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Multiphase Power Conversion for Portable and Point-of-Load *Boost* Applications

by David Salerno

Introduction

Multiphase power converters offer the advantages of higher efficiency, smaller size and lower capacitor ripple currents over their single phase counterparts. The higher effective switching frequency and phased ripple currents significantly reduce the size and cost of the filter capacitors and lower output ripple, while allowing the use of several small inductors. This has made them popular in many high current buck (step-down) applications, especially where space is a concern. With the LTC3425, the industry's first multiphase monolithic boost converter, you can achieve the same performance and size benefits in boost (step-up) applications.

This 4-phase synchronous boost converter can deliver over 12W of power in a smaller size, with higher efficiency and lower output ripple than is achievable with a comparable single-phase boost converter. The LTC3425 can startup with as little as 1V, and operate with inputs up to 4.5V, making it suitable for a variety of input voltage applications. The output voltage range is 2.4V to 5.25V, and peak current capability is over 5A.

Multiphase Converters: They're Not Just for Buck Applications Anymore

The high frequency (up to 8MHz) 4-phase architecture allows the use of small, low cost inductors rather than a single large, bulky inductor, and requires much less output filter capacitance than the equivalent

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Figure 1. How does a multiphase boost converter improve on its single phase counterpart? First of all, a multiphase topology saves space and simplifies layout by removing bulky, hard-to-place components and replacing them with easier-to-fit, low profile components. Inductor and output capacitor size comparison of single-phase and 4-phase circuits.

Issue Highlights

Our cover article introduces the LTC3425, the industry's first multi–phase monolithic boost converter, offering the advantages of small size and low output ripple currents.

This 4-phase synchronous boost converter can deliver over 12W of power in a smaller size, with higher efficiency and lower output ripple than is achievable with a comparable single-phase boost converter.

Featured Devices

This issue features a variety of exciting new devices which simplify application designs and improve their performance.

Hot Swap, Hot Plug

The LTC4240 provides a controlled on-off switch for four hot swappable board power supply voltages, allowing the board to be safely inserted or removed from a live CompactPCI (CPCI) slot without disturbing the system power supplies. The LTC4240 includes an I^2C-compatible interface that allows software control and monitoring of device function and power supply status.

The new LTC4302 addressable 2-wire bus buffer addresses problems associated with live insertion of I/O cards in data processing and communications systems. The LTC4302 provides a bridge between any two physically separate 2-wire buses. Masters on the buses can command the LTC4302 to connect and disconnect the backplane and card buses through software. By using multiple LTC4302s in a system and connecting and disconnecting them appropriately, users can assign the same address to multiple devices in the system.

The LT4254 is designed to turn on a board's supply voltage in a controlled manner, allowing the board to be safely inserted or removed from a live backplane having a supply voltage from 10.8 to 36 volts. The device features programmable inrush current control, current foldback, programmable 1% tolerance on the undervoltage and overvoltage thresholds, overcurrent protection, and a power good output signal that indicates when the output supply voltage is ready.

Power

The LTC1980 is a single-device that manages both battery charging and generation of the regulated system bus voltage.

The LT3464 is the only micropower boost converter in the industry to combine a 36V NPN power switch, power Schottky diode, and output disconnect into an 8-lead ThinSOT™. This unprecedented level of integration saves several external components while offering true output disconnect, making it possible to generate outputs of up to 34V with a zero current shutdown while using a mere 40mm^2 of board area.

Amplifiers and Filters

The LT6011 dual precision op amp fits into a 3×3mm^2 DFN package, which is so small it doesn't even have leads. The LT6011 also offers low voltage, micro-power operation. A quad version, the LT6012, is also available in the SO-14 and GN16 packages.

The LT6600 is a family of fully differential amplifiers with an integrated 4th order lowpass filter. Each device in the family features a fixed cutoff frequency (2.5MHz, 10MHz and 20MHz) and operates with power supplies ranging from 3V to 10V. The LT6600s have a proprietary architecture that minimizes noise and distortion while maximizing speed.

Two new families of single, dual, and quad amplifiers, the 1mA LT6220/1/2 and the 3mA LT1803/4/5, provide high speed amplification on supplies as low as 2.5V. The 85MHz LT1803 series is optimized for large signals, with a 100V/µs slew rate, while the 60MHz LT6220 series focuses on excellent DC performance at low current, with a maximum input offset specification of only 350µV.

LTC in the News...

On July 22, Linear Technology Corporation announced its financial results for its fiscal year ending June 29, 2003. According to Robert H. Swanson, Chairman of the Board and CEO, "Fiscal 2003 was a successful, but challenging year for Linear, as we reported solid growth in a modestly improving economic environment. We benefited from the demand for our high-performance analog solutions in many new diverse applications in wireless, high-end consumer, automotive, security, computer, military, industrial/medical and networking. Our sales increased 18%, and operating income improved 31%. We had an increase in net income of 20%, which was impacted by a decrease in interest income due to the dramatic decrease in interest rates.

Although the overall demand for our products is improving, our customers continue to be cautious in their ordering patterns. While accurately forecasting short-term results remains a challenge, we expect low single digit growth in sales and profits in the September quarter, which is in line with our normal summer quarter patterns."

The Company reported net sales for the year of $606,573,000 and net income of $236,591,000. Diluted earnings were $0.74 per share. A cash dividend of $.06 per share was paid on August 20, 2003 to stockholders of record on August 1, 2003.

Design Ideas and Cameos

Starting on page 30 are five new Design Ideas, and at the back are three New Device Cameos. See www.linear.com for complete device specifications and applications information.

CompactPCI is a trademark of PCI Industrial Computer Manufacturers Group. I^2C is a trademark of Philips Electronics N.V.

LTC3425, continued from page 1

Table 1. Comparison of typical inductors used for 1-phase and 4-phase designs

	Inductor	Inductance (µH)	Required Qty	Total Area (mm^2)	Height (mm)	Max Total DC Resistance (Ω)	Total Peak Current Rating (A)
1-Phase	Coilcraft DO3316	2.2	1	122	5.21	0.012	7.0
4-Phase	TDK RLF5018T	2.7	4	29.12 × 4 = 116.5	2.05	0.033/4 = 0.0083	1.8 × 4 = 7.2

single-phase circuit. This is ideal for space-constrained boards, Point-of-Load regulators, and portable devices that demand the use of low-profile components. For example, in a 2-cell NiCd or NiMH to 3.3V/2A boost application, the peak inductor current required for a single-phase design is nearly 5A.

Figure 1 shows the size difference between a typical single inductor that would be required to handle this current, and the inductors that could be used in a 4-phase design. Figure 1 also compares the output capacitors required to achieve the same output ripple voltage in single-phase and 4-phase applications. Table 1 shows specifications for the inductors pictured in Figure 1—not only are the four small inductors much thinner, but they also have a lower combined DC resistance for improved efficiency.

Easy to Use

Designing a converter using the LTC3425 is no different than designing a traditional single phase boost converter. All the power switches are internal, so the 4-phase operation is transparent. Current limit and switching frequency for all four phases are each programmed by a single resistor, as in single phase designs. Setting the output voltage and compensating the loop are also no different than in other familiar designs.

Circuit Description: Four Independent Power Stages

Each of the four phases has an NMOS and a PMOS power switch, and controls its own inductor current using a peak current mode control loop, consisting of a current comparator with adaptive slope compensation and a reverse current comparator for discontinuous mode operation. In discontinuous mode, an internal resistor is placed across the inductor when the synchronous rectifier turns off, damping any high frequency ringing.

A single error amplifier is used for all four phases, and controls the peak current required to maintain regulation. Referring to Figure 2, the loop compensation components are connected between COMP and GND. Soft-start time is set by the C_{SS} capacitor, which ramps the current limit up to its final value during startup

Each V_{OUT} pin should have its own ceramic filter capacitor located as close as possible to the V_{OUT} and GND pins in that phase. These are typically 0805 size parts. The pinout of the LTC3425 lends itself to a tight symmetrical layout of the power components. With the 4-phase architecture, low output voltage ripple is achieved using only the four small ceramic capacitors, even at load currents of 2A or more. An optional bulk capacitor on V_{OUT} can be added to improve transient response with dynamic loads. This can be a ceramic, tantalum, or an OSCON-style capacitor.

The output disconnect feature effectively eliminates the PMOS body diode between the switch node and V_{OUT} during shutdown, allowing V_{OUT} to discharge to zero volts, all while achieving less than 1µA shutdown current. The disconnect feature also blocks unwanted current flow between V_{IN} and V_{OUT}, eliminating the large in-rush currents during startup that are inherent to most boost converters.

The internal oscillator, programmed by a resistor from R_T to GND, generates four internal clock pulses, each phase shifted by 90°. The switching frequency can be set from 100kHz to as high as 2MHz per phase, for an effective frequency of 8MHz as seen at the output filter cap. Maximum duty cycle for each phase is set to 90%. A sync input and oscillator output are provided for synchronizing the converter to a system clock, or synchronizing two converters together. Note that the sync input and clock output are at four times the switching frequency of each phase.

In Burst Mode® operation, only phase A is active, reducing switching and quiescent losses for maximum efficiency. In this mode, phase A operates with a fixed peak inductor current of 0.6A typical. Drawing just 12µA of quiescent current in Burst Mode operation allows the LTC3425 to operate with high efficiency during very light load conditions.

2-Cell to 3.3V/2.2A Boost Application, with Automatic Burst Mode Operation

Figure 2 shows a typical application circuit using the LTC3425 to boost from two NiCd or NiMH cells to 3.3V. This design can supply over 2A of load current with efficiencies up to 94% while switching at 1MHz per phase (4MHz output ripple frequency). Maximum component height is a slim 2.05mm. High efficiency is maintained over a very wide load range, as shown in Figure 3.

A key feature of the LTC3425 is the programmable automatic Burst Mode operation, which allows the user to set the load current where the converter enters Burst Mode operation, extending the efficiency at light load. This is ideal for systems where the mode cannot be controlled manually by the host. Since the Burst Mode circuit monitors average output current (rather than peak inductor current), the mode threshold is not affected by input

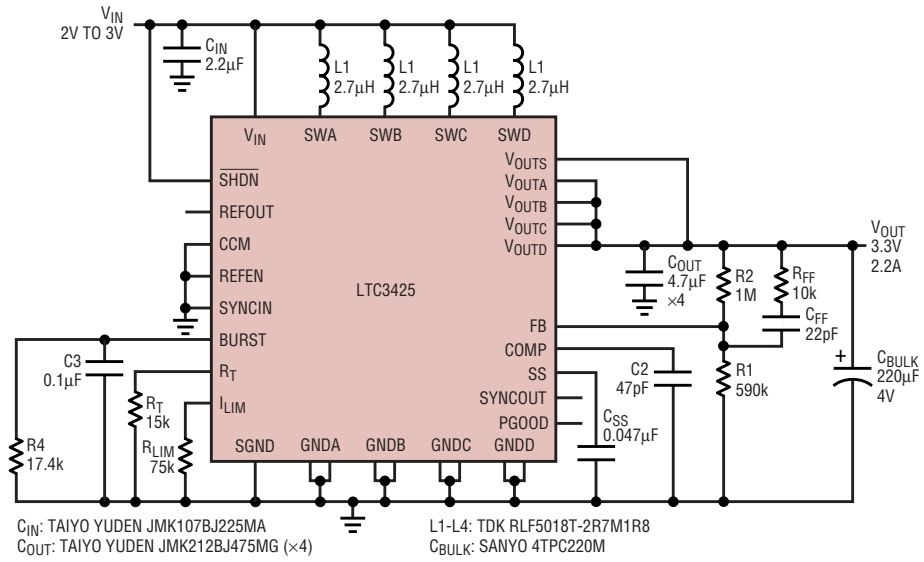

Figure 2. 2-cell to 3.3V boost application

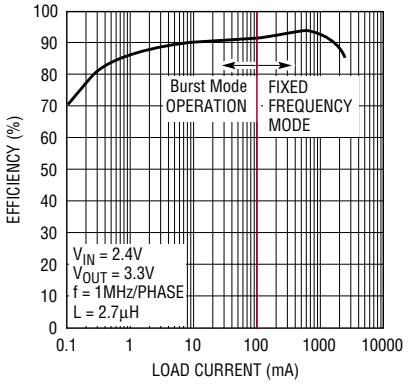

Figure 3. Efficiency vs load of 3.3V boost, using automatic Burst Mode operation

voltage variation. In this example, the Burst Mode threshold is set by R4 to 100mA. When the average load current drops below 100mA, the part enters Burst Mode operation, when the load current increases again, it leaves Burst Mode operation and returns to fixed frequency operation. Capacitor C3 filters the switching ripple at the Burst pin.

Because of the bulk capacitor on V_{OUT} in this example, only a single compensation capacitor is required. The feed-forward network, consisting of R_{FF} and C_{FF}, reduces output ripple in Burst Mode operation and further improves transient response during load steps. It also lowers the high frequency impedance at the FB pin, allowing the use of large value feedback resistors for maximum light load efficiency.

3.3V/Li-Ion to 5V/2.4A Boost Application with Active Clamp

Figure 4 shows the LTC3425 in a 5V boost application. This circuit can deliver 5V at 2.4A from a single Li-Ion cell, or from a 3.3V supply. That's 12W of output power in a 475mm^2 (0.74in^2) footprint with a component height of only 2.5mm. As Figure 5 shows, the efficiency peaks at 95%. Output ripple

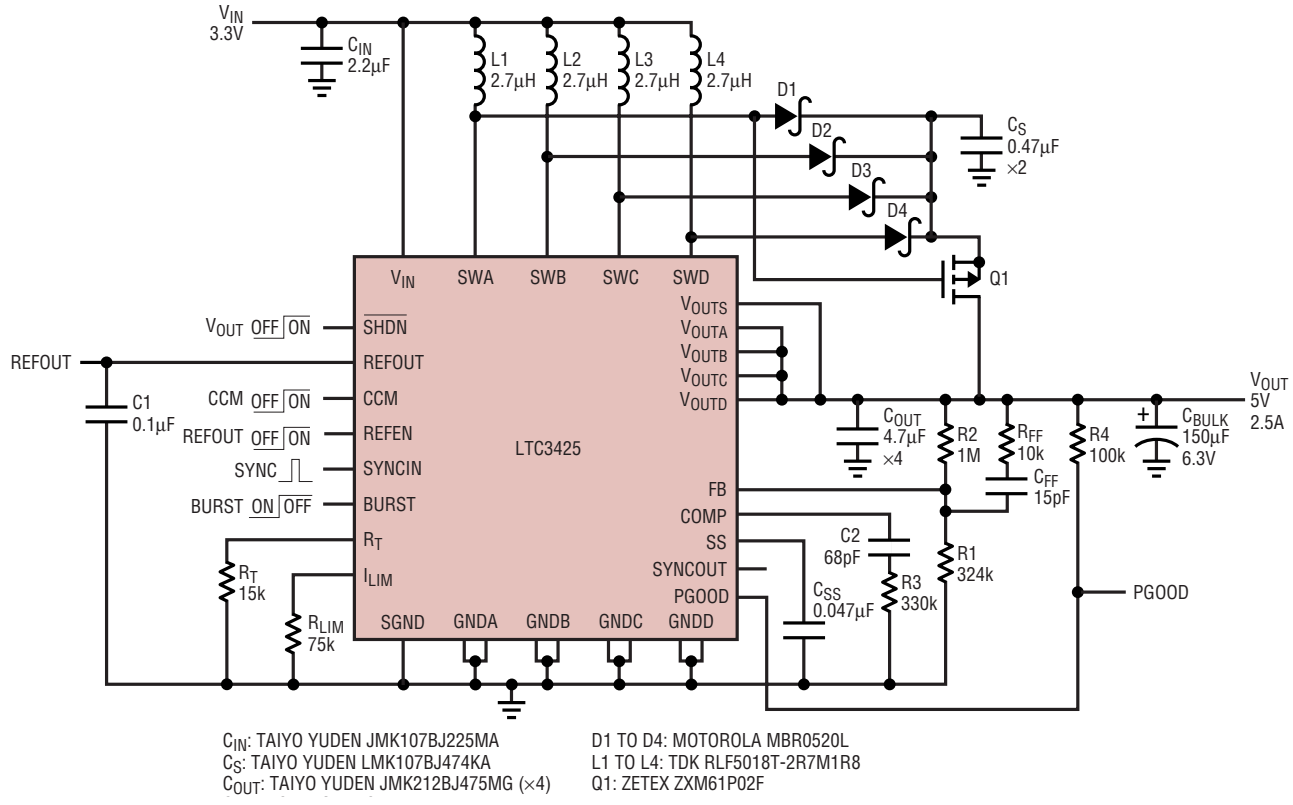

Figure 4. Li-Ion/3.3V to 5V boost application with active clamp

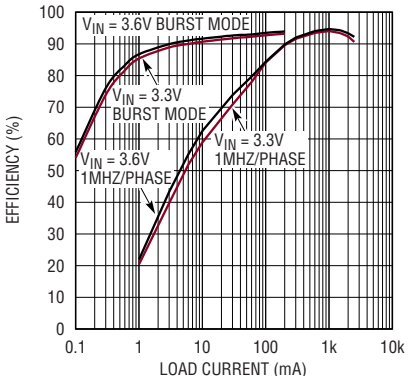

Figure 5. Efficiency vs load of 5V boost

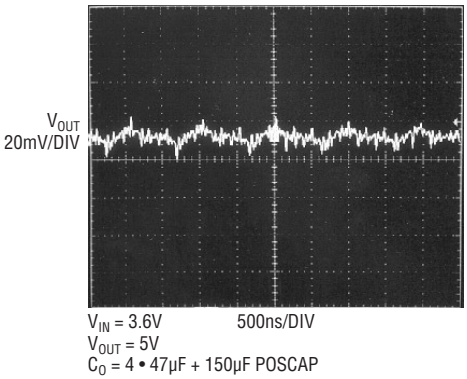

Figure 6. Li-Ion to 5V output voltage ripple at 2.5A load

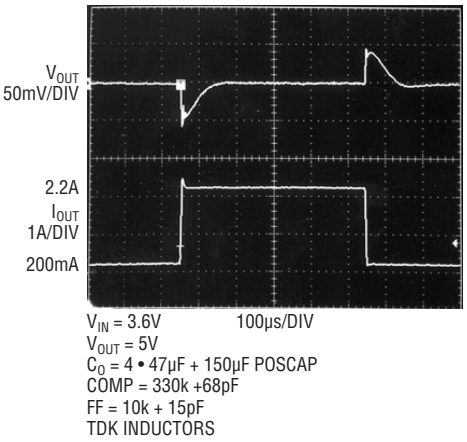

Figure 7. Step response for a 2A load step

at full load, shown in Figure 6, is less than 20mV$_{P-P}$.

In this application, Schottky diodes are used as part of an active clamp to limit the peak voltage seen at the switch nodes during the anti-cross-conduction time between the turn-on and turn-off of the internal NMOS and PMOS switches. The use of the external SOT-23 P-channel MOSFET (Q1) and 0.47µF capacitors (C_S) preserves the output disconnect feature of the LTC3425, allowing V_{OUT} to go to 0V in shutdown and limits the inrush current. If output disconnect is not required, Q1 and C_S can be eliminated, and the Schottky diodes can be tied directly from SW to V_{OUT}.

This circuit also illustrates the features and flexibility of the LTC3425. There is a 1.22V, short circuit protected reference output that can be turned on or off (for higher efficiency at very light loads), a sync input for synchronizing the internal oscillator to an external clock, and an open-drain Power Good output that monitors the output voltage.

The CCM input allows the user to force continuous conduction mode, which eliminates pulse skipping at light loads for noise sensitive applications. When CCM is pulled high, the synchronous rectifier stays on until a reverse inductor current of about 0.6A is sensed. Note that this lowers the efficiency at light load, and should only be used during fixed frequency mode operation.

In this example, the BURST pin is used to manually command either fixed frequency or Burst Mode operation. This can improve the transient response by bringing the converter out of Burst Mode operation prior to a large load step. A scope photo of the output step response, while operating in fixed frequency mode, is shown in Figure 7.

Low Cost, Very Low Profile 5W Boost Application Using All Ceramic Caps

Many portable applications have strict limitations on component height. This can be a challenge for a power converter, since the inductor and filter capacitors are usually among the tallest components. The LTC3425's 4-phase architecture is ideal for these applications. An example of a two cell to 3.3V/1.6A boost converter with a component height of only 1.55mm is shown in Figure 8.

In this design, the only output filter capacitors needed are the four 0805

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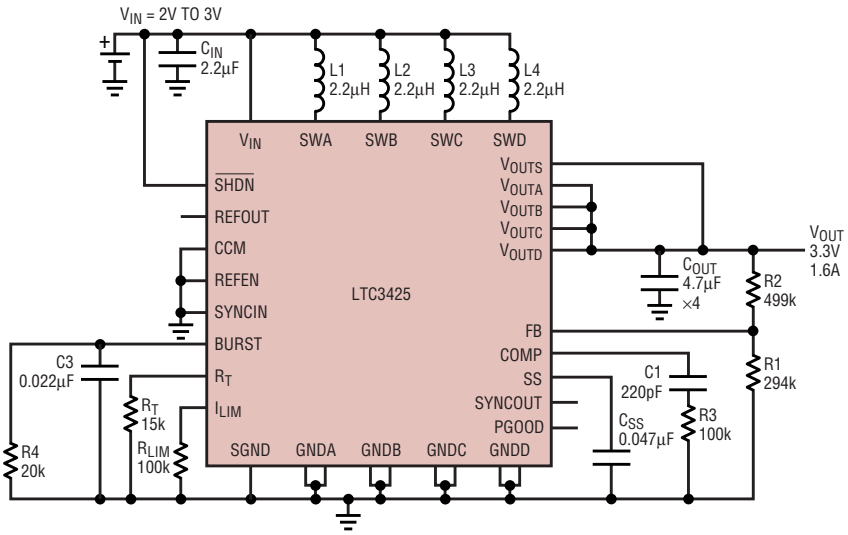


Figure 8. Low cost, 1.55mm profile 3.3V boost example

Figure 9. 3.3V boost demo (circuit shown in Figure 8)

CompactPCI Hot Swap Controller with I²C Interface, Bus Precharge and On-Chip LOCAL_PCI_RESET# Logic

by Victor Fleury

Introduction

When a board is plugged into the live backplane of a host system (hot swapped), the bulk bypass capacitors of the board can draw large inrush currents as they charge. These transient currents can damage connectors or create glitches on the backplane, potentially causing other boards in the system to inadvertently reset. To prevent such large inrush currents, the bulk bypass capacitors on the plug-in board must be isolated during the Hot Swap sequence.

The LTC4240 provides a controlled on-off switch for four hot swappable board power supply voltages, allowing the board to be safely inserted or removed from a live CompactPCI (CPCI) slot without disturbing the system power supplies. The LTC4240 includes an I²C-compatible interface that allows software control and monitoring of device function and power supply status.

