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LTC1733: Thermal Regulation Maximizes Lithium-Ion Battery Charging Rate Without Risk of Overheating

by Trevor Barcelo

Introduction

Linear battery chargers are typically smaller, simpler and less expensive than switcher-based solutions, but they have one major disadvantage: excessive power dissipation. When the input voltage is high and the battery voltage is low (discharged battery) a linear charger could generate enough heat to damage itself or other components. Typically, such conditions are temporary—as the battery voltage rises with its charge—but it is worst case situations that one must account for when determining the maximum allowable values for charge current and IC temperature. To solve this problem, the LTC1733 employs internal thermal feedback to regulate the charge current and limit the die temperature. This feature translates to faster charge times, because a designer can program a high charge current (to minimize charging time) without the risk of damaging the LTC1733 or any other components. The need for thermal over-design is also eliminated. To further improve heat transfer, the LTC1733 is housed in a thermally enhanced 10-pin MSOP package. For simplicity, the LTC1733 provides a complete lithium-ion charger solution requiring only three external components, as shown in Figure 1.

An internal power MOSFET allows charge current to be programmed up

continued on page 3

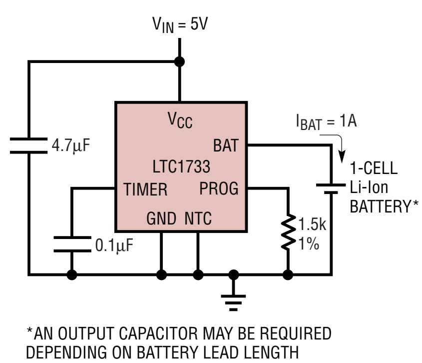

Figure 1. Standalone Li-Ion battery charger

Issue Highlights

Our cover article introduces the LTC1733 battery charger, which employs internal thermal feedback to regulate the charge current and limit the die temperature. This feature translates to faster charge times, because a designer can program a high charge current (to minimize charging time) without the risk of damaging the LTC1733 or any other components. The need for thermal over-design is also eliminated. To further improve heat transfer, the LTC1733 is housed in a thermally enhanced 10-pin MSOP package. For simplicity, the LTC1733 provides a complete lithium-ion charger solution requiring only three external components.

The remainder of the Design Features section presents a variety of power products:

The LT3420 is a power IC that is designed for charging large-valued capacitors to high voltages, such as those used for the strobe flashes of digital and film cameras. Using the LT3420, only a few external components are necessary to create a complete solution, which saves valuable board space in ever shrinking camera designs.

The LTC3411 DC/DC converter provides features that shrink total solution size enough to fit into the latest portable electronics. Its capable of switching frequencies as high as 4MHz, allowing the use of smaller and less costly capacitors and inductors to complete the circuit. It also saves space by placing the switcher and MOSFETs in a small monolithic package.

The LTC3701 is another space saving power product. It is an efficient, low input voltage, dual DC/DC controller that fits into tight spaces. It uses 2-phase switching techniques to reduce required input capacitance (saving space and cost) and increase efficiency. The versatile LTC3701 accepts a wide range of input voltages, from 2.5V to 9.8V, making it useful for single lithium-ion cell and many multicell systems. It can provide output voltages as low as 0.8V and output currents as high as 5A.

The LTC4400-1 and LTC4401-1 provide RF power controller solutions for the latest cellular telephones. They feature very small footprints, low power consumption and wide frequency ranges while minimizing adjacent channel interference by carefully controlling RF power profiles. The LTC4400-1 and LTC4401-1 are both available in a low profile 6-pin ThinSOT package, and require few external parts. For example, when used with a directional coupler, only two resistors and two capacitors are required. Both devices require minimal power to operate, typically 1mA when enabled and 10µA when in shutdown.

Starting on page 21 are nine new Design Ideas covering a variety of applications, from a lightweight portable altimeter, to a way to create a VCO from the LTC6900 precision oscillator, to a simple way to create two lowpass filters out of a single filter IC. See page 21 for a complete list of the Design Idea articles.

At the back are seven New Device Cameos. See www.linear.com for complete device specifications and more applications information.

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.

LTC in the News...

On April 16, Linear Technology Corporation announced its financial results for the 3rd quarter of fiscal year 2002. According to Robert H. Swanson, Chairman of the Board and CEO, "For the first time in several quarters, all of the critical financial trends showed healthy improvement. Sales and profits grew sequentially 7% and 12% respectively over the previous quarter. Bookings, which exceeded sales, grew in all major geographical and major end markets. Even at these reduced sales levels from last year, we continue to be strongly profitable with a 40% after tax return on sales. In January, we discontinued production in our oldest wafer fabrication plant. The associated costs had been previously provided for in past financial statements, and therefore, no special one-time charges were required.

