMER$ to source 200 μ A into a timing capacitor C_T . C_T eventually charges to a 4V threshold and the part latches off. If the fault goes away before C_T reaches 4V, C_T slowly discharges (5 μ A). A low impedance short can glitch the voltage across R_S above 200mV. This triggers a fast comparator that asserts a hard pulldown on the MOSFET gate to quickly bring the voltage across R_S

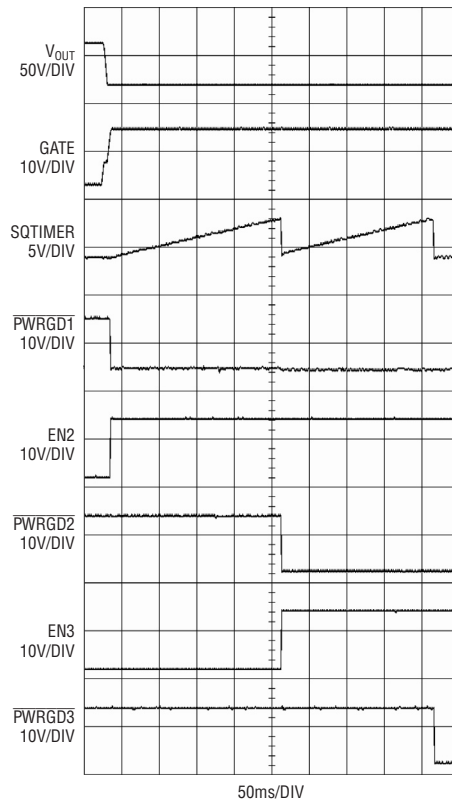


Figure 6. The LTC4253A controlling the turn on of three load modules using Load Feedback Sequencing

back below 200mV. This effectively limits the initial transient fault current. An analog current limit loop then controls the voltage across R_S to 60mV until $TIMER$ reaches 4V (see Figure 8).

R_D in Figure 1 allows the MOSFET drain to pump current into the $DRAIN$ pin internally clamped at around

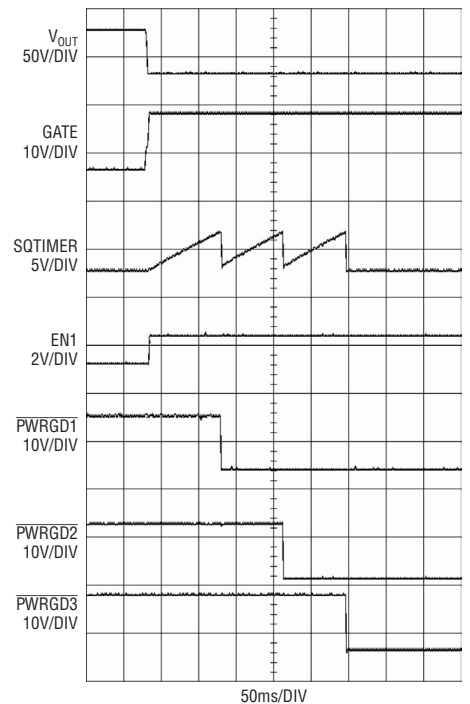



Figure 7. The LTC4253A-ADJ controlling the turn on of four load modules using Timing Sequencing

6V. This current is multiplied up by eight times and added to the 200 μ A circuit breaker $TIMER$ pullup current. This adds a component to the circuit breaker timeout period that is linearly proportional to the V_{DS} of the MOSFET, thus allowing the MOSFET to be designed to function closer to its SOA limits under different conditions.

Conclusion

The LTC4253A and LTC4253A-ADJ inherit the proven capabilities of Linear Technology's -48V Hot Swap family and add enhanced features. Chief among these is a highly flexible and 1% accurate undervoltage and overvoltage detection capability. Additional features include enhanced slew rate controlled inrush current profile and the ability to sequence up to four load modules. The LTC4253A is available in a 16-pin SSOP package and is completely pin compatible to the LTC4253. The LTC4253A-ADJ is available in a 20-pin SSOP package as well as a 20-pin 4mm x 4mm QFN package. 

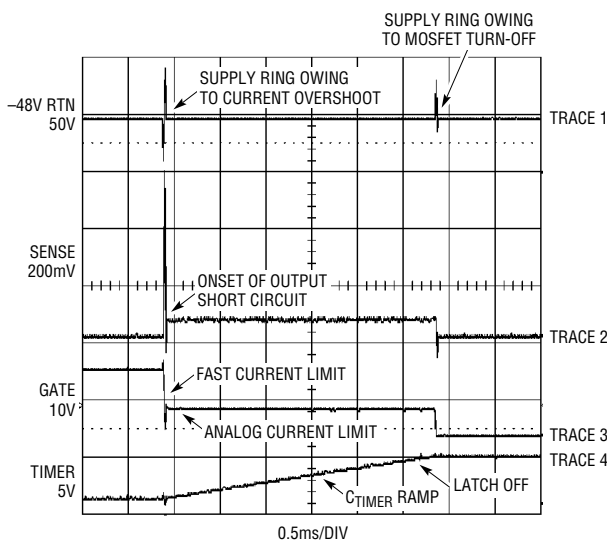


Figure 8. Output short-circuit waveforms

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Simplify High-Resolution Video Designs with Fixed-Gain Triple Multiplexers

by Jon Munson

Introduction

The LT6555 and LT6556 triple video multiplexers offer up to 750MHz performance in compact packages, requiring no external gain-setting resistors to establish a gain of two or unity. A single integrated circuit, in a choice of either 24-lead SSOP or 24-contact QFN (4mm x 4mm), performs fast switching between a pair of three-channel video sources, such as RGB or component HDTV.

The LT6555 provides a built-in gain of two that is ideal for driving back-terminated cables in playback or signal routing equipment. The LT6556 provides a unity-gain function, in the same footprints, that is ideal as an input selector in high-performance video displays and projectors.

The three video channels exhibit excellent isolation between themselves (50dB typical at 100MHz) and the inactive inputs (70dB typical at 100MHz) for the highest quality video transmission. Excellent channel-to-channel gain-matching preserves high fidelity color balance.

The increasing popularity of the UXGA professional graphics format (1600 x 1200), which generates a whopping 200-megapixel-per-second flow, has put exceptional demands on the frequency response of video amplifiers. For instance, pulse-amplitude

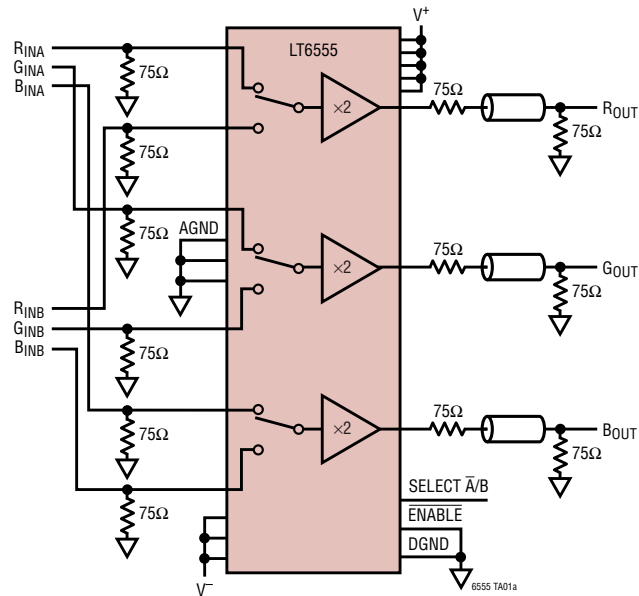


Figure 1. The LT6555 in an RGB cable driving multiplexer circuit

waveforms like those of RGB baseband video, generally require reproduction of high-frequency content to at least the 5th harmonic of the fundamental frequency component, which is 2.5 times the video pixel rate, accounting for the 2-pixels-per-fundamental-cycle relationship. This means that UXGA requires a flat frequency response to beyond 0.5GHz! The wide bandwidth performance of the LT6555 and LT6556 makes them ideally suited to such high performance video applications.

response and crosstalk anomalies can plague the circuit development process. The LT6555 and LT6556 conveniently solve these problems by providing internal factory-matched resistors and an efficient 3-channel, 2-input group, flow-through layout arrangement.

Figure 1 shows the typical RGB cable driver application of an LT6555, and its excellent frequency and time response plots are shown in Figures 2 and 3 (as implemented on demo

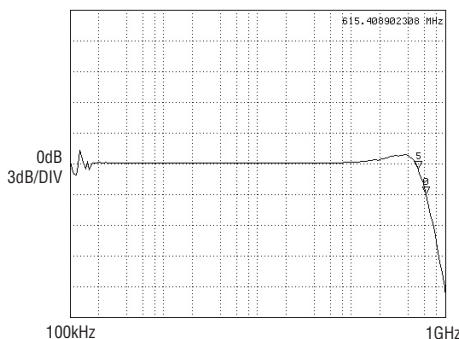


Figure 2. Wide frequency response of circuit in Figure 1

Easy Solution for Multi-Channel Video Applications

Baseband video generated at these higher rates is processed in either native red, green and blue (RGB) domain or encoded into component luma plus blue and red chroma channels (YPbPr); three channels of information in either case. With frequency response requirements extending to beyond 500MHz, amplifier layouts that require external resistors for gain setting tend to be real-estate inefficient, and frequency

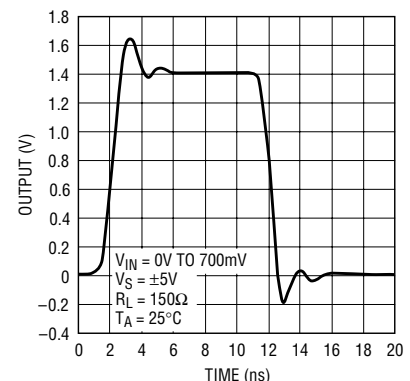


Figure 3. Fast pulse response of circuit in Figure 1

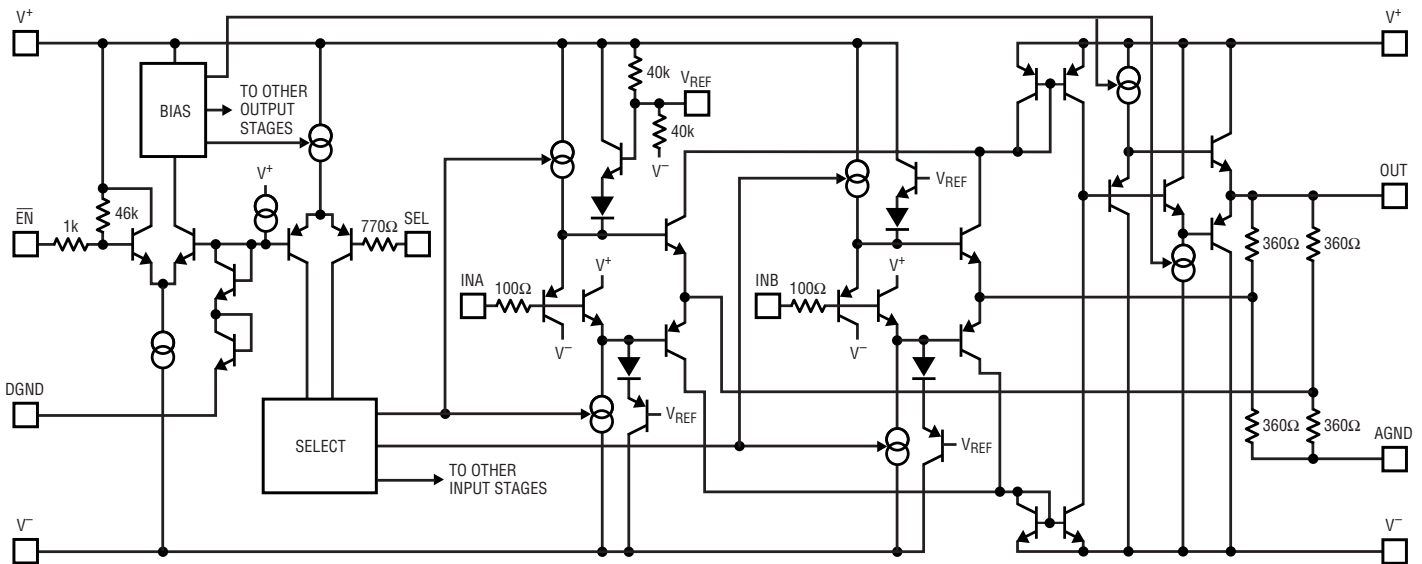


Figure 4. Simplified internal circuit functionality of the LT6555 and LT6556

circuit 892A-A). Frequency markers in Figure 2 show the small-signal -0.5dB response beyond 500MHz and -3dB response above 600MHz . The LT6556, when used to drive high impedances, provides bandwidth to 750MHz , though the LT6556 demo circuit 892A-B uses 75Ω back termination (rather than $1\text{k}\Omega$), resulting in performance similar to the LT6555.

Taking a Look at the Internal Details

The LT6555 and LT6556 integrate three independent sections of circuitry that form classic current-feedback amplifier (CFA) gain blocks, but with switchable input sections, all implemented on a very high-speed fabrication process. The diagram in Figure 4 shows the equivalent internal circuitry (one LT6555 section shown).

Feedback resistors are provided on-chip to set the closed-loop gain to either unity or two, depending on the part. The nominal feedback resistances are chosen to optimize flat frequency response. The LT6555 is intended to drive back-terminated 50Ω or 75Ω cables (for effective loading of 100Ω to 150Ω respectively), while the LT6556 is designed to drive ADCs or other high impedance loads (characterized with $1\text{k}\Omega$ as a reference loading condition).

Common to all three CFAs in each part is a bias control section with a power-down command input. The input select logic steers bias current to the appropriate input circuitry, enabling the input function of the selected signal. The shutdown function includes an internal on-chip pull-up resistance to provide a default disable command, which when invoked, reduces typical power consumption to less than $125\mu\text{A}$ for an entire three-channel part. During shutdown mode the amplifier outputs become high impedance, though in the case of the LT6555, the feedback resistor string to AGND is still present. The parts come into full-power operation when the enable input voltage is brought within 1.3V above the DGND pin. The typical on-state supply current of about 9mA per amplifier provides for ample cable-drive capacity ($>40\text{mA}$) and ultra-fast 2.2V per nanosecond slew rate performance.

Expanding MUX Input Selection

The power-down feature of the LT6555 and LT6556 may be used to control multiple ICs in a configuration that provides additional input selections. Figure 5 shows a simple 4-input RGB selecting cable driver using two LT6555 devices with the enable pins driven by complementary logic signals. The

shared-output connections between the devices need to be kept as short as possible to minimize printed-circuit parasitics that might affect frequency response. This circuit would be ideal in an A/V control-unit for driving the component-video output, for example. The same basic expansion concept applied to an LT6556 pair would be ideal at the input section of a four-source HD video display.

Operating with the Right Power Supplies

The LT6555 and LT6556 require a total power supply of at least 4.5V , but depending on the input and output swings required, may need more to avoid clipping the signal. The LT6556, having unity gain, makes the analysis simple—the maximum output swing is $(V^+ - V^- - 2.6)\text{V}_{\text{P-P}}$ and governed only by the output saturation voltages. This means a total supply of 5V is adequate for standard video ($1\text{V}_{\text{P-P}}$). For the LT6555, extra allowance is required for load-driving, so the output swing is $(V^+ - V^- - 3.8)\text{V}$. This means a total supply of about 6V is required for the output to swing $2\text{V}_{\text{P-P}}$, as when driving cables. For best dynamic range along with reasonable power consumption, a good choice of supplies would be $\pm 3\text{V}$ for the LT6556 and $+5\text{V}/-3\text{V}$ for the LT6555.