Hot Swap features include:

- ❑ PRECHARGE output for biasing I/O connector pins during board insertion and extraction
- ❑ Circuit breakers on all four supplies with 35µs overcurrent glitch filters
- ❑ Foldback current limit to reduce power dissipation while charging large capacitive loads and during short circuit conditions
- ❑ Supports backplanes with and without bypass capacitors

I²C read and write functions include:

- ❑ Under a fault condition, determine which supply created the fault
- ❑ Read the maximum allowed board power consumption: PRSNT1#, PRSNT2#
- ❑ Cycle board power, reset the board after a fault condition
- ❑ Ignore faults on the +12V and –12V supplies

Typical Hot Swap Application

Figure 1 shows a CPCI Hot Swap application. Transistors Q1 and Q2 isolate 3.3V and 5V backplane power supplies from the plug-in board's bulk capacitance. The currents through Q1 and Q2 are sensed by R1 and R2. Resistors R3 and R4 prevent high frequency oscillations in Q1 and Q2. R5 and C1 stabilize the 3.3V and 5V current limit loop. During a fault condition, R5 also serves to isolate C1 from the fast internal pull down resistor. Capacitors C7 and C8 are 0.01µF, per the CPCI

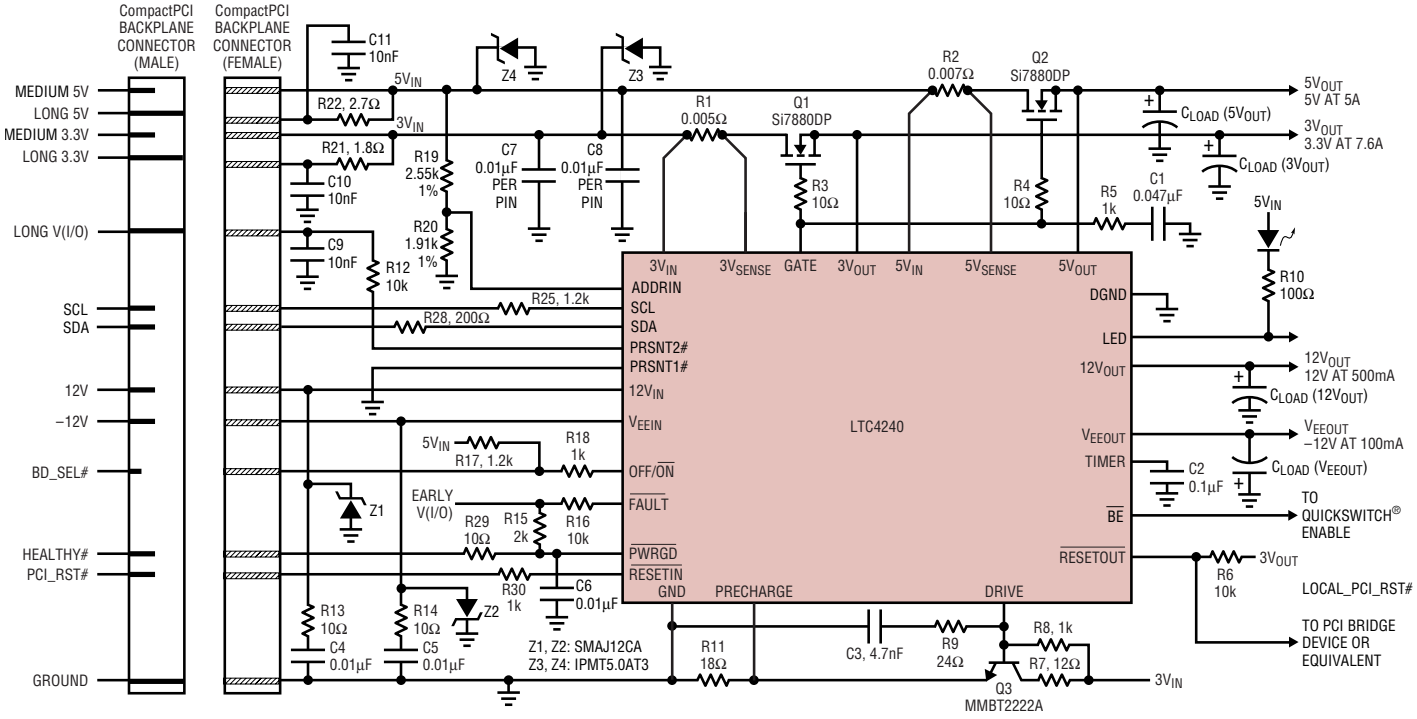


Figure 1. Typical CompactPCI application

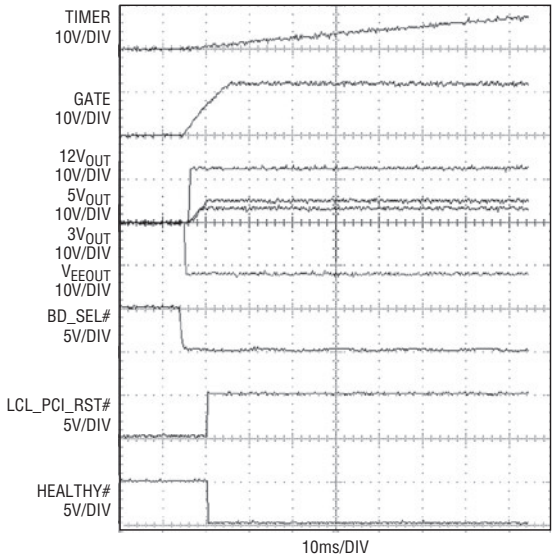


Figure 2. Typical power-up sequence

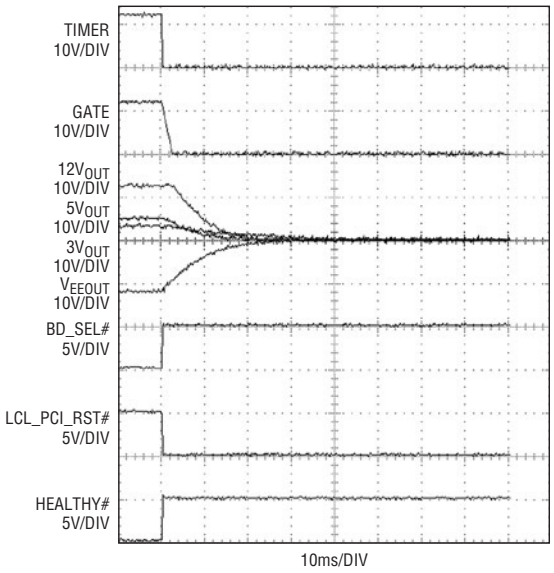


Figure 3. Power-down sequence

Hot Swap specification. On-chip power transistors isolate the –12V and +12V supplies. Transistor Q3 and its associated components form the pre-charge circuit.

CompactPCI Connection Pin Sequence

The staggered lengths of the CPCI male connector pins ensure that all power supplies are physically connected before back-end power is allowed to ramp (BD_SEL# asserted low). The long pins, which include 5V, 3.3V, V(I/O) and GND, mate first. The short pins, which includes BD_SEL# (OFF/$\overline{ON}$), mate last. The 3.3V and 5V long pins must be connected to the LTC4240 in order for the 1V PRECHARGE voltage to be available during early power. The following is a typical hot-plug sequence:

- ❑ ESD clips make contact.
- ❑ Long power and ground pins make contact and the 1V PRECHARGE becomes valid. Power is applied to the pull-up resistors connected to $\overline{FAULT}$, $\overline{PWRGD}$, and OFF/$\overline{ON}$ pins. The status LED is lit, indicating that the plug-in board is in the process of being connected (LOCAL_PCI_RST# is asserted). All power switches are off.
- ❑ Medium length pins make contact. There are six 5V and eight 3.3V medium length pins, bringing the 5V total to eight pins and the 3.3V total to ten pins. The CPCI specification limits the DC current to 1A/pin. The I^2C latch is initialized to allow seamless

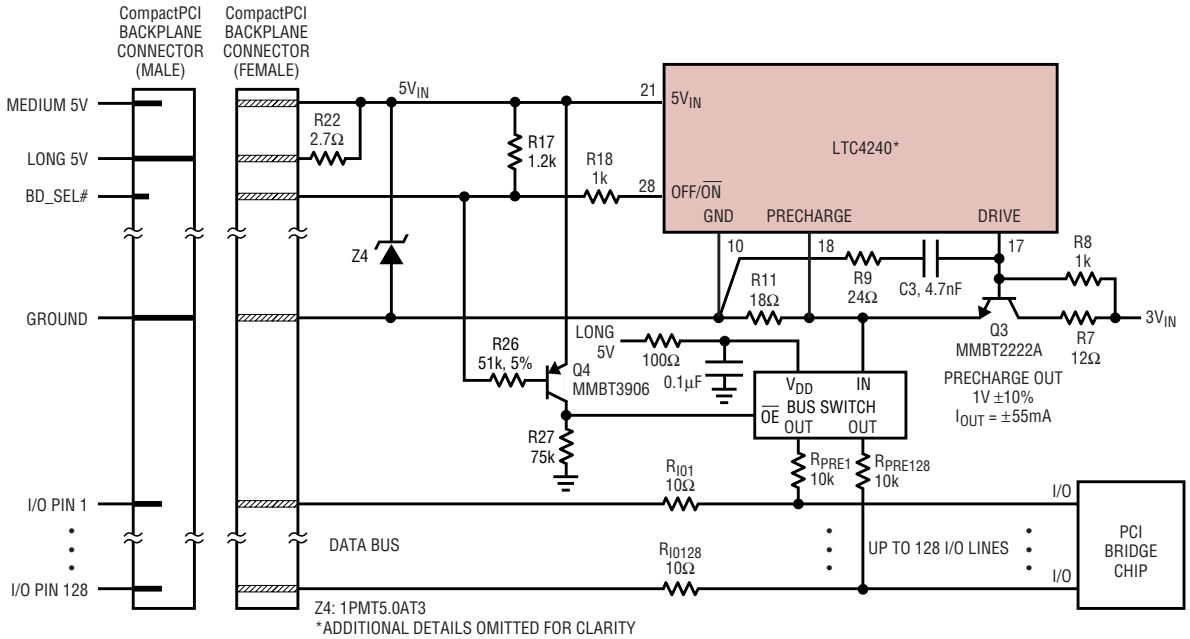


Figure 4. PRECHARGE bus switch application circuit for 3.3V and universal Hot Swap boards

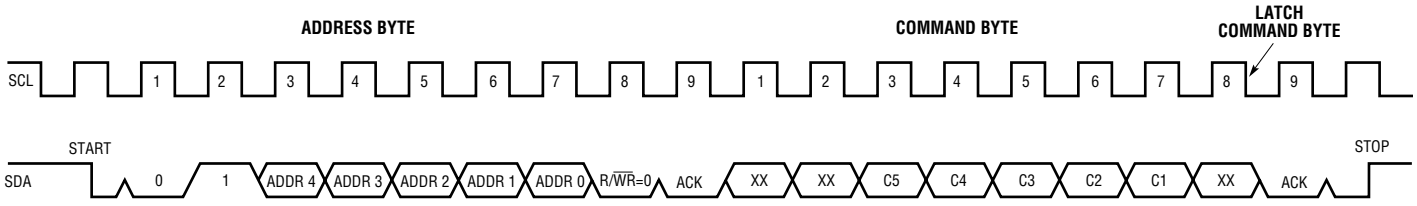
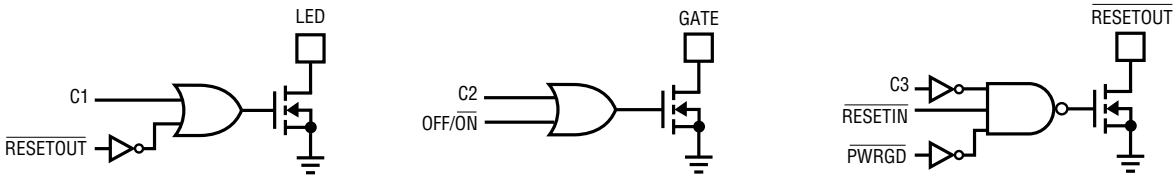


Figure 5. Send byte timing diagram



C1 TURNS ON THE EXTERNAL STATUS LED INDEPENDENT OF RESETOUT.

C2 PULLS DOWN THE GATE OF THE EXTERNAL N-CHANNEL SWITCHES. IT ALSO TURNS OFF THE $12V_{IN}$ AND V_{EEIN} INTERNAL POWER SWITCHES.

C3 IS USED TO SET LOCAL_PCI_RST# (RESETOUT).

Figure 6. Send byte command byte logic.

CPCI Hot Swap operation.

The +12V and –12V supply pins make contact at this stage. Zener clamps Z1 and Z2 plus shunt RC snubbers R13-C4 and R14-C5 help protect the +12V and –12V supply inputs, respectively, from large transient voltages during hot insertion and short circuit conditions. The signal pins also connect at this point. This includes the HEALTHY# signal connecting to the PWRGD pin, and the PCI_RST# signal connecting to the RESETIN pin.

- Short pins make contact last. BD_SEL# signal connects to the OFF/ON pin, thus starting the electrical connection process. If the BD_SEL# signal is grounded on the backplane, the electrical connection process begins immediately. The electrical connection process can be interrupted at any time via the I²C serial interface.

Power-Up Sequence

Figure 2 shows a typical power-up timing sequence. The connection sequence is triggered by a high to low transition on the BD_SEL# signal or by a power cycling executed by the I²C interface. A 65µA current source charges the gate nodes of the external power transistors. The power-up voltage rate of the 3VOUT and 5VOUT is approximately given by: dV/dt = 65µA/C1 or as determined by the current limit and the load capacitances.

Concurrently, an 11.5µA current source charges up the TIMER pin capacitance. Current limit faults are ignored until the voltage at the TIMER pin reaches 5.5V. Once all output supply voltages have crossed their power good thresholds, the HEALTHY# signal is pulled low (green LED turns on) and LOCAL_PCI_RST# is free to follow PCI_RST# and bit C3 of the I²C command latch.

Table 1. Send byte definition

Bit	HIGH	LOW
C5	Ignore V_{EEOUT} Faults	Does not Ignore V_{EEOUT} Faults
C4	Ignore $12V_{OUT}$ Faults	Does not Ignore $12V_{OUT}$ Faults
C3	Sets RESETOUT Low	Does not Set RESETOUT Low
C2	Turns off all switches Overrides OFF/ON Pin	Does not turn off all switches Does not override OFF/ON pin
C1	Turns on LED	Does not turn on LED

Controlled Turn-Off Allows Safe Extraction

Figure 3 shows a typical power-down timing sequence. When either BD_SEL# or bit C2 of the I²C command latch is set high, a 200µA current source discharges the capacitance on the gates of the external FETs. The internal +12V and –12V power switches also turn off. The four power switches are turned off slowly to avoid glitching the power supplies. Internal resistors discharge the output load capacitors. Once the power-down sequence is complete, the status LED lights up and the CPCI card can then be safely removed from the slot.

Disconnecting PRECHARGE Resistors

Universal Hot Swap and 3.3V signaling boards use a 50k, or larger, resistor to precharge the I/O lines. Since leakage currents at the I/O lines can be as high as 10µA, a 10k biasing resistor is allowed, but must be disconnected during normal operation. Figure 4 shows an application circuit that connects the PRECHARGE voltage to the I/O lines during insertion, but disconnects the resistors once the BD_SEL# pin makes contact.

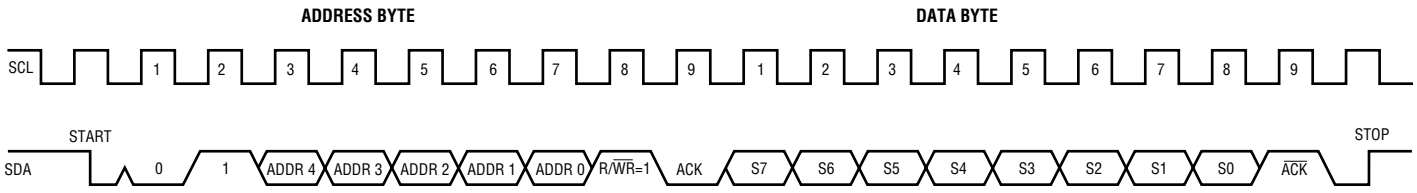


Figure 7. Receive byte timing

Table 2. Receive byte definition

S7	Logic State of the PRSNT2# Pin
S6	Logic State of the PRSNT1# Pin
S5	Logic State of the $\overline{\text{PWRGD}}$ Pin
S4	Logic State of the $\overline{\text{RESETOUT}}$ Pin
S3	Logic State of the $\overline{\text{RESETIN}}$ Pin
S2	FAULTCODE1 (See Table 3)
S1	FAULTCODE0 (See Table 3)
S0	Logic State of the $\overline{\text{FAULT}}$ pin

Table 3. Supply causing fault

FAULTCODE0	FAULTCODE1	$\overline{\text{FAULT}}$	Supply Causing Fault
LOW	LOW	LOW	$3V_{IN}$
LOW	HIGH	LOW	$5V_{IN}$
HIGH	LOW	LOW	$12V_{IN}$
HIGH	HIGH	LOW	V_{EEIN}
X	X	HIGH	None

Control and Monitor Card Power with I²C Interface

The LTC4240 incorporates an I²C compatible 2-wire (SCL, SDA) interface that allows the user to easily query and control the status of the LTC4240. A single analog pin selects 1 of 32 allowed addresses. The LTC4240 supports send byte and receive byte commands. Figure 5 and Table 1 depict the timing and bit definition of the send byte command. Figure 6 schematically outlines some of the command bit functions. Figure 7 shows the timing of the receive byte command. Tables 2 and 3 define the data byte. If a fault occurs, the FAULTCODE bits can be used to determine which supply generated the fault.

Conclusion

The LTC4240 provides a comprehensive solution to CompactPCI Hot Swap applications. An integrated I²C-compatible interface allows software control and monitoring of device function and power supply status. The LTC4240 control functions allow the plug-in board to be safely inserted or removed from a live CompactPCI slot without disturbing the system power supplies or I/O lines.

LTC3425, continued from page 5

size, 4.7µF ceramics, with a height of 1.35mm. Output voltage ripple is under $50mV_{P-P}$ at full load. The four low-cost inductors are only 1.55mm high, with a 3.2mm by 2.5mm footprint. The entire 5W power converter can fit into a 20mm by 16mm space, as seen in Figure 9.

2- or 3-Phase Operation

For cost-sensitive applications or for reduced board area with lower maximum current capability, the LTC3425 can be used as a 2- or 3-phase converter by simply de-populating one or two of the inductors. Figure 10 illustrates the typical efficiency difference between 2-, 3- and 4-phase operation. In Burst Mode, there is no efficiency penalty, since only phase A is used.

Conclusion: Good Things Do Come in Small Packages

The examples here illustrate the performance, flexibility, small size and ease-of-use of the LTC3425. The synchronous 4-phase architecture achieves high efficiency over a wide range of loads while enabling the use of low-profile components. The four-to-one reduction in output ripple current makes it possible to achieve very low output voltage ripple using small, lower cost ceramic capacitors. Users can choose between automatic or manual Burst Mode operation, pulse skipping mode or forced continuous conduction mode for noise sensitive applications. All these features, along with output disconnect, soft-start, 1µA shutdown current, anti-ringing control, thermal shutdown, a buffered reference output and a Power Good output are packed in a small 5mm by 5mm, thermally enhanced QFN package.

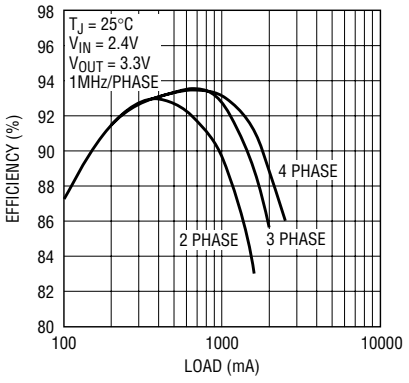


Figure 10. Typical efficiency with 2, 3 and 4 phases (fixed frequency mode)

Low Noise, Micropower Precision Op Amp Swings Outputs from Rail to Rail

by Kris Lokere and Glen Brisebois

Introduction

Applications that measure temperature, location or light using thermocouples, hall-effect sensors, or precision photodiodes can benefit from an op amp with offset voltage of less than 100µV, an input bias current in the picoamps, and thermal drift of less than 1µV/°C. Op amps that meet these stringent requirements are available, and have been for some time, but they tend to come in relatively large packages, do not work with low supply voltages, and sometimes consume milliamps of supply current.

The LT6011 dual precision op amp fits into a 3×3mm^2 DFN package, which is so small it doesn't even have leads. The LT6011 also offers low voltage, micro-power operation. A quad version, the LT6012, is also available in the SO-14 and GN16 packages.

Low voltage operation is useless if the outputs of the op amp can't swing nearly from rail to rail. Many older op amps clip if the output is driven closer than 1V from either V_{CC} or ground. This is not a problem in systems with split ±15V supplies, but in a system with a single 2.7V supply (the minimum for the LT6011), the requirement to stay 1V away from either rail leaves less than 1V for the sensor signal, severely reducing the dynamic range. The LT6011 allows the outputs to swing to 40mV from either supply rail, making it practical in low supply voltage applications.

In portable instrumentation, medical applications, or in sophisticated systems that measure hundreds or even thousands of variables simultaneously, the power consumption of the precision op amp is important. This is especially so because designers usually want to burn as much of the available power in the sensor itself, since this tends to reduce noise. The LT6011 suits all these applications because the supply current is less than 150µA per amplifier—instead of the milliamps of other precision amplifiers. In addition, micropower operation has the obvious benefits of increasing battery run time and reducing system heat dissipation, thus simplifying system design and improving the system reliability.

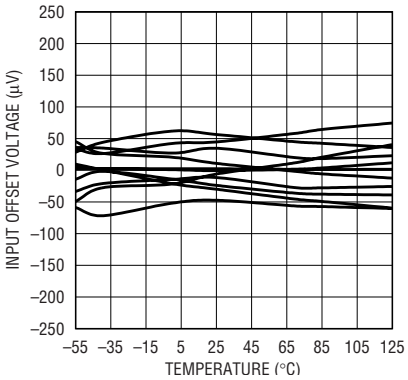


Figure 1. The LT6011 typical input offset drift of 0.2µV/°C is close to V_{OS}/298K, the theoretical minimum for a bipolar differential pair with resistive load. Performance of a representative sample of amplifiers shown.

Performance

Table 1 shows the precision specifications for the LT6011. While input offset voltage is an obvious measure of an amplifier's precision, other specifications can affect the overall precision of the application, and thus, should be considered when choosing a precision op amp. The LT6011 is carefully designed so that its low input offset voltage is not corrupted by noise, offset

Table 1: LT6011 specifications that impact precision operation

Parameter	LT6011A/LT6012A		LT6011/LT6012	
Available Packages	S8/S14	DFN/GN16	S8/S14	DFN/GN16
Input Offset Voltage (max)	60µV	85µV	75µV	125µV
Input Offset Drift (max)	0.8µV/°C	1.3µV/°C	0.8µV/°C	1.3µV/°C
Input Bias Current (max)	300pA		900pA	
Input Noise Voltage, 0.1Hz to 10Hz	$0.4\mu V_{P-P}$			
Input Noise Voltage Density (1kHz)	$14nV/\sqrt{Hz}$			
Common-Mode Rejection Ratio (min) V_{CM} = 1V to 3.8V , V_S = 5V	107dB			
Power Supply Rejection Ratio (min) V_S = ±1.35V to ±18V	112dB			
Open-Loop Voltage Gain (min) V_{OUT} = ±13.5V, V_S = ±15V, 10k load	1000V/mV			

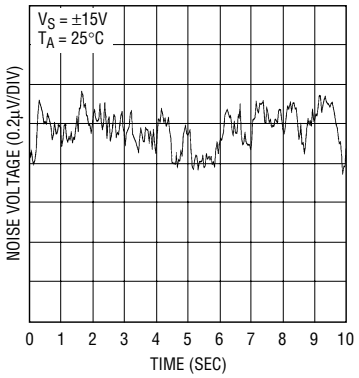


Figure 2. As a result of low 1/f noise, the total LT6011 input noise is only 0.4µV_{P-P} in the 0.1Hz to 10Hz frequency band.

current, temperature drift or common-mode voltage.