Looking forward, we have seen a broad based increase in our bookings activity throughout the quarter. However, our backlog, while improving, is still low, and global economic and political conditions continue to be tenuous. Therefore, confidently and accurately forecasting future financial results remains difficult. However, based on analysis of the data available to us, we believe excess inventory of our product at customers continues to be worked off and we expect bookings to continue to improve. Consequently, we forecast sales and profits to grow sequentially in the 8% to 10% range from the quarter just completed."

The Company reported net sales of $130,155,000 and net income of $51,480,000 for the quarter ended March 31, 2002. Diluted earnings were $0.16 per share.

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LTC1733, continued from page 1

to 1.5A, with 7% accuracy, to ensure a fast and complete charge. The internal MOSFET also eliminates the need for an external current sense resistor or blocking diode. The final float voltage is pin selectable to either 4.1V or 4.2V with 1% accuracy to prevent dangerous overcharging or reduced battery capacity due to undercharging.

Following battery manufacturers' guidelines, the LTC1733 includes a programmable charge termination timer and thermistor input for temperature qualified charging. Status outputs include C/10 charge detection to indicate a near end-of-charge condition, wall adapter present detection to determine whether charging may proceed or not, charge current monitoring for gas gauging, and fault detection for identifying bad cells. Low battery charge conditioning (trickle charging) safely charges an over-discharged cell, and automatic recharge ensures that the battery is always fully charged. To conserve battery power, the LTC1733 battery drain current drops to less than 5µA when a wall adapter is not present or when the part is shutdown.

Charging a Battery

To charge a single cell Li-ion battery, the user must apply an input voltage (typically, a wall adapter) of at least 4.5V to the V_{CC} pin. The $\overline{ACPR}$ pin will subsequently pull low to indicate that the input voltage condition has been met. Furthermore, a 1% resistor must be connected from PROG to GND to program the nominal charge current to $1500V/R_{PROG}$. The $\overline{CHRG}$ pin will then pull low to indicate that a charge cycle has commenced. A capacitor connected between the TIMER pin and GND will program the charge termination time to 3 hours per 100nF.

If the BAT pin voltage is below 2.48V at the beginning of a charge cycle then the charge current will be one-tenth of the programmed value in order to safely bring the cell voltage high enough to allow full charge current. If the cell is damaged, and the voltage does not rise above 2.48V

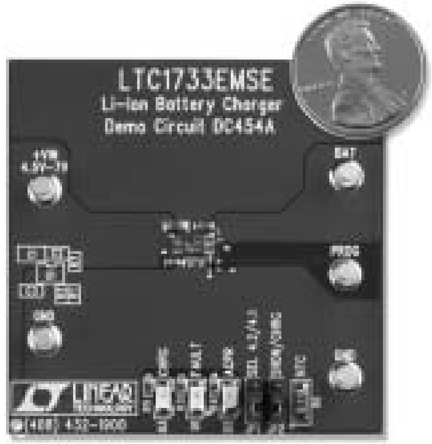


Figure 2. Full featured single cell Li-Ion charger

within one-quarter of the programmed termination time, the charge cycle will terminate, and the $\overline{FAULT}$ status output will latch low indicating a bad cell. All three of these status output pins, $\overline{ACPR}$, $\overline{CHRG}$ and $\overline{FAULT}$, have enough current sinking capability to light an LED.

Once the battery voltage rises above 2.48V (which typically occurs soon after the start of a charge cycle), the LTC1733 will provide a constant current to the battery as programmed by R_{PROG}. The LTC1733 will remain in constant-current mode until the BAT pin voltage approaches the selected final float voltage (4.1V for SEL = 0V and 4.2V for SEL = V_{CC}). At this point the part enters constant-voltage mode.

In constant-voltage mode, the LTC1733 begins to decrease the charge current to maintain a constant voltage at the BAT pin rather

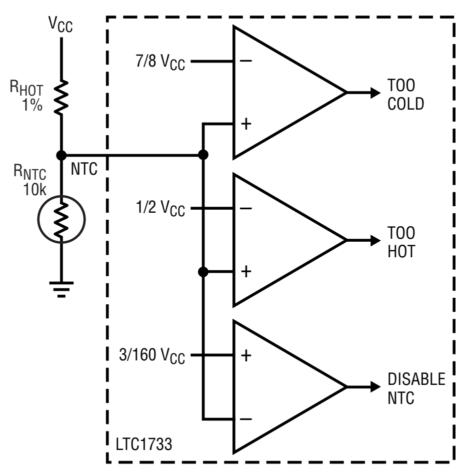


Figure 3. Temperature qualification circuitry

than a constant current out of the BAT pin. When the current drops to 10% of the full-scale programmed charge current, an internal comparator latches off the strong pull-down at the $\overline{CHRG}$ pin and connects a weak current source (about 25µA) to ground to indicate a near end-of-charge (C/10) condition.