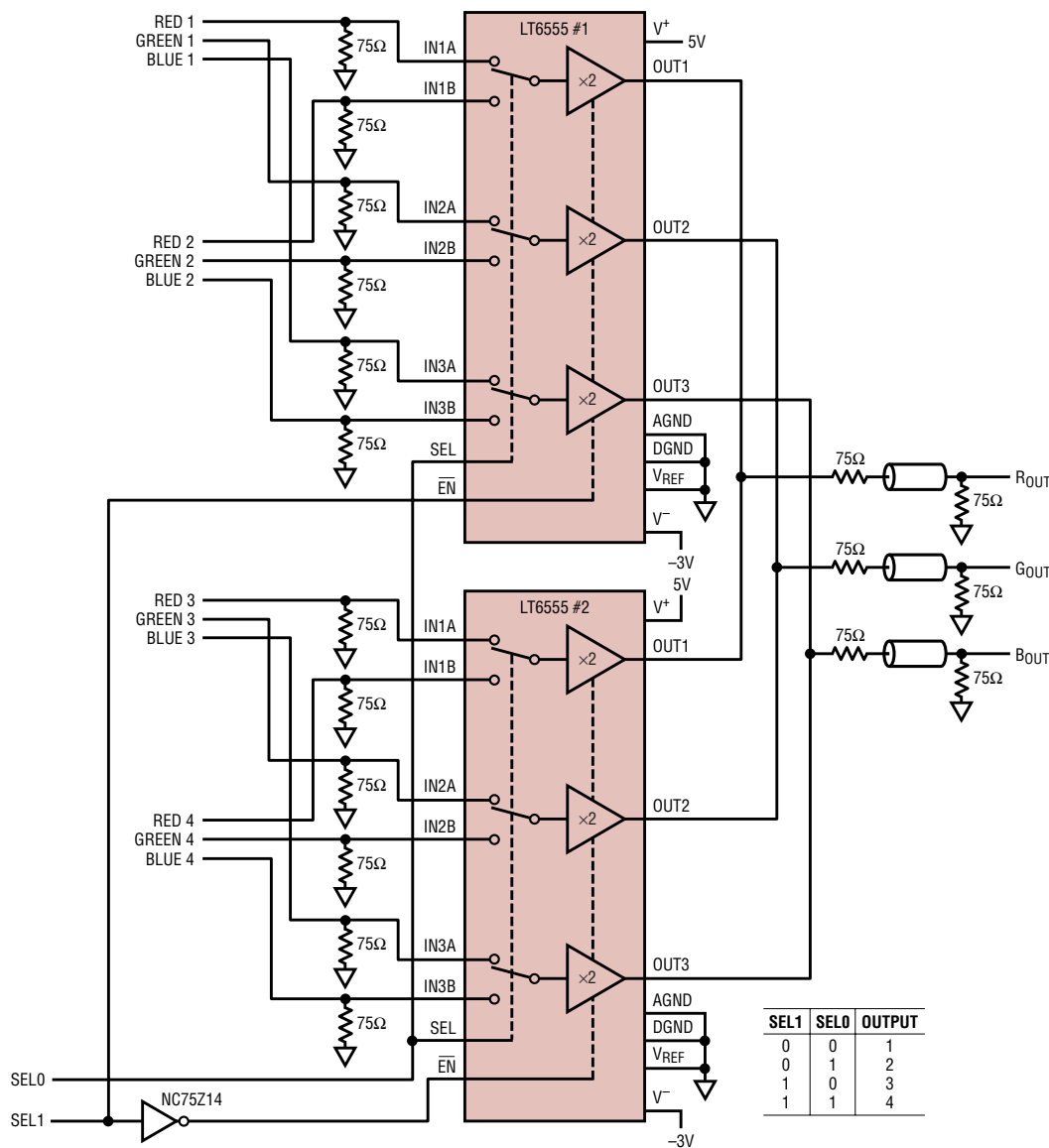


Figure 5. A 4-to-1 video multiplexer using the shutdown feature for expansion

Since many systems today lack a negative supply rail, a small LTC1983-3 solution can be used to generate a simple -3V rail for local use, as shown in Figure 6. The LTC1983-3 solution is more cost effective and performs

at high frequencies better than AC-coupling and resistor network biasing techniques that might otherwise be employed. For example, Figure 7 shows the typical AC-coupling networks used when operating from a

single supply. With six input networks and three large output capacitors required, the AC-coupled method uses more board space and adds parasitics to the signal path that can degrade frequency response.

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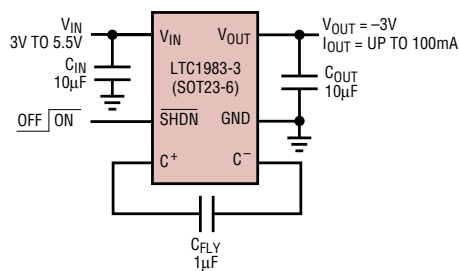


Figure 6. Generating a local -3V supply with four tiny components

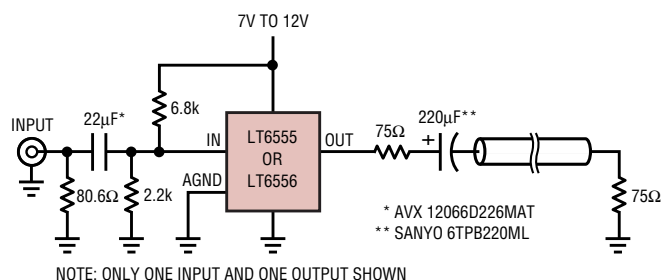


Figure 7. AC-coupling techniques for single-supply operation

High Efficiency, Monolithic Synchronous Buck-Boost LED Driver Drives up to 1A Continuous Current

Introduction

The LTC3454 is a synchronous buck-boost DC/DC converter, designed for driving a single high power LED with regulated currents up to 1A from a single Li-Ion battery. Switching converters are typically used to regulate a voltage, but LEDs require constant current to generate predictable light output. The LTC3454 uses an autozero transconductance error amplifier in its regulation loop to accurately control LED current. The LED current can be set to one of four levels, including shutdown, using two external resistors and dual enable pins. In shutdown no current is drawn.

The wide V_{IN} range of a Lithium-Ion battery (2.7V to 4.2V) requires that a converter be able to both step-up and step-down the input voltage when the LED forward voltage is within the range of the battery discharge profile. The LTC3454 LED driver efficiently performs step-up and step-down conversion via four internal switches. The regulator operates in synchronous buck, synchronous boost or buck-boost mode, depending on V_{IN} and the LED forward voltage. Transitions between modes are automatic and smooth.

The LTC3454 operates at a high fixed frequency of 1MHz, which re-

duces inductor size and eases output filtering.

Application

Figure 1 shows the LTC3454 driving a high power LED in torch and flash modes. Only six external components are required in this application. Efficiency, P_{LED}/P_{IN} , greater than 90% is possible over the entire usable range of a Li-Ion battery (see Figure 2).

by Aspiyan Gazder

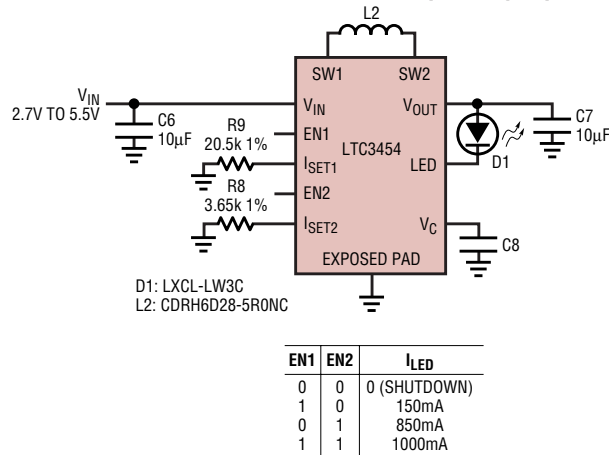


Figure 1. LTC3454 used in a typical application

The LTC3454 has two enable pins that control two current setting amplifiers. A resistor connected from an I_{SET} pin to GND programs the LED current to:

$$I_{LED} = 3850 \cdot \frac{0.8}{R_{ISET}}$$

when the current setting amplifier is enabled via its EN pin. When both enable pins are asserted, the net LED

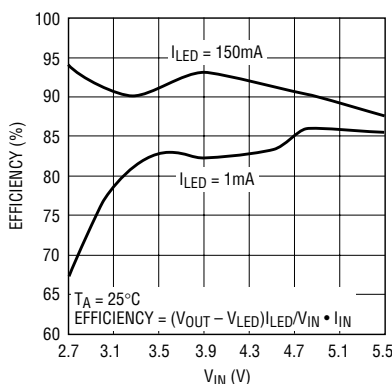


Figure 2. Efficiency for circuit of Figure 1

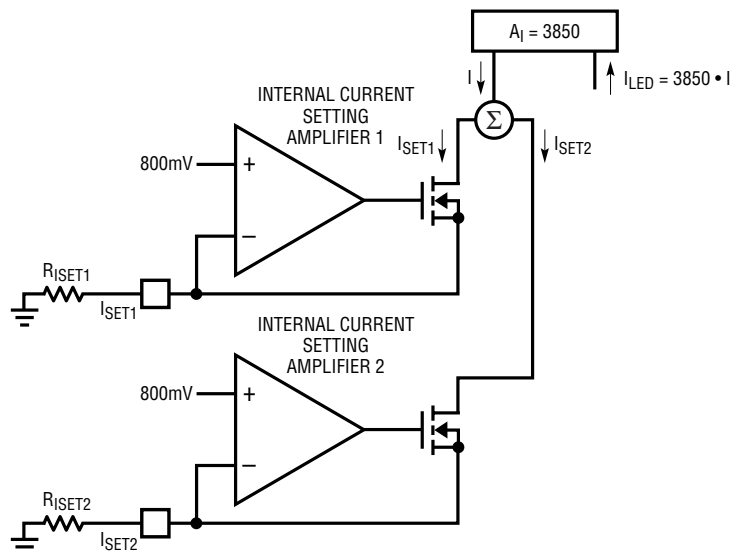


Figure 3. Two current setting amplifiers give the user the flexibility to choose more than one non-zero current level.

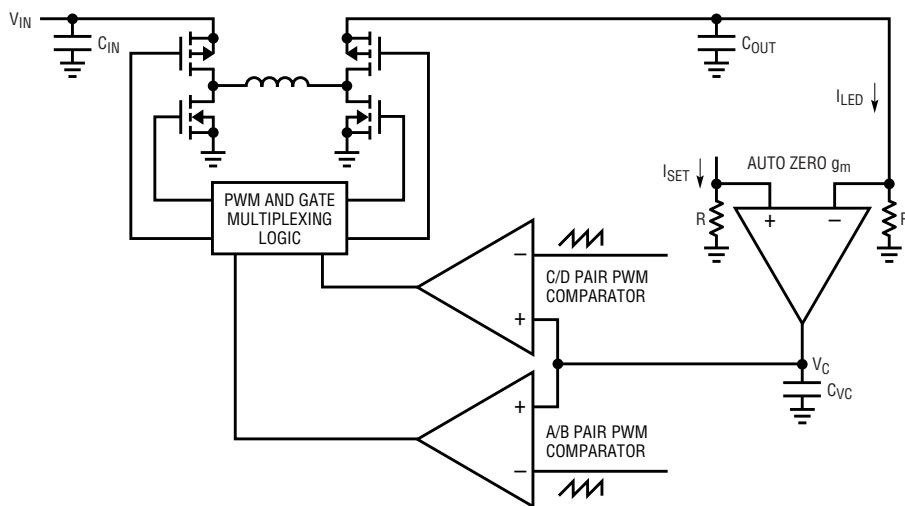


Figure 4. An auto-zeroing transconductance amplifier maintains loop regulation.

current is the sum of each individually programmed current. Figure 3 shows schematically how the LED current is programmed.

Autozeroing Transconductance-Amplifier-Based Current Regulation

The LTC3454 employs an auto-zeroing transconductance amplifier in its regulation loop, as shown in Figure 4. The autozero amplifier topology nullifies any offset at its input, allowing an accurate LED current to be achieved with very low common mode input voltage levels, resulting in high P_{LED}/P_{IN} efficiency. The regulation voltage present at the LED pin can be as low as 100mV at 100mA of LED current.

Synchronous Buck-Boost DC/DC Converter

The LTC3454 can drive an LED at up to 1A continuous current. LEDs that can be driven with such high current typically have forward voltage drops of 3.3V – 3.6V. When powered from a single Li-Ion battery (2.7V to 4.2V), as in the case of handheld battery powered applications, neither a pure buck nor a pure boost solution can efficiently regulate the LED current. A pure buck would dropout at lower battery voltages, causing a lower than programmed LED current to flow. At high battery voltages, a pure boost converter would regulate a higher output voltage than necessary, result-

ing in low efficiency. The buck-boost converter can efficiently regulate LED current over the entire Li-Ion battery range.

The autozero amplifier topology nullifies any offset at its input, allowing an accurate LED current to be achieved with very low common mode input voltage levels, resulting in high P_{LED}/P_{IN} efficiency.

The control voltage, V_C , determines the region of operation of the buck-boost converter. The gate drives of the internal power switches A, B, C and D are controlled by the logic block (Figure 4). A patented gate drive multiplexing scheme enables smooth

transition between buck and boost modes and through the four-switch region. In buck mode, the duty cycles on gate drives of switches A and B are controlled while switch D is turned on continuously. In boost mode, duty cycles of switches C and D are controlled, while switch A is on continuously.

Using synchronous rectifier switches B and D instead of catch diodes helps improve efficiency. This scheme requires that the synchronous rectifier switch and the main switch are not turned on simultaneously. A cross conduction delay prevents this condition from occurring. The LTC3454 has a break before make time of approximately 30ns. During this time the current conduction path is completed through the body diodes of the switches. In the case of forward current flow from the SW1 pin to the SW2 pin through the inductor, the body diode of NMOS switch B conducts in buck mode. The SW1 node is pulled a diode drop below ground. Likewise, in boost mode, the body diode of PMOS switch D conducts during the switch C and switch D switching, but node SW2 now flies above V_{OUT} by a diode drop. Body diodes of the main switches A and C conduct during reverse current flow. Figure 5 shows the switch waveforms in the buck-boost mode.

The LTC3454 has both forward and reverse current limiting—requiring no external sense resistors. If the peak input current exceeds approximately 3.4A, forward current limit is tripped and switches B and D are turned on for the rest of the cycle. The reverse current limit is tripped when current flowing from switch D through the inductor to the SW1 node exceeds approximately 250mA and switches A and C are turned on for the rest of the cycle.

Robust Design: Can Tolerate Open and Shorted LED Conditions

If the LED faults as an open circuit, the regulation loop drives V_C higher, which has the effect of raising the output voltage. A safety amplifier—a

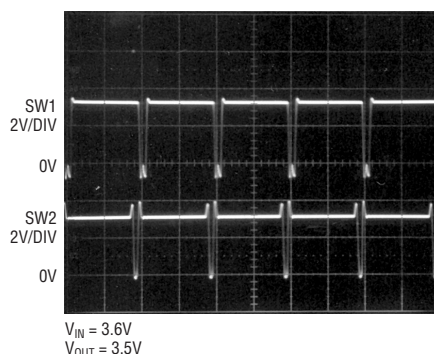


Figure 5. Switching waveforms in buck-boost mode

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Constant Current from 3A DC/DC Converter with 2 Rail-to-Rail Current Sense Amplifiers

by Daniel Chen

Introduction

Traditional DC/DC converters use voltage feedback for constant output voltage regulation. There are many applications, however, that need to regulate a constant output current. Driving LEDs in series is one such application. The LT3477 combines a traditional voltage feedback loop and two unique current feedback loops to operate as a constant-current, constant-voltage source. It is a current mode, 3A DC/DC converter with dual rail-to-rail 100mV current sense amplifiers that can be configured as a buck mode or buck-boost mode LED driver. It is versatile enough to also be configured as an input-output current limited boost, SEPIC or inverting converter. Both current sense voltages can be adjusted independently using the I_{ADJ1} and I_{ADJ2} pins.

With two identical precision current sense amplifiers, the LT3477 can provide an accurate input current limit as well as an accurately regulated output

current. With an input voltage range of 2.5V to 25V, the LT3477 works from a variety of input sources. The 42V switch rating allows an output voltage of up to 41V to be generated, easily

The unique feature of the three-feedback-loop topology (two current and one voltage) is that it can support constant voltage and/or constant current applications.

driving up to ten white LEDs in series. The buck mode LED configuration is capable of driving multiple ten-LED strings in parallel if external current mirroring circuitry is added.

The switching frequency is adjustable from 200kHz to 3.5MHz, set by

a single resistor. The available high operating frequencies allow the use of low profile inductors and capacitors—important in applications where space is a premium. The wide available range makes it possible to optimize size and efficiency for your application.

How It Works

Figure 1 shows a block diagram of the LT3477. The voltage error amplifier has both FBP and FBN pins to allow a positive or negative output configuration. With the addition of two current feedback control loops, amplifier A3 becomes a summing point for three feedback loops. Depending on configuration, any of the loops can take over feedback control by sourcing or sinking current at the V_C node. The unique feature of the three-feedback-loop topology (two current and one voltage) is that it can support constant voltage and/or constant current applications.