The LT6011 typical input offset drift of 0.2µV/°C is close to V_{OS}/298°K, the theoretical minimum for a bipolar differential pair with ideal resistive loads. In order to achieve this low drift, all internal base currents must be balanced. In addition, the LT6011 superbeta input devices are particularly insensitive to packaging stresses. Figure 1 shows V_{OS} drift for a representative sample of LT6011 amplifiers.

Whenever microvolt levels of DC input precision are required, low frequency noise can corrupt the readings. At low frequencies, total noise is often dominated by process-dependent 1/f noise, which is inadequately captured by looking at the higher frequency noise density spec. In the 0.1Hz to 10Hz frequency band, total LT6011 input noise is only 0.4µV_{P-P} (Figure 2). Higher frequency noise, to the extent that the application does not filter it out, can be calculated from the 14nV/√Hz white-noise density.

Input bias current can easily be as important to precision as offset voltage, especially when using higher impedance sensors, or when large feedback resistors are needed to maintain low power. For a 10k total source impedance, the 300pA maximum input bias current of the LT6011 causes only 3µV of error. The LT6011 features internal base current cancellation, which makes the positive and negative input bias current uncorrelated. It is therefore not necessary to try to balance the input impedances.

To get a better appreciation of how low a 300pA current is, consider that sloppy board design can easily generate leakage current much larger than that. For example, if an input trace at 0V (on a PCB) would run next to a supply trace of 15V, then even 10GΩ of parasitic resistance would cause an extra 1500pA of input current.

Finally, in order to maintain input accuracy over operating conditions, you must consider the effects of common-mode voltage, power supply voltage, and output swing. Divide any changes in the supply voltage by the PSRR to see how much the input offset changes. Similarly, divide the changes in the input common-mode voltage by the CMRR. At a 5V supply, the 107dB CMRR spec translates a 2.8V ΔV_{CM} to 12.5µV of offset change (worst case over temperature). When the output of the op amp swings, the gain-induced input error is calculated from the open-loop gain spec. For the LT6011, this translates to 12µV worst-case over temperature (3V swing at 5V supply with 2k load).

Applications

Hall Sensor Amplifier

The circuit of Figure 3 shows the LT6011 applied as a low power Hall sensor amplifier. The magnetic sensitivity of a Hall sensor is proportional to the DC bias voltage applied to it. With a 1V bias voltage, the sensitivity of this Hall sensor is specified as 4mV/mT of magnetic field. At that level of DC bias, however, the 400Ω bridge consumes 2.5mA. Reducing the bias voltage would reduce the power consumption, but it would also reduce the sensitivity. This is where the beauty of precision micropower amplification becomes especially apparent. First, though, lets look at the operation of the circuit.

The LT1790-1.25 micropower reference provides a stable reference voltage. Resistive ladder 7.87k:100k attenuates this to about 90mV across the 7.87k, and the LT1782 acts as a buffer. When this 90mV is applied as bias across the Hall bridge, the current is only 230µA. This is less than 1/10 of the original value. (Just imagine if all your batteries could last 10 times longer than they do.)

But, as mentioned earlier, the sensitivity is now likewise reduced by the same factor, down to 0.4mV/mT. The way back to high output voltage is to take gain with a precision micropower amplifier. The LT6011 is configured as a differential gain block in a gain of 100. Such high gains, and even higher, are permissible and advantageous using an LT6011 because of its exceptional precision and low input drift. The output sensitivity of the circuit is raised to a whopping 40mV/mT, with a total supply current budget of about 600µA. (Here in Milpitas, California, the Earth's 50µT field is about 60 degrees from horizontal, and causes a 2mV output shift in the circuit.)

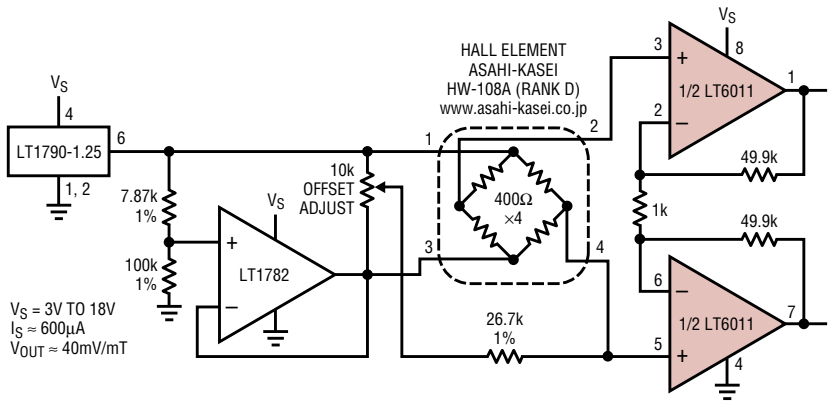


Figure 3. Amplifying the Hall Sensor voltage with a low-power precision amplifier allows you to burn less power in the sensor—reducing total current consumption by a factor of 4, while achieving 10 times the sensitivity.

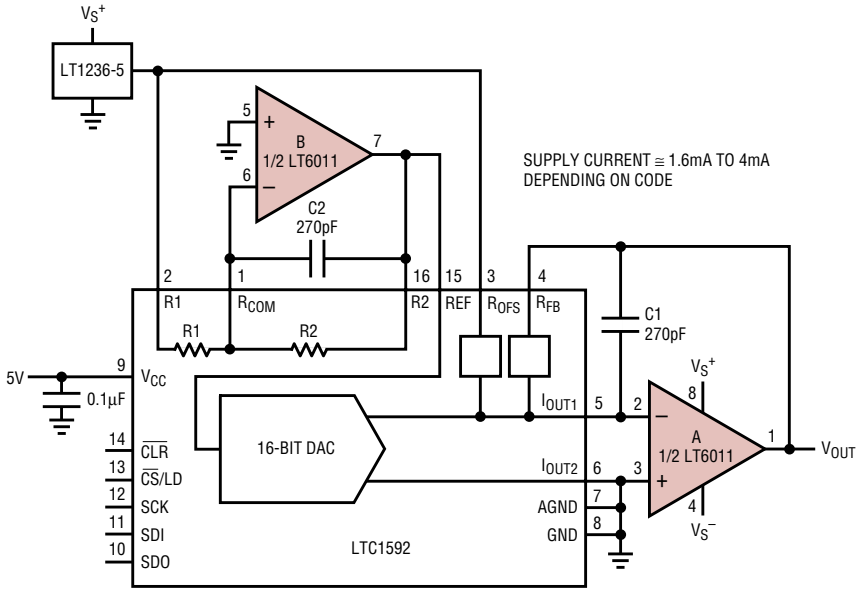


Figure 4. Functioning as reference inverter and I-to-V converter for this DAC, the LT6011 maintains 16-bit precision without adding much to the total supply current.

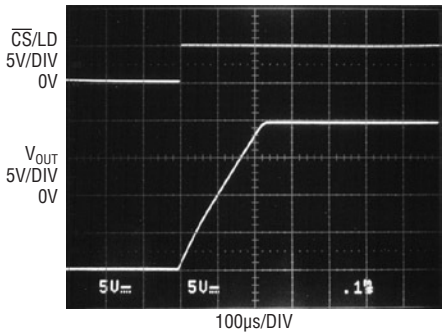


Figure 5. The output of the circuit in Figure 4 settles in less than 250µs.

DAC I-to-V Converter

Figure 4 shows the LT6011 applied as both a reference amplifier and I-to-V converter with the LTC1592 16-bit DAC. Whereas faster amplifiers such as the LT1881 and LT1469 are also suitable for use with this DAC, the LT6011 is desirable when power consumption is more important than speed. The total supply current of this application varies from 1.6mA to 4mA, depending on code, and is almost entirely dominated by the DAC resistors and the reference.

The DAC itself is powered only from a single 5V supply. Op amp B of the LT6011 inverts the 5V reference using the DAC's internal precision resistors R1 and R2, thus providing the DAC with a negative reference allowing bipolar output polarities. Op amp A provides the I-to-V conversion and buffers the final output voltage. The precision required of the I-to-V converter function is critical because the DAC output resistor network is obviously very code dependent, so the noise gain that the op amp sees is also code dependent. An imprecise op amp in this function would have its input errors amplified almost chaotically versus code.

Since the outputs of the LT6011 swing to within 40mV of either supply rail, the supply voltages to the amplifier need to be only barely wider than the desired ±5V DAC outputs.

The large signal time domain response of the circuit is shown in Figure 5.

How it Works

The simplified schematic in Figure 6 shows how the op amp achieves its precision input performance and rail-to-rail output capabilities.

The overall architecture features three gain stages, providing very high open loop voltage gain of 1MV/V.

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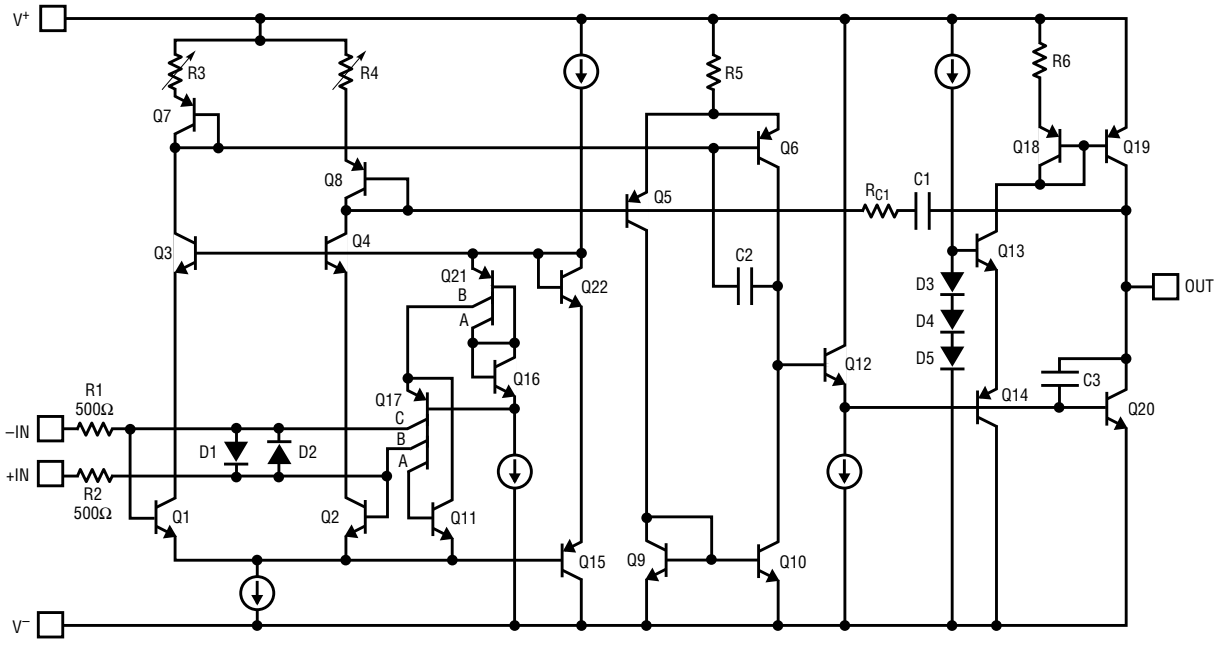


Figure 6. This simplified schematic shows how the LT6011 achieves its precision input performance and rail-to-rail output capabilities.

Addressable Bus Buffer Provides Capacitance Buffering, Live Insertion and Nested Addressing in 2-WireBus Systems

by John Ziegler

Introduction

The reliability of data processing, data storage and communications systems depends in part on how well the system is monitored. To this end, input/output (I/O) cards contain circuitry to monitor parameters such as temperature, fan speed and system voltages. These circuits often communicate through 2-wire serial buses, such as SMBus or I^2C.

Several practical problems can occur in these systems, especially as they become large. First, data bus rise time specifications become difficult to meet, as the SDA (serial data) and SCL (serial clock) capacitances of each I/O card add directly to those of the backplane, while the pull-up impedance remains constant. Second, cycling power whenever a new I/O card is installed is not an option in many uninterruptible systems. Third, a device's address is often dictated by the function that it performs. If an existing system already contains a temperature sensor, for example, then inserting a new I/O card with a temperature sensor into the system could result in multiple devices having the same address. Finally, the likelihood of a device becoming confused and holding the bus lines low increases as the number of devices on the bus increases.

The new LTC4302 addressable 2-wire bus buffer addresses all of these problems. The LTC4302 provides a bridge between any two physically separate 2-wire buses. SDAIN and SCLIN connect to one 2-wire bus, e.g., a backplane; and SDAOUT and SCLOUT connect to a second 2-wire bus, e.g. an I/O card. Masters on the buses can command the LTC4302 to connect and disconnect the backplane and card buses through software. By using multiple LTC4302s in a system and connecting and disconnecting them appropriately, users can assign the same address to multiple devices in the system. If a large bus becomes stuck low, the CONN pin of the LTC4302 can be used to isolate portions of the bus, helping the master locate the source of the problem quickly.

When it connects the backplane and card buses, the LTC4302 provides bidirectional buffering, keeping the backplane and card capacitances isolated from each other. Therefore, users can design several LTC4302s into a large system to break its large capacitance into several smaller pieces, while still passing the SDA and SCL signal to all devices at the same time. Finally, because the LTC4302's SDA and SCL pins default to a high impedance state even when no V_{CC} voltage is applied, an LTC4302 can be inserted onto a live 2-wire bus without corrupting it. Two general purpose input-outputs (GPIOs) are available on the LTC4302-1. The LTC4302-2 replaces one GPIO with a second supply voltage V_{CC2}, providing level shifting between systems with different supply voltages.

Circuit Operation

Startup

A block diagram for the LTC4302-1 is shown in Figure 1. During power-up, the LTC4302 starts in an undervoltage lockout (UVLO) state, ignoring any activity on the SDA and SCL pins until the V_{CC} voltage rises above 2.5V (typical). This ensures that the LTC4302 does not try to function until it has sufficient bias voltage. During this time, the 1V pre-charge circuitry is active and forces 1V through 100kΩ nominal resistors to the SDA and SCL pins. The 1V pre-charge minimizes the disturbance caused by the LTC4302's SDA and SCL pins in hot-swapping applications. When the LTC4302 is being inserted into a live backplane, the backplane SDA and SCL bus voltages may be anywhere from 0V to the bus pull-up supply voltage. Pre-charging the SDA and SCL pins to 1V minimizes the worst-case backplane-to-pin voltage differential at the moment of connection, thus minimizing the amount of disturbance on the backplane.

Once the LTC4302 comes out of undervoltage lockout, the pre-charge circuitry is shut off, and the 2-Wire Digital Interface circuitry (shown in Figure 1) is activated. The master on SDAIN and SCLIN can then address the LTC4302 and utilize its various features, which include the Backplane-to-Card Connection circuitry (also referred to below as "Connection Circuitry"), Rise Time Accelerators, and GPIO circuits. These features default to a high-impedance state: the Connection Circuitry and Rise Time Accelerators are inactive, and the GPIOs are in open-drain output mode with their pull-down devices turned off. The LTC4302 is in the default state whenever it is in UVLO or when the CONN voltage is low.

Connection Circuitry

When the connection circuitry is activated, the functionality of the SDAIN and SDAOUT pins is identical. A low forced on either pin at any time results in both pins being low. SDAIN and SDAOUT enter a logic high state only when all devices on both SDAIN and SDAOUT turn their pull-downs off. The same is true for SCLIN and SCLOUT.

This important feature ensures that clock stretching, clock arbitration and the acknowledge protocol always work, regardless of how the devices in the system are connected to the LTC4302.

Another key feature of the connection circuitry is that, while it joins the two buses together, it still maintains electrical isolation between them, thus providing capacitance buffering for both sides. This means that devices on the backplane must drive only the backplane capacitance plus the low capacitance of the LTC4302 (around 10pF). The LTC4302 drives the capacitance of the rest of the I/O card. Likewise, devices on the card must only drive the capacitance of the card plus the low capacitance of the LTC4302. The LTC4302 drives the capacitance on the backplane. The LTC4302 is capable of driving capacitive loads ranging from 0pF to 1000pF on all of its data and clock pins.

Rise Time Accelerators

Masters on the bus may activate rise time accelerators on the backplane side (SDAIN and SCLIN), the card side (SDAOUT and SCLOUT), neither or both. When activated, the accelerators switch in 2mA of pull-up current at V_{CC} = 2.7V and 9mA at V_{CC} = 5.5V during bus rising edges to quickly slew the SDA and SCL lines once their DC voltages exceed 0.6V and the initial rise rate on the pin exceeds 0.8V/µs. Figure 2 shows the rise time reduc-

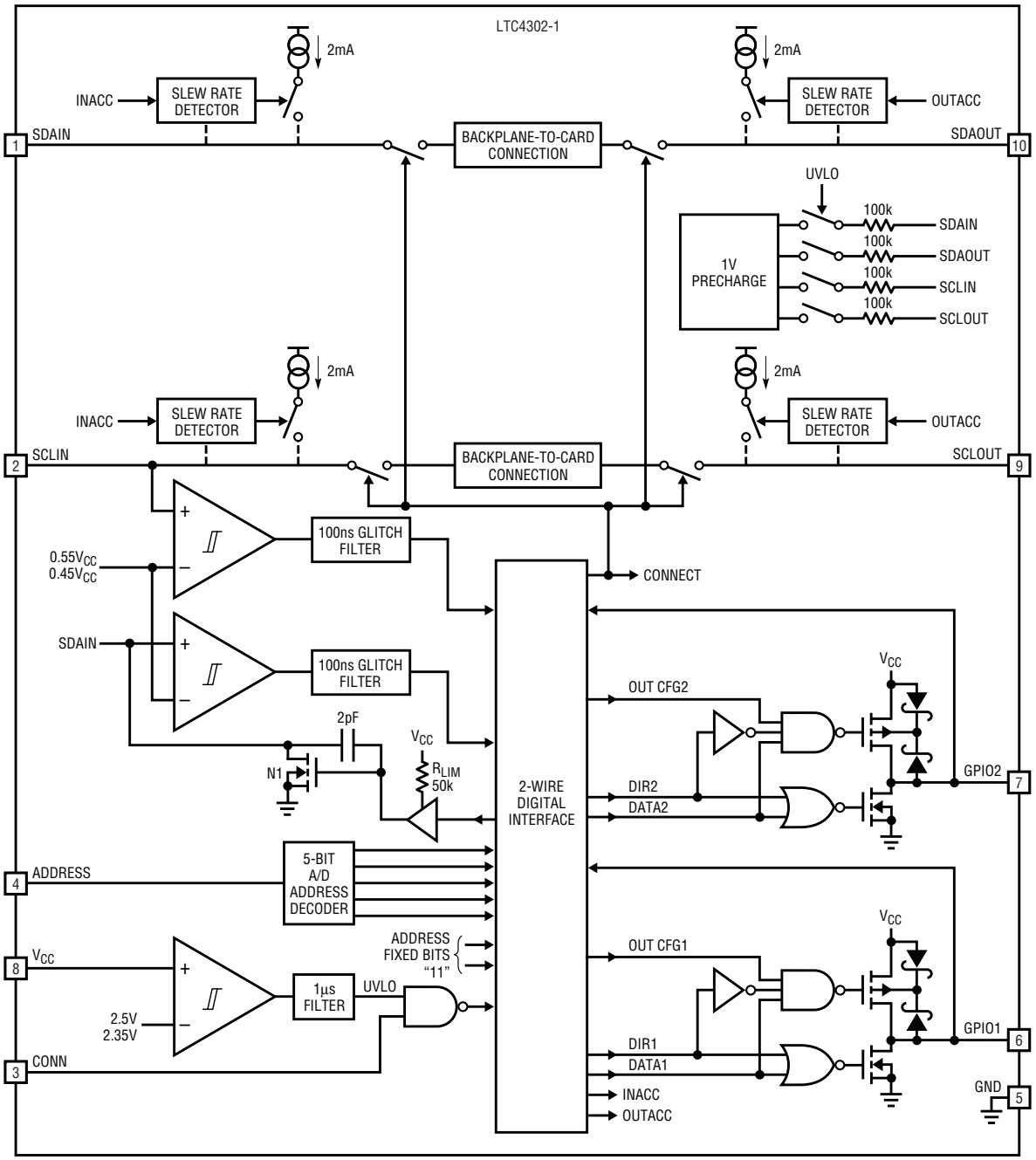
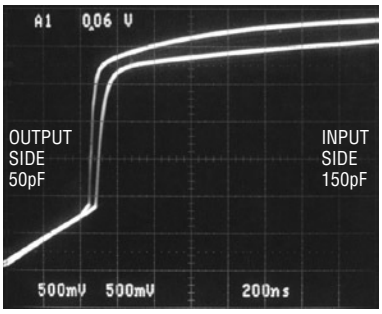


Figure 1. Addressable 2-wire bus buffer block diagram



200ns/DIV

Figure 2. Rise time accelerators reduce rise time for C_{BUS} = 50pF, C_{BUS} = 150pF.

tion achieved for V_{CC} = 3.3V and bus equivalent capacitances of 50pF and 150pF.

General Purpose Input/Outputs (GPIOs)

The LTC4302 -1 provides two GPIOs that can be configured as input, open-drain outputs, or push-pull outputs. In push-pull mode, at V_{CC} = 2.7V, the typical pull-up impedance is 670Ω and the typical pull-down impedance is 35Ω, making the GPIO pull-downs capable of driving LEDs. In open-drain output mode, the logic high is provided by connecting a pull-up resistor from the GPIO pin to an external supply voltage. This supply voltage can range from 2.2V to 5.5V, independent of the V_{CC} voltage. The LTC4302-2 replaces one GPIO with a second supply voltage pin V_{CC2} and therefore provides a single GPIO.

Single ADDRESS Pin Provides 32 Addresses

The LTC4302 saves valuable board space by providing 32 unique addresses from a single ADDRESS pin. A resistive divider between V_{CC} and ground sets an analog voltage on the ADDRESS pin. An internal A/D converter translates the ADDRESS voltage into a 5-bit digital word, which sets the five LSBs of the address. The two MSBs are hard-wired to "11."

CONN Reset Pin

Grounding the CONN pin resets the LTC4302 to its high-impedance default state: the output side is disconnected from the input side, the rise time accelerators on both sides are disabled, and the GPIOs are set in open-drain output mode with the NMOS open-drain pulldown turned off. Grounding the CONN pin also disables the 2-Wire Digital Interface circuitry, preventing masters on the bus from communicating with the LTC4302. When the CONN voltage is brought back high, the LTC4302 remains in its default state.

When an SDA or SCL line is stuck low, masters can use the CONN pins of the LTC4302s in the system to find the source of the problem. A master drives one CONN pin low at a time while monitoring the stuck bus. When the line returns high, the master then knows that the stuck device is on the other side of the last LTC4302 whose CONN pin was driven low.

Live Insertion and Removal, and Capacitance Buffering Application

The application shown in Figure 3 highlights the live insertion and removal, and the capacitance buffering features of the LTC4302. Assuming that a staggered connector is available, make ground and V_{CC} the longest pins to guarantee that SDAIN and SCLIN receive the 1V pre-charge voltage before they connect. Make SDAIN and SCLIN medium length pins to ensure that they are firmly connected while CONN is low. Make CONN the shortest pin and connect a weak resistor from CONN to ground on the I/O card. This ensures that the LTC4302 remains in a high impedance state while SDAIN and SCLIN are making connection during live insertion. During live removal, having CONN disconnect first ensures that the LTC4302 enters a high impedance state in a controlled manner before SDAIN and SCLIN disconnect.