Unlike battery chargers that terminate when the current reaches C/10, the LTC1733 continues to charge the battery after the C/10 point, as long as the termination time has not elapsed, to ensure that the battery is fully charged. Terminating charging at C/10 can leave a battery charged to only 90% to 95% capacity, while charging past C/10 and terminating based on time can charge a battery to 100% capacity. Upon termination, the $\overline{CHRG}$ pin assumes a high impedance state.

Recharging a Battery

The LTC1733 has the ability to recharge a battery assuming that the battery voltage has been charged above 3.95V (SEL = 0V) or 4.05V (SEL = V_{CC}) during the initial charge cycle. Once above these thresholds, a new charge cycle begins if the battery voltage drops below 3.9V (SEL = 0V) or 4.0V (SEL = V_{CC}) due to either a load on the battery, or the self-discharge current of the battery. The recharge circuit integrates the BAT pin voltage for a few milliseconds to prevent transients from restarting the charge cycle. This feature ensures that the battery remains charged even if left connected to the powered charger for very long periods of time.

Thermal Regulation

An additional key feature of the LTC1733 is the internal thermal regulation loop. If high power operation and/or high ambient temperature conditions cause the junction temperature of the LTC1733 to approach 105°C, the charge current is automatically reduced to maintain the junction temperature at roughly 105°C (board temperatures typically remain below about 85°C). This is called constant-temperature mode.

This feature allows the user to program a charge current based on typical operating conditions and eliminates the need for the complicated thermal over-design necessary in many linear charger applications. Worst-case conditions are automatically taken care of by the LTC1733. In addition to protecting the LTC1733, this feature eliminates “hot spots” on the board, thereby protecting surrounding components. The thermal shutdown features of other battery chargers simply turn off the charger at very high temperatures (typically, in excess of 130°C). This junction-temperature-based type of shutdown allows both the battery charger and the surrounding board to get extremely hot, so even though the shutdown “protection” exists, the application must be painstakingly designed to avoid reaching the thermal shutdown temperature under all scenarios. The LTC1733 simplifies thermal design by automatically balancing charge current, power dissipation and operating temperature.

To further improve the thermal performance of the LTC1733, it is packaged in a 10-pin thermally enhanced MSOP package. The application board pictured in Figure 2 occupies just 76mm^2 of board space and can dissipate over 2W of power at room temperature. That equates to a maximum charge current of about 1.5A, with a 5V input supply. This assumes that a Li-ion battery spends most of its time at 3.7V during charge. In fact, this is a conservative assumption, since a typical Li-ion battery will rise above 3.8V within the first few minutes of charging. The powerful thermal features of the LTC1733 and the 7% accuracy of the programmed charge current allow very fast and accurate charging of single cell Li-ion batteries.

PROG Current Monitor

For gas gauging applications, the PROG pin provides very accurate information regarding the current flowing out of the BAT pin. The relationship is given by:

$$I_{BAT} = (V_{PROG}/R_{PROG}) \bullet 1000$$

During constant-current mode, the PROG pin voltage is always 1.5V, indicating that the programmed charge current is flowing out of the BAT pin. In constant-temperature or constant-voltage mode, the BAT pin current is reduced and can be determined by measuring the PROG pin voltage and applying the above formula. The PROG pin, along with the three open-drain status outputs ($\overline{\text{ACPR}}$, $\overline{\text{CHRG}}$, and $\overline{\text{FAULT}}$), inform the user of exactly what the LTC1733 is doing at all times.

NTC Thermistor

In addition to the programmable timer and low battery charge qualification, the LTC1733 adds temperature qualified charging to the list of battery manufacturer recommended safety features. The battery temperature is measured by placing a negative temperature coefficient (NTC) thermistor close to the battery pack. Using the circuitry shown in Figure 3, the LTC1733 can temporarily suspend the internal timer and stop charging when the battery temperature falls below 0°C or rises above 50°C. To perform this function, R_{HOT} should be chosen to be the value of the selected NTC thermistor at 50°C. This will ensure that the internal comparator’s trip point of $1/2V_{CC}$ corresponds to an NTC temperature of 50°C. Furthermore, the selected NTC thermistor should have a value at 0°C that is as close to seven times the value at 50°C as possible. A 7:1 cold to hot NTC ratio ensures that the internal comparator’s trip point of $7/8V_{CC}$ corresponds to an NTC temperature of 0°C. The hot and cold comparators each have approximately 2°C of hysteresis to prevent oscillation about the trip point. In addition, the NTC function can be disabled without any external components by simply grounding the NTC pin.