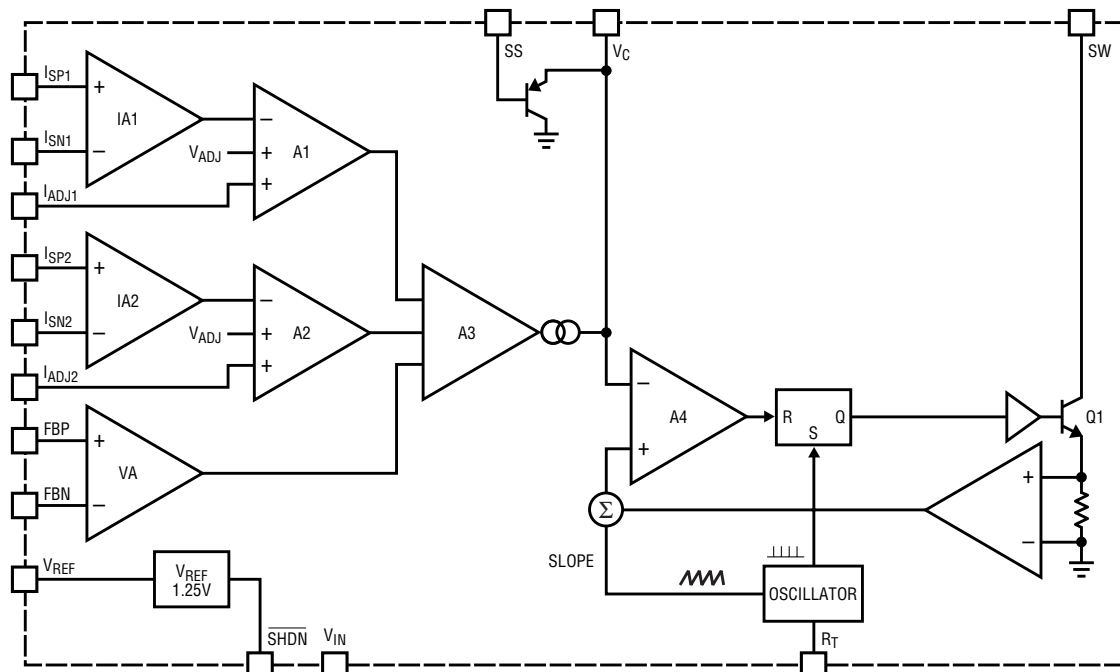


Figure 1. LT3477 block diagram

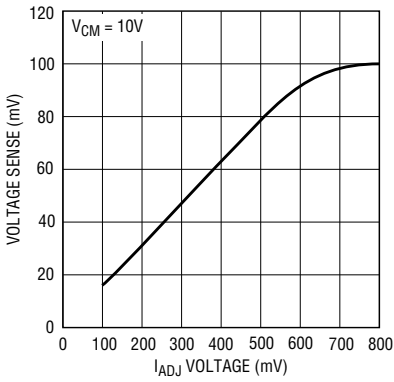


Figure 2. Current sense amplifier voltage sense level vs I_{ADJ} pin voltage

Current sense levels are adjustable via sense resistors at the I_{ADJ1} and I_{ADJ2} pins. The default sense voltage is 100mV for each current sense amplifier if the I_{ADJ1} and I_{ADJ2} pins are tied to a potential higher than 650mV. If the potentials at the I_{ADJ1} and I_{ADJ2} pins are lower than 625mV, the LT3477 linearly adjusts the current sense level. Figure 2 shows the voltage sense level vs the I_{ADJ} pin voltage. For LED drivers, I_{ADJ1} and I_{ADJ2} pins can be used to adjust LED current levels. Rail-to-rail current sense amplifiers allow flexible current sense schemes.

Applications

Buck Mode High Current LED Driver

Figure 3 shows a typical application to drive high current LEDs. Traditionally,

LED drivers use a grounded current sense resistor to regulate current, but the LT3477 current sense amplifiers work in a high side sense scheme, so the sensed voltage for current feedback no longer needs to be ground referred. In buck mode configuration, the sense resistor is placed right at the input supply. The LEDs are placed between the sense resistor and the inductor and the

Schottky diode is connected between the SW and PV_{IN} nodes. With high side current sense, the boost converter is effectively converted into a buck LED converter, which increases the part's power handling capability. In addition, the V_{IN} pin, which provides the chip operating current, can be tied to a lower voltage level such as 3.3V. As a result, the power consumption on the chip itself is also reduced, thus improving overall efficiency. Over 90% efficiency can be readily achieved with a wide range of inductor and frequency selections.

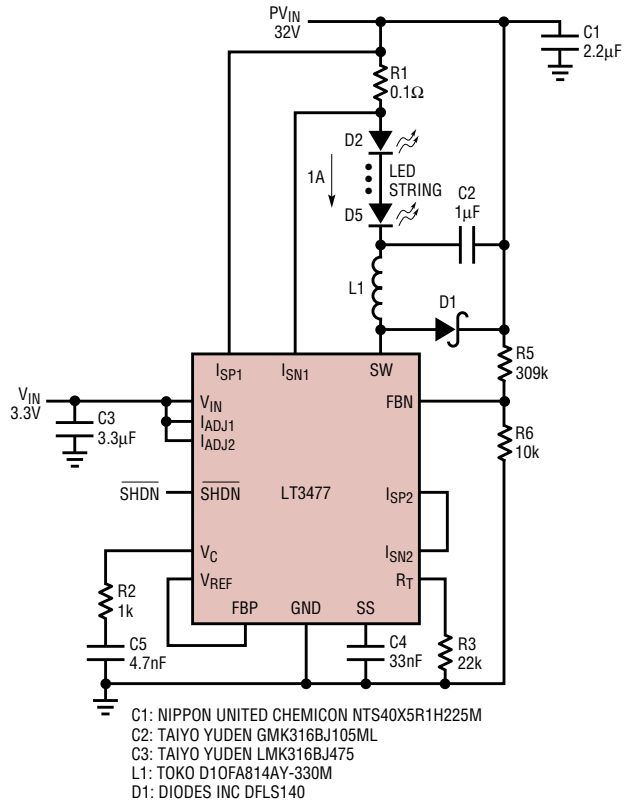


Figure 3. Buck mode high current LED driver

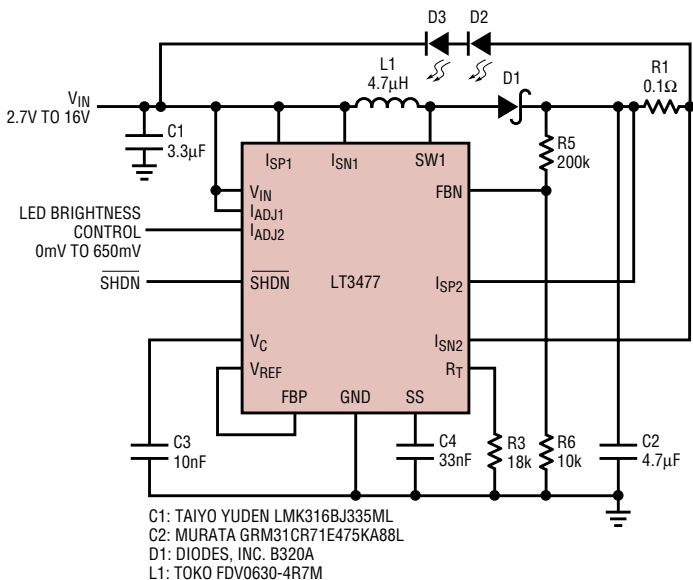


Figure 4. Buck-boost LED driver

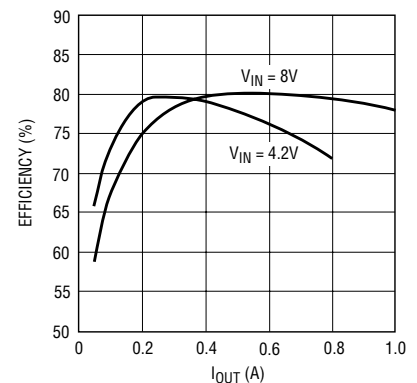


Figure 5. Buck-boost LED driver efficiency

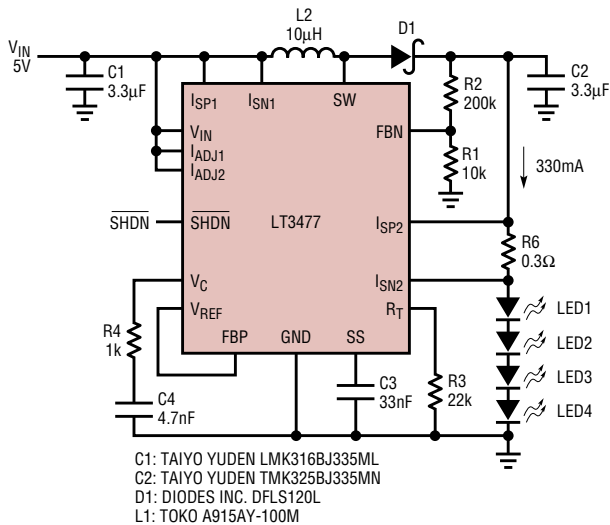


Figure 6. 4W LED driver

Buck-Boost LED Driver

In some applications, the input voltage might be comparable to the total LED voltage drop or the input voltage might fluctuate to higher or lower than the total LED voltage drop. A buck-boost LED driver works well in this type of application. Figure 4 shows the LT3477 buck-boost LED driver. The cathode end of the LED string is tied back to the input voltage, which allows it to operate from a wide input voltage range. R5 and R6 in Figure 4 are used for open LED protection. Figure 5 is the efficiency measured for this circuit.

330mA LED Driver with Open LED Protection

LT3477 can also be used for LED driver applications using a conventional boost topology with the current sense amplifier for current regulation. Figure 6 shows a typical application circuit, and Figure 7 shows the efficiency. Figure 6 uses a high side current sense configuration for feedback control. The current sense amplifier could also be used for a grounded current sense for this application, if desired, so the output can be tied to the LED string directly. Isp2 would be tied to the cathode side of the LEDs, and Isn2 is tied to ground.

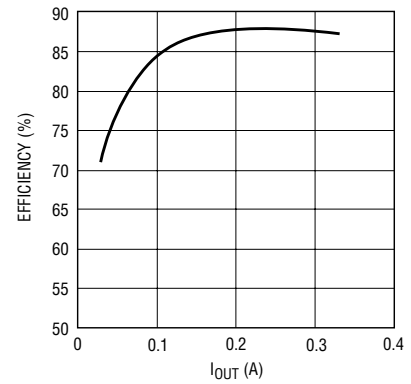


Figure 7. 4W LED driver efficiency

Voltage feedback is used for open LED protection.

5.5V SEPIC Converter with Short-Circuit Protection

Certain applications demand a converter output that is DC-isolated from the input. SEPICs (single-ended primary inductance converters) provide the solution. Figure 8 is an implementation which provides a 5.5V output with complete short-circuit protection. The current sense amplifier used for current sense not only provides excellent short-circuit protection, but also helps soft start the output. The accurate output current limit ensures the maximum current is set at 670mA. When the load demands more, the output voltage will droop while the 670mA output current is maintained. Efficiency is shown in Figure 9.

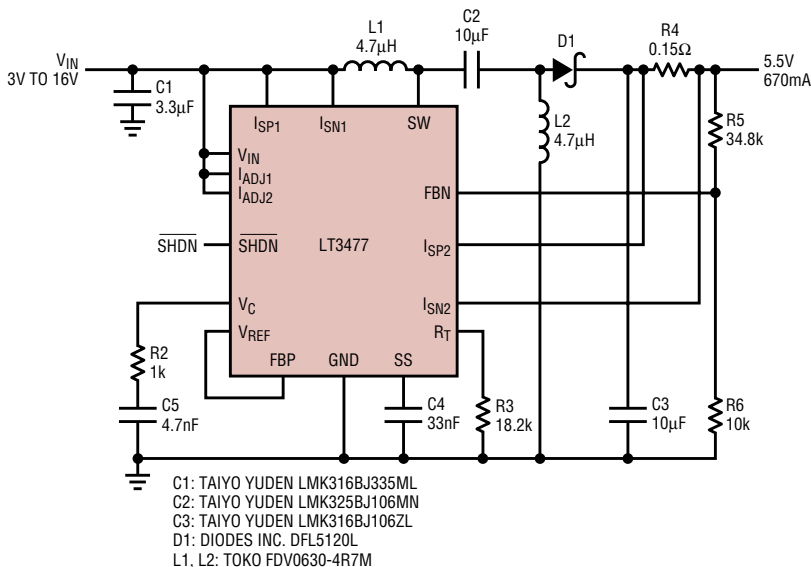


Figure 8. 5.5V SEPIC converter with short-circuit protection

Cuk Converter

The LT3477 provides pins for both inputs to the voltage error amplifier, which enables negative output voltages. Figure 10 is an implementation

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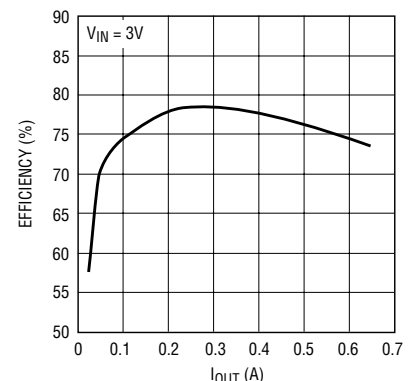


Figure 9. 5.5V SEPIC converter with short-circuit protection efficiency

4-Channel I²C Multiplexer Provides Address Expansion, Bus Buffering and Fault Management

by John Ziegler

Introduction

As data processing, mass storage and communications systems have grown, the size and complexity of the subsystems employed to transfer information such as temperature, fan speed, system voltages and Vital Product Data (VPD, board identification, for example) have grown in proportion. This information is most often transferred through two-wire serial buses, such as I²C or SMBus.

Several practical problems can arise in the design of these systems, especially as they become large. First, many devices, such as Small Form Factor Pluggable optical modules (SFPs) have hard-wired I²C addresses, preventing the use of multiple such devices due to address conflict. Second, as the variety of devices increases and more I/O cards are hot-swapped into and out of a system, the likelihood of an I²C device becoming confused and holding the bus low increases. Third, bus timing specifications become difficult to meet with increasing equivalent bus capacitance. In addition to these large system issues, cycling power whenever a new I/O card is installed is not an option in uninterruptible systems of any size.

The LTC4306 4-channel 2-wire bus multiplexer/switch with bus buffers addresses all of these issues (see Table 1 for a short list of features). A master on the upstream 2-wire bus (SDAIN, SCLIN) can connect to any combination of downstream buses through the LTC4306's bus buffers and multiplexers/switches. As a result, the same device address can be used on multiple downstream buses. The buffers provide capacitive isolation between the upstream and downstream buses, allowing for partitioning of the system loading. Rise time accelerators further aid in overcoming capacitance limitations. Stuck Low Timeout circuitry

Table 1. Some features of the LTC4306

Feature	Benefits
4 Selectable Downstream Buses	<input type="checkbox"/> Maximum flexibility of bus configurations <input type="checkbox"/> Nested addressing when used a MUX
Disconnect from Stuck Bus	<input type="checkbox"/> Frees masters to resume upstream communications
2-Wire Bus Buffers	<input type="checkbox"/> Breaks up capacitance
Buffer Supply Independence	<input type="checkbox"/> Level-shifting: 2-Wire buses can be pulled up to supply voltages ranging from 2.2V to 5.5V, independent of the LTC4306 V _{CC} voltage
Slew Limited Rise Time Accelerators	<input type="checkbox"/> Aid in reducing rise time <input type="checkbox"/> Allow larger bus pull-up resistors for better noise margin <input type="checkbox"/> Drive long cables with no reflection issues
2-Wire Bus Hot Swap	<input type="checkbox"/> Prevents 2-wire bus corruption during live insertion and removal from backplane
Fault Reporting	<input type="checkbox"/> Helps master find and resolve system faults efficiently
Mass Write Address	<input type="checkbox"/> Issue one command to all LTC4306s at the same time

disconnects the upstream bus from the downstream buses when the bus is low for a programmed length of time, freeing the upstream bus to resume communications. Finally, any of the LTC4306's 2-wire bus pins can be hot-swapped into and out of a live system without corrupting it. The LTC4306 works with supply voltages ranging from 2.7V to 5.5V.

General Operation

A block diagram for the LTC4306 is shown in Figure 1, and a description of its register contents is given in Table 2. The UVLO comparator prevents the LTC4306 from receiving commands until the V_{CC} voltage rises above 2.5V (typical). This ensures that the LTC4306 does not try to function until it has sufficient bias voltage. When ENABLE is brought below 1V, the LTC4306 is reset to its default

high-impedance state and ignores any attempts at communication on its 2-wire buses. When ENABLE is brought back above 1.1V, masters may resume communication with the LTC4306.

Disconnecting from a Stuck Bus

The LTC4306 disconnects the upstream bus from the downstream buses when the 2-wire bus is stuck low for a programmed period of time. Masters are then free to resume communications on the upstream bus, assuming the source of the problem resides on a downstream bus. The Stuck Low Timeout circuitry monitors the two common internal nodes of the downstream SDA and SCL switches and runs a timer whenever either of the internal node voltages is below 0.52V. The timer is reset whenever both internal voltages are above 0.6V.