Note that if the I/O card were plugged directly into the backplane, the card capacitance would add directly to the backplane capacitance, making rise and fall time requirements difficult to meet. Inserting a LTC4302 on the edge of the card, however, isolates the card capacitance from the backplane. The LTC4302 drives the capacitance of everything on the card, and the backplane must drive only the capacitance of the LTC4302, which is less than 10pF. As more I/O cards are added and the system grows, placing a LTC4302 on the edge of each

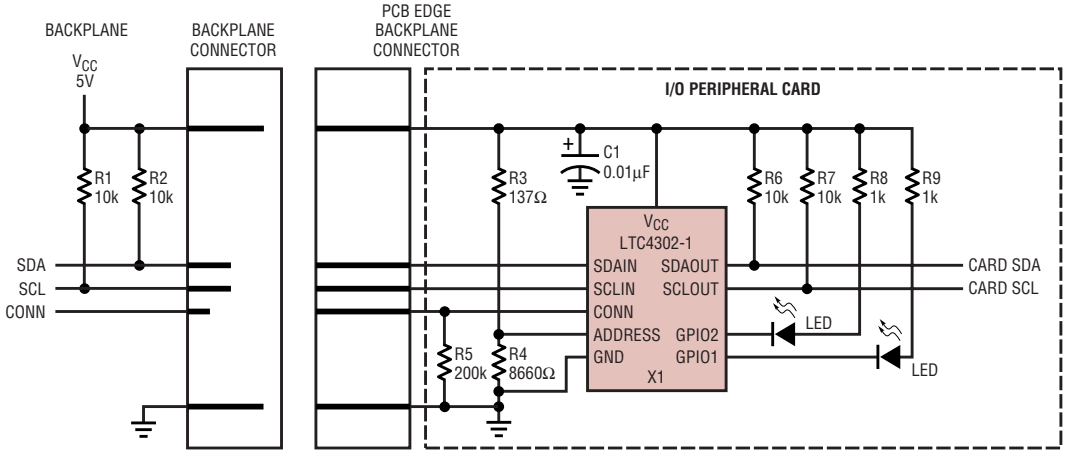


Figure 3. LTC4302-1 in a live insertion and capacitance buffering application

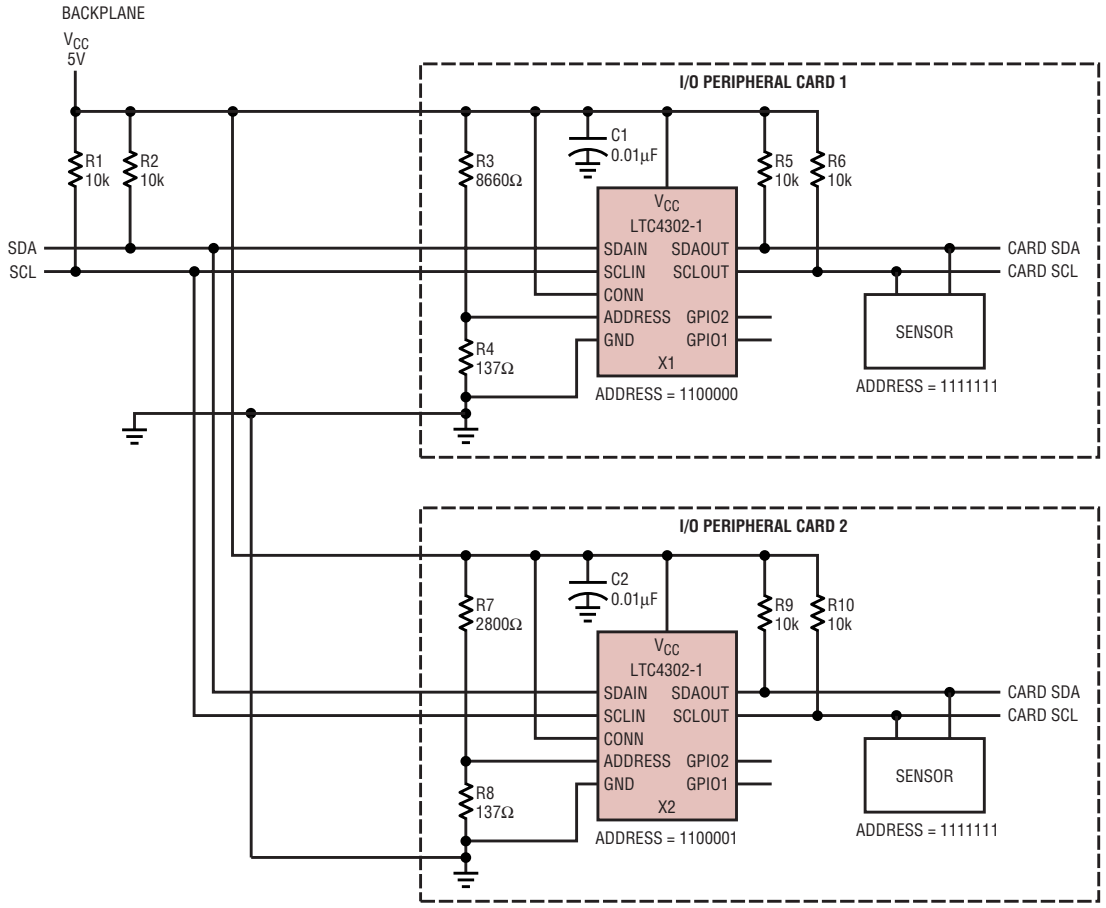


Figure 4. LTC4302-1 in a nested addressing application

card breaks what would be one large, unmanageable bus into several manageable segments, while still allowing all segments to be active at the same time. Moreover, the LTC4302's rise time accelerators provide strong pull-up currents during bus rising edges, so that even heavily loaded bus lines meet system rise time requirements with ease.

Address Expansion with Nested Addressing

Figure 4 illustrates how the LTC4302 can be used to expand the number of devices in a system by using nested addressing. Note that each I/O card contains a sensor device having address 1111111. If the two cards were plugged directly into the backplane, the two sensors would require two different addresses. However, each LTC4302 isolates the devices on its card from the rest of the system until it is commanded to connect. If masters use the LTC4302s to connect only one I/O card at a time, then each I/O card can have a device with address 1111111 and no problems occur.

5.5V to 3V Level Translator and Power Supply Redundancy (LTC4302-2)

Systems requiring different supply voltages for the backplane side and the card side can use the LTC4302-2 as shown in Figure 5. The pull-up resistors on the card side connect from SDAOUT and SCLOUT to V_{CC2}, and those on the backplane side connect from SDAIN and SCLIN to V_{CC}. The LTC4302-2 functions for voltages ranging from 2.7V to 5.5V on both V_{CC} and V_{CC2}. There is no constraint on the voltage magnitudes of V_{CC} and V_{CC2} with respect to each other.

This application also provides power supply redundancy. If either the V_{CC} or V_{CC2} supply voltage falls below its UVLO threshold, the LTC4302-2 disconnects the backplane from the card, so that the side that is still powered can continue to function.

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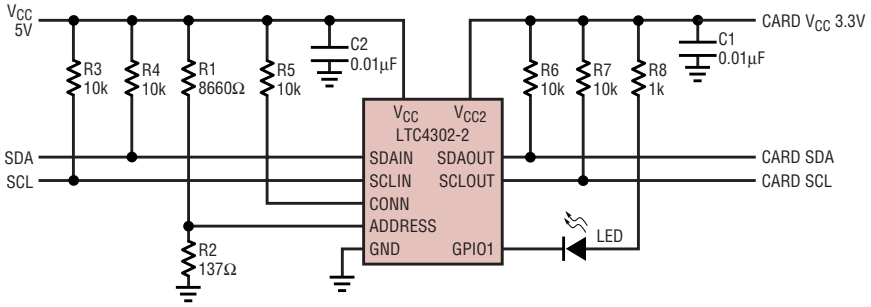


Figure 5. 5V to 3.3V level translator application

Versatile Hot Swap Controller with Open Circuit Detect, Foldback Current Limiting and Much More

by Mark Belch

Introduction

When a circuit board is inserted into a live backplane, the input capacitors on the board can draw high inrush currents from the backplane power bus as they charge. The inrush current can permanently damage the connector pins and board components as well as glitch the system supply, causing other boards in the system to reset. The new LT4254 provides a compact and robust solution to eliminate these hot plugging issues.

The LT4254 is designed to turn on a board's supply voltage in a controlled manner, allowing the board to be safely inserted or removed from a live backplane having a supply voltage from 10.8V to 36V. The device features programmable inrush current control, current foldback, programmable 1% tolerance on the undervoltage and overvoltage thresholds, overcurrent protection, and a power good output signal that indicates when the output supply voltage is ready.

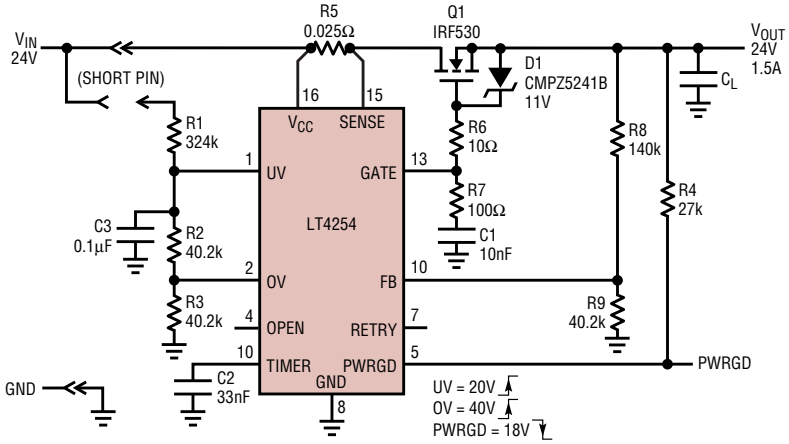


Figure 1. 24V, 1.5A application

Power-Up Sequence

Figure 1 shows a typical LT4254 application. An external N-channel MOSFET pass transistor (Q1) is placed in the power path to control the turn-on/turn-off characteristics of the supply voltage. Capacitor C1 controls the GATE slew rate, R7 provides compensation for the current control loop and R6 prevents high frequency oscillations in Q1.

When the power pins first make contact, transistor Q1 is held off. The V_{IN} and GND connector pins should be longer than the pin that goes to R1 so they connect first and keep the LT4254 off until the board is completely seated in its connector. When the voltage on the V_{CC} pin is between the externally programmed undervoltage and overvoltage thresholds, transistor Q1 is turned on (Figure 2). The voltage at the GATE pin rises with a slope equal to 35µA/C1 and the supply inrush current is $I_{INRUSH} = C_L \bullet 35\mu A/C1$, where C1 is the total load capacitance, provided the part is not in current limit. When the FB pin voltage goes above 4.45V, the PWRGD pin goes high.

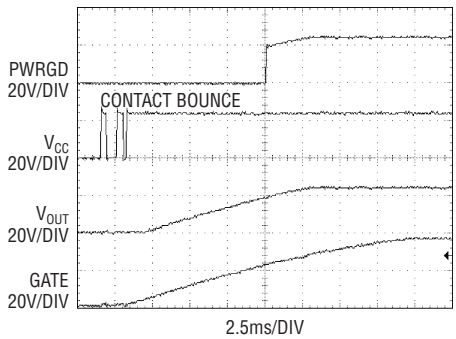


Figure 2. Startup waveforms

Short-Circuit Protection

The LT4254 features a programmable foldback current limit with an electronic circuit breaker that protects against short circuits or excessive load currents. The current limit is set by placing a sense resistor (R5) between V_{CC} and SENSE. To limit excessive power dissipation in the pass transistor and to reduce voltage spikes on the input supply during short-circuit conditions at the output, the current folds back as a function of the output voltage, which is sensed internally on the FB pin. When the voltage at the FB pin is 0V, if the part goes into current limit, the current limit circuitry drives the GATE pin to force a constant 12mV drop across the sense resistor.

Under high current (but not short-circuit) conditions, as the FB voltage increases linearly from 0V to 2V, the voltage across the sense resistor increases linearly from 12mV to 50mV (see Figure 3). With FB above 2V, a

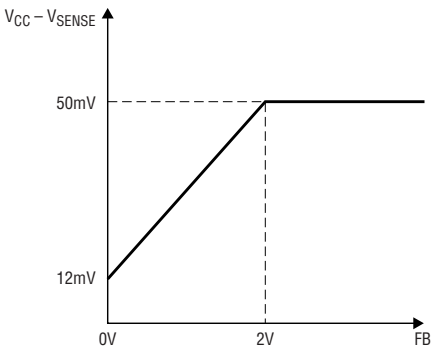


Figure 3. Current limit sense voltage vs FB pin voltage

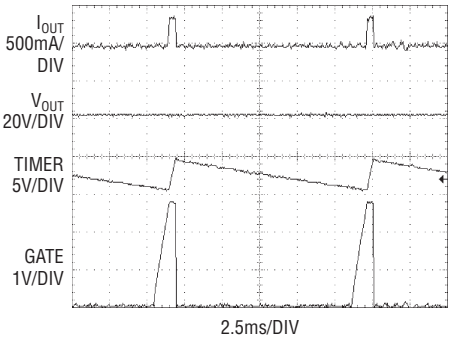


Figure 4. Retry waveforms

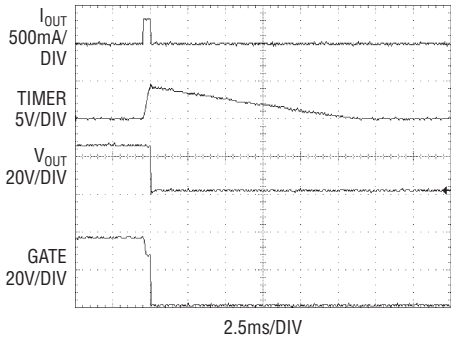


Figure 5. Latch off waveforms

constant 50mV is maintained across the sense resistor.

During startup, a large output capacitance can cause the LT4254 to go into current limit. The current limit level when V_{OUT} is low is only one quarter of the current limit level under normal operation, and it is time limited, so careful attention is needed to insure proper start up. The maximum time the LT4254 is allowed to stay in current limit is defined by the TIMER pin capacitor.

The current limit threshold (during normal operation) is I_{LIMIT} = 50mV/R5, where R5 is the sense resistor. For a 0.025Ω sense resistor, the current limit is set at 2A and folds back to 480mA when the output is shorted to ground. For a 24V application, MOSFET dissipation under short circuit conditions is reduced from 48W to 11.5W.

The LT4254 also features a variable overcurrent response time. The time required for the part to regulate the GATE pin voltage is a function of the voltage across the sense resistor, R5. This helps to eliminate sensitivity to current spikes and transients that might otherwise unnecessarily trigger a current limit response and increase MOSFET dissipation.

Current Limit TIMER

The TIMER pin provides a method for programming the maximum time the part is allowed to operate in current limit. When the current limit circuitry is not active, the TIMER pin is pulled to GND by a 3µA current source. When the current limit circuitry becomes active, a 120µA pull-up current source is connected to the TIMER pin and the voltage rises with a slope equal to 117µA/C_{TIMER}. Once the desired maximum current limit time is chosen, the capacitor value is:

$$C(nF) = 25 \bullet t(ms)$$

If the TIMER pin reaches 4.65V (typ), the internal fault latch is set causing the GATE to be pulled low and the TIMER pin to be discharged to GND by the 3µA current source. The LT4254 does not turn on again until the voltage at the TIMER pin falls below 0.65V (typ).

Undervoltage and Overvoltage Detection

The LT4254 uses the UV (undervoltage) and OV (overvoltage) pins to monitor V_{CC} and allow the user the greatest flexibility for setting the operational thresholds. Figure 1 also shows the UV and OV level programming via a 3-resistor divider (R1, R2, and R3). The UV and OV pins are internally connected to an analog window comparator.

If the UV pin goes below 3.6V or the OV pin rises above 4V, the GATE pin will be immediately pulled low until the UV/OV pin voltages return to the normal operating voltage window (4V and 3.65V on UV and OV, respectively).

Automatic Restart and Latch Off Operation

The RETRY pin can be configured either to latch off the LT4254 or force it into a hiccup mode after an overcurrent fault condition.

If the RETRY pin is floating, when the voltage at the TIMER pin ramps back down to 0.65V (typ), the LT4254 turns Q1 on again. If the short-circuit condition at the output still exists, the cycle will repeat itself indefinitely. The duty cycle under short-circuit conditions is 3%, which limits the power dissipated by Q1, preventing overheating (see Figure 4).

If the RETRY pin is grounded, the LT4254 latches off after a current fault (see Figure 5). After the part latches off, it can be commanded to start back up by cycling the UV pin to ground and then back high. This command can only be accepted after the TIMER pin discharges below the 0.65V typ threshold (to prevent overheating transistor Q1).

Power Good Detection

The LT4254 includes a comparator for monitoring the output voltage. The output voltage is sensed through the FB pin via an external resistor string. If the FB pin goes above 4.45V, the comparator's output releases the PWRGD pin so it can be externally pulled up. The comparator's output (PWRGD pin) is an open collector capable of operating from a pull-up voltage as high as 36V, independent of V_{CC}.

Open MOSFET Detection

The LT4254 can be used to detect the presence of an open MOSFET through the OPEN and PWRGD pins. When the voltage across the sense resistor is less than 3.5mV, the open collector pull-down device is shut off allowing the OPEN pin to be externally pulled high.

An open MOSFET condition is inferred when the OPEN pin is high and the PWRGD pin is low (after the part has completed its start-up cycle). This condition can be falsely signaled during start-up if the load is not activated until after PWRGD goes high; or if the inrush current during start-up is too small, OPEN will not go low until the load is enabled (by the PWRGD signal) as shown in Figure 6. During this time, the OPEN pin is high and the PWRGD pin is low falsely signaling an open MOSFET condition (unless this start-up period is ignored). To avoid this false indication, the OPEN and PWRGD pins should not be polled for a period of time, $T_{STARTUP}$, given by:

$$T_{STARTUP} = (3 \bullet V_{CC} \bullet C1)/35\mu A$$

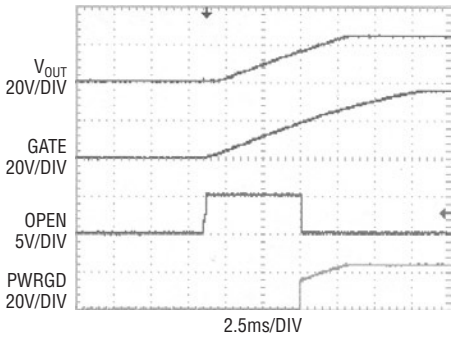


Figure 6. Normal MOSFET start-up waveforms

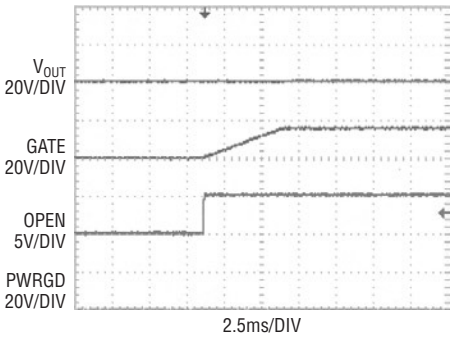


Figure 7. Open MOSFET start-up waveforms.

For example, in Figure 6, $T_{STARTUP}$ is equal to 3 times the typical start-up time, which is 25.5ms (3 × 8.5ms).

This can be accomplished either by using a microcontroller and not polling the logic signals during $T_{STARTUP}$ or by placing an RC filter on the OPEN pin. Once the OPEN voltage exceeds the monitoring logic threshold (signaling an undercurrent condition lasting longer than the start-up period), and PWRGD is low (signaling that the output is not high after the start-up period has finished), an open MOSFET condition is indicated.

Figure 7 shows the typical waveforms for an actual open MOSFET condition. Since the MOSFET is open, V_{OUT} and PWRGD never go high. OPEN goes high as soon as the part is powered up while GATE is clamped to the external Zener voltage above V_{OUT}.

Another condition that can cause a false open MOSFET indication is if the LT4254 goes into current limit during start-up. This causes $T_{STARTUP}$ to be longer than anticipated. Also, if the LT4254 stays in current limit long enough for the TIMER pin to fully charge up to its threshold, the LT4254 will either latch off (RETRY = 0) or go into the current limit hiccup mode (RETRY = floating). In either case, an open MOSFET condition will be falsely signaled. If the LT4254 does go into current limit during start-up, C1 can be increased (to reduce inrush current).

GATE Pin

The GATE pin is clamped to a maximum of 12V above the V_{CC} voltage. This clamp is designed to withstand the internal charge pump current. An external Zener diode must be used if the possibility exists for an instantaneous low resistance short from V_{OUT} to GND. When the input supply voltage is between 12V and 15V, the minimum gate drive voltage is 4.5V, and a logic level MOSFET must be used. When the input supply voltage is higher than 20V, the gate drive voltage is at least 10V, and a standard threshold MOSFET is recommended.

Conclusion

The LT4254's comprehensive set of advanced protection and monitoring features make it applicable in a wide variety of Hot Swap solutions. It can be programmed to control the output voltage slew rate and inrush current. It has programmable undervoltage and overvoltage protection, and monitors the output voltage via the PWRGD pin. The part also indicates if an open MOSFET condition exists. The LT4254 provides a simple and flexible Hot Swap solution with the addition only a few external components.

LT6011, continued from page 12

Differential pair Q1 and Q2, together with load resistors R3 and R4, form a first gain stage. The PNPs Q5 and Q6, and current mirror Q9 and Q10 form the second gain stage. The output stage is designed to be able to both source and sink much larger currents than the stage biasing current. The current-sinking device NPN Q20 is driven directly by Q12, while the current-sourcing PNP Q19 is driven through level-shifting bias network Q13 and Q14.

The level-shifter works as follows: The fixed current flowing into diodes D3–D5 establishes a bias voltage at the base of Q13. As the base of Q14 is driven lower, the V_{BE} of both Q13 and Q14 increases. This increases their current, which flows through Q18/R6 and is mirrored as sourcing current in Q19. Since only collectors are connected to the output, a mere 40mV V_{CE} saturation voltage limits the output swing to either supply rail.

Input devices Q1 and Q2 are super-beta transistors. Their lightly doped base region results in a current gain of more than 1000. In addition, the already low base current is internally compensated by a base current-cancellation circuit. Current mirror Q21 biases Q11 with the exact same current as the input devices. Q17 measures the base current of Q11 and feeds this same current back into the bases of Q1 and Q2. The resulting input bias current is limited only by mismatch and is typically just 20pA.

The input offset voltage of the amplifier is a result of mismatch in Q1/Q2 as well as R3/R4. These internal load resistors are trimmed at the factory to cancel out the total offset voltage to less than 60µV (A-grade). A high degree of balance is maintained through the second stage, which virtually eliminates second-order temperature drift contributions.

Conclusion

Rail-to-rail output swing to supplies as low as 2.7V, power consumption as low as 400µW, and availability in tiny packages make the LT6011 op amp the ideal precision op amp for low voltage, low power, or space constrained applications.