Conclusion

The LTC1733 is a full-featured, standalone Li-ion battery charger. In its simplest form, the LTC1733 only requires three external components and can safely and accurately charge high-capacity batteries very quickly with up to 1.5A of charge current. An NTC thermistor and a few LEDs can be added to take advantage of the safety and status features.

LT3420 Charges Photoflash Capacitors Quickly and Efficiently While Using Minimal Board Space

by Albert Wu

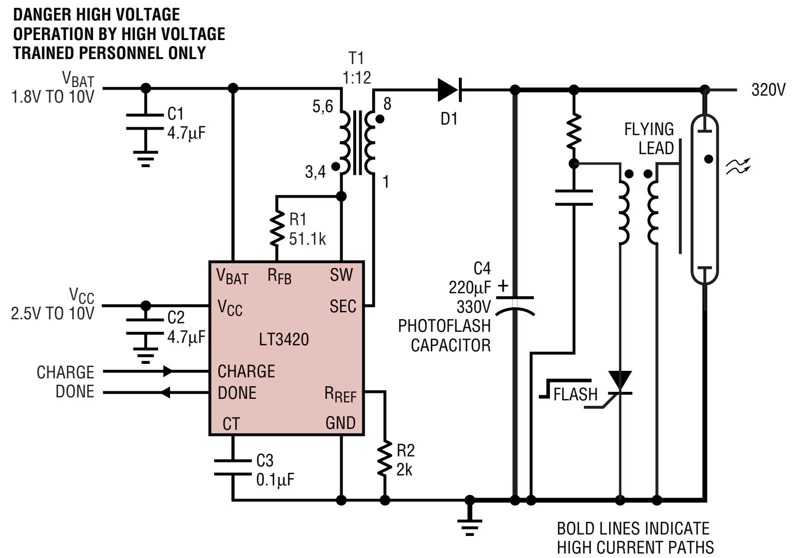
Introduction

The LT3420 is a power IC, designed primarily for charging large-valued capacitors to high voltages, such as those used for the strobe flashes of digital and film cameras. These capacitors are generally referred to as photoflash or strobe capacitors and range from values of a hundred microfarads to a millifarad, with target output voltages above 300V. The photoflash capacitor is used to store a large amount of energy, which can be released nearly instantaneously to power a xenon bulb, providing the light necessary for flash photography. Traditional solutions for charging the photoflash capacitor, such as the self-oscillating type, are extremely inefficient. More modern techniques use numerous discrete devices to implement a flyback converter but require a large board area and suffer from high peak currents, reducing battery life. The LT3420 incorporates a low resistance integrated switch and utilizes a new patent-pending control technique to solve this difficult high voltage power problem. Using the LT3420, only a few external components are necessary to create a complete solution, which saves valuable board space in ever shrinking camera designs. Efficiency of the LT3420 is high, typically greater than 75%, while the peak current of the part is well controlled, important features for increasing battery life.

Overview

Figure 1a shows a photoflash application for the LT3420. To generate the high output voltage required, the LT3420 is designed to operate in a flyback switching regulator topology. The LT3420 uses an adaptive on-time/off-time control scheme resulting in excellent efficiency and precise control of switching currents. The LT3420 can charge a 220µF capacitor from 50V to 320V in 3.5s from a 5V input, as shown in Figure 1b. Charge time decreases with higher V_{IN}, as shown in Figure 1c. 50V is used as the starting point in calculating charge time since the xenon bulb will self exinguish at this voltage, halting any further voltage drop on the photoflash capacitor.

In Figure 1a, the circuitry to the right of C4 shows a typical way to generate the light pulse once the photoflash capacitor is charged. When the SCR is fired, the flying lead placed next to the xenon bulb reaches many kilovolts in potential. This ionizes the



C1, C2: 4.7µF, X5R or X7R, 10V
C4: RUBYCON 220µF PHOTOFLASH CAPACITOR (858) 496-8990
T1: TDK SRW10EPC-U01H003 FLYBACK TRANSFORMER (408) 392-1400
D1: GENERAL SEMICONDUCTOR GSD2004S SOT-23 DUAL DIODE. DIODES CONNECTED IN SERIES (516) 847-3000