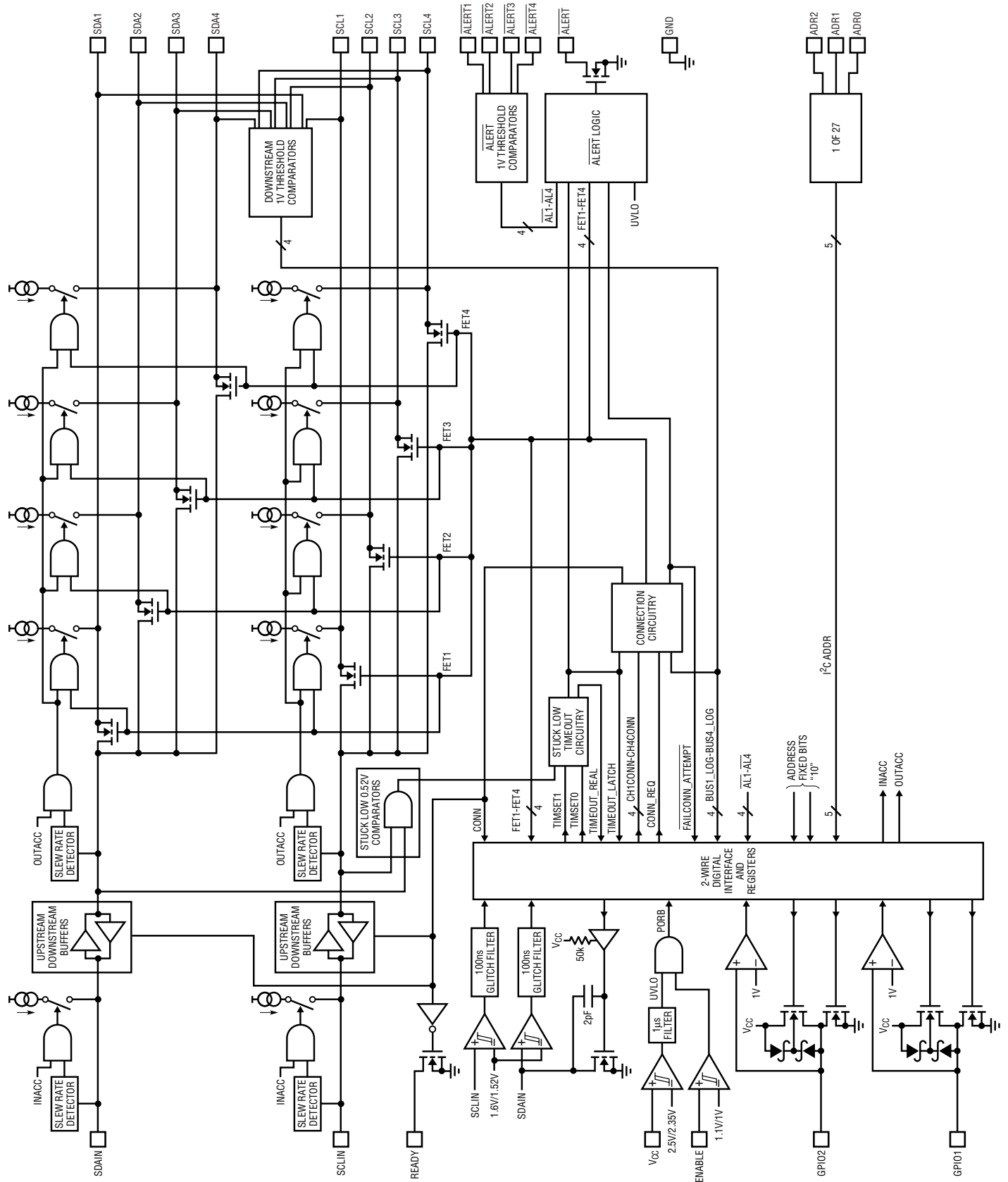


Figure 1. A block diagram of the LTC4306

Table 2. LTC4306 Register Contents

Register	Contents
0	Gives logic state of ALERT1#–ALERT4# pins, and present and latched states of Stuck Low Timer. Indicates whether upstream bus is connected to any downstream buses and whether any failed attempts at connection occurred.
1	Activates/deactivates upstream and downstream rise time accelerators. Writes and reads logic states of GPIO pins.
2	Configures behavior mode of GPIOs. Enables/disables Mass Write feature. Programs Stuck Low Time. Sets requirements on downstream bus logic states for connection to upstream bus.
3	Connects upstream bus to any combination of 4 downstream buses. Masters can read logic state of the downstream buses before connecting to them.

cables. In addition, given the strong drive provided by the accelerators, system designers can choose large resistor pull-ups to minimize bus logic low voltages, thereby maximizing logic low noise margin.

Fault Information Aids Diagnosis

After a fault occurs and the LTC4306 pulls the ALERT# pin low, the LTC4306 works with the master to resolve the fault simply and quickly. The LTC4306 stores specific fault information in read-only register 0. Faults stored include a stuck low bus, faults on the downstream buses, and a failed attempt to connect to a downstream channel.

If the source of the problem is on a connected downstream bus, the master can communicate directly with the offending device. In this case, the LTC4306 acts transparently, with the master and offending device communicating directly via the LTC4306's bus buffers.

In all other cases, the LTC4306 communicates with the master on the upstream 2-wire bus to resolve the fault. After the master broadcasts the Alert Response Address (ARA), the LTC4306 responds with its address on SDAIN and releases ALERT#. The LTC4306 also releases ALERT# if it is addressed by the master. The master determines the source of the fault by reading register 0. After the master solves the problem, it writes a dummy byte to register 0 (which is a read-only register) to reset the fault detection circuitry.

Using register 2, masters can set times of 7.5ms, 15ms, or 30ms, or they can choose to disable the timeout feature.

2-Wire Bus Buffers and Multiplexer Switches Provide Capacitance Buffering and Level Shifting

Masters write to register 3 to connect to any combination of downstream channels. The 2-Wire Bus Buffers provide capacitive isolation between the upstream SDAIN, SCLIN bus and the downstream buses. Thanks to this feature, masters can include LTC4306s at various points in their system to break one large bus into several smaller buses. When any downstream bus is connected, the LTC4306 allows the READY pin to be pulled to a logic high by an external resistor.

By default, the LTC4306 only connects to downstream buses that are high. Attempts to connect to a low downstream bus fail and cause the LTC4306 to pull the ALERT# pin low to indicate a fault. Masters can override this feature by writing to register 2 and instructing the LTC4306 to execute connection commands regardless of the downstream logic state.

The upstream and downstream bus pull-up supply voltages can range from 2.2V to 5.5V, independent of the LTC4306 V_{CC} voltage—the LTC4306 therefore provides level-shifting between buses having different pull-

up voltages. To guarantee proper operation when connecting multiple downstream channels at once, make sure that the LTC4306 V_{CC} voltage is less than or equal to all downstream pull-up voltages to maintain channel-to-channel isolation during logic highs.

Rise Time Accelerators Reduce Rise Times

By writing to Register 2, masters may activate the rise time accelerators on the upstream bus, downstream bus, neither or both. When activated, the accelerators turn on in a controlled manner and source current into the buses to make them rise at a typical rate of 100V/μs during positive bus transitions. These strong pull-up currents allow users to build large, heavily capacitive systems while still meeting rise time specifications, but are also slew limited for driving long

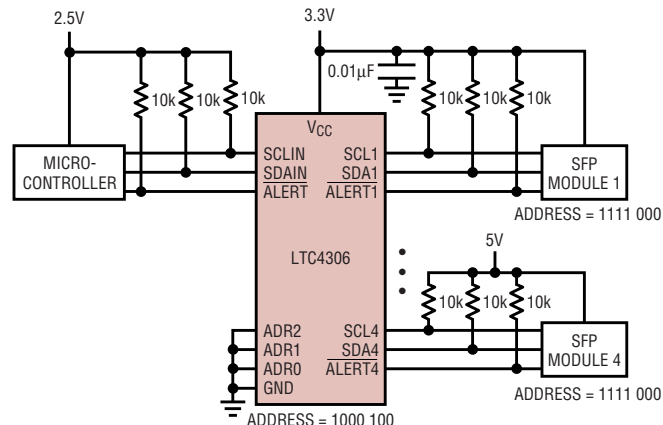


Figure 2. A circuit illustrating the nested addressing and level shifting features of the LTC4306

Nested Addressing and Level-Shifting

The circuit shown in Figure 2 illustrates the nested addressing, level-shifting and capacitance buffering features of the LTC4306. For simplicity, only channels 1 and 4 are shown. Note that the backplane, card 1 and card 4 are pulled up to three different supply voltages. Also, the SFP modules have the same address, but no conflict occurs as long as channels 1 and 4 are never active at the same time.

2-Wire Bus Hot Swapping with the LTC4306 Located on the Backplane

Figure 3 shows a circuit with the LTC4306 located on the backplane and an I/O card plugging into downstream channel 4. Again, channels 2 and 3 are omitted for simplicity. Before plugging and unplugging the card, make sure that channel 4 is not connected to the upstream bus, so that any transaction occurring on the upstream bus is not disturbed. The pull-up resistors on

SDA4 and SCL4 are shown on the backplane, but they may be located on the I/O card, as long as masters on the backplane do not connect to channel 4 when no card is present. The pull-up resistor on ALERT4# must be located on the backplane, to prevent false fault reporting when the I/O card is not present.

2-Wire Bus Hot Swapping with the LTC4306 Located on an I/O Card

In Figure 4 the LTC4306 resides on the edge of an I/O card having four separate downstream buses. Connect a 200kΩ resistor from ENABLE to ground and make ENABLE the short-est pin on the connector. This ensures

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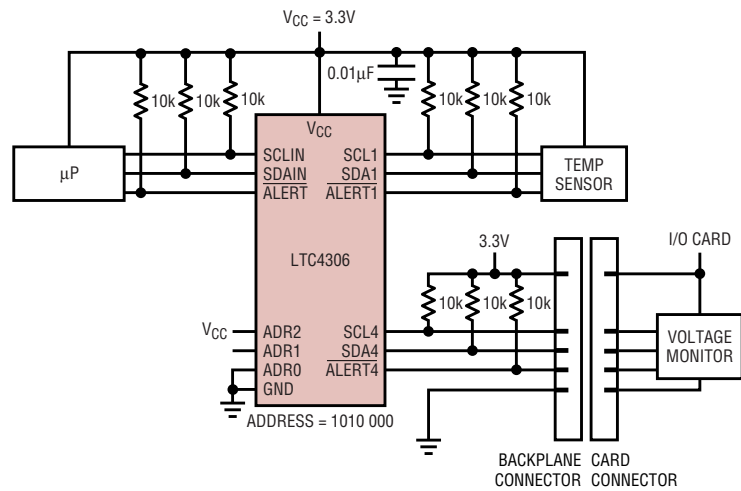


Figure 3. A 2-Wire Bus hot-swapping application circuit with the LTC4306 resident on the backplane

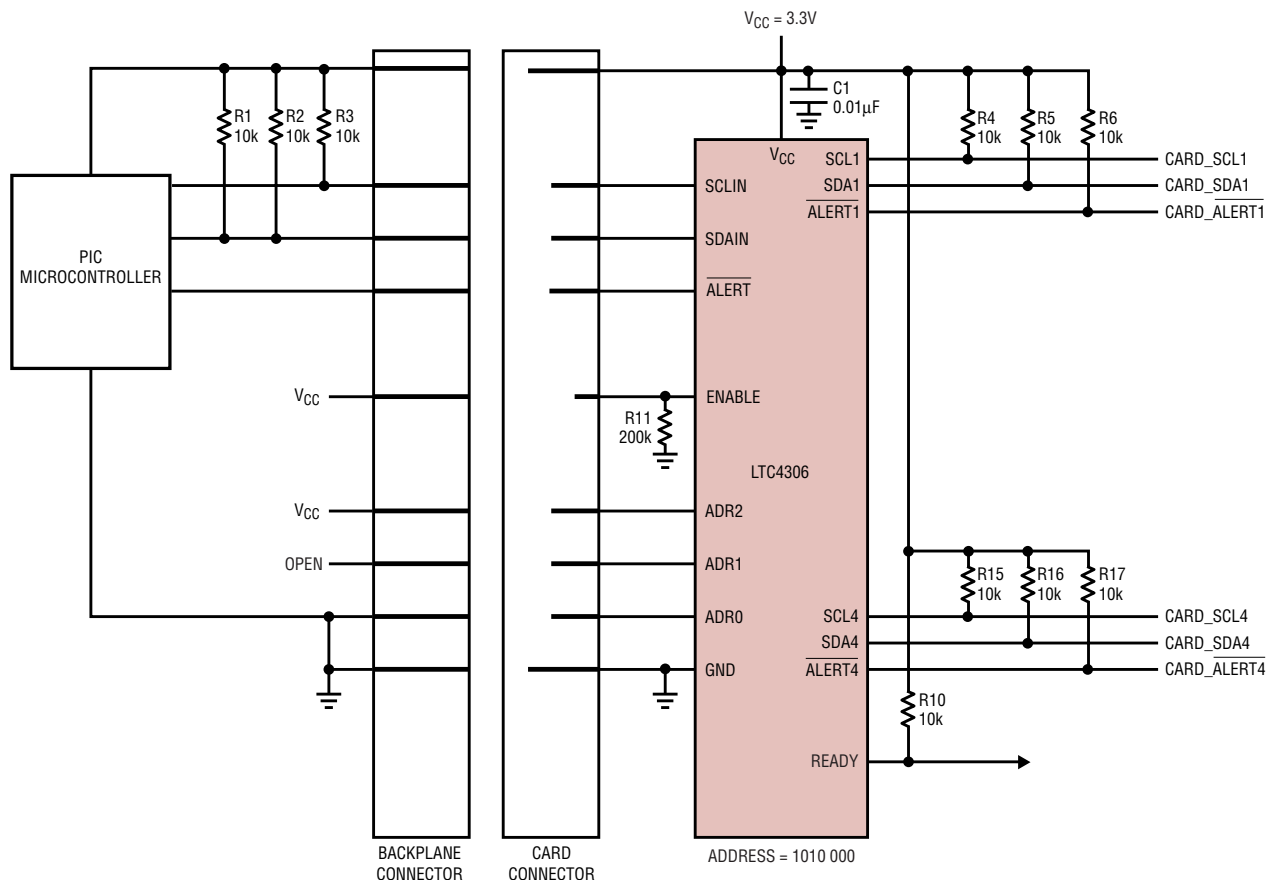


Figure 4. A 2-Wire Bus hot-swapping application circuit with the LTC4306 resident on the I/O card

Lithium Ion Battery Charger Allows Choice of Termination Method and Includes 100mA Adjustable Low Dropout Regulator

by Fran Hoffart

Introduction

Lithium ion batteries, including lithium ion polymer, come relatively close to being the perfect battery: high energy density, lightweight, low self-discharge, high voltage (compared to other cells), no memory problem, low maintenance, and best of all, they are simple to charge. Of course, there are some disadvantages too, but let us leave that for later in this article.

Since many hand held products can operate from a single Li-Ion cell, many single cell chargers use a linear, rather than a switching topology. Linear chargers are simpler than switchers and comparably efficient at the low input-to-output voltage differential typical of portable devices.

This article presents a simple stand-alone 1A battery charger that combines many desirable charger features and an LDO regulator in a tiny 3mm x 3mm low profile DFN package. Also, a brief discussion on lithium ion battery pros and cons, and charging methods are discussed.

Lithium Ion Batteries Are Simple to Charge

There are several recommended methods for charging Li-Ion cells. One method is to apply a current limited constant voltage to the battery for three hours, then stop. Using this method, the battery will be 100% charged after 3 hours, provided the charge current is set between approximately C^1 and $C/2$.

A second, similar method is to apply a current limited constant voltage to the battery while monitoring the charge current. During the first portion of the charge cycle, the charger is in constant current mode, with the battery voltage slowly rising as the

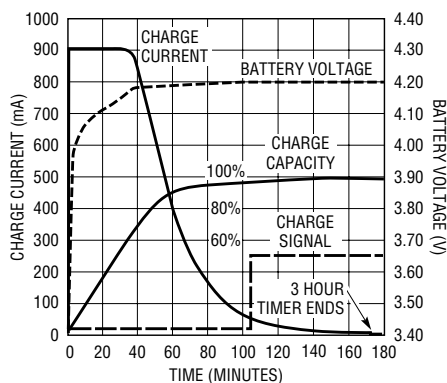


Figure 1. Charge cycle of a 900mAh Li-Ion cell charged at 1C using timer termination

battery accepts charge. As the battery voltage approaches the programmed constant (float) voltage, the charge

current begins to drop exponentially. When the charge current drops to a sufficiently low value, the charger stops charging. Depending on the minimum charge current selected, the battery is between 95% and 100% charged. Since Li-Ion batteries are unable to absorb an overcharge, all charge current must stop when the battery becomes fully charged.

A Charger and an LDO Regulator in One Small DFN Package

The LTC4063 is a complete single cell Li-Ion battery charger that provides the user a choice of charge termination methods and includes an adjustable low dropout 100mA linear regulator. In addition to the usual constant-current/constant-voltage charge algorithm, other desirable features include power limiting that reduces the charge current under high ambient temperature and/or high power dissipation conditions. This allows the charger to provide higher charge currents under normal conditions and still provide safe charging under abnormal conditions such as high ambient temperature, high input voltage or low battery voltage.

The LTC4063 contains many common features of other Li-Ion chargers including trickle charge for low battery, auto recharge, charge current monitor, charge status output, capable of charging from USB power, low battery drain current when V_{IN} is removed and precision ($\pm 0.35\%$) battery float voltage accuracy.

What sets this linear charger apart from other single cell chargers is the selectable charge termination and the onboard voltage regulator.

DESIGN IDEAS

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Termination can be based on either total time, which is programmable, or minimum charge current which is also programmable, or the charge cycle can be stopped by the user via the charge enable pin.

The low dropout regulator, which is powered from the battery, is adjustable from 1V to almost 4.2V and can provide up to 100mA to a load. A low 15μA operating quiescent current and 2.5μA shutdown current extend battery life.