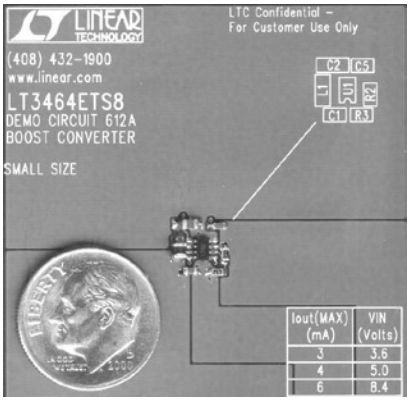
Micropower SOT-23 Boost with Integrated Schottky Diode Provides Output Disconnect and Short Circuit Protection

by Leonard Shtargot

Introduction

The LT3464 is the only micropower boost converter in the industry to combine a 36V NPN power switch, power Schottky diode, and output disconnect into an 8-lead ThinSOT. This unprecedented level of integration saves several external components while offering true output disconnect, making it possible to generate outputs of up to 34V with a zero current shutdown while using a mere 40mm^2 of board area (refer to Fig. 1).

In addition to component savings, the LT3464 offers a low typical switch current limit of 115mA and fast switching, a combination that allows the use of a tiny chip inductor and tiny ceramic capacitors. The LT3464 also features Burst Mode control (see Figure 2), which results in highly efficient operation over a wide range of load currents and a low quiescent current of only 25µA typical. The CTRL pin of the LT3464 acts much like a dial on a lab power supply—it allows the output voltage to be varied, which is useful in applications for purposes such as LCD contrast adjustment.



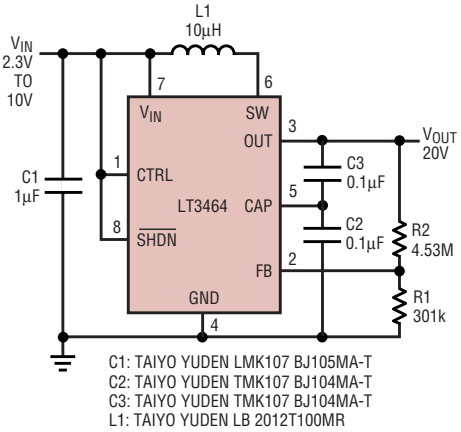


Figure 1. The integrated Schottky diode and output disconnect transistor result in a tiny solution occupying as little as 40mm^2.

The LT3464's small size and high efficiency make it an especially attractive power solution for portable electronics requiring long battery life and compact circuitry. See Figure 3 for a simplified block diagram of the LT3464.

Output Disconnect

In a simple boost circuit (Figure 4) there exists a DC path from the input supply (V_{IN}) through the inductor and diode to the load (V_{OUT}), effectively leaving the load connected to V_{IN} during shutdown. The resulting current drain during shutdown is unacceptable in many applications, requiring the addition of several external components to isolate the load from V_{IN}. To save space and complexity, the LT3464 is equipped with a PNP that completely

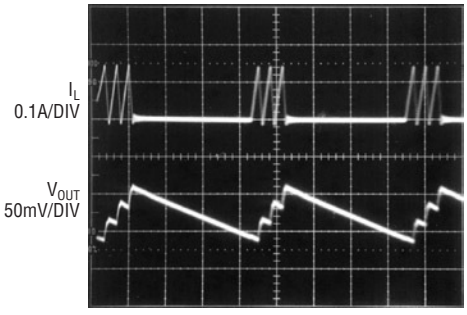


Figure 2. Burst Mode waveforms showing low output ripple. The LT3464 consumes only 25µA typical when not switching.

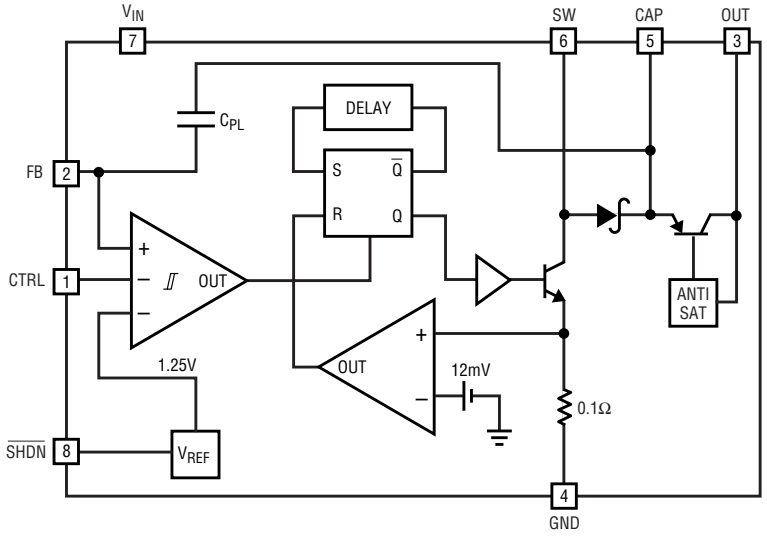


Figure 3. LT3464 block diagram showing integrated NPN switch, Schottky diode and output disconnect PNP.

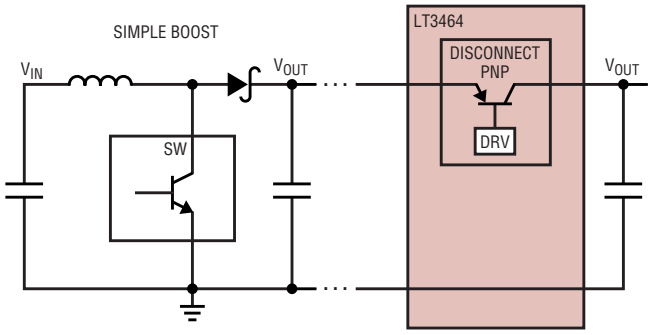


Figure 4. A simple boost circuit with LT3464's output disconnect allows the complete solution to draw less than 0.5µA during shutdown. The output disconnect is designed with a 25mA current limit to protect the circuit during short circuit conditions.

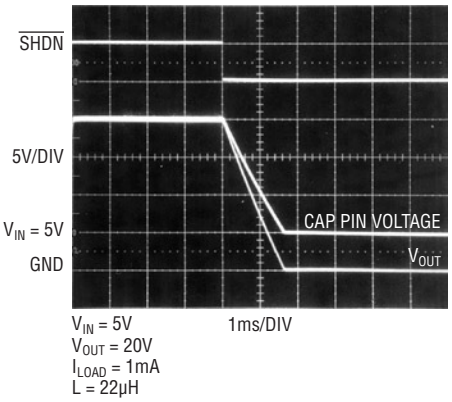


Figure 5: The output disconnect isolates the output load from the input supply during shutdown. CAP Pin voltage is the output of the Schottky in Figure 4.

disconnects the load from the Schottky diode during shutdown (see Figures 4 and 5). During normal operation, the control circuitry turns on the PNP and keeps it just out of saturation, resulting in low $V_{CE(SAT)}$ and low quiescent current. In addition, the disconnect circuit has a built in current limit of 25mA to protect the chip during a short-circuit at the output. This feature allows the LT3464 to tolerate an indefinite short, but care must be taken to avoid exceeding the maximum junction temperature.

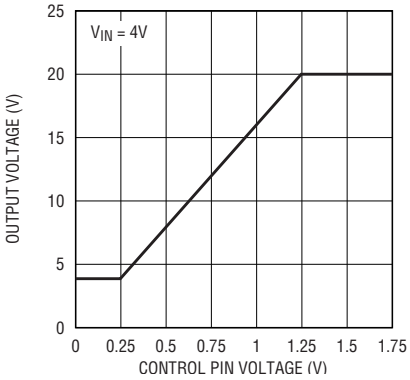


Figure 6: Using the CTRL Pin as an auxiliary reference input to control the output voltage

Using the CTRL Pin

The LT3464 features an auxiliary reference input that provides an easy way to vary the output voltage for purposes such as LCD contrast adjustment or display dimming. When the CTRL pin held at or above 1.25V, the LT3464 uses the internal 1.25V reference, but when a voltage lower than 1.25V is applied to the CTRL pin, that voltage becomes the new reference. Figure 6 shows the output voltage versus the CTRL pin voltage for a 20V output circuit. Note that the LT3464 will not regulate the output to a voltage lower than the input.

LT3464 ±20V Dual Output Converter

Figure 7 shows a single-inductor dual-output converter for applications that require both a positive and negative voltage. The positive output is generated by a simple boost set up, whereas the negative output is generated using an inverting charge pump. Although well regulated, the negative output will have a slight offset from the positive output because the external diodes have a different on voltage when compared to the integrated Schottky diode.

1-Cell Li-Ion to 16V Boost Converter

Figures 8 and 9 show that the LT3464 performs well in applications that need a high output voltage at a relatively

continued on page 24

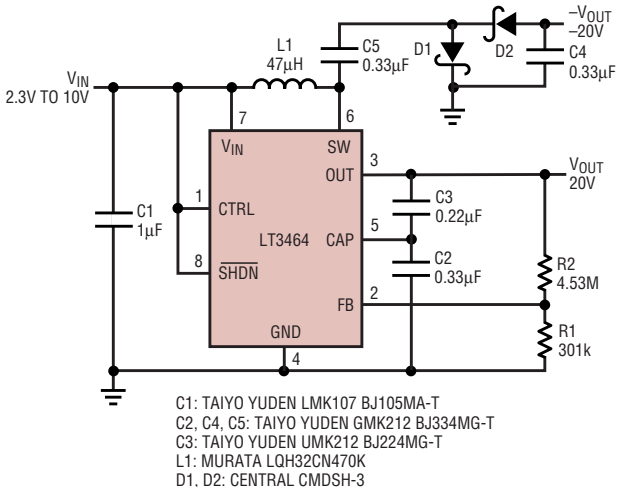


Figure 7. ±20V Dual output converter

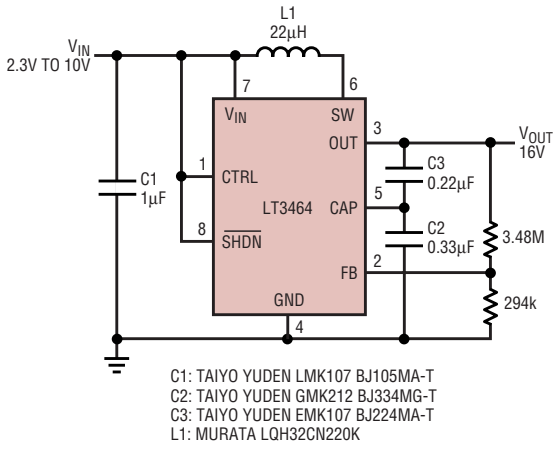


Figure 8. Li-Ion to 16V boost converter

Amplifier with Integrated Filter Offers the Best High Speed, Low Noise Interface for Differential DACs and ADCs

by Michael Kultgen

Introduction

Differential signal paths are becoming a popular way to improve system performance. A differential signal has twice the amplitude for a given supply level. Interference from other components, digital clocks for example, become common mode signals and are rejected by the amplifiers in the differential signal chain. Hence, with a differential signal path, dynamic range can be maintained while the supply voltage is reduced.

Differential output op amps are one means of providing gain, buffering, and filtering in these signal paths, but often at the price of added complexity and board real estate. A circuit using a differential output op amp typically requires twice the number of resistors and capacitors as the corresponding single-ended circuit. Therefore, there is a need for a compact, high performance means to process differential signals. The solution is the new LT6600.

The LT6600 is a family of fully differential amplifiers with an integrated 4th order lowpass filter. Each device in the family features a fixed cutoff frequency (2.5MHz, 10MHz and 20MHz) and operates with power supplies ranging from 3V to 10V. The LT6600 is packaged in an SO-8, and is pin-for-pin compatible with other commercially available high-speed differential op amps (Figure 1). Like their industry counterparts, the LT6600 amplifiers can accept single-ended or differential input signals, translate common mode voltages, and have a common mode input range that extends to ground. But, unlike these other amplifiers, the LT6600s have a proprietary architecture that minimizes noise and distortion while maximizing speed. Furthermore, the custom lowpass filter

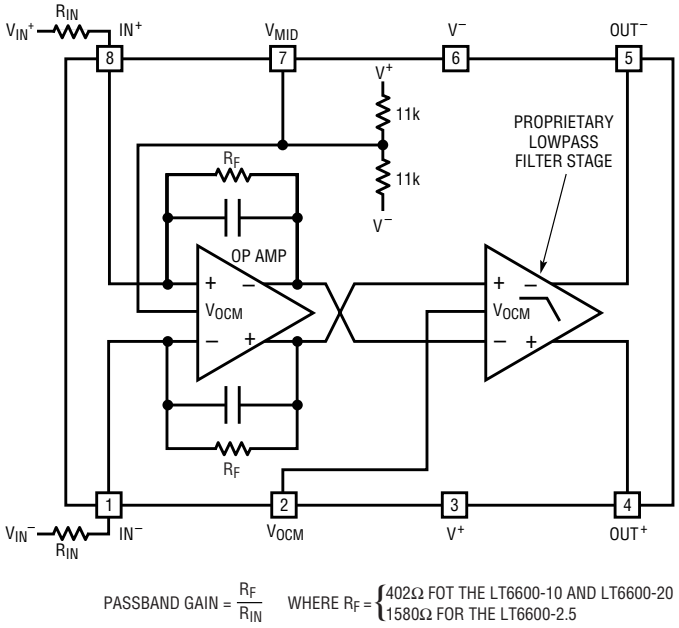


Figure 1. LT6600 block diagram

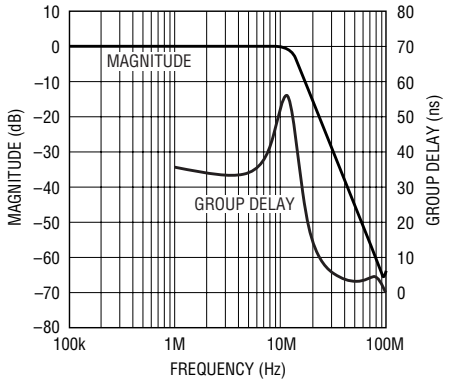


Figure 2. The frequency response of the LT6600-10

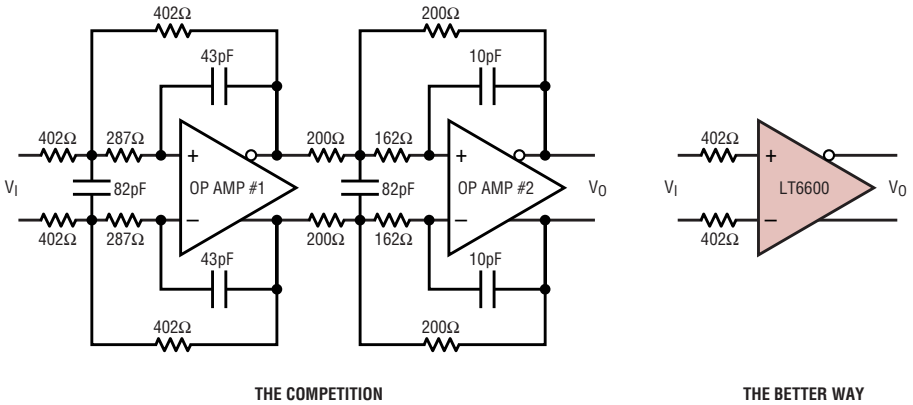


Figure 3. The LT6600 vs the competition

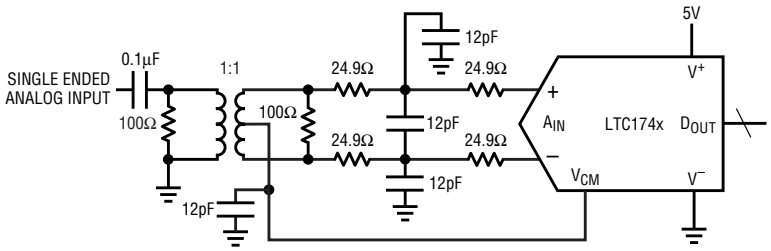


Figure 4. Using a transformer to drive a differential ADC

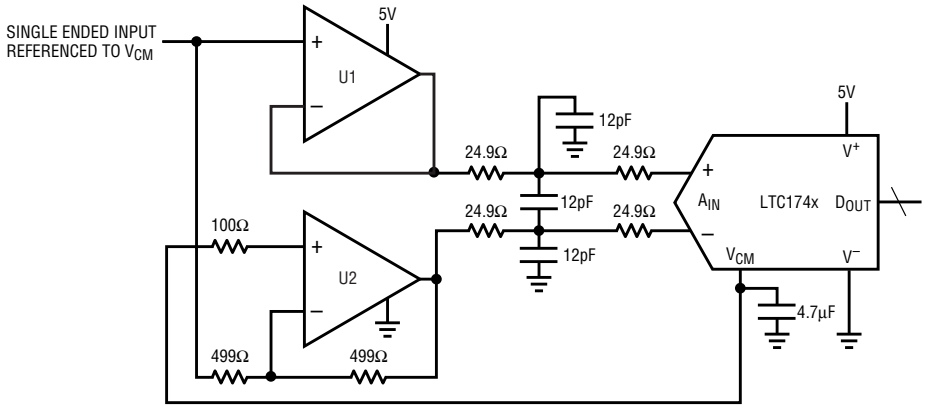


Figure 5. Using single-ended op amps to drive a differential ADC

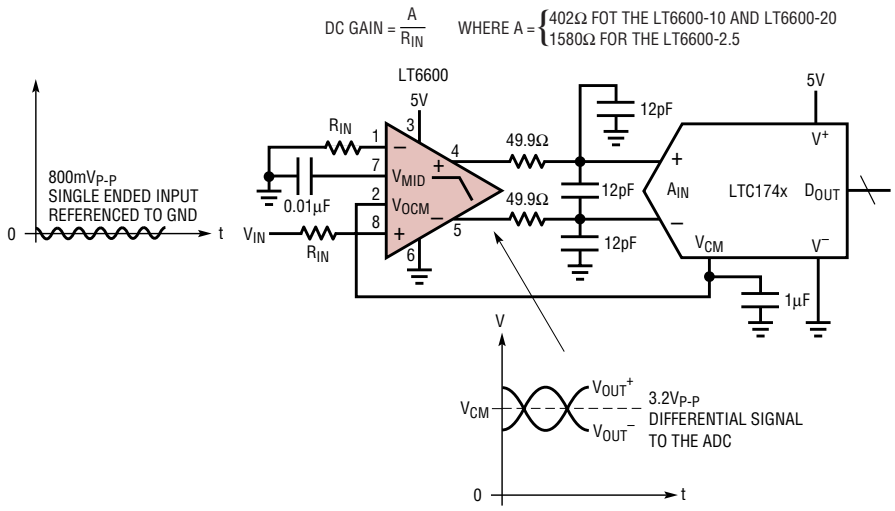


Figure 6. Using the LT6600 to drive a differential ADC

response provides 30dB attenuation at 3 times the cutoff frequency with low delay distortion (Figure 2).

Quiet, Compact and Easy to Use

The LT6600s are also the most compact anti-alias/smoothing filter solutions available. These integrated filter-amplifiers need only two external resistors to set the gain. The precision response is completely determined by the monolithic filter. By contrast, a discrete active-RC design would require 18 precision resistors and capacitors, as well as a second operational amplifier package (Figure 3). The first available devices in the family are the LT6600-2.5, the LT6600-10, and the LT6600-20, with fixed filter bandwidths of 2.5MHz, 10MHz, and 20MHz respectively.

The LT6600-2.5 offers true 14-bit performance with noise and distortion components below –86dB for 1MHz $1V_{RMS}$ inputs. The LT6600-10 has a total noise of $56\mu V_{RMS}$ in a 10MHz bandwidth with harmonics below –74dB for 5MHz $2V_{P-P}$ signals. With the gain set to +12dB, the LT6600-20 has $42\mu V_{RMS}$ total input referred noise in a 20MHz bandwidth and shares many features of the -2.5 and -10. Each member of the family has a lowpass response with less than 0.5dB of passband ripple. These combinations of low noise, low distortion and controlled frequency response are practically impossible to duplicate with discrete designs.

The LT6600s are specified and tested for both single 3V supply operation and ±5V supply operation. This flexibility, combined with the low external parts count and the wide input common mode range, makes the LT6600 extremely easy to use.

A Great Solution for Differential ADC and DAC Interfaces

Differential output amplifiers have gained popularity in systems that have differential input ADC converters. Often the signal to be processed by the converter is single-ended, low amplitude, and from a large source impedance. For the converter to realize its full range and accuracy, it must be presented with a larger differential signal and a common mode level near mid supply. There are two common solutions to this problem. Figure 4 shows a transformer coupling circuit and Figure 5 shows an operational amplifier circuit. The circuit shown in Figure 5 is preferable when size and DC response are important, when gain is needed, or when buffering is required. The circuit of Figure 4 is useful for wide bandwidth applications.

Figure 6 shows how simple the differential converter interface is with the LT6600. This circuit retains all of the benefits of Figure 5's circuit (gain, DC response, and single-ended to differential conversion) with the added feature of selective anti-alias filtering. Furthermore, the single-ended analog input can have a common mode level that differs from the ADC converter (unlike the circuit of Figure 5). The LT6600 automatically translates the common

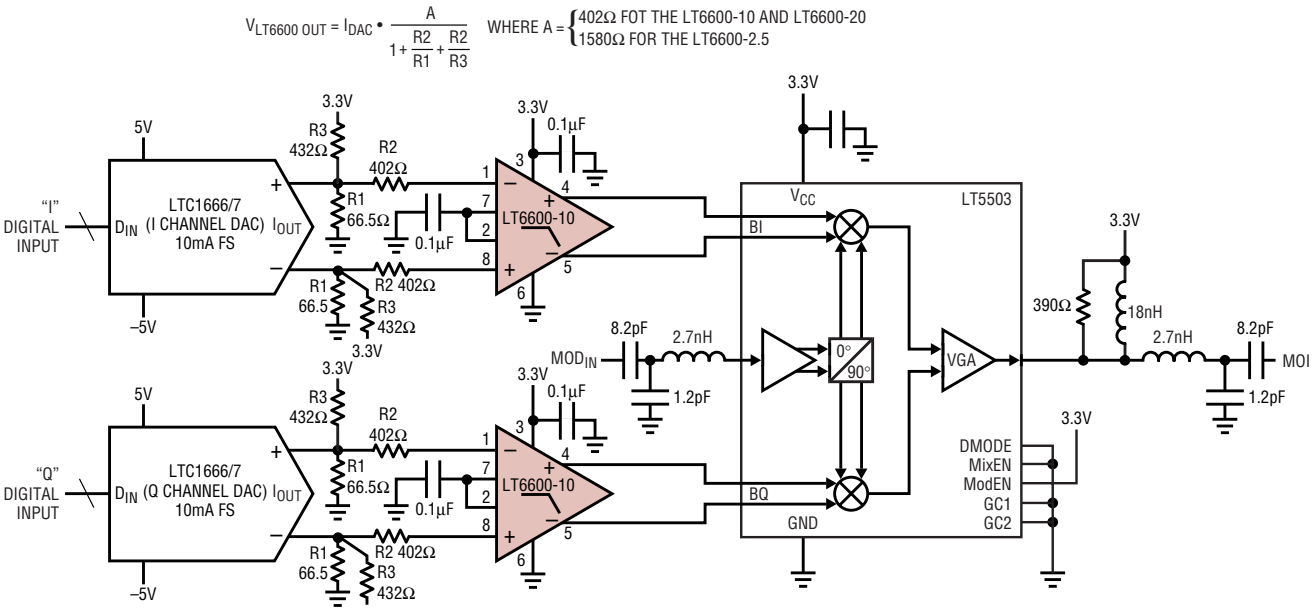


Figure 7. Using the LT6600 as a transimpedance amplifier and smoothing filter in a base station application.

mode level when it converts the single-ended input to differential. In Figure 6, the input signal is referenced to ground and the signal presented to the ADC is referenced to V_{CM}.