Figure 1a. 320V photoflash capacitor charging circuit

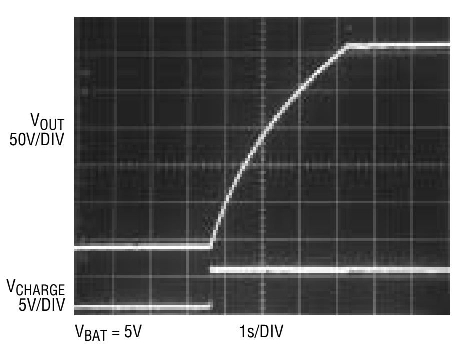


Figure 1b. Charging waveform

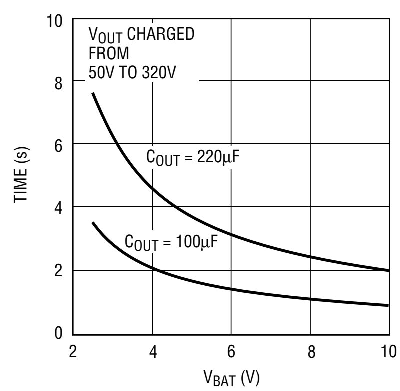


Figure 1c. Charge time

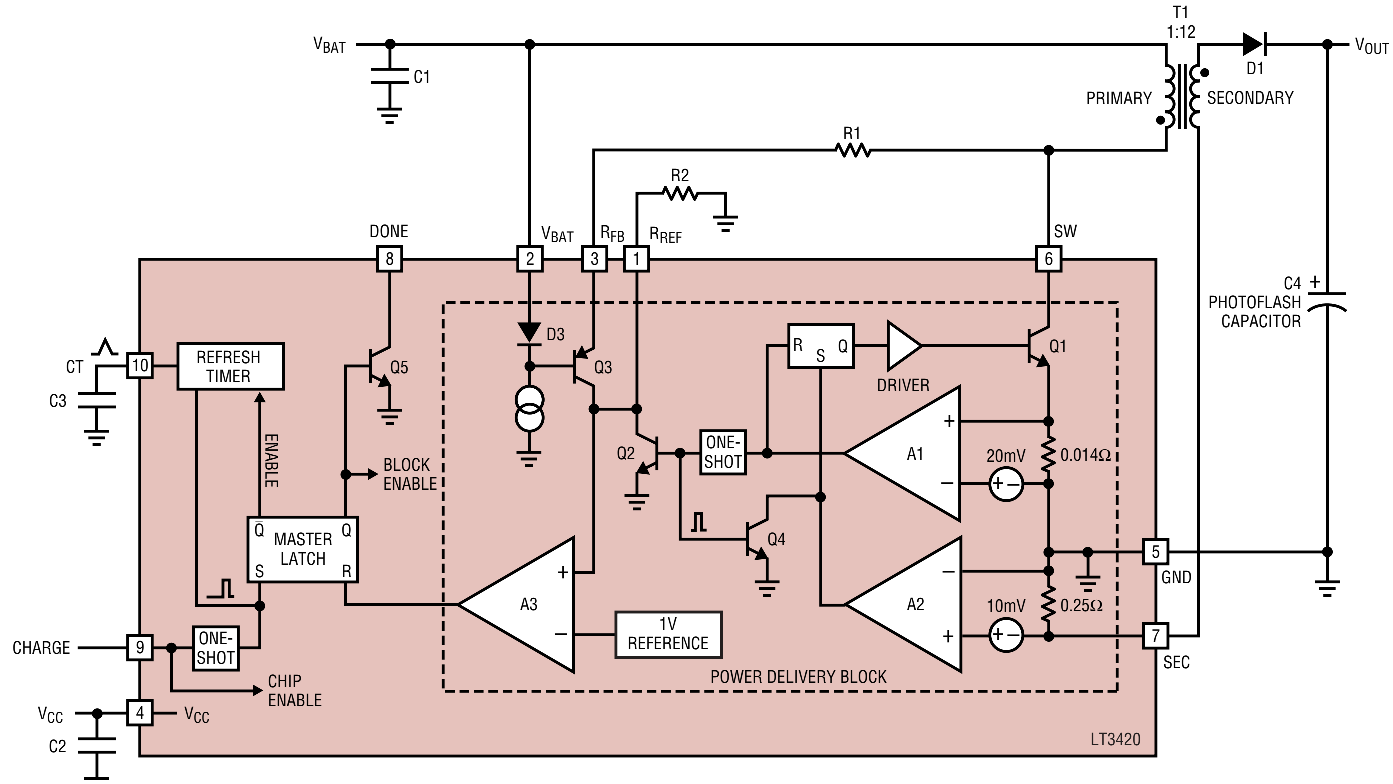


Figure 2. Simplified block diagram of the LT3420

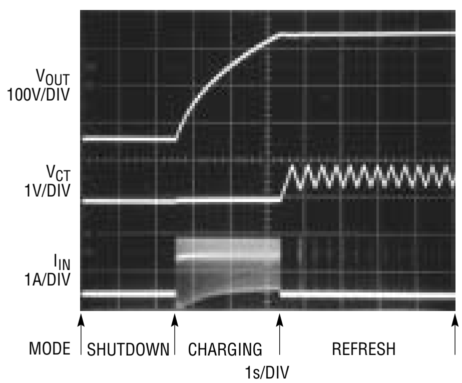


Figure 3. The three operating modes of the LT3420: shutdown, charging, and refresh of the photoflash capacitor