Charge Termination Methods: Which One to Use

The first portion of a charge cycle consists of forcing a constant current (typically 1C) into the battery until the cell voltage approaches the programmed float voltage (typically 4.2V ±1% or better) at which time the charge current begins to drop. For a depleted battery this occurs after approximately 30 minutes with the battery state of charge at approximately 55% of full capacity. Since the charge current drops rather quickly in the constant voltage portion of the charge cycle, the battery requires another 2 hours to bring the battery up to a 100% charge level. Unfortunately, there is not much that can be done to speed up this portion of the charge cycle without exceeding the recommended charge voltage.

Some chargers utilize a Negative Temperature Coefficient (NTC) thermistor that is located near or inside the battery pack to measure battery temperature. This protects the battery by not allowing a charge cycle to begin if the battery temperature is less than 0°C or greater than 50°C. During a normal charge cycle, there is very little temperature rise for Li-Ion batteries.

Figure 1 shows a LTC4063 charge cycle for a 900mAh Li-Ion polymer battery charging at a 1C rate. The curves show the relationship between the charge current, battery voltage, charge capacity and the $\overline{\text{CHRG}}$ output signal. Since the timer termination method was selected, the charge cycle ended after approximately 172 minutes with the battery at 100% charge

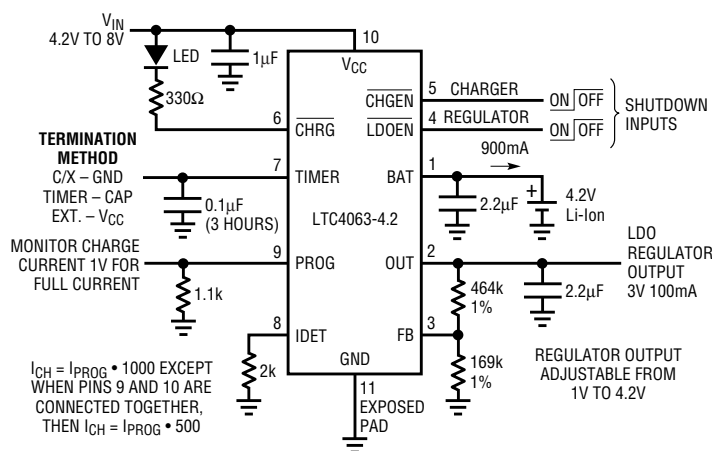


Figure 2. Complete single cell Li-Ion charger with timer termination, 50mA minimum charge current detection and 3V 100mA LDO voltage regulator

level. (Note: the charge current near the end of the charge cycle is a very low 6mA). Also shown in Figure 1 is the $\overline{\text{CHRG}}$ open drain output signal, which was programmed to go high when the charge current dropped below 50mA (I_{DETECT} threshold) or approximately C/20.

Had the minimum charge current termination method been selected rather than the timer method, the charge cycle would have ended when the $\overline{\text{CHRG}}$ signal went high (after 105 minutes). At that point the battery is approximately 97% charged, and it would take another hour of charging for the last 3%. The programmable I_{DETECT} current threshold level of the LTC4063 has excellent accuracy, even at current levels as low as 5mA. Programming a low I_{DETECT} current and selecting minimum current termination would result in the charge cycle ending at approximately the same time as timer termination.

Which termination is better? From the previous paragraph, it appears that it may not make much difference because by selecting a low I_{DETECT} current level, the two methods can be made virtually identical. Minimum

charge current termination can have an advantage in a situation where different charge current levels may need to be selected during a charge cycle, or when charging a battery that still has a partial charge, the charge cycle can be very short. But timer termination may be better if a load that is greater than the programmed I_{DETECT} current level is permanently connected to the battery. In that situation, the charge cycle may never terminate. Also, in timer termination, if the battery does not reach the recharge threshold of 4.1V when the timer ends, the timer is reset and a new charge cycle begins.

A Quick Primer on Rechargeable Li-Ion Batteries

Within the lithium ion family of batteries there are several formulations: mainly lithium cobalt oxide or lithium manganese oxide as the positive electrode, and either coke or graphite as the negative electrode. The electrolyte is a liquid in cylindrical cells or a solid or a gel in Li-Ion polymer cells. Since no liquid is used in the polymer cells, the cell package can consist of an inexpensive lightweight foil pouch

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About Battery Capacity and Charge Current

The correct charge current is always related to a battery's capacity, or simply "C". The letter "C" is a term used to indicate the manufacturers stated battery discharge capacity, which is measured in mAh. For example, a 900mAh rated battery can supply a 900mA load for one hour before the cell is depleted. In the same example, charging the battery at a C/3 rate would mean charging at 300mA.

Low Ripple Micropower SOT-23 Buck Regulator with Integrated Boost and Catch Diodes Accepts Inputs to 40V

by Leonard Shtargot

Introduction

The LT3470 is a micropower buck regulator that integrates a 300mA power switch, catch diode and boost diode into a low profile 8-Pin ThinSOT package (see Figure 1). The combination of single cycle Burst Mode and continuous operation allows the use of tiny inductor and capacitors while providing a low ripple output to loads of up to 200mA. With its wide input range of 4V to 40V and low quiescent current of 26 μ A (12V in to 3.3V out) the LT3470 can regulate a wide variety of power sources, from 2-cell Li-Ion batteries to unregulated wall transformers and lead acid batteries.

5V, 200mA from 40V Consumes Less than 1mW at No Load

Figure 2 shows a 5V, 200mA supply that accepts inputs from 5.5V to 40V. While the output is in regulation and with no load the power loss is lower than 1mW. The LT3470 can also be put in a shutdown mode that reduces the input current to <1 μ A by pulling the $\overline{\text{SHDN}}$ pin low. When always-on operation is desired, the $\overline{\text{SHDN}}$ pin can be tied to V_{IN} .

The LT3470 uses a control system that offers low (<10mV) ripple at the

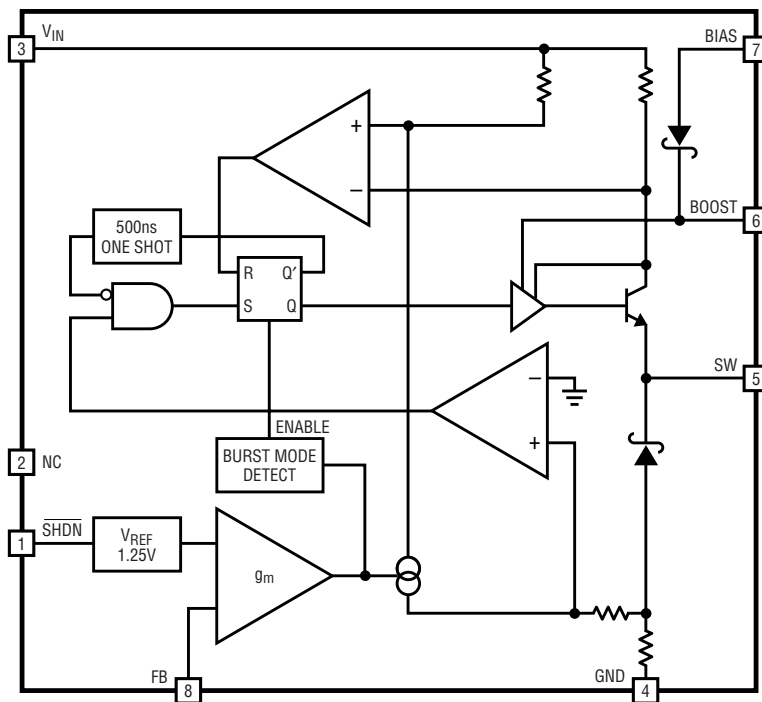


Figure 1. The block diagram of the LT3470 shows the integrated boost and catch Schottky diodes. Inductor current is kept under control at all times by monitoring the V_{IN} current as well as the catch diode current, thereby providing short circuit protection even if $V_{\text{IN}} = 40\text{V}$.

output while keeping quiescent current to a minimum. When output load is light, the LT3470 remains in sleep mode while periodically waking up for single switch cycles to keep the output in regulation. The current limit of these single switch cycles is about

100mA, which keeps output ripple to a minimum. At greater output loads the LT3470 no longer enters sleep mode, and instead serves the peak switch current limit (up to 300mA) to regulate the output. See Figure 3 for operating waveforms.

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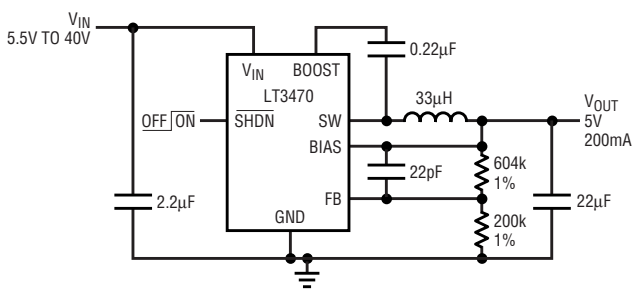
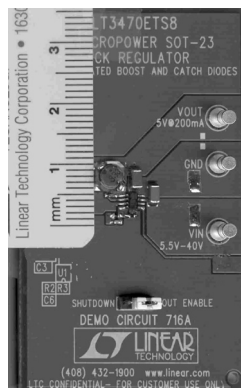
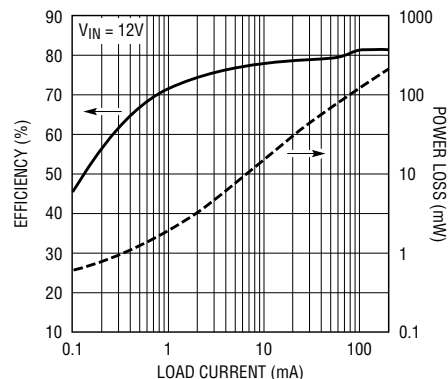


Figure 2. The LT3470 uses a minimum of board space and external components while delivering wide output range and high efficiency. This buck regulator supplies up to 200mA at 5V from inputs up to 40V. Input power loss is below 1mW when there is no output load.



Taking Full Advantage of Very Low Dropout Linear Regulators

by Joe Panganiban

Introduction

Linear regulators are generally considered inefficient step-down DC/DC converters, but low dropout linear regulators (LDOs) can be a good fit in many handheld battery applications where low power and efficient power conversion are critical. The lower the dropout voltage, the more efficient the LDO solution. Generally, LDOs with a very small dropout voltage come at the expense of increased package size and higher quiescent current. The LTC3035 overcomes these tradeoffs by offering a very low dropout voltage without sacrificing small solution size or low power.

The LTC3035 is a micropower, VLDO™ (very low dropout) linear regulator, which operates from input voltages between 1.7V and 5.5V. The device is capable of supplying 300mA of continuous output current with an ultra-low dropout voltage of 45mV typical (see Figure 1). The output voltage is externally adjustable over a wide voltage range, spanning between 0.4V and 3.6V.

The LTC3035 is ideal for battery-powered applications where low power, low dropout, low noise, and small solution size are essential. Under no-load conditions, the chip draws only

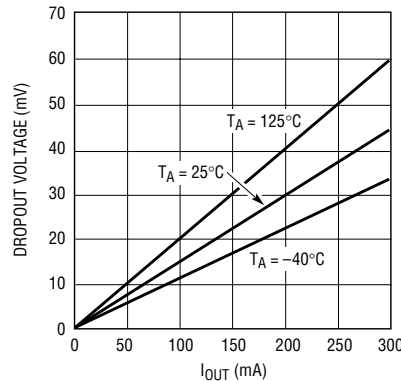


Figure 1. Typical dropout voltage versus load current

100µA from the V_{IN} supply, and drops to 1µA when in shutdown. The LDO is stable for all ceramic capacitors down to 1µF. Other features include output short-circuit protection, reverse output current protection, and thermal overload protection, all available in a tiny 3mm × 2mm DFN package.

Low Dropout from an NMOS Pass Device

Conventional LDOs integrate a P-type transistor (either PNP or PMOS) as the power pass device to deliver current from the input supply to its output. The LTC3035, instead, incorporates an NMOS transistor as its pass element

in a source-follower configuration. This architecture allows for several performance advantages over conventional P-type LDOs, such as greater V_{IN} power supply rejection, lower dropout voltage, and better transient response characteristics, while maintaining a smaller solution size.

Using an NMOS pass device is not entirely transparent. In order to achieve low dropout performance using an NMOS pass device, the LDO circuitry must be capable of driving the NMOS gate above the V_{IN} supply. This implies that a separate higher voltage supply is necessary to power the LDO circuitry. For many applications, the luxury of an extra higher supply is simply unavailable. The LTC3035 overcomes this problem by including a built-in charge pump that generates a higher BIAS supply from the V_{IN} input to power its LDO circuitry. The charge pump requires only a 0.1µF flying capacitor and a 1µF bypass capacitor for operation. The value of the generated BIAS supply is adaptively controlled to provide sufficient gate drive over the full V_{IN} operating range, optimizing the current carrying capabilities and dropout characteristics of the VLDO regulator.

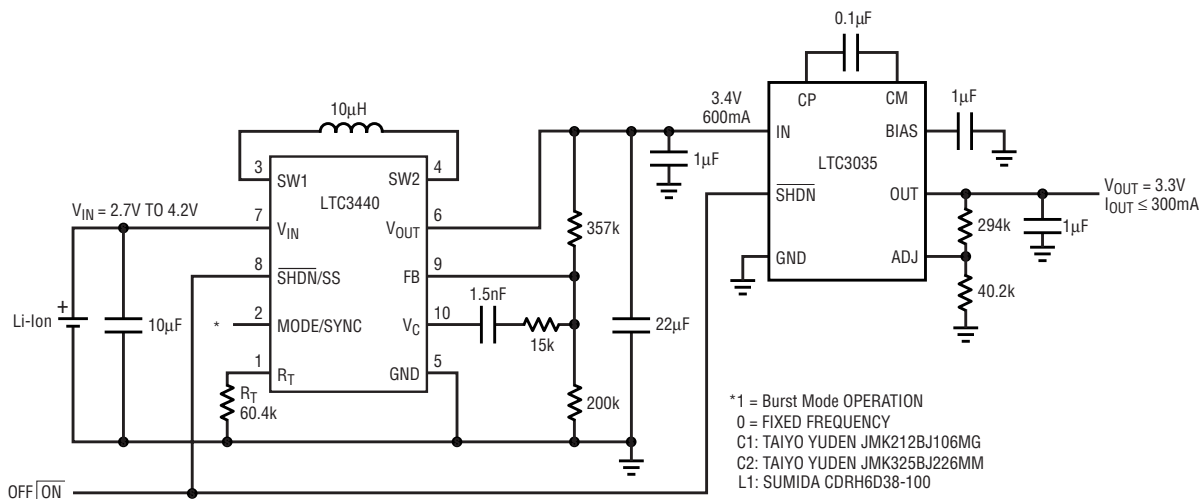


Figure 2. A high-efficiency and low-noise lithium-ion to 3.3V solution

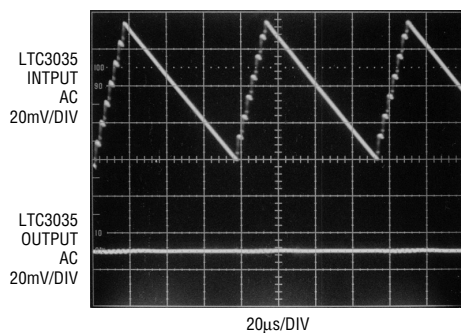


Figure 3. Input and output waveforms to the LTC3035 in the Li-Ion to 3.3V application, showing its excellent ripple rejection ($I_{OUT} = 25mA$, LTC3440 in Burst Mode®)

High Efficiency, Low Noise Li-Ion to 3.3V

Figure 2 shows a high efficiency and low noise lithium-ion to 3.3V solution. The LTC3440, a buck-boost converter, converts the Li-Ion battery voltage to an efficient intermediate voltage (3.4V) at the input of the VLDO. The LTC3035 then regulates this intermediate voltage down to 3.3V, providing a lower noise output voltage. Figure 3 shows the input and output waveforms of the LTC3035 at 25mA of output current, illustrating its excellent power supply rejection characteristics for a lower noise solution.

For optimum total efficiency, the input to output voltage differential across the LDO should be as small as possible, since the magnitude of the dissipated power equals the product of the voltage differential and the output current. Because of the LTC3035's

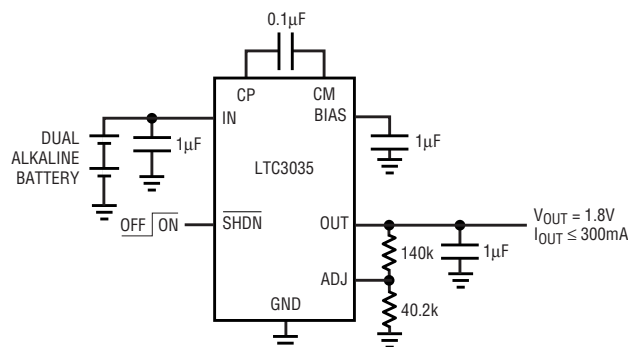


Figure 4. A very low dropout dual-alkaline to 1.8V application

very low dropout voltage, its input voltage can be programmed to only 100mV above the 3.3V output and still maintain regulation at 300mA. Conventional LDOs with higher dropout voltages force greater input and output voltage differentials, effectively reducing efficiency by the same ratio.