To illustrate the excellent dynamic range of the LT6600, consider Figure 6 with a 1MHz input signal of $800mV_{PP}$ amplified by an LT6600-2.5. With R_{IN} = 402Ω, the amplifier provides 12dB of voltage gain. The signal presented to the ADC converter is $3.2V_{P-P}$. The distortion components will be at least 82dB below the fundamental, and the signal-to-noise ratio will be 81dB in a 5MHz bandwidth.

The differential output DAC is another application where the LT6600 excels. Figure 7 shows the LT6600 acting as a transimpedance amplifier and a 4th order smoothing filter, in a base station application. The input common mode range of the LT6600 accommodates the compliance range of the DAC. The output common mode voltage of the LT6600 is set to optimize the performance of the LT5503 direct I/Q modulator. The resistors between the DAC and the LT6600 allow the user to adjust the transimpedance gain. The LT6600 and LT5503 are operating on a 3.3V power supply.

To illustrate the optimized filtering of the LT6600, consider the case where the DAC in Figure 7 has a sample rate of 50Msps and the baseband signal information extends to 10MHz. By using an LT6600-10, the attenuation of the images near 40MHz would be more than 50dB (filter response plus sin(x)/x attenuation). The excellent rejection in the stopband is combined with low delay distortion in the passband (Figure 2), making for an outstanding DAC smoothing solution.

Conclusion

The LT6600 differential filter-amplifiers are the most compact ADC anti-aliasing and DAC smoothing solutions available in the 2.5MHz to 20MHz range. The combination of low noise, low distortion, and precision response are impossible to replicate with discrete designs. The LT6600 is pin compatible with standard differential output op amps and performs all of the same functions. The LT6600 improves the design of any system requiring differential signal buffering and filtering.

LT3464, continued from page 21

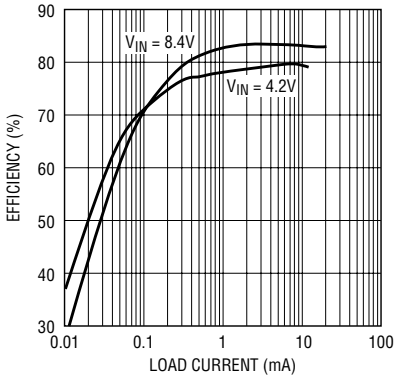


Figure 9. Efficiency for the circuit in Figure 7

low current. As shown in Figure 9, high efficiency is maintained with low output currents.

Conclusion

The LT3464 in the ThinSOT package produces an ultra compact boost solution featuring high efficiency, low quiescent current, true output disconnect, and low external parts count.

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

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Ask for the pertinent data sheets and Application Notes.

Low Voltage Amplifiers Give Choice of Accuracy or Speed

by Frank Johnston, Glen Brisebois and Danh Tran

Introduction

Two new families of single, dual and quad amplifiers, the 1mA LT6220/1/2 and the 3mA LT1803/4/5, provide high speed amplification on supplies as low as 2.5V. The 85MHz LT1803 series is optimized for large signals, with a 100V/µs slew rate, while the 60MHz LT6220 series focuses on excellent DC performance at low current, with a maximum input offset specification of only 350µV. Both series have rail-to-rail output stages that can swing to within 20mV of the rails, and rail-to-rail inputs that can be used anywhere within the supplies. The supply range is 2.5 to 12V. DC accuracy is insured by trimming of the input offset voltage and cancellation of input bias current. Figures 1 and 2 show the results of a proprietary bias current cancellation circuit.

The devices are available in small packages; singles in the SOT-23, and duals in the DFN as well as the larger SO packages.

Performance

Table 1 summarizes the performance of the devices. Note that input offset voltage and input bias current are specified and guaranteed with the common mode voltage near each rail. Histograms of input offset voltage are shown in Figures 3 and 4. The large signal transient response is shown in Figures 5 and 6. The response is clean with no aberrations.

Circuit Description

Figure 7 shows a simplified schematic of the amplifiers. The circuit is composed of three distinct stages: an input stage, an intermediate stage, and an output stage. The input stage consists of two differential amplifiers, a PNP stage (Q1 and Q2) and an NPN stage (Q3 and Q4), that are active over different portions of the input common mode range. The intermediate stage is a folded cascode configuration formed by Q8, Q9, Q11 and Q12, which provides most of the voltage gain. A pair of complementary common emitter devices, Q14 and Q15, creates an output stage which can swing from rail to rail.

In the input stage, devices Q18 and Q19 act to cancel the bias current of the PNP input pair. When Q1 and Q2 are active, the current in Q16 matches the current in Q1 and Q2, thus the base current of Q16 is nominally equal to the base current of Q1/Q2. The base current of Q16 is mirrored by devices Q17, Q18 and Q19 to each input. The cancellation is effective for common mode voltage

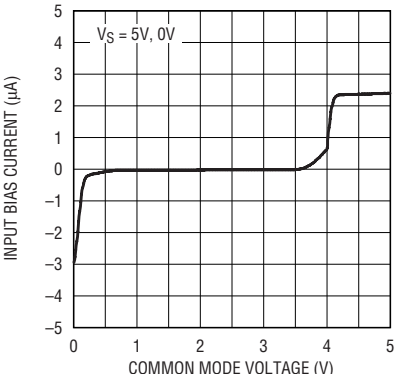

Figure 1. LT1803 I_B vs common mode voltage

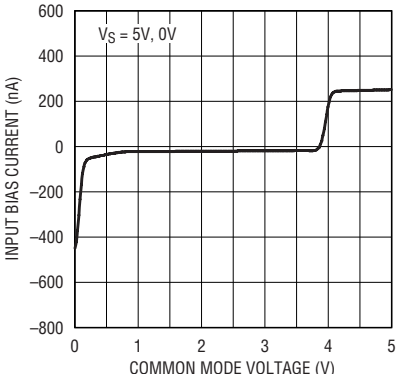

Figure 2. LT6220 I_B vs common mode voltage

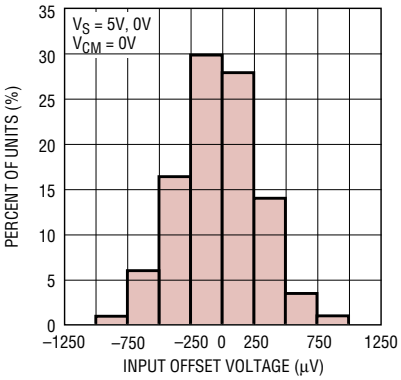

Figure 3. LT1803 offset voltage distribution

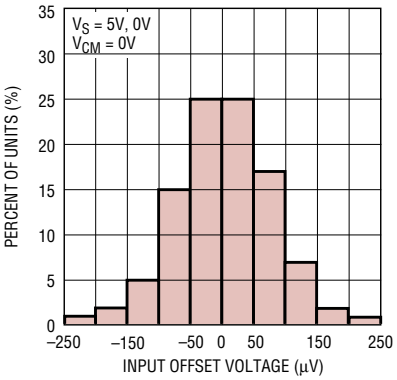

Figure 4. LT6220 offset voltage distribution

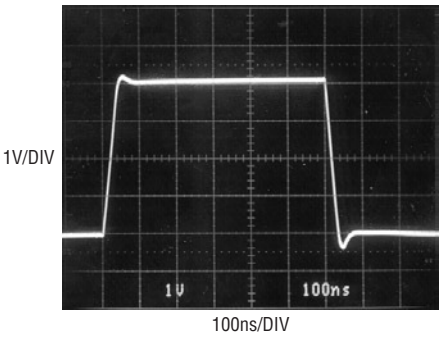

Figure 5. LT1803 slew rate

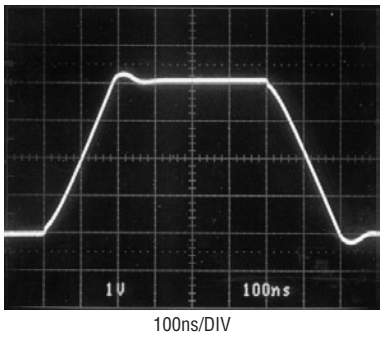

Figure 6. LT6220 slew rate

Table 1. Specifications of LT1803 and LT6220 at 25°C, V_S = 5V, 0V

Parameter	Conditions	LT1803	LT6220	Units
–3dB Bandwidth	A_V = 1	60	55	MHz
Gain-Bandwidth Product		85	60	MHz
Slew Rate	R_L = 1k	100	20	V/µs
Supply Current		3.0	1.0	mA (Max)
Operating Supply Range		2.5 to 12	2.5 to 12	V
Input Offset Voltage	V_{CM} = V⁻, SO-8	2	0.35	mV (Max)
	V_{CM} = V⁻,SOT-23	5	0.85	
Input Bias Current	V_{CM} = V⁻+1V	750	150	nA (Max)
	V_{CM} = V⁺	5500	600	
CMRR		66	85	dB (Min)
PSRR	V_S = 2.5V to 10V, V_{CM} = 0V	68	86	dB (Min)
Input Voltage Noise	f = 10kHz	21	10	nV/√Hz
Harmonic Distortion	V_S = 5V, A_V = 1, R_L = 1k, V_O = 2V_{P-P}, f_C = 500kHz (LT6220) f_C = 1MHz (LT1803)	–75	–75	dBc
A_{VOL}	V_S = 5V, V_O = 0.5V to 4.5V, R_L = 1k	20	30	V/mV (Min)
Output Voltage Swing LOW	I_L = 0mA	60	40	mV (Max)
	I_L = 15mA (LT1803) I_L = 20mA (LT6220)	300	650	
Output Voltage Swing HIGH	I_L = 0mA	60	40	
	I_L = 15mA (LT1803) I_L = 20mA (LT6220)	600	900	

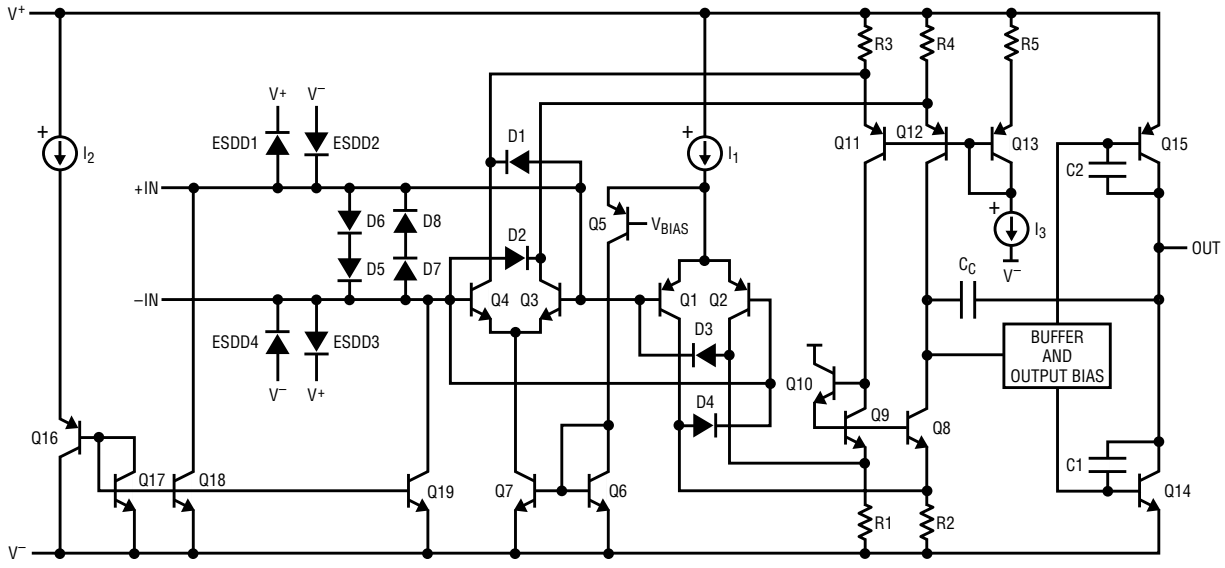

Figure 7. Simplified schematic

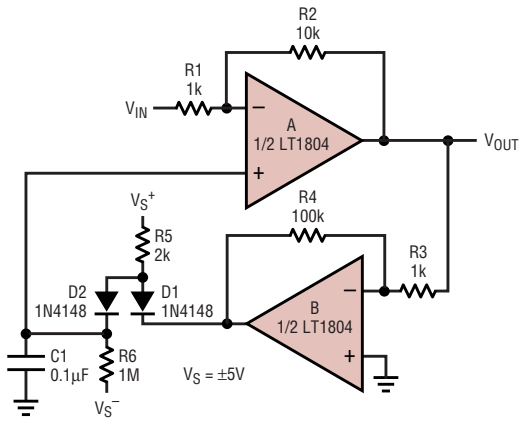

Figure 8. Inverting DC restore

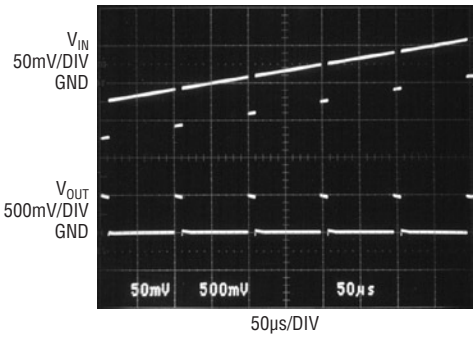

Figure 9. Inverting DC restore oscillograph

greater than the saturation voltage of Q18 and Q19, about V^- + 0.2V, up to the voltage that the PNP devices switch off, about V^+ – 1.3V.

Inverting Amplifier with DC Restore

The circuit of Figure 8 shows half of the LT1804 used as a gain of –10 amplifier, and the other half as a DC restore. This type of circuit is often associated with photomultiplier tubes and photodiodes (see Figure 10), which are inherently unipolar but can be subject to annoying DC components caused by amplifier offsets, dark current, and the presence of residual light. The oscillograph in Figure 9 shows how effectively the circuit rejects the varying DC level which the incoming negative going pulses are riding on (top trace). The pulses are inverted and restored to a 0V ground reference (bottom trace)[1]. Note that this is not the same as AC coupling, which would simply center the average output waveform around ground. The DC restore function has three operating regions: output high or "ignore," output low or "restore," and output zero or "lock."

When the output is high, op amp B's output falls low turning D1 on and D2 off. This leaves the pull down R6 in place, so the voltage on C1 falls slowly. Therefore the positive output voltage also starts to fall back towards ground via op amp A. The long time constant of R6 • C1 is what makes this the "ignore" function. Of course, it does not fully ignore positive outputs, as that would make the circuit useless. The point is, though, that a positive signal indicates the presence of signal, so the restore circuit does little to zero the output.

When the output is low, however, we have a situation where we have negative light. Since that situation is unlikely to exist outside of academic circles, we can assume that we have zero light, or at least are at the light signal floor, and would like to set that as the DC reference level. To that end, op amp B's output goes high turning D1 off and D2 on. This now leaves R5 to pull up on C1 about 500 times harder than R6 had pulled down. Again, through op amp A, this positive going voltage causes the output to rise quickly. When the output reaches 0V, op amp B detects this and pulls down on D1, stopping the restore function and entering "lock" mode.

In lock mode, light is presumably absent, and the output is held close to 0V. D1 and D2 are both on, and although they are running different currents, the resultant mismatch voltage is relatively small output referred because of the high gain around op amp B. Should a negative going "light present" pulse occur at V_{IN}, the circuit goes again into ignore mode.

Photodiode Amplifier

The circuit in Figure 10 is a stepped gain transimpedance photodiode amplifier. At low signal levels, the circuit has a high 100kΩ transimpedance gain, but at high signal levels the circuit automatically and smoothly changes to a low 3.1kΩ gain. The benefit of a stepped gain approach is that it maximizes dynamic range,

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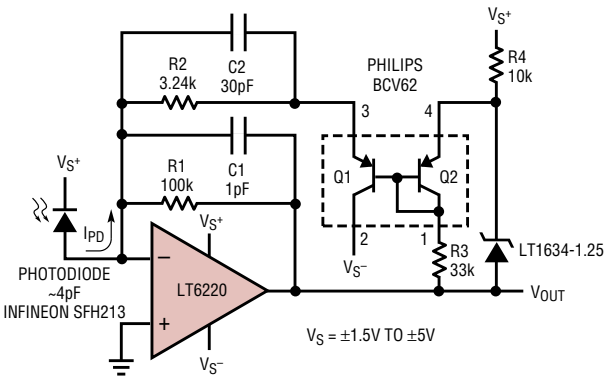

Figure 10. Photodiode amplifier

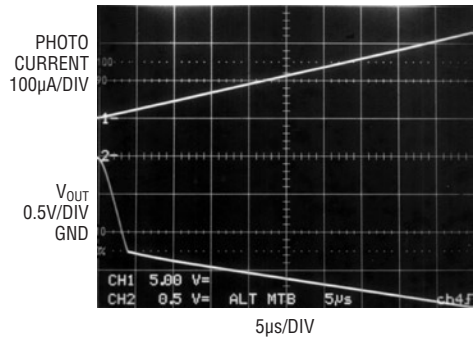

Figure 11. Stepped gain photodiode amplifier oscillograph

Feature-Rich Battery Charger that Manages Both Battery Charging *and* Bus Voltage Regulation

by John Shannon

Introduction

Until now power management in portable devices has required a mix of major components to fulfill the basic functions of battery charging and generation of system supply voltages. A typical solution requires at least two major devices (and associated external components): one charger IC for charging the battery and another IC to supply a regulated system bus voltage from a constantly changing battery voltage. The LTC1980 is a single-device solution that manages both battery charging and generation of the regulated system bus voltage.

Powerful Features

The LTC1980, in simple terms, controls the power flow between the AC adapter, a battery and the system bus. The basic LTC1980 circuit is a synchronous flyback converter. In such a configuration, power can flow either way through the converter, a fact that is exploited to charge or discharge the battery, depending on the power needs of the system.

The battery charger portion of the LTC1980 is a full-featured, constant current, constant voltage, Li-Ion charger with timer termination. The LTC1980 can be set up for either 1- or 2-cell, and 4.1V or 4.2V chemistries. This switch mode charger maintains high efficiency over a wide range of input voltages. The flyback topology allows any input voltage to generate any output voltage, unlike buck or boost topology chargers that require the input voltage to be always higher or always lower than the battery voltage.

Charging (AC Power Present)

If the AC adapter is present and has sufficient voltage then the LTC1980 enters charge mode. In charge mode power flows from the adapter to both the system bus and the battery. The charger uses a constant current, constant voltage algorithm that is suitable for Li-Ion cells. Deeply discharged batteries are trickle charged with a low current until the battery voltage exceeds the trickle charge threshold, at which point full current charging commences. The switch mode operation of the charger typically keeps efficiency above 80%, which results in less heat generation compared to a linear charger. Adapter power also flows directly to the system bus via a linear regulator. The efficiency of the linear converter is simply the ratio of the system bus voltage to the adapter voltage, so losses are minimized if the adapter voltage is close to the desired system bus voltage.

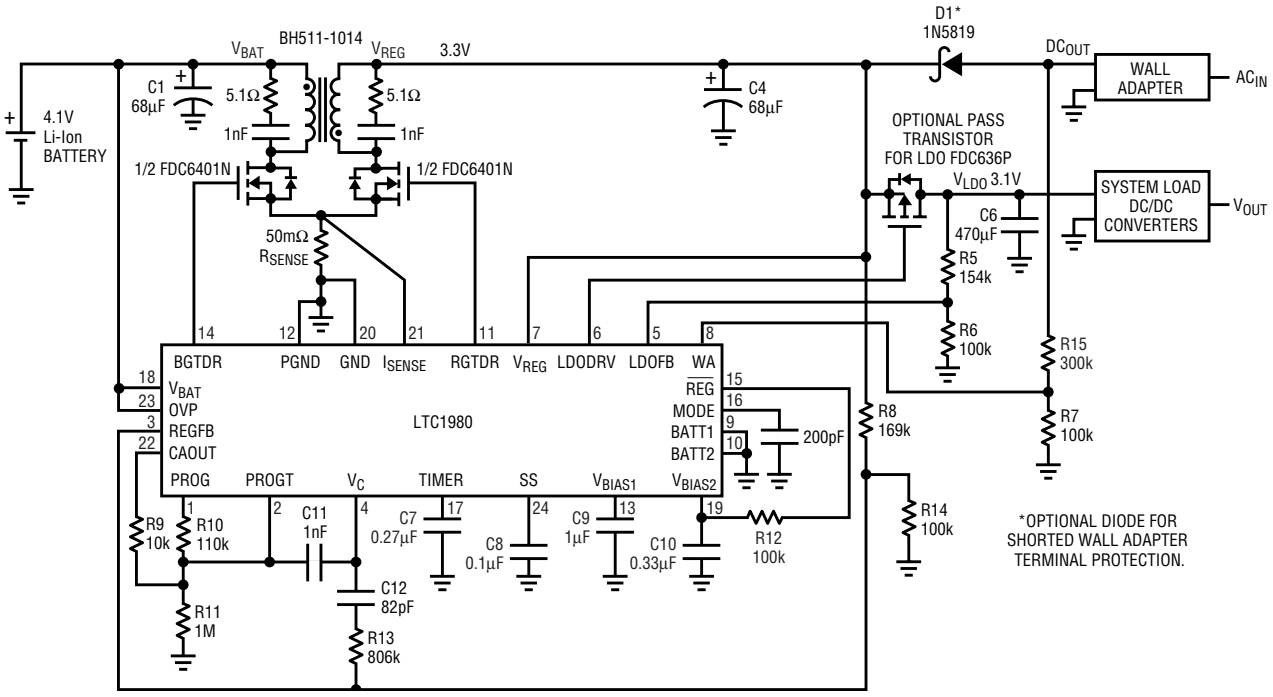

Figure 1. Typical application for single Li-Ion cell

Discharging (Battery Mode)

When the adapter input falls, so that the system bus voltage requirements can no longer be met, the LTC1980 switches to the regulator mode. In this mode the LTC1980 no longer functions as a battery charger. It instead acts as a battery discharger. Power flows "backwards" from the battery to the linear regulator. The output voltage of the flyback, which is input to the linear regulator, should be as low as possible in order to maximize efficiency and battery run time. The efficiency of the battery to system bus voltage conversion can be as high as 88%.

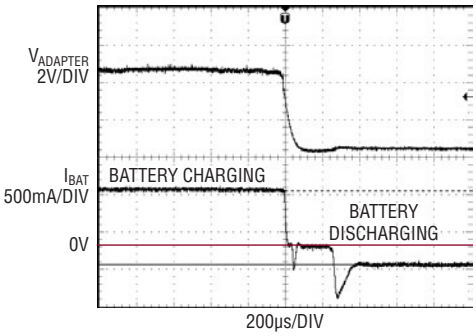

Figure 2. Adapter voltage and battery current (adapter removal)

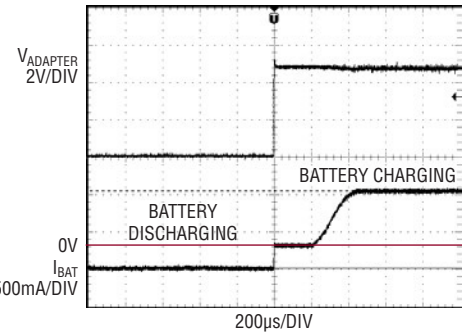

Figure 3. Adapter voltage and battery current (adapter insertion)

The Linear Regulator

A low dropout regulator, using an external P-FET as the pass element, regulates the system bus voltage. The linear regulator takes its power from the output of the AC adapter. Dissipation in the linear regulator is lowest when the AC adapter voltage is near the system bus voltage. When the system is in battery discharging mode, the voltage input to the linear regulator is the output of the synchronous flyback converter. This voltage should be set to be only a percent or two above the required output voltage (allowing for IR drops in the pass element). This prevents saturating the gate drive to the pass element and will aid in transient recovery.