gas inside the bulb forming a low impedance path across the bulb. The energy stored in the photoflash capacitor quickly flows through the Xenon bulb, producing a burst of light. It is important to implement the ground routing shown in Figure 1a, because during a flash, hundreds of amps can flow in the traces indicated by bold lines. Improper ground routing can result in erratic behavior of the circuit.

Figure 2 shows a simplified block diagram of the LT3420. At any given instant, the Master Latch determines which one of two modes the LT3420 is in: "Power Delivery" or "Refresh." In Power Delivery Mode, the circuitry enclosed by the smaller dashed box is enabled, providing power to charge photoflash capacitor C4. The output voltage is monitored via the flyback pulse on the primary of the transformer. Since no output voltage divider is needed, a significant source of power loss is removed. In fact, the only DC loading on the output capacitor is due to inherent self-leakage of the capacitor and minuscule leakage from the rectifying diode. This results in the photoflash capacitor being able to retain most of its energy when the LT3420 is in shutdown.

Once the target output voltage is reached, the power delivery mode is terminated and the part enters the refresh mode. In refresh mode, the power delivery block is disabled, reducing quiescent current, while the refresh timer is enabled. The refresh timer simply generates a user programmable delay, after which the part reenters the power delivery mode. Once in the power delivery mode, the

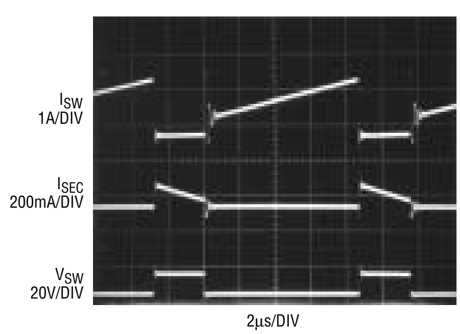


Figure 4a. Switching waveforms with V_{OUT} = 100V, V_{CC}=V_{BAT} = 3.3V

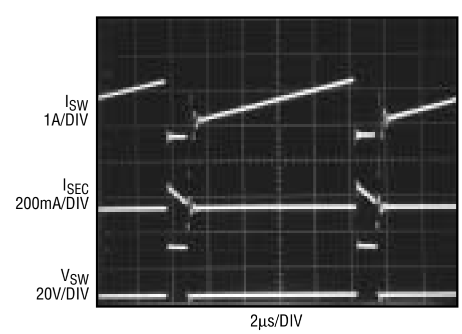


Figure 4b. Switching waveforms with V_{OUT} = 300V, V_{CC}=V_{BAT} = 3.3V

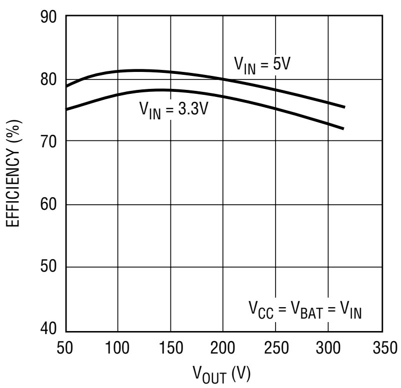


Figure 5. Efficiency for the circuit in Figure 1

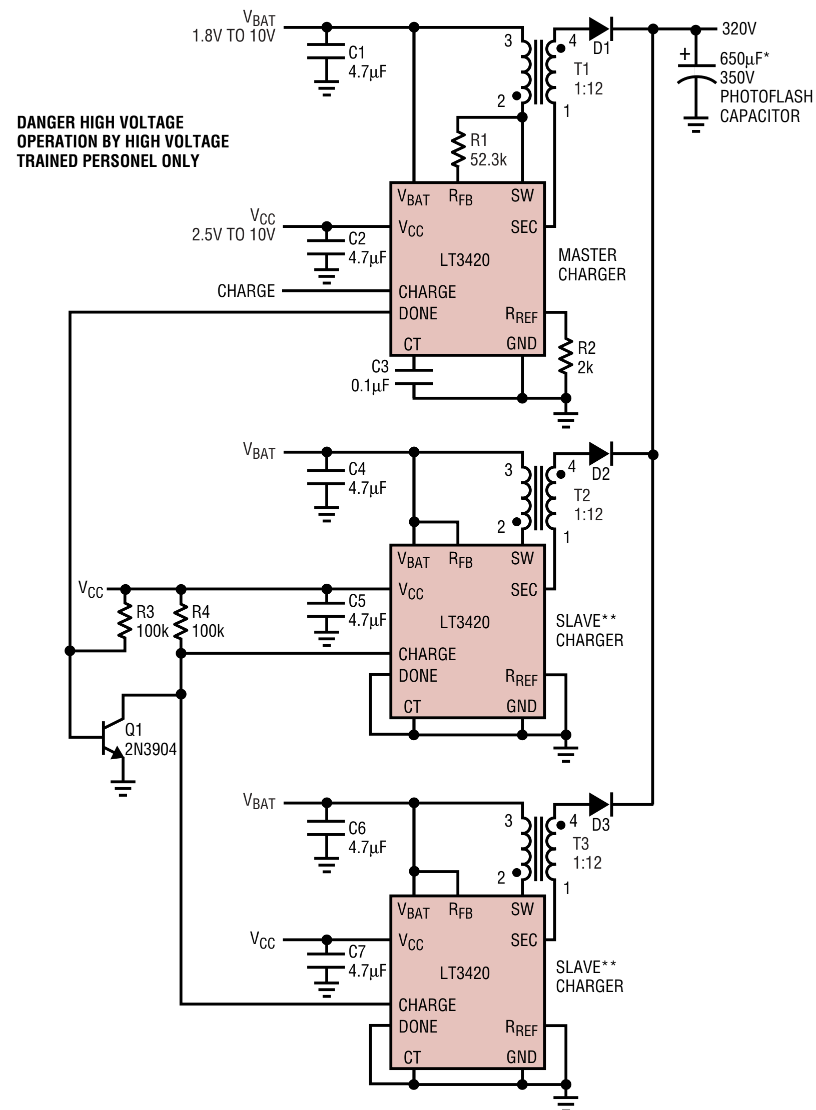
LT3420 will again provide power to the output until the target voltage is reached. Figure 3 is an oscillogram showing both the initial charging of the photoflash capacitor and the subsequent refresh action. The upper waveform is the output voltage. The middle waveform is the voltage on the CT pin. The lower waveform shows the input current. The mode of the part is indicated below the photo.