Double Alkaline to 1.8V LDO

Handheld applications using two alkaline batteries in series demand low power solutions that use as much of the battery's operating voltage range as possible. In Figure 4, two series alkaline batteries are regulated down to provide a 1.8V supply taking advantage of the LTC3035's excellent dropout characteristics.

The dropout voltage and maximum output current capabilities of typical low power LDOs using P-type transistors suffer as the input voltage supply decreases, since the power transistor's overdrive reduces. With input and

output voltages near 1.8V, conventional low power LDOs may have dropout voltages over 200mV, if they can deliver 300mA of output current at all. Using the LTC3035, the battery voltage can discharge much further to only about 50mV above the 1.8V output before the LDO begins to drop out at 300mA. Allowing the battery to discharge longer essentially extends the battery life for the application when compared to solutions that use higher dropout LDOs.

Conclusion

The very low dropout characteristics of the LTC3035 can be exploited in battery-powered applications to obtain higher efficiency and increased battery life. Its very low dropout voltage, excellent power supply rejection, low-quiescent current, and small solution size make the LTC3035 an ideal choice for many low power, handheld battery applications.

LT3470, continued from page 34

The fast cycle-by-cycle current limit of the LT3470 keeps the switch and inductor currents under control at all times. In addition, the LT3470 uses hysteretic mode control where the switching frequency automatically adjusts to accommodate variations in

V_{IN} and V_{OUT} . This means that the part switches at a slower frequency when the output is in short circuit or when V_{IN}/V_{OUT} ratio is high. This ensures that the LT3470 can handle a short circuit at the output even if $V_{IN} = 40V$ and the inductor value is small. It

is, however, important to choose an inductor that does not saturate excessively at currents below 400mA to guarantee short circuit protection.

Conclusion

The LT3470 is a small buck regulator with a unique combination of features that make it a great choice in applications requiring small size, high efficiency across a wide range of currents, and low output ripple. It can deliver up to 200mA from inputs as high as 40V using only an inductor, four small ceramic capacitors, and two resistors while consuming only 26µA during no load operation.

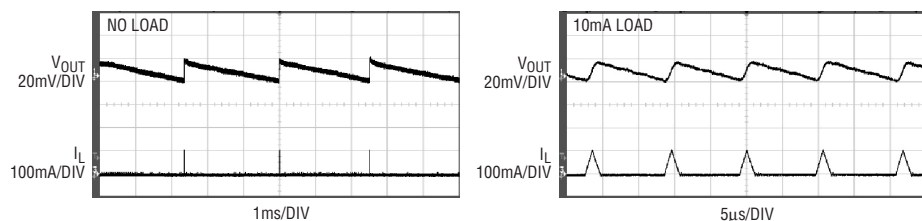


Figure 3. Operating waveforms show the output voltage ripple remains at 10mV in BurstMode operation, while requiring only a 22µF ceramic output capacitor.

Op Amp Selection Guide for Optimum Noise Performance

by Glen Brisebois

Introduction

Linear Technology continues to add to its portfolio of low noise op amps. This is not because the physics of noise has changed, but because low noise specifications are being combined with new features such as rail-to-rail operation, shutdown, low voltage, and low power operation. Op amp noise is dependent on input stage operating current, device type (bipolar or FET) and input circuitry. This selection guide is intended to help you identify basic noise tradeoffs and select the best op amps, new or old, for your application.

Quantifying Resistor Thermal Noise and Op Amp Noise

The key to understanding noise tradeoffs is the fact that resistors have noise. At room temperature, a resistor R has an RMS voltage noise density (or “spot noise”) of $V_R = 0.13\sqrt{R}$ noise in nV/\sqrt{Hz} . So a 10k resistor has $13nV/\sqrt{Hz}$ and a 1M resistor has $130nV/\sqrt{Hz}$. Rigorously speaking, the noise density is given by the equation $V_R = \sqrt{4kTR}$, where k is Boltzman’s constant and T is the temperature in degrees Kelvin. This dependency on temperature explains why some low noise circuits resort to super-cooling the resistors. Note that the same resistor can also be considered to have a noise current of $I_R = \sqrt{4kT/R}$, or a noise power density $P_R = 4kT = 16.6 \cdot 10^{-21} W/Hz = 16.6$ zeptoWatts/Hz independent of R. Selecting the right amplifier is simply finding which one will add the least amount of noise above the resistor noise.

Don’t be alarmed by the strange unit “ \sqrt{Hz} ”. It arises simply because noise power adds with bandwidth (per Hertz), so noise voltage adds with the square root of the bandwidth (per root Hertz). To make use of the specification, simply multiply it by the square

root of the application bandwidth to calculate the resultant RMS noise within that bandwidth. Peak-to-peak noise, as encountered on an oscilloscope for example, will be about 6 times the total RMS noise 99% of the time (assuming Gaussian “bell curve” noise). Do not rely on the op amp to limit the bandwidth. For best noise performance, limit the bandwidth with passive or low noise active filters.

Op amp input noise specifications are usually given in terms of nV/\sqrt{Hz} for noise voltage, and pA/\sqrt{Hz} or fA/\sqrt{Hz} for noise current, and are therefore directly comparable with resistor thermal noise. Due to the fact that noise density varies at low frequencies, most op amps also specify a typical peak-to-peak noise within a “0.1Hz to 10Hz” or “0.01Hz to 1Hz” bandwidth. For the best ultra low frequency performance, you may want to consider an zero drift amplifier like the LTC2050 or LTC2054.

Summing the Noise Sources

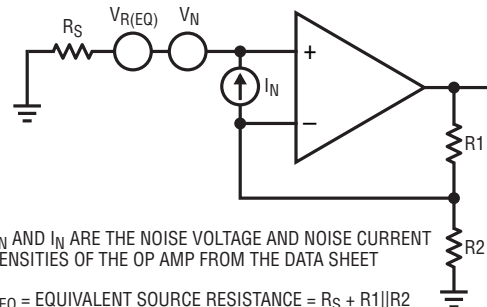
Figure 1 shows an idealized op amp and resistors with the noise sources presented externally. The equation for the input referred RMS sum of all

the noise sources, $V_{N(TOTAL)}$, is also shown. It is this voltage noise density, multiplied by the noise gain of the circuit ($NG = 1 + R1/R2$) that appears at the output.

From the equation for $V_{N(TOTAL)}$ we can draw several conclusions. For the lowest noise, the values of the resistors should be as small as possible, but since R1 is a load on the op amp output, it must not be too small. In some applications, such as transimpedance amplifiers, R1 is the only resistor in the circuit and is usually large. For low R_{EQ} , the op amp voltage noise dominates (as V_N is the remaining term), while for very high R_{EQ} the op amp current noise dominates (as I_N is the coefficient of the highest order R_{EQ} term). At middle values of R_{EQ} , the resistor noise dominates and the op amp contributes little significant noise. This is the $R_{OPTIMUM}$ of the amplifier and can be found by taking the quotient of the op amp’s noise specs: $V_N/I_N = R_{OPT}$.

Selecting the Best Op Amps

Figure 2 shows plots of voltage noise density of the source resistance and of various op amps at three different



V_N AND I_N ARE THE NOISE VOLTAGE AND NOISE CURRENT DENSITIES OF THE OP AMP FROM THE DATA SHEET

$R_{EQ} = \text{EQUIVALENT SOURCE RESISTANCE} = R_S + R1 || R2$

$V_{R(EQ)} = 0.13\sqrt{R_{EQ}}$ IS RESISTOR THERMAL NOISE IN nV/\sqrt{Hz}

EXPRESS V_N , $V_{R(EQ)}$ AND $I_N \cdot R_{EQ}$ IN nV/\sqrt{Hz}

$V_{N(TOTAL)} = \sqrt{V_N^2 + V_{R(EQ)}^2 + (I_N \cdot R_{EQ})^2}$
= THE TOTAL INPUT REFERRED NOISE IN nV/\sqrt{Hz}

Figure 1. The op amp noise model. V_N and I_N are op amp noise sources (correlated current noise is not shown). $V_{R(EQ)}$ is the voltage noise due to the resistors.

frequencies. Each point labelled by an op amp part number is that part's voltage noise density plotted at its R_{OPT} .

Use the graph with the most applicable frequency of interest. Find your source resistance on the horizontal axis, and mark that resistance at the point where it crosses the resistor noise line. This is the "source resistance point." The best noise performance op amps are under that point, the further down the better.

For all candidate op amps, draw a horizontal line from your source

resistance point all the way to the right hand side of the plot. Op amps beneath that line will give good noise performance, again the lower the better. Draw another line from the source resistance point down and to the left at one decade per decade. Op amps below that line are also good candidates.

If you still can't find any candidates, then you have a very low source impedance and should use op amps that are closest to the bottom. In such cases,

paralleling of low noise op amps is also an option.

Conclusion

Noise analysis can be a daunting task at first and is unfamiliar territory for many design engineers. The greatest influence on overall noise performance is the source impedance associated with the signal. This selection guide helps the designer, whether novice or veteran, choose the best op amps for a given source impedance. **LT**

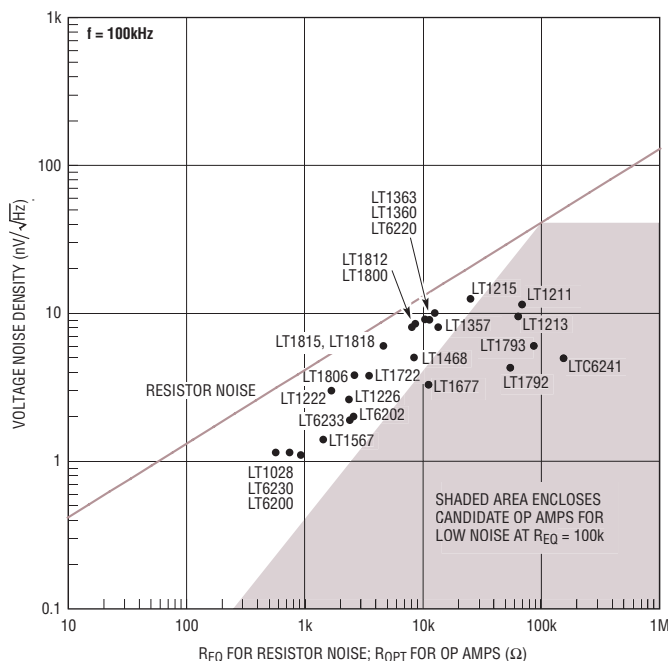
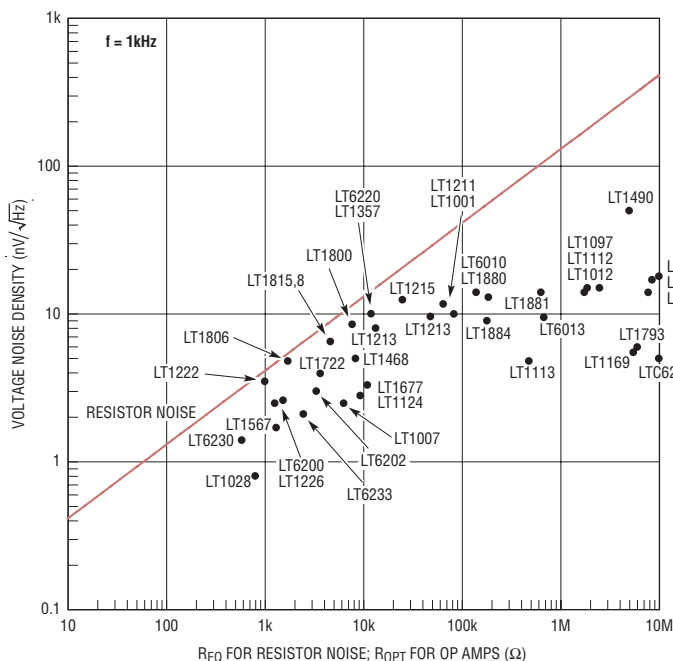
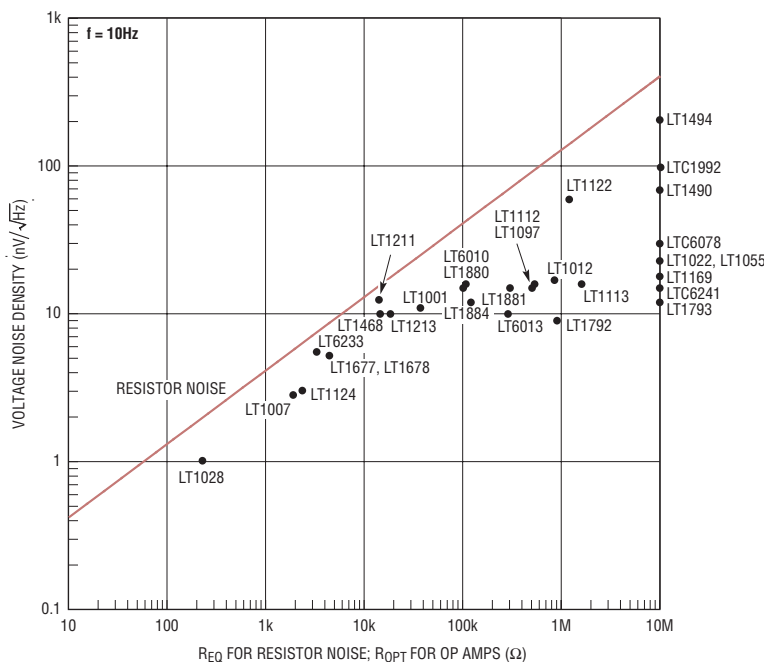


Figure 2. Use these three plots to find the best low noise op amps for your application. See "Selecting the Best Op Amps" in the text.

Multi-Output Supply Drives White LEDs, Provides LCD or OLED Bias in a 3mm × 3mm DFN Package

Introduction

Many of today's cell phones, PDAs and digital still cameras contain a high-resolution TFT-LCD display and sometimes an additional secondary OLED (Organic Light-Emitting Diode) display. OLED displays are fast becoming the secondary display of choice because they are brighter, thinner and more responsive than equivalent LCDs. The LT3466-1 is a dual switching regulator designed to meet the power supply requirements of small displays, including LCD-bias, white LED backlight and OLED displays.

The LT3466-1 integrates a full featured white LED driver and a boost converter in a low profile 3mm × 3mm DFN package. It provides space and component savings with integrated 44V power switches and Schottky diodes. The LED driver can be configured to drive up to 10 white LEDs in series from a single Li-Ion battery. The white LED driver features a low 200mV reference for programming the LED current, thereby minimizing the power loss in the current setting resistor for better efficiency. The boost converter can be used for generating the main LCD bias voltages, or for providing the OLED bias supply. The boost converter achieves ±1.5% out-

by Gurjit Thandi

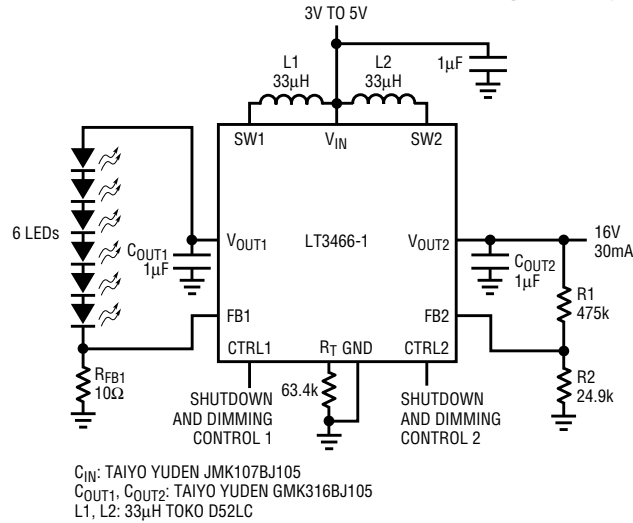


Figure 1. The LT3466-1 powers a main LCD backlight and a secondary OLED display. It provides a 20mA drive for the six-white-LED LCD backlight and a 16V output for the OLED display.

put voltage accuracy by the use of an internal precision 0.8V reference.

The LT3466-1 also provides independent dimming and shutdown control of the two converters. The operating frequency of LT3466-1 can be set with an external resistor over

a 200kHz to 2MHz range. Additional features include output overvoltage protection, internal compensation and internal soft-start. The LT3466-1 operates from a wide input voltage range of 2.7V to 24V, making it suitable for a variety of applications.