Figure 1 shows a typical application circuit for charging a single 4.1V Li-Ion cell. The adapter voltage can vary from 4V to 9V, demonstrating one key advantage of the flyback topology. Figures 2 and 3 show battery current and adapter voltage during the transition from battery charging (adapter present) to regulator mode (battery discharging). The load on the linear regulator is 200mA, supported either by the battery or the adapter. When the adapter is present the battery is charged at about 650mA. Once the wall adapter is removed the battery is discharged as power flows back through the synchronous flyback converter to support the 200mA load on the linear regulator.

Conclusion

The LTC1980 manages both battery charging *and* system voltage regulation, which is typically the work of two separate devices and their corresponding external circuitry. This feature combined with the fact that the design of the LTC1980 also allows for battery voltages either above or below the adapter voltage, greatly simplifies the task of integrating a battery and adapter into a portable device.

LT1803 and LT6220, continued from page 27

which is very useful on limited supplies. Put another way, in order to get 100kΩ sensitivity and still handle a 1mA signal level without resorting to gain reduction, the circuit would need a 100V negative voltage supply.

The operation of the circuit is quite simple. At low photodiode currents (below 10µA) the output and inverting input of the op amp are no more than 1V below ground. The LT1634 in parallel with R3 and Q2 keep a constant current though Q2 of about 20µA. R4 maintains quiescent current through the LT1634 and pulls Q2's emitter above ground, so Q1 is reverse biased and no current flows through R2. So for small signals, the only feedback path is R1 (and C1) and the circuit is a simple transimpedance amplifier with 100kΩ gain.

As the signal level increases though, the output of the op amp goes more negative. At 12.5µA of photodiode current, the 100kΩ gain dictates that the LT6220 output is about 1.25V below ground. At that point, however, the emitter of Q2 is at ground, and the base of Q1 is one V_{be} below ground. Thus, Q1 turns on and photodiode current starts to flow through R2. The transimpedance gain is therefore now reduced to R1||R2, or about 3.1kΩ. The circuit response is shown in Figure 11. Note the smooth transition between the two operating gains, as well as the linearity of both regions.

Conclusion

The LT1803 series and LT6220 series deliver exceptional performance, and the rail-to-rail inputs and outputs of these devices maximize signal dynamic range while simplifying design for single supply systems. The LT1803 series and the LT6220 series feature reduced supply current, lower input offset voltage, lower input bias current, and higher DC gain than other devices with comparable bandwidth, which is critical in circuits having high input impedance, such as active filters, or in circuits having precision requirements, such as current sensing amplifiers. The LT1803 and LT6220 series are offered in a variety of small packages including a 3mm × 3mm dual fine pitch leadless package with the standard dual op amp pinout and also in the SOT-23 package for a single amplifier. The combination of speed, DC accuracy and low power makes the LT1803 series and the LT6220 series a preferred choice for battery powered, low voltage signal conditioning.

Notes

[1] A DC bias on op amp B's + input could set the output restore to some other reference voltage.

Synchronous, Phase Modulated, Full Bridge Converter Targets Isolated High Power Applications

by Kurk Mathews

Introduction

In networking and telecom equipment, power supplies provide isolated low voltage outputs from the 48V input supply rail, with the added requirements of high efficiency (to reduce heat dissipation) and low component height. These requirements become even more difficult to meet with increased power levels, because of the corresponding component power dissipation and increased transformer size. Simple power supply topologies give way to more complex single and two-stage approaches that focus on transformer and semiconductor utilization.

One such approach, the phase modulated full bridge converter, is a popular choice for high power supplies. Unfortunately, until now, there has not been a controller that offers both the flexibility in timing control and features (such as synchronous rectifier outputs) to manage a variety of high power applications. The new LTC3722-1 current-mode controller is designed specifically with these issues in mind, providing a full-featured controller for high power, phase modulated, full bridge converters.

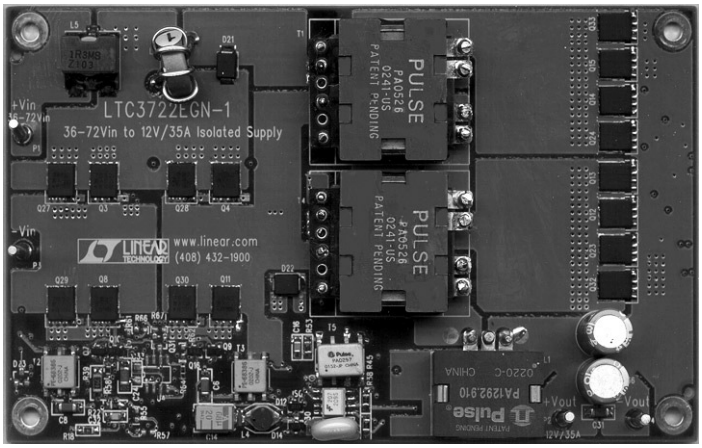

Figure 1. LTC3722-1 36–72V input to 12V/35A isolated power supply

12V Isolated Converter

Figures 1 and 2 show a 36V–72V input to isolated 12V at 35A supply using the LTC3722EGN-1 (24-lead SSOP) and LTC4440 high side drivers. The low profile design features surface mount power MOSFETs and planar transformers (less than 0.4" high) configured in a parallel/series configuration. With a 48V input, this circuit has a typical full load (35A) efficiency of 93%. The PCB board shown in Figure 1 is 3"× 5", and with 200 linear feet per minute of airflow provides full load operation from 48V input to 50°C ambient without the use of a heat sink. The actual PCB area can be reduced further (depending on airflow and ambient temperature) when used as part of a large system board.

Even with 12V output, the synchronous rectifiers reduce the power dissipation in the secondary rectifiers. The output of 12V is chosen here because it is a good intermediate bus voltage, but the circuit is easily modified to meet other input or output voltages. For example, applications not requiring the full input voltage range or tight regulation could further optimize transformer and semiconductor utilization resulting in increased efficiency.

Operation

The start-up of the circuit in Figure 2 begins with C14 trickle charging via R29 and Q41's base-collector junction until U2's V_{IN} pin reaches 10.2V (the internal shunt regulator voltage). Assuming the undervoltage lockout pin (UVLO) is above a 5V threshold, switching begins. C14 keeps the U2's V_{IN} pin above its 6.0V shutdown threshold until a bias winding on T4 (along with D12, D14 and L4) takes over. U4 and U5 provide the level translation for the two high side switches. U2's OUTE and OUTF pins provide synchronous timing signals to the output rectifiers gate drive transformer T5 and the LTC1693 gate driver. The loop is closed by U3, the LT1431 programmable reference, and optocoupler ISO1.

Operation of the phase modulated full bridge converter is similar to a conventional full bridge converter in

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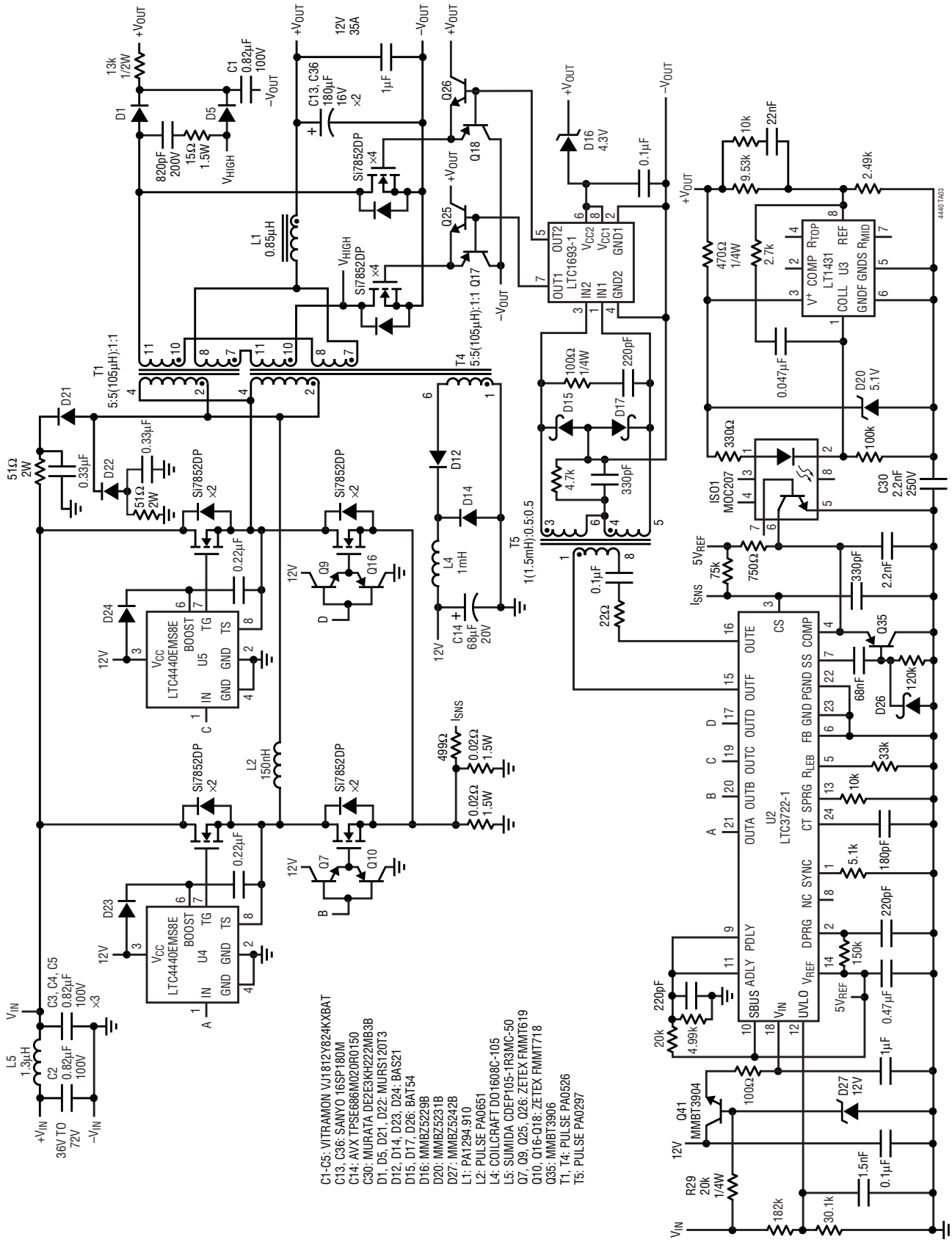

Figure 2. LTC3722-1 36–72V input to 12V/35A isolated synchronous full bridge power supply

Monolithic Buck-Boost Converter Provides 1A at 3.3V without Schottky Diodes

by Mark Jordan

Introduction

The power density and small form factor of lithium-Ion batteries makes them the power source of choice for many portable devices. A SEPIC converter topology is a popular way to provide a regulated bus voltage that falls within the 2.7V to 4.2V battery range, but a SEPIC converter has some flaws. It offers mediocre efficiency and requires coupled inductors and a high current flyback capacitor. The LTC3441 1A buck-boost converter offers a compact and efficient alternative that requires only a single inductor and very few external components.

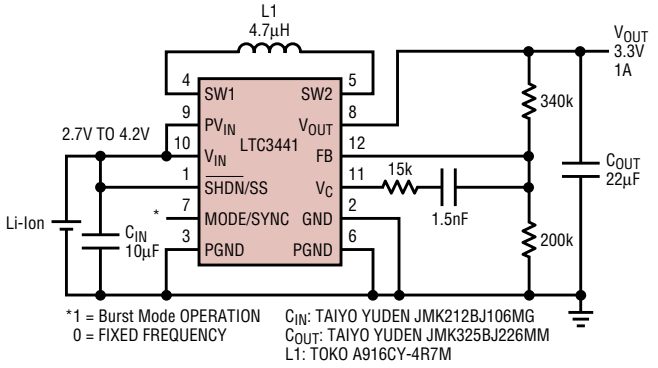

Figure 1. Li-Ion to 3.3V at 1A boost converter

Inside the LTC3441

The LTC3441 patented control technique provides smooth and continuous transfer from buck, buck-boost and boost modes while maintaining a constant frequency at no load. The operating frequency is factory set to 1MHz and can be synchronized up to 1.7MHz. For light loads, the part offers user controlled Burst Mode operation to maximize battery life, drawing only 25µA of quiescent current. To limit inrush current at start-up, an external RC network can be connected to the SHDN/SS pin to control output voltage rise time.

The LTC3441 is available in a small 3mm by 4mm low thermal resistance 12-lead DFN package.

Single Inductor Li-Ion to 3.3V/1A Converter

Figure 1 shows a 3.3W converter powered from a single Lithium-Ion battery. The single inductor topology of the LTC3441, along with all ceramic capacitors, minimizes critical board real estate. Dominant pole compensation is shown as a simple means to compensate the converter's transient response. For applications requiring optimum transient response an additional pole/zero pair to broaden the loop will achieve the desired results. Figure 2 shows that the converter can achieve 95% peak efficiency without the use of Schottky diodes.

Not Just a Buck-Boost

The LTC3441 can also be configured as a boost converter with output disconnect as shown in Figure 3. The 5V at 600mA converter from a Lithium-Ion battery demonstrates peak efficiencies of over 94%. Input current at start-up is also controlled by the LTC3441, reducing the load burden on the battery.

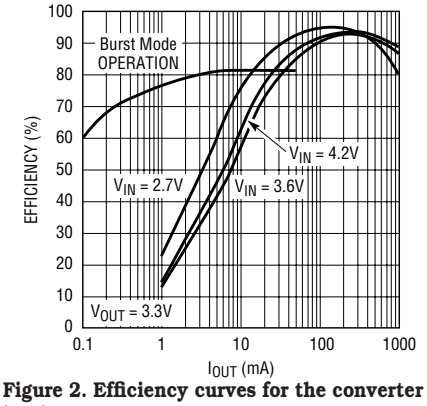

Figure 2. Efficiency curves for the converter in Figure 1

The Schottky diode limits the voltage spikes on the SW2 pin.

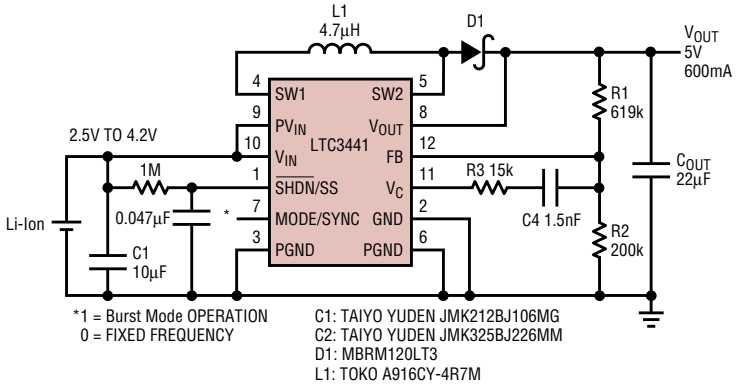

Figure 3. Li-Ion to 5V at 600mA boost converter with output disconnect

Inductorless, Efficient Step-Down DC/DC Converter Provides Dual Low Noise Outputs in Space-Constrained Designs

by Bill Walter

Introduction

Linear Technology's new LTC3252 switched capacitor step-down DC/DC converter squeezes dual adjustable outputs into a space saving 3mm by 4mm DFN package. Each output is programmable within a range of 0.9V to 1.6V, is capable of 250mA of current, and operates from a single 2.7V to 5.5V supply. To keep the converter footprint small, the LTC3252 operates at high frequency, allowing the use of tiny low cost ceramic capacitors—no inductors are required.

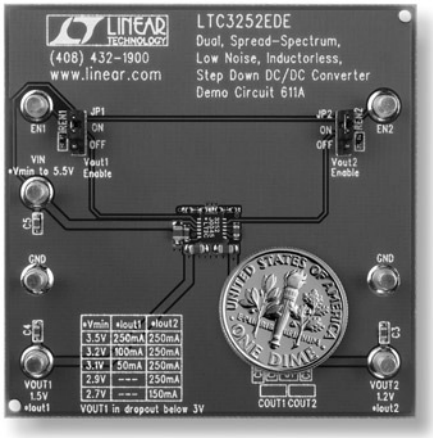

Figure 1. A complete dual output Li-Ion converter

Improve Efficiency and Save Space

The 2-to-1 switched capacitor fractional conversion architecture of the LTC3252 is twice as efficient as a linear regulator, which translates to battery run times that are double that of an LDO. Five tiny ceramic capacitors and four surface mount resistors are all that are required for operation.

Reduce Noise

The LTC3252 employs a unique spread spectrum architecture that continually switches, which not only provides a low input and output noise, but also significantly reduces EMI (Electro-Magnetic Interference). Regulation is achieved by sensing the output voltage and regulating the amount of charge transferred per cycle. This method of regulation provides much lower input and output ripple than that of conventional switched capacitor charge pumps. The spread spectrum feature of the LTC3252 randomly modulates the charge transfer rate between 1.0 MHz and 1.6MHz on a cycle-by-cycle basis. Modulating the frequency in this manner virtually eliminates high frequency harmonic EMI that can be conducted into other circuits.

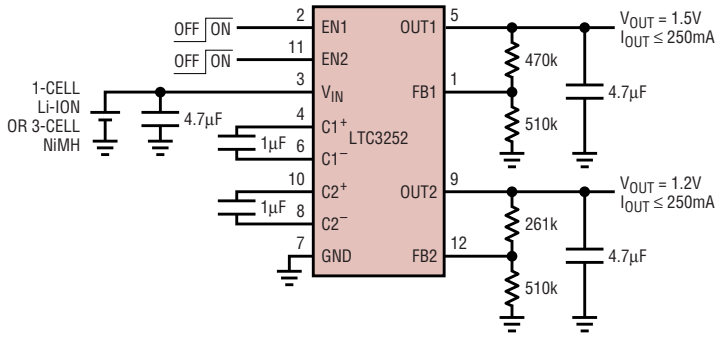

Figure 2. Space saving, low noise, inductorless dual output DC-DC converter: Li-Ion to 1.5V/250mA and 1.2V/250mA

Increase Battery Run Time

The LTC3252 also features Burst Mode® operation, which allows the LTC3252 to achieve high efficiency even with lightly loaded outputs. While in Burst Mode operation the LTC3252 delivers a minimum amount of charge for a few cycles then goes into a low current state until the output drops enough to require another burst of charge. A current sense circuit is used to detect when the required output current of both outputs drops below about 30mA. When this occurs, the oscillator shuts down and the part goes into a low current operating state. The LTC3252 remains in the low current operating state until either output has dropped enough to require another burst of current. The current transferred to the output is limited by internal circuitry, thus providing a nearly fixed output ripple of about $12mV_{P-P}$. The unloaded operating current of the part is just 35µA with one output enabled and 60µA with both outputs enabled.

Circuit Protection Features

The LTC3252 has built-in short-circuit current limiting as well as over temperature protection. During a short-circuit condition the part automatically limits the output current to approximately 500mA. The LTC3252 shuts down and stops all charge transfer when the IC temperature exceeds approximately 160°C. Under normal operating conditions, the part should not go into thermal shutdown but the function is included to protect the IC from excessively high ambient temperatures, or from excessive power dissipation inside the IC (i.e., over-current or short circuit). The

continued on page 37

Using Current Sensing Resistors with Hot Swap Controllers and Current Mode Voltage Regulators

by Eric Trelewicz

Introduction

Current mode switching regulators and Hot Swap controllers—such as the LTC1622 regulator and the LTC4210 Hot Swap controller—use a sub-50mV voltage across a sense resistor in a high-current (Amps to tens of Amps) path to control the current. Failure to properly Kelvin sense the current across the sense resistor is the most common cause of circuit malfunction in current mode power supplies and Hot Swap circuits.

Problems usually arise when the layout of the circuit does not take into account the high currents and small resistances involved. For instance, a low value precision sense resistor in the mΩ range is typically used to measure current in Hot Swap controllers and current mode switching regulators. A typical 0.003Ω, 1W sense resistor in a 2512 surface mount package is only 0.125 inches wide and 0.250 inches long. Consider the same length copper trace with a typical thickness of 0.0014 inches (1oz copper laminate). The resistance across a quarter inch of copper trace is 0.0009Ω at room temperature. Adding a quarter inch of copper to the measurement path induces a sense measurement error of 0.0009Ω/0.003Ω, or 30%! The circuit simply wouldn't work, because it would prematurely trip the current limit.

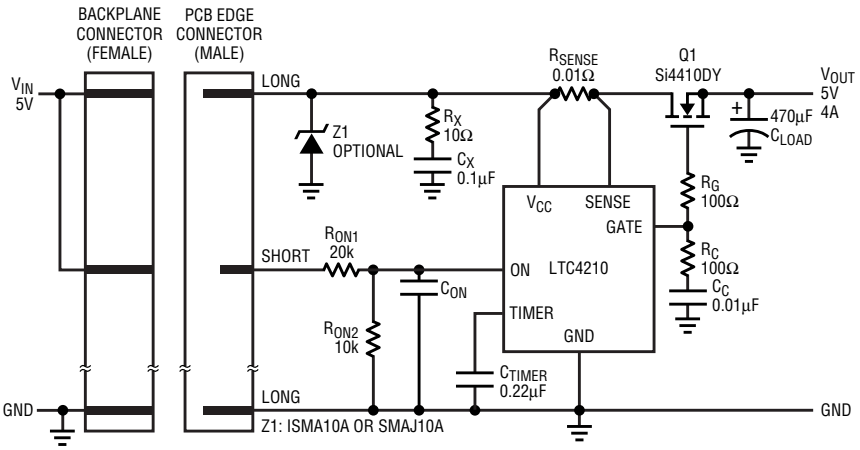

Figure 1. This Hot Swap circuit controls current by sensing the voltage across a small value resistor, R_{SENSE}—a method called Kelvin sensing. The layout of this circuit is important for accurate current sensing.

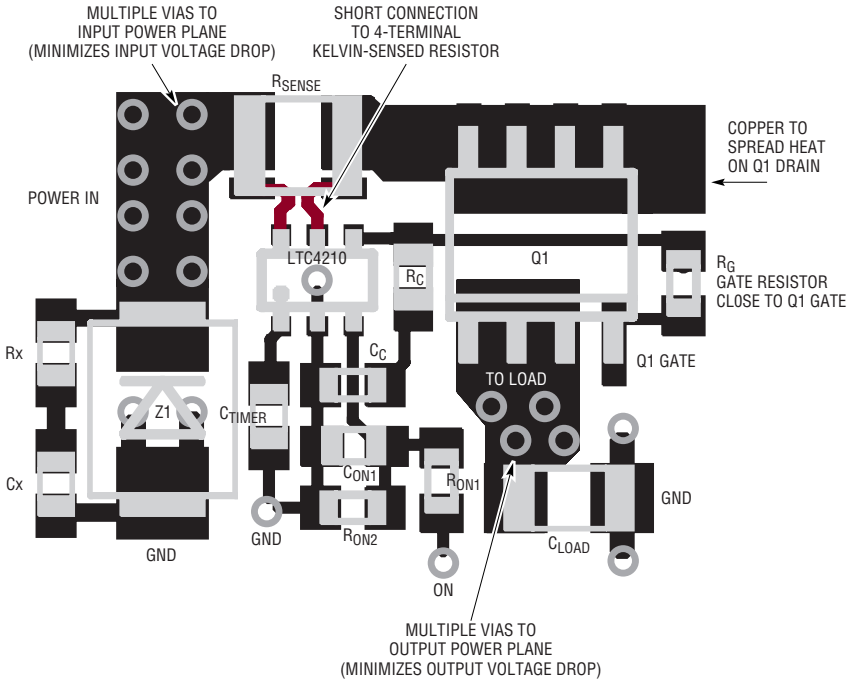

Figure 2. Example of proper Kelvin sense connections. The important features shown here are: short connections to terminals of Kelvin sense resistor, expanded copper to heat sink Q1, and multiple vias connecting power in and the load to pass FET Q1 for minimal voltage drop.