The user can defeat the refresh timer and force the part into power delivery mode by toggling the CHARGE pin high then low, then high again. The low-to-high transition on the CHARGE pin fires a one-shot that sets the master latch, putting the part in power delivery mode. Bringing CHARGE low puts the part in shutdown. The refresh timer can be programmed to wait indefinitely by simply grounding the CT pin. In this configuration, the LT3420 will only reenter the power delivery mode by toggling the CHARGE pin.

In power delivery mode, the LT3420 operates by adaptively controlling the switch on-time and off-time. The switch on-time is controlled so that the peak primary current is 1.4A (Typical). The switch off-time is controlled so the minimum secondary current is 40mA (Typical). With this type of control scheme, the part always operates in the CCM (Continuous Conduction Mode), resulting in rapid charging of the output capacitor. A side benefit of this scheme is that the part can survive a short circuit on the output indefinitely. Figure 4a and 4b show the relevant currents during the power delivery mode when V_{OUT} is 100V and 300V respectively. Notice how the on-time and off-time are automatically adjusted to keep the peak current in the primary and secondary of the transformer constant as V_{OUT} increases.

Measuring Efficiency

Measuring the efficiency of a circuit designed to charge large capacitive loads is a difficult issue, particularly with photoflash capacitors. The ideal way to measure the efficiency of a capacitor charging circuit would be to find the energy delivered to the output capacitor ($0.5 \bullet C \bullet V^2$) and divide it by the total input energy. This method does not work well here because photoflash capacitors are far from ideal. Among other things, they have relatively high leakage currents, large amounts of dielectric absorption, and significant voltage coefficients. A much more accurate, and easier, method is to measure the efficiency as a function of the output



C1, C2, C4, C5, C6, C7: 4.7µF, X5R or X7R, 10V
T1-T3: PULSE PA0367 FLYBACK TRANSFORMER (619) 674-8100
D1-D3: GENERAL SEMICONDUCTOR GSD2004S SOT-23 DUAL DIODE. DIODES CONNECTED IN SERIES (516) 847-3000
Q1: 2N3904 OR EQUIVALENT
* CAN CHARGE ANY SIZE PHOTOFLASH CAPACITOR
** USE AS MANY SLAVE CHARGERS AS NEEDED.