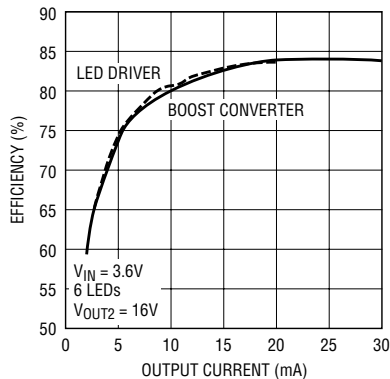


Figure 2. Efficiency versus load current for the circuit in Figure 1

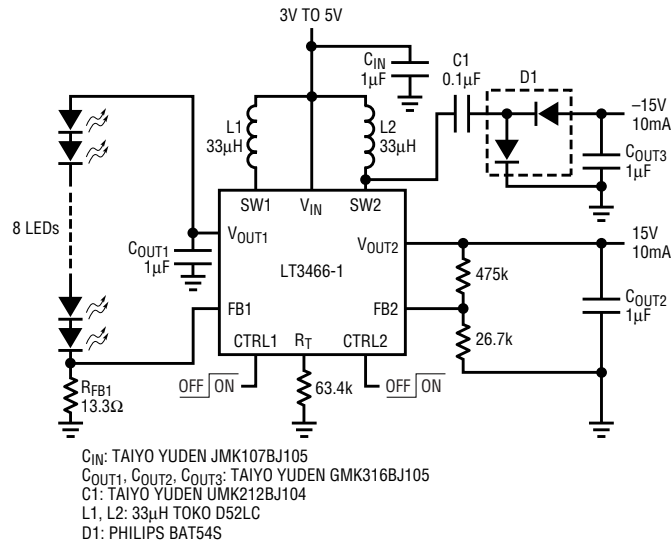


Figure 3. High efficiency, Li-Ion powered complete TFT-LCD supply (bias and backlighting). The LT3466-1 drives eight white LEDs at 15mA to provide the backlight and generates dual, ±15V, outputs for the LCD-bias.

Dual Display Power Supply for Cell Phones

A typical application for the LT3466-1 is as a driver for dual displays in cell phones. Present day, clam-shell cell phones typically use a color TFT-LCD main display and a secondary OLED display. Figure 1 shows the LT3466-1 powering the main LCD backlight and the secondary OLED display. The

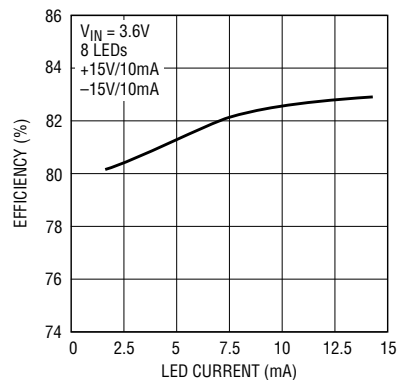


Figure 4. Efficiency versus LED current for the circuit in Figure 3. The circuit achieves greater than 83% efficiency driving eight LEDs at 15mA from 3.6V input.

LT3466-1 drives 6 white LEDs at 20mA for backlighting the main LCD panel and generates 16V output for powering the OLED. The LT3466-1 allows for independent dimming control of the main and secondary displays via the respective CTRL1 and CTRL2 pins. Figure 2 shows the efficiency versus output current for both the LED driver and the boost converter. The typical efficiency at 3.6V input supply is 84% with the white LEDs and the OLED driven at 20mA.

Low Cost, Complete LCD Bias and White LED Backlighting Solution for Small TFT Displays

Small, active-matrix, TFT-LCD displays, used in cell phones, PDAs and other handheld devices generally require four to ten white LEDs for providing the backlight and fixed +15V and -15V supply voltages to bias the LCD. Figure 3 shows LT3466-1 powered complete TFT-LCD supply with minimal external components

and high efficiency. The LT3466-1 drives eight white LEDs at 15mA and generates 15V boost output powered from a single Li-Ion supply. A discrete charge pump produces the secondary output of -15V. As seen in Figure 4, the circuit achieves greater than 83% efficiency driving eight LEDs at 15mA from 3.6V input.

Conclusion

The LT3466-1 integrates a full featured white LED driver and a boost converter in a space saving 3mm x 3mm DFN package. Integrated power switches and Schottky diodes reduce the overall system cost and size making it an excellent fit for handheld applications. Features like internal compensation, soft-start, Open LED protection enables LT3466-1 to provide complete TFT-LCD supply (bias and white LED backlight) for handheld devices with minimal external components and high efficiency.

LT3477, continued from page 27

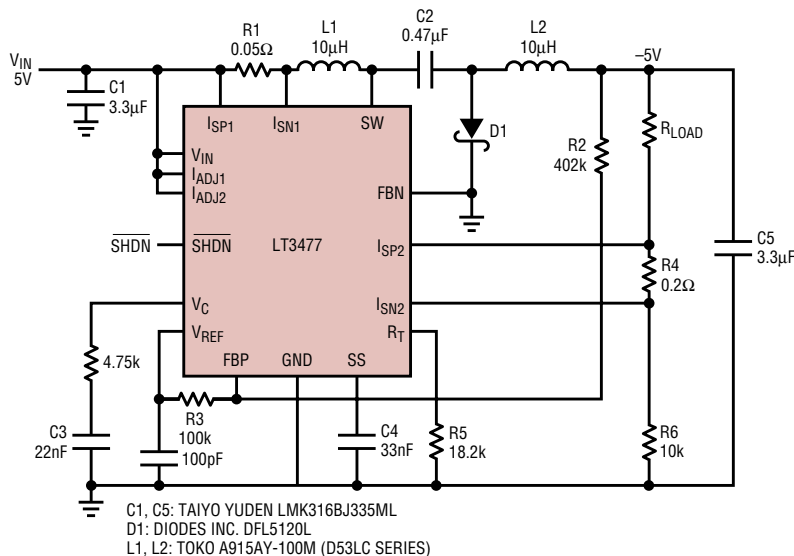


Figure 10. Negative output voltage Cuk converter.

using a Cuk topology for 5V to -5V conversion. The first current sense amplifier is used for input current limit, and the second current sense amplifier is used for ground rail current sense to accurately limit the load current at 500mA. Even though the two current sense amplifiers are used,

efficiency up to 81% at 500mA output load can still be achieved. Figure 11 shows the efficiency.

Conclusion

The rail-to-rail constant-current/constant-voltage operation of the LT3477 makes the device an ideal choice for

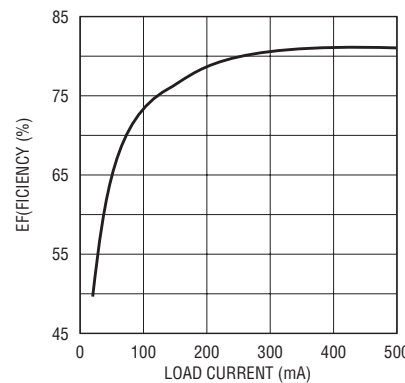


Figure 11. Efficiency of the Cuk converter.

a variety of constant-current designs, including negative outputs. The dual current-sense amplifiers allow flexible configuration for input current limit, constant output current and fail-safe protection, along with excellent output voltage regulation. A wide input voltage range and the ability to produce outputs up to 42V make the LT3477 extremely versatile.

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Single Cell Step-Up DC/DC Converter Features 400mA Switch Current in an SC70 Package

by Dave Salerno

Introduction

The LTC3525 raises the bar for boost converter performance and power capability in an SC70 package. It is an inductor-based synchronous step-up (boost) DC/DC converter that operates from input voltages as low as 1V, boosting them to 3.3V or 5V. Its powerful internal 400mA switch allows the LTC3525 to deliver up to 150mA of load current with efficiency up to 94%. To further save space, it requires only three external components (two small ceramic capacitors and a small inductor), so a complete solution fits into spaces previously reserved only for charge pump designs.

The LTC3525-3.3 and LTC3525-5 are both packaged in the 2mm × 2mm × 1mm SC70 package, and operate over an input range of 0.8V to 4.5V. This flexibility makes them suitable for compact applications powered by 1 to 3 alkaline/NiMH cells, or a single Li-ion battery. The 3.3V version can even maintain regulation with input voltages exceeding the output voltage.

Small But Full-Featured Solution

Despite its diminutive SC70 package, the LTC3525 includes many sophisticated features, such as:

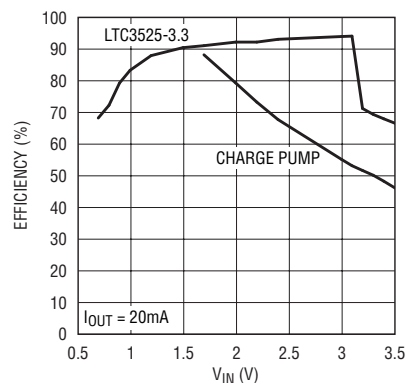


Figure 3. Comparison of efficiency versus input voltage for LTC3525-3.3 and an equivalent charge pump based boost circuit

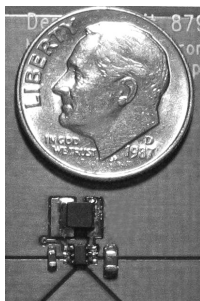


Figure 1. Typical application using the LTC3525-3.3/-5

output disconnect, inrush current limiting, low output voltage ripple, synchronous rectification, single cell capability, anti-ring control and less than 1 μ A shutdown current. It also features overcurrent protection and thermal shutdown, enabling it to sustain an indefinite short circuit without damage.

External component selection is easy, since most applications require just a 1 μ F ceramic input capacitor for local decoupling, a 10 μ F ceramic output filter capacitor and a 10 μ H inductor (although any value from 4.7 to 15 μ H can be used). Be sure to use only X5R or X7R style capacitors, keeping them close to the pins of the IC.

The LTC3525 is enabled by pulling the $\overline{\text{SHDN}}$ pin up to any voltage between 1V and 5V, regardless of input or output voltage.

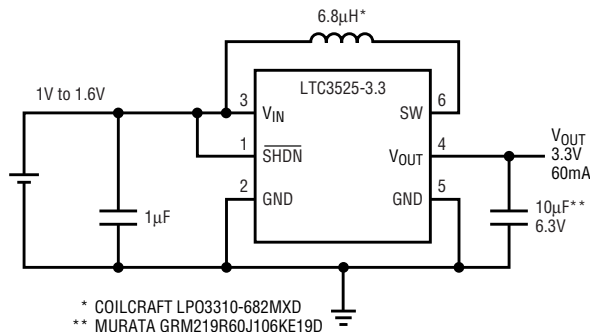


Figure 4. Single cell to 3.3V converter delivers 60mA of load current in a 1mm profile

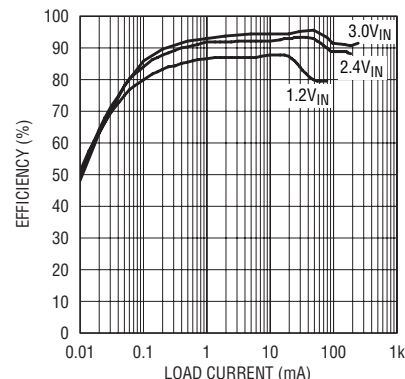


Figure 2. Efficiency versus load for the LTC3525-3.3

High Efficiency Over a Wide Range of Input Voltages & Load Currents

The LTC3525 uses a proprietary, patent pending technique of adaptively adjusting peak inductor current as a function of load and input voltage. This technique provides optimum efficiency at light to medium loads, while enabling it to supply heavier load currents that are beyond the capability of other solutions of this size.

The LTC3525's low quiescent current of only 7 μ A on V_{OUT} allows it to maintain impressive efficiency down to extremely light loads, as shown in Figure 2, and over a broad range of input voltages. By comparison, the efficiency of a charge pump design varies widely as the battery voltage changes, as illustrated in the graph in

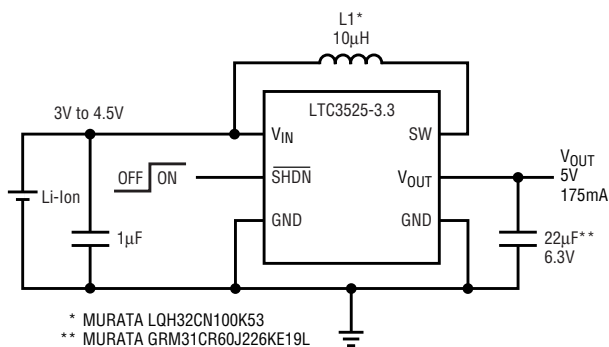


Figure 5. Li-ion to 5V converter delivers 175mA of load current with <0.5% ripple

Figure 3. Note that the charge pump design requires an input voltage of at least 1.7V to generate a regulated 3.3V output. Comparable inductor-based solutions require larger packages and more external components, making them unsuitable to applications where board space is at a premium, or too expensive for cost sensitive applications.

Single Cell to 3.3V Converter with 1mm Profile

A single alkaline or nickel cell to 3.3V converter, using the LTC3525-3.3, is shown in Figure 4. This application uses an inductor and output capacitor chosen to achieve a 1mm profile. It delivers 60mA of load current from a single cell, and 140mA from two cells, while fitting into a 5mm x 7mm footprint. The ability of the converter to operate with input voltages below 1V allows it to use all the available energy in the battery, and also prevents the converter from shutting off in the event that a load transient causes a momentary drop in input voltage.

Li-ion/3-Cell to 5V Converter Delivers Over 175mA with Low Output Ripple

The LTC3525 has been designed for very low output ripple with minimal output capacitance. In most applications, a 10µF ceramic capacitor will yield less than 1% peak-to-peak output ripple. By using a 22µF capacitor, the output ripple can be reduced to less than 0.5% of V_{OUT} , making it suitable for many noise sensitive applications that previously required a larger, more expensive fixed frequency converter.

The circuit in Figure 5, which occupies a space of just 6mm x 6mm, supplies 5V at 175mA or more from a Li-ion battery (or three alkaline or nickel cells). With a 22µF output capacitor, the output ripple is only 22mV_{P-P} at light load, and less than 50mV_{P-P} at full load, as shown in Figure 6. The efficiency peaks at 93% and remains above 85% over three decades of load current, as shown in Figure 7. This solution could also be used to provide 5V at 200mA in a 3.3V powered system. The entire solution fits in a 1.8mm profile.

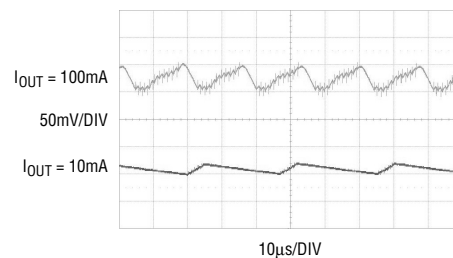


Figure 6. Output voltage ripple of the 5V converter at min and max load

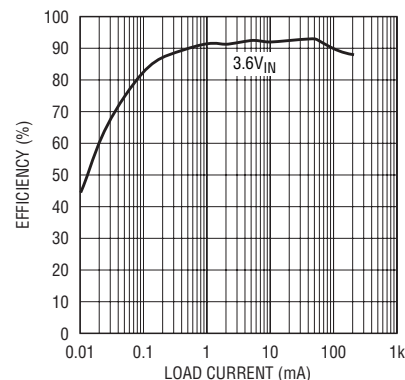


Figure 7. Efficiency versus load for the Li-ion to 5V converter

Conclusion

Many of today's battery powered portable devices, such as MP3 players, medical instruments and digital cameras can benefit from the small size, simplicity and extended battery life offered by the LTC3525. Its tiny, low profile SC70 package and minimal external part count make it a viable, high performance alternative to less efficient charge pump designs. Its 400mA switch current and low output ripple allow it to replace more expensive fixed frequency converters in cost-sensitive applications.

LTC4306, continued from page 31 that ENABLE remains at a constant logic low while all other pins are connecting, so that the LTC4306 remains in its default high impedance state and ignores connection transients on SDAIN and SCLIN during connection. In addition, make the ALERT# connector pin shorter than the V_{CC} pin, so that V_{CC} establishes solid contact with the I/O card pull-up supply pin and powers the pull-up resistors on ALERT1#-ALERT4# before ALERT#

makes contact. When disconnecting, ENABLE breaks contact first, resetting the LTC4306 to its default state, so that it causes minimal disturbance on the SDAIN and SCLIN bus as the card disconnects.

Conclusion

The LTC4306 eases the practical design issues associated with large 2-wire bus systems. It serves as a multiplexer to provide nested addressing. It disconnects buses when they are

stuck low. It breaks a large capacitive bus into smaller pieces and allows I/O cards to be hot-swapped into and out of live systems. It logs faults, reports to the master, and works with the master to resolve faults efficiently.



Tiny DC/DC Buck Controller Provides High Efficiency and Low Ripple

by Theo Phillips

Introduction

To secure a foothold in today's congested circuit boards, a power controller must deliver the most functionality in the smallest package. With a blend of popular features squeezed into a SOT-23 or 3mm × 2mm DFN, the LTC3772 makes a power supply designer's job easy. This versatile DC-DC controller supports a wide input voltage range, 2.5V to 9.8V, and maintains high efficiency over a variety of output current levels. Its 550kHz switching frequency trims solution size by permitting the use of small passive components. Its No R_{SENSE}^{TM} constant frequency architecture also eliminates the need for a sense resistor.

Circuit Description

Figure 1 shows a typical application for the LTC3772. This circuit provides a regulated output of 2.5V from a typical input voltage of 5V, but it can also be powered from any input voltage between 2.75V and 9.8V (depending on the voltage rating of the P-channel power MOSFETs). This wide input range makes the LTC3772 suitable for a variety of input supplies, including 1- and 2-cell Li-Ion and 9V batteries, as well as 3.3V and 5V supply rails. The internal soft-start ramps the output voltage smoothly from 0V to its final value in 1ms (Figure 2).