Pitfalls Lie in the Layout

The printed circuit board layout process is full of pitfalls, especially when an auto-router is part of the process. When one terminal of the sense resistor is the power plane, the sense pin on the IC and the terminal of the current sense resistor can end up connected across a significant span of copper, with uncontrolled current flow from other circuits on the board flowing between the resistor and IC connections. Excess voltage drop and noise coupling are a prescription for circuit malfunction. With many designs outsourced to PCB design houses, the circuit designer faces a formidable task of controlling the layout.

What can be done?

Several resistor manufacturers[1] now sell 4-terminal Kelvin current sense

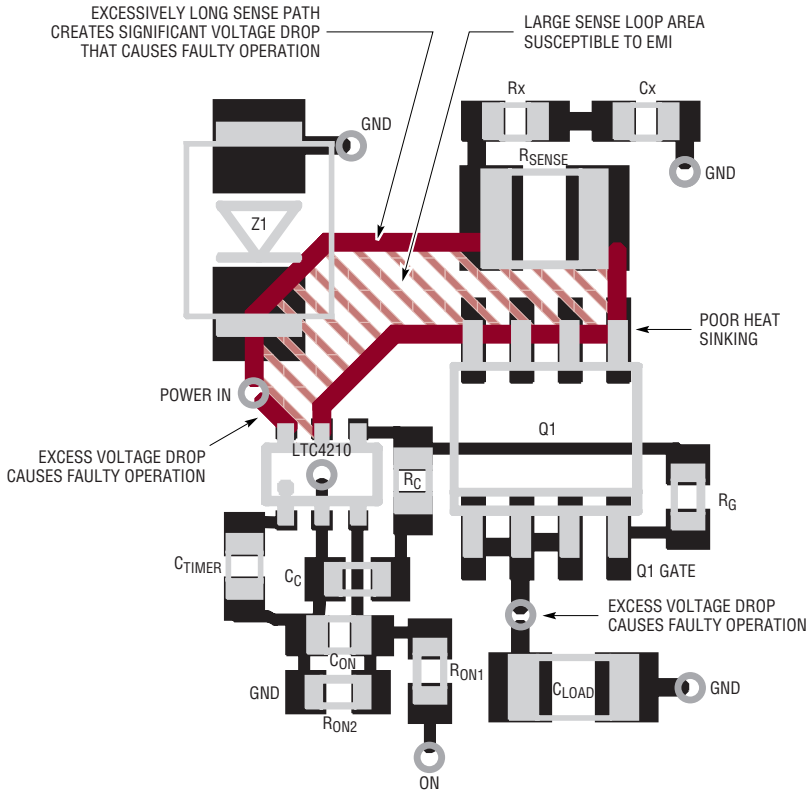

Figure 3. Example of layout that can reduce the accuracy of Kelvin sensing. The problems shown here include: excess length of thin high resistance track in series with sense resistor, inadequate heat sinking on Q1, and an insufficient number of vias for input power and output load connections.

resistors. The use of a 4-terminal resistor forces the auto router to make a correct Kelvin connection to the current sense resistor. But this alone is not enough. High speed switch mode power supplies have a high dI/dt path that can inductively couple with the sense loop and also cause malfunction. To minimize inductive coupling, the Kelvin sense circuit must exhibit minimal loop area.

Setting the Proper Constraints in an Auto-Router

Set the auto-router constraints to route the Kelvin sense connections as a differential pair to keep the connections side by side and close together. Use maximum length constraints to prevent the connections from wandering too far from the direct path. Constrain the connection to the component layer on a multi-layer PC board to prevent unwanted vias in this critical connection path. Although the proper choice of sense resistor and layout constraints can mitigate many of the PCB layout pitfalls, in the end it's up to the designer to carefully check the layout.

Notes

[1] Some sources of 4-terminal Kelvin sensed resistors include:
- www.Caddock.com
- www.IMS-Resistors.com
- www.IRCtt.com
- www.Vishay.com

LTC3722, continued from page 30

that power is delivered when diagonal switches are on. It differs in that during the free-wheeling portion of the switching cycle, either the top or bottom switches of the bridge remain on. This provides for recovery of parasitic energy and zero-voltage turn-on transitions for the primary switches. The LTC3722-1 can be configured to provide adaptive (with programmable time-out) or fixed delay control for zero voltage switching operation. In adaptive DirectSense™ mode, the turn-on timing adjusts automatically by sensing the transition voltages on the bridge legs, eliminating external trims. This provides accurate zero voltage transition timing with changes in input voltage, output load and circuit parasitics. Fixed (or manual) delay control is also available, which allows for fixed transition delays or even custom dynamic timing schemes. The LTC3722-1 also features adjustable synchronous rectifier timing.

Conclusion

The new LTC3722-1 current-mode controller provides a wealth of features targeted at high power isolated full bridge applications, including flexible timing control, synchronous rectifier outputs, under-voltage lockout, programmable slope compensation and current mode leading edge blanking.

LTC4302, continued from page 16

Summary

The LTC4302-1/LTC4302-2 addressable 2-wire bus buffers ease the practical issues associated with complex 2-wire bus systems. They allow I/O cards to be hot-plugged into live systems and break one large capacitive bus into several smaller ones, while still passing the SDA and SCL signals to every device in the system. They can also connect and disconnect different bus segments at different times, providing nested addressing capability and easing the debugging process during stuck low situations.

Complete USB Solution Provides PowerPath Control and Input Current Limiting while Charging a Li-Ion Battery

by Trevor Barcelo

Introduction

An increasing number of portable handheld devices are using the Universal Serial Bus (USB) to exchange data with a host computer. Now, more portable device designs are taking advantage of the power-supply provisions of the USB specification—hosts can supply up to 500mA at a nominal 5V—to power the portable device and charge its batteries. The most effective use of the USB-limited 500mA requires a USB-compliant battery charger and well-designed PowerPath™ controls.

USB Power Specifications

USB power specifications require that a Li-Ion battery charger is able to operate at input voltages as low as 4.75V (ignoring resistive drops in the cable and connectors which further reduce this value to 4.4V), has a low current standby mode, and that it limits the total current drawn from the USB power port to 500mA.

The LTC4056 provides a unique feature allowing it to work well with USB power. The undervoltage charge current limiting function automatically reduces charge current if the input supply voltage drops to approximately 4.575V. This feature prevents resistive drops from lowering the USB voltage below approximately 4.575V. Typically, if a battery charger is connected to a particularly resistive USB cable, the USB voltage drops below the charger's undervoltage lockout threshold as soon as charge current is turned on. This causes the charger to turn off, the USB supply voltage rebounds, and the cycle repeats. If an LTC4056 is connected to a particularly resistive USB cable, the charge current is reduced to ensure that the input voltage does not drop below the undervoltage lockout threshold. In fact, charge current is adjusted to maintain the USB voltage at or above the undervoltage charge current limit voltage of approximately 4.575V.

The TIMER/$\overline{\text{SHDN}}$ pin of the LTC4056 can be used to reduce its supply current to about 40µA allowing it to meet the USB low current standby mode specification.

The final USB requirement is that the device cannot draw more the 500mA from the USB port. Careful PowerPath design is required to *efficiently* meet this requirement.

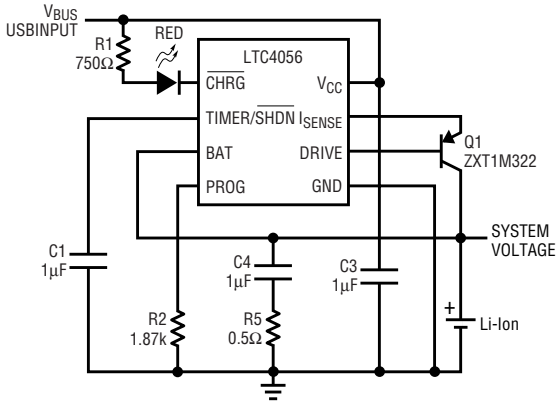

Figure 1. Basic USB charger solution

Basic Solution

Figure 1 shows a basic USB solution. This solution draws system power directly from the battery as it is charging. Since the battery charger is the only system on the device that draws current directly from the USB port, the current limit is set simply by programming the battery charge current. For instance, if the battery charger is programmed to charge at 490mA and the system load is 140mA, then 350mA is effectively used to charge the battery.

This design works within the confines of the USB specification, but it does not maximize efficiency. Here's why. Consider an application with a buck switching regulator that needs to provide a 1.8V supply at 300mA. Powering the buck regulator directly from the battery (at a nominal voltage of 3.85V) would require approximately 140mA, leaving 350mA to charge the battery (assuming the same 490mA charge current as above). On the other hand, powering the buck regulator from the USB supply (nominally 5V) would require just 110mA leaving 380mA to charge the battery, 30mA more. This additional 10% charge current can reduce the battery charge time by about 10%.

PowerPath Solution

Figure 2 shows an application allowing the peripheral to draw current from the USB supply when it is present, and otherwise, draw current from the battery. The LTC4056 actually regulates the current output from the I_{SENSE} pin (rather than the BAT pin current). This feature allows the 500mA maximum USB power port consumption to be easily enforced by tying *all* system loads to the I_{SENSE} pin and programming the charger to

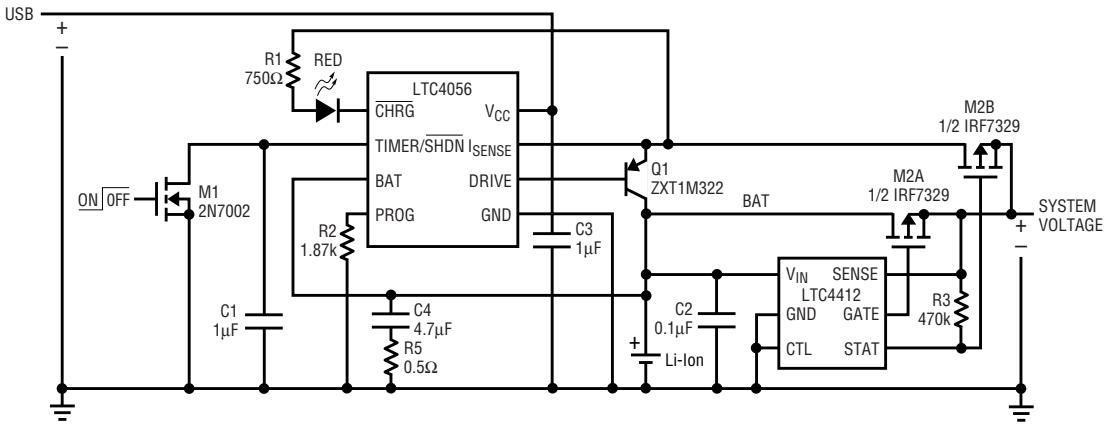

Figure 2. Complete USB PowerPath control and battery charger solution

supply just under 500mA (in Figure 2 the charger is programmed for 490mA). The total impedance between the V_{CC} pin and I_{SENSE} pin is typically 0.2Ω, so the maximum drop is just 100mV (at 500mA) allowing the peripheral device to operate at a voltage significantly higher than a single Li-Ion battery when the USB supply is present.

It is important to keep in mind that the LTC4056 can only control charge current. If the system load is less than 500mA, then the LTC4056 simply reduces the charge current by an amount equal to the system load current. For instance, if the system load is 110mA, then the charge current is reduced from 490mA to 380mA to keep the total USB input current at 490mA, thereby meeting the specification. The 110mA system load, however, is now being provided at approximately 5V rather than the battery voltage. Assuming a nominal battery voltage of 3.85V, the circuit in Figure 2 can provide approximately 23% more power to the system than the circuit in Figure 1 for a given battery charge current.

Of course, if the system load is increased beyond 500mA the LTC4056 will reduce the charge current to zero, and all of the system load will be provided by the USB input. This scenario violates the USB power specification. In order to avoid this situation, it is important to ensure that the system load never exceeds 500mA.

Note that even the $\overline{CHRG}$ LED is connected to I_{SENSE}. This is a good example of a peripheral load current. When the LTC4056 is charging a battery, the $\overline{CHRG}$ pin is pulled low drawing 4mA to 5mA through R1. This load current reduces the amount of current delivered to the battery by an equal amount.

To ensure that the system voltage is always present (even when USB power is not), the LTC4412 provides automatic switchover of the system load between a battery and the USB input supply. This feature reduces the current drain on the battery to just a few microamps when a USB input is present. Figure 2 shows a dual FET solution to minimize voltage drop between the USB input voltage and the system voltage (a Schottky diode can also be used in place of M2B). The combination of the LTC4412 and M2A forms an ideal diode from BAT to SYSTEMVOLTAGE. M2B serves as a switch that is ON when the ideal diode is not conducting and OFF otherwise. Therefore, as long as the USB input is present and the voltage on the I_{SENSE} pin is higher than BAT, M2B is ON and M2A is OFF. As soon as the USB input supply drops below the battery voltage, M2A turns on and acts as an ideal diode.

Conclusion

The USB specification allows for up to 500mA of current to be delivered from the USB port. Portable devices are increasingly using the USB power provided by a host computer to power the device system bus and to charge batteries. When used in a PowerPath control configuration, the LTC4056 makes the most of this 500mA to efficiently charge the battery, even while the system draws power from the USB.

LTC3252, continued from page 33

charge transfer reactivates once the junction temperature drops back to approximately 150°C. The LTC3252 can cycle in and out of thermal shutdown, without latch-up or damage, until the fault condition is removed.

The EN1 and EN2 pins are used to individually enable OUT1 and OUT2 respectively. When both EN pins are low the outputs become high impedance and all control circuitry is disabled leaving only a few nanoamps of supply current. The LTC3252 includes a soft-start feature that limits the inrush currents required to charge the output capacitor when an output is enabled, thereby minimizing input supply transients caused by the power on phase of the IC. The soft-start is implemented whenever an output is brought out of shutdown.

Conclusion

The LTC3252 is well suited for medium to low power step-down applications requiring multiple low noise outputs in a small footprint. It is an especially good match for single cell Li-Ion and multi-cell NiMH/NiCd battery powered applications and where EMI is a concern.

New Device Cameos

Data/Clock Hot Swap Devices Provide Capacitance Buffering, Level Translation on 2-Wire Bus Systems

The LTC4300A-1 and LTC4300A-2 enable I/O card insertion into a live backplane without corruption of the data and clock lines (SDA and SCL) for I^2C and SMBus systems. The LTC4300A is typically used at the edge of a peripheral card, with two of its pins, SDAOUT and SCLOUT, connected to the data and clock buses on the card. When the card is plugged into a live backplane, two other pins, SDAIN and SCLIN connect to the data and clock buses on the backplane. Control circuitry provides a glitch-free connection by preventing the backplane buses from being connected to the card buses until data transactions on both sides are complete.

Once the backplane is connected to the card, the functionality of the SDAIN and SDAOUT pins is identical. A low forced on either pin at any time results in both pin voltages being low. SDAIN and SDAOUT enter a logic high state only when all devices on both SDAIN and SDAOUT force a high. The same is true for SCLIN and SCLOUT. This important feature ensures that clock stretching, arbitration, synchronization, and data acknowledge always work, regardless of how devices in the system are tied to the LTC4300A-1 or LTC4300A-2.

Another important feature of the connection circuitry is that, while it joins the two buses together, it still maintains electrical isolation between them, thus providing capacitance buffering for both sides. This means that devices on the backplane must drive only the backplane capacitance plus the low capacitance of the LTC4300A (around 10pF). The LTC4300A drives the capacitance of the rest of the I/O card. Likewise, devices on the card must only drive the capacitance of the card plus the low capacitance of the LTC4300A. The LTC4300A drives the capacitance on the backplane. The LTC4300A is capable of driving capacitive loads ranging from 0pF to 1000pF on all of its data and clock pins.

Other features of the part include: rise time accelerator circuitry, which allows the use of weaker DC pull-up currents while still meeting rise time requirements; and pre-charge circuitry, which initializes the data and clock buses to 1V before card insertion. The LTC4300A2-1 features a digital ENABLE input pin, which disconnects the backplane from the card and forces the part into a low current mode; and an open drain READY output pin, which indicates whether the backplane and card sides are connected together. The LTC4300A-2 features independent power supply inputs for the backplane and card, enabling level translation between 3.3V and 5V systems. The LTC4300A-1 and LTC4300A-2 are both available in a small 8-pin MSOP package.

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

1-800-4-LINEAR

Ask for the pertinent data sheets and Application Notes.

High Speed, High Voltage, High Side Gate Driver

The LTC4440 is a high frequency, high side N-channel MOSFET gate driver that is designed to operate in applications with input voltages up to 80V. It is able to withstand, and continues to function through 100V input supply transients, making it ideally suited for use in telecommunications power supplies.

The LTC4440 receives a ground-referenced, low voltage input logic signal that is internally level-shifted to drive an external high-side MOSFET. The input stage has TTL/CMOS compatible thresholds with 350mV of hysteresis. The LTC4440 has propagation delays from input to output of less than 30ns, making it ideal for high frequency applications.

Designed to drive standard threshold MOSFETS, the LTC4440 operates with a bootstrapped supply voltage of 8V to 15V. Its powerful drive capability reduces switching losses in MOSFETs with high gate capacitances. Its pull-up has a peak output current of 2.4A, and its pull-down has an output impedance of 1.6Ω. The LTC4440 can drive a 1000pF load with a 10ns rise time and 7ns fall time with 12V of gate drive.

To protect the external MOSFET from inadequate supply voltages, the LTC4440 features both high-side and low-side undervoltage lockout circuits that disable the external MOSFET when activated.

This rugged architecture, high speed, and powerful drive capability are available in a 6-lead low profile ThinSOT package or a thermally enhanced 8-lead MSOP package.

Programmable Supply Current, Rail-to-Rail Output, Current Feedback Amplifier

The LT6210 is a current feedback amplifier with externally programmable supply current and bandwidth ranging from 10MHz at 300µA to 200MHz at 6mA. It features a low distortion rail-to-rail, C-Load™-stable output stage, 700V/µs slew rate and a minimum output current drive of 75mA.

The LT6210 operates on supplies as low as a single 3V and up to either 12V or ±6V. The I SET pin allows for the optimization of quiescent current for specific bandwidth, distortion or slew rate requirements. Regardless of supply voltage, the supply current is programmable from just 300µA to 6mA with an external resistor or current source.

The LT6210 is manufactured on Linear Technology's proprietary low voltage complementary bipolar process and is available in the low profile (1mm) 6-lead ThinSOT package.

DESIGN TOOLS

2003 Databooks

Linear Technology is pleased to announce the availability of seven databooks organized by product family. This set supersedes all previous Linear databooks. Each databook contains all related product data sheets, selection guides, QML/space information, package information, appendices, and a complete reference to all of the other family databooks.

For more information, or to obtain any of the databooks, contact your local sales office (see the back of this magazine), or visit www.linear.com.

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Linear Applications Handbook, Volume I—Almost a thousand pages of application ideas covered in depth by 40 Application Notes and 33 Design Notes. This catalog covers a broad range of real world linear circuitry. In addition to detailed, systems-oriented circuits, this handbook contains broad tutorial content together with liberal use of schematics and scope photography. A special feature in this edition includes a 22-page section on SPICE macromodels.

Linear Applications Handbook, Volume II—Continues the stream of real world linear circuitry initiated by Volume I. Similar in scope to Volume I, this book covers Application Notes 40 through 54 and Design Notes 33 through 69. References and articles from non-LTC publications that we have found useful are also included.

Linear Applications Handbook, Volume III—This 976-page handbook includes Application Notes 55 through 69 and Design Notes 70 through 144. Subjects include switching regulators, measurement and control circuits, filters, video designs, interface, data converters, power products, battery chargers and CCFL inverters. An extensive subject index references circuits in Linear data sheets, design notes, application notes and *Linear Technology* magazines.

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www.linear.com—Customers can quickly and conveniently find and retrieve the latest technical information covering the Company's products on Linear's website. Located at www.linear.com, the site allows searching of data sheets, application notes, design notes, *Linear Technology* magazine issues and other LTC publications. The LTC website simplifies searches by providing three separate search engines. The first is a quick search function that provides a complete list of all documentation for a particular word or part number. There is also a product function tree that lists all products in a given product family. The most powerful, though, is the parametric search engine. It allows engineers to specify key parameters and specifications that satisfy their design requirements. Other areas within the site include a sales office directory, press releases, financial information, quality assurance documentation, and corporate information.

Linear Direct Online Store—The Linear Online Store at www.linear.com now offers a simple way to order LTC products factory direct. The new store accepts major credit cards and allows customers to create personalized accounts where they can check order history, shipment information and reorder products. The maximum quantity per order has increased to 500.

Brochures

Power Management & Wireless Solutions for Handheld Products—The solutions in this product selection guide solve real-life problems for cell phones, digital cameras, PDAs and other portable devices. Circuits are shown for Li-Ion battery chargers, battery managers, USB support, system power regulation, display drivers, white LED drivers, photoflash chargers, DC/DC converters, SIM and smart card interfaces, photoflash chargers, and RF PA power supply and control. All solutions are designed to maximize battery run time, save space and reduce EMI where necessary—important considerations when designing circuits for handheld devices.

Software

SwitcherCAD™ III/LTC SPICE—LTC SwitcherCAD III is a fully functional SPICE simulator with enhancements and models to ease the simulation of switching regulators. This SPICE is a high performance circuit simulator and integrated waveform viewer, and also includes schematic capture. Our enhancements to SPICE result in much faster simulation of switching regulators than is possible with normal SPICE simulators. SwitcherCAD III includes SPICE, macromodels for 80% of LTC's switching regulators and over 200 op amp models. It also includes models of resistors, transistors and MOSFETs. With this SPICE simulator, most switching regulator waveforms can be viewed in a few minutes on a high performance PC. Circuits using op amps and transistors can also be easily simulated. Download at www.linear.com

FilterCAD™ 3.0—FilterCAD 3.0 is a computer aided design program for creating filters with Linear Technology's filter ICs. FilterCAD is designed to help users without special expertise in filter design to design good filters with a minimum of effort. It can also help experienced filter designers achieve better results by playing "what if" with the configuration and values of various components and observing the results. With FCAD, you can design lowpass, highpass, bandpass or notch filters with a variety of responses, including Butterworth, Bessel, Chebychev, elliptic and minimum Q elliptic, plus custom responses. Download at www.linear.com

SPICE Macromodel Library—This library includes LTC op amp SPICE macromodels. The models can be used with any version of SPICE for analog circuit simulations. These models run on SwitcherCAD III/LTC SPICE.

Noise Program—This PC program allows the user to calculate circuit noise using LTC op amps, determine the best LTC op amp for a low noise application, display the noise data for LTC op amps, calculate resistor noise and calculate noise using specs for any op amp.

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