Figure 6. This professional grade charger uses multiple circuits in parallel to quickly charge large photoflash capacitors.

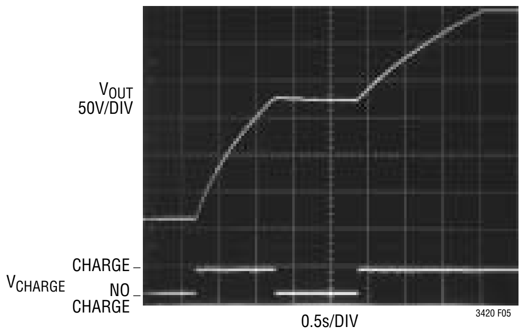


Figure 7. Halting the charge cycle at any time

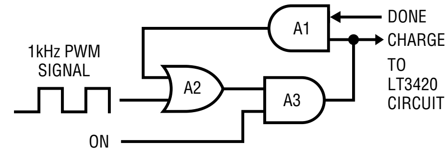


Figure 8. Simple logic for adjustable input current

voltage. In place of the photoflash capacitor, use a smaller, high quality capacitor, reducing errors associated with the non-ideal photoflash capacitor. Using an adjustable load, the output voltage can be set anywhere between ground and the maximum output voltage. The efficiency is measured as the output power (V_{OUT} • I_{OUT}) divided by the input power (V_{IN} • I_{IN}). Figure 5 shows the efficiency for the circuit in Figure 1, which was measured using this method. This method also provides a good means to compare various charging circuits since it removes the variability of the photoflash capacitor from the measurement. The total efficiency of the circuit, charging an ideal capacitor, would be the time average of the given efficiency curve, over time as V_{OUT} changes.

Standard Transformers

Linear Technology Corporation has worked with several transformer manufacturers (including TDK, Pulse and Sumida) to provide transformer designs optimized for the LT3420 that are suitable for most applications. Please consult with the transformer manufacturer for detailed information. If you wish to design your own transformer, the LT3420 data sheet contains a section on relevant issues.

Professional Photoflash Charger

Figure 6 shows a professional grade charger designed to charge large (>500µF) photoflash capacitors quickly and efficiently. Here, multiple LT3420 circuits can be used in parallel. The upper most circuit in the figure is the master charger. It operates as if it were the only charger in the circuit. The DONE signal from this charger is inverted by Q1 and drives the CHARGE pin of all the other slave chargers. Notice that grounding the RREF and CT pins disables the control circuitry of the Slave chargers. The charging time for a given capacitor is inversely proportional to the number of chargers used. Three chargers in parallel takes a third of the charging time as a single charger applied to the same photoflash capacitor. This circuit can charge a 650µF capacitor from 50V to 320V in 3.5s from a 5V input.

Interfacing to a Microcontroller

The LT3420 can be easily interfaced to a microcontroller. The CHARGE and DONE pins are the control and mode indicator pins, respectively, for the part. By utilizing these pins, the LT3420 can be selectively disabled and enabled at any time. The microcontroller can have full control of the LT3420. Figure 7 shows the LT3420 circuit being selectively disabled when the CHARGE pin is driven low midway through the charge cycle. This might be necessary during a sensitive operation in a digital camera. Once the CHARGE pin is returned to the high state, the charging continues from where it left off.

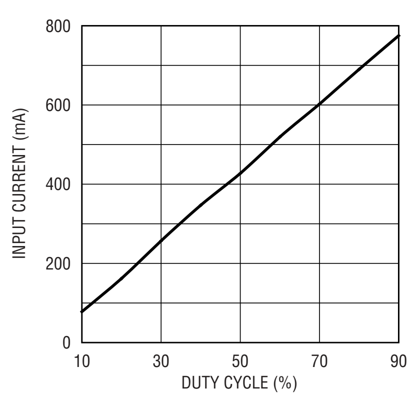


Figure 9. Input current as duty cycle is varied

Adjustable Input Current

With many types of modern batteries, the maximum allowable current that can be drawn from the battery is limited. This is generally accomplished by active circuitry or a polyfuse. Different parts of a digital camera may require high currents during certain phases of operation and very little at other times. A photoflash charging circuit should be able to adapt to these varying currents by drawing more current when the rest of the camera is drawing less, and vice-versa. This helps to reduce the charge time of the photoflash capacitor, while avoiding the risk of drawing too much current from the battery. The input current to the LT3420 circuit can be adjusted by driving the CHARGE pin with a PWM (Pulse Width Modulation) signal. The microprocessor can adjust the duty cycle of the PWM signal to achieve the desired level of input current. Many schemes exist to achieve this function. Once