At low load currents ($\leq 10\%$ of I_{MAX}), the LTC3772 enters Burst Mode operation. Compared with other power saving schemes, this variant of Burst Mode operation surrenders a modicum of efficiency to obtain very low output voltage ripple. Typically producing just 30mV for a typical application using ceramic output capacitors, the LTC3772 is ideal for noise-sensitive portable applications. Figure 3 illustrates inductor current and output voltage waveforms for Burst Mode operation.

The LTC3772 uses the drain to source voltage (V_{DS}) of the power

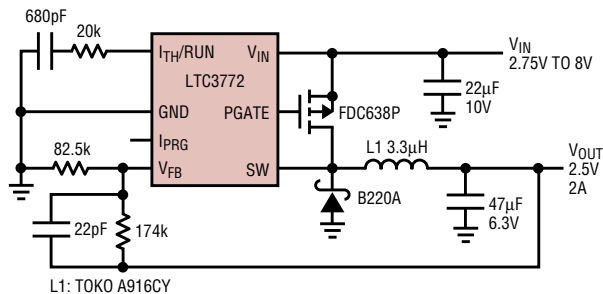


Figure 1. Typical application delivering 2.5V at 2A

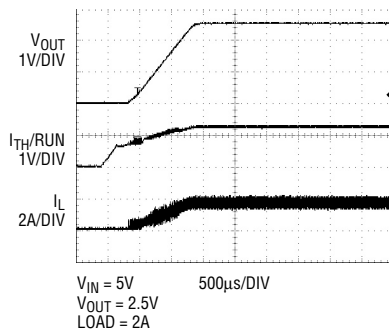


Figure 2. The output voltage rises smoothly without requiring a soft-start capacitor as seen in this startup waveform for the converter in Figure 1.

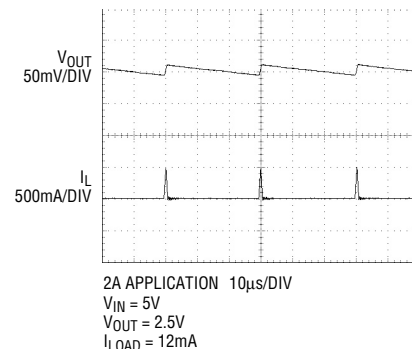


Figure 3. The LTC3772's Burst Mode operation maintains light load efficiency while holding output voltage ripple to just 20mV in this application.

P-Channel MOSFET to sense the inductor current. The maximum load current that the converter can provide is determined by the $R_{DS(ON)}$ of the MOSFET, which is a function of the input supply voltage (which supplies the gate drive). The maximum load current can also be changed using the

current limit programming pin I_{PRG} , which sets the peak current sense voltage across the MOSFET to one of three states; each voltage is associated with its own inductor current limit. With I_{PRG} floating, the circuit of Figure 1 can reliably provide 2.5V at 2A from a 3.3V input supply. Efficiency for this circuit exceeds 93%, as shown in Figure 4. In drop out, the LTC3772 can operate at 100% duty

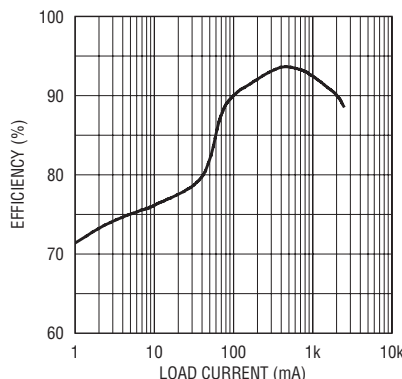


Figure 4. Efficiency vs load current for the converter in Figure 1, with input of 3.3V

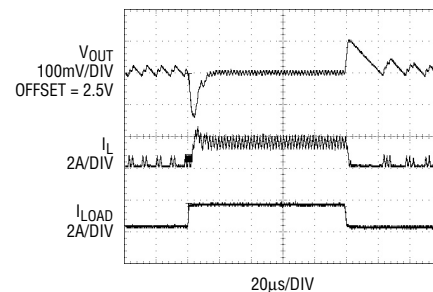


Figure 5. Transient performance of the converter in Figure 1, with input of 5V

cycle, providing maximum operating life in battery-powered systems.

OPTI-LOOP Compensation

To meet stringent transient response requirements, some switching regulators use many large and expensive output capacitors to reduce the output voltage droop during a load step. The LTC3772, with OPTI-LOOP compensation, is stable for a wide variety of output capacitors, including tantalum, aluminum electrolytic, and ceramic capacitors. The ITH pin of the LTC3772 allows users to choose the proper component values to compensate the loop

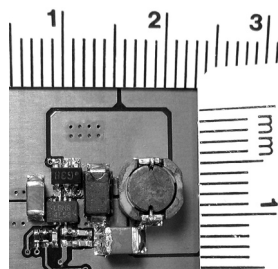



Figure 6. A typical LTC3772 application occupies just 1.5 square centimeters.

so that the transient response can be optimized with the minimum number of output capacitors. Figure 4 shows a transient response for the circuit in

Figure 1, using just one 47 μ F output capacitor. The response is quite fast, even though it involves a transition from Burst Mode operation to continuous conduction mode.

Conclusion

For single-output designs with load currents as high as 5A from input voltages up to 9.8V, the LTC3772 delivers the most popular features of PFET controllers in a very small package. With small ancillary components and no sense resistor, the overall solution is unmatched where board space is at a premium. 

LTC4063, continued from page 33

which can be made in various shapes including very thin cells, ideal for cell phones and other small handheld devices. Although the discharge characteristics and performance of the different types of Li-Ion cells vary, the charging characteristics are essentially the same.

Rechargeable lithium battery technology is relatively new, and because of that, many improvements in future battery performance are almost guaranteed. Different materials, chemicals and construction will undoubtedly produce a battery that is ever closer to that perfect battery.

The recommended charge voltage is a compromise between cell capacity, cell life and cell safety. Higher charge voltages increase the mAh cell capacity, but shorten the cell lifetime. There are also upper limits that must be adhered to for safety reasons. The most common charge voltage is 4.2V \pm 1% although future battery designs may have a slightly higher voltage. In applications that favor cycle life over cell capacity, a lower charge voltage greatly increases cycle life. Shallow rather than deep discharge cycles increase cycle life as well. The end of life for a Li-Ion battery is typically when its capacity drops to 80% of its rating.

One lesser known fact about Li-Ion batteries is their aging characteristics.

Li-Ion batteries have a limited lifetime whether they are stored or in daily use. The permanent capacity loss, especially for lithium manganese chemistries, increases with charge level and temperature. For example, storing a battery at a 40% charge level at 25°C for a year could result in a permanent capacity loss of 4%, whereas if stored at a 100% charge level, the permanent capacity loss would be close to 20%. Stored at 100% charge level at 40°C could produce a permanent capacity loss up to 35% after one year. Of course, further improvements in Li-Ion battery technology will surely minimize aging


Li-Ion batteries cannot absorb overcharge. Charge current must be completely stopped when the battery reaches full charge. Overcharge can cause internal lithium metal plating, which is a safety concern. Also, Li-Ion batteries should not be discharged below 2.5V to 3V, depending on battery chemistry, as internal copper plating can form causing a short circuit.

Battery Pack Protection: What Is It?

Most manufacturers of Li-Ion batteries will not sell batteries unless they include built in battery pack protection circuitry for safety and to prolong battery life. The circuitry includes a FET switch in series with the battery

that turns off in the event of an over voltage, under voltage, over current and over temperature condition when either charging or discharging the battery. A prolonged overvoltage when charging can result in the battery overheating, bursting or even exploding. When discharging, the pack protection disconnects the battery if the battery voltage drops below a predetermined threshold level or if the battery current exceeds a preset limit. Without pack protection, Li-Ion batteries can easily be damaged or worse, can cause damage to other circuitry or bodily injury.

Conclusion

The LTC4063 Li-Ion battery charger provides the user with an excellent combination of packaging (3mm \times 3mm DFN), high charge current (1A), tight float voltage (0.35%), low I_{DETECT} current capability (5mA), choice of termination and an integrated 100mA LDO regulator. Two other chargers share similar charging characteristics but differ on features. The LTC4061 has no regulator but includes a NTC temperature qualification input, a USB current select input and an additional status output. The LTC4062 replaces the LDO regulator with a programmable comparator and reference and also includes a USB current select input. 

New Device Cameos

Unprecedented Power Density from Breakthrough 10A DC/DC μ Module

A breakthrough DC/DC power converter combines the best features of two heretofore separate design approaches. It mixes the design simplicity of a power module with the power densities of a high performance IC to create a device that is unprecedented in ease-of-use, versatility and power density.

The first of many μ Modules™ provides designers a complete 10A switching power supply in a tiny (15mm x 15mm) footprint, low profile (2.8mm) land grid array (LGA) package. The LTM[®]4600 is a synchronous switch mode DC/DC step-down regulator with built-in inductor, supporting power components and compensation circuitry. By simplifying power system development, this new high-density power supply reduces development time for a broad range of systems, including network routers, blade servers, cellular base stations, medical diagnostic equipment, test instrumentation and RAID systems.

The LTM4600 accommodates a wide input voltage range of 4.5V to 28V. The high level of integration and synchronous current mode operation allows the LTM4600 to deliver superior transient response and up to 10A continuous current (14A peak) at up to 92% efficiency. It simplifies power supply design and construction, requiring only input and output bulk capacitors and a single resistor to set the output voltage within a range of 0.6V to 5V.

The LTM4600 DC/DC μ Module is a complete stand-alone surface-mount power supply that can be handled and assembled like a standard integrated circuit. Moreover, its low profile design permits the LTM4600 to be soldered onto the back side of a circuit board, freeing up valuable board space.

The LTM4600 DC/DC μ Modules are self-protected against overvoltage

and short circuit conditions. Its fast transient response minimizes required bulk output capacitance. Furthermore, two LTM4600s can be operated in parallel, increasing load current capability to 20A. The LTM4600 is offered in two versions: standard and high input voltage. The LTM4600EV operates from 4.5V to 20V, whereas the LTM4600HVEV has an operating voltage range from 4.5V to 28V.

The LTM4600IV and LTM4600HVIV are tested and guaranteed to operate over the -40°C to 85°C temperature range.

16-bit, 130Msps ADC Delivers 100dBc SFDR for High Performance Receivers and Instrumentation

The LTC2208 is a 130Msps, sampling 16-bit A/D converter designed for digitizing high frequency, wide dynamic range signals with input frequencies up to 700MHz. The input range of the ADC can be optimized with the PGA front end.

The LTC2208 is perfect for demanding communications applications, with AC performance that includes 78dBFS Noise Floor and 100dB spurious free dynamic range (SFDR). Ultra low jitter of 70fs_{RMS} allows undersampling of high input frequencies with excellent noise performance. Maximum DC specs include ± 4 LSB INL, ± 1 LSB DNL (no missing codes).

The digital output can be either differential LVDS or single-ended CMOS. There are two format options for the CMOS outputs: a single bus running at the full data rate or demultiplexed buses running at half data rate. A separate output power supply allows the CMOS output swing to range from 0.5V to 3.6V.

The ENC⁺ and ENC⁻ inputs may be driven differentially or single-ended with a sine wave, PECL, LVDS, TTL or CMOS inputs. An optional clock duty cycle stabilizer allows high performance at full speed with a wide range of clock duty cycles.

The LTC2208 packages an extensive feature set in a 9mm x 9mm QFN package delivering low power consumption at 1250mW without the need for heat sinking. Most importantly, both the power consumption and total solution size with integrated bypass capacitance are less than half that of the nearest competitor.

The LTC2208 family includes speed grades of 130Msps, 105Msps, 80Msps, 65Msps, 40Msps, 25Msps and 10Msps all with superior SFDR and SNR performance. In addition to the 16-bit parts, 14-bit versions of this family will also be available. All devices are supported with demo boards for quick device evaluation.

Tiny Controller Makes it Easy to Rapidly Charge Large Capacitors

The LT3750 is a current-mode flyback controller optimized for charging large value capacitors to a predetermined target voltage. This target voltage is set by the turns ratio of the flyback transformer and just two resistors in a simple, low voltage network, so there is no need to connect components to the high voltage output. The charging current is set by an external sense resistor and is monitored on a cycle-by-cycle basis.

The device is compatible with a wide range of control circuitry, being equipped with a simple interface consisting of a CHARGE command input bit and an open drain DONE status flag. Both of these signals are compatible with most digital systems, yet are tolerant to voltages as high as 24V.

The architecture balances a high degree of integration with the flexibility of leaving key parameters definable by the user. This leaves only a few issues to consider in order to complete the design: Input capacitor sizing, transformer design, and output diode selection.

The LT3750 is available in a space saving, 10-lead MSOP package.

Performance of 50µA CMOS Amplifier Rivals Best Bipolar Op Amps with 0.7µV/°C Drift

The LTC6078/LTC6079 are dual/quad, low offset, low noise operational amplifiers with low power consumption and rail-to-rail input/output swing.

Input offset voltage is trimmed to less than 25µV and the CMOS inputs draw less than 50pA of bias current. The low offset drift, excellent CMRR, and high voltage gain make it a good choice for precision signal conditioning.

Each amp draws only 54µA current on a 3V supply. The micropower, rail-to-rail operation of the LTC6078/LTC6079 is well suited for portable instruments and single supply applications.

The LTC6078/LTC6079 are specified on power supply voltages of 3V and 5V from -40°C to 125°C. The dual amplifier LTC6078 is available in 8-lead MSOP and 10-lead DFN packages. The quad amplifier LTC6079 is available in 16-lead SSOP and DFN packages.

Dual and Quad, 1.8V, 13µA Precision Rail-to-Rail Op Amps

The LT6001 and LT6002 are dual and quad precision rail-to-rail input and output operational amplifiers. Designed to maximize battery life in always-on applications, the devices operate on supplies down to 1.8V while drawing only 13µA quiescent current. The low supply current and low voltage operation is combined with precision

specifications—for instance, input offset is guaranteed less than 500µV. The performance on 1.8V supplies is fully specified and guaranteed over temperature. A shutdown feature in the 10-lead dual version can be used to extend battery life by allowing the amplifiers to be switched off during periods of inactivity.

The LT6001 is available in the 8-lead MSOP package and a 10-lead version with the shutdown feature in a tiny, dual fine pitch leadless package (DFN). The quad LT6002 is available in a 16-lead SSOP package and a 16-lead DFN package. These devices are specified over the commercial and industrial temperature range. 

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LTC2950/51, continued from page 4

Figure 5 shows an actual ESD event. Note the arc onto the \overline{PB} pin. The ESD strike fed directly onto the pin; there were no series resistors or parallel capacitors. This strike did not damage the pin, nor did it generate any leakage.

LTC2950-1 and LTC2950-2 Versions

The LTC2950-1 (high true EN) and LTC2950-2 (low true \overline{EN}) differ only by the polarity of the EN/ \overline{EN} pin. Both versions allow the user to extend the amount of time that the \overline{PB} must be held low in order to begin a valid power on/off sequence. An external capacitor placed on the ONT pin adds additional time to the turn-on time. An external capacitor placed on the


OFFT pin adds additional time to the turn-off time. If no capacitor is placed on the ONT (OFFT) pin, then the turn on (off) duration is given by an internally fixed 32ms timer. The LTC2950 fixes the \overline{KILL} turn off delay time ($t_{\overline{KILL}(\text{OFF DELAY})}$) at 1024ms (the amount of time from interrupting the µP to turning off power).

LTC2951-1 and LTC2951-2 Versions

The LTC2951 fixes the turn on debounce time at 128ms. The turn off debounce time is the same as the LTC2950: 32ms internal plus the optional additional external when a capacitor is placed on the OFFT pin. The \overline{KILLT} pin in the LTC2951-1 and LTC2951-2 provides extendable \overline{KILL}

turn off timer, $t_{\overline{KILL}(\text{OFF DELAY, ADDITIONAL})}$, by connecting an optional external capacitor on the \overline{KILLT} pin. The default power down delay time is 128ms, $t_{\overline{KILL}(\text{OFF DELAY})}$.

Conclusion


The LTC2950/LTC2951 is a family of micro-power (6µA), wide input voltage range (2.7V to 26.4V) push button controllers. The parts lower system cost and preserve battery life by integrating flexible push button timing, a high voltage LDO, and a simple µP interface that provides intelligent power up and power down. The device is available in space saving 8-lead 3mm × 2mm DFN and ThinSOT™ packages. 

LTC3454, continued from page 24

transconductance amplifier with sink only capability—takes control of the regulation loop and prevents V_{OUT} runaway. The V_{OUT} threshold at which this happens is approximately 5V.

If the LED faults as a short circuit, the regulation loop continues to regulate the output current to its programmed current level.

Conclusion

The LTC3454 adds to Linear Technology's family of LED drivers. High efficiencies can be achieved over the entire Li-Ion range with a minimal number of external components. Additionally, it draws zero current when in shutdown, helping conserve battery life in hand held battery powered applications. The LTC3454 is available in a low profile small footprint 3mm × 3mm DFN package. 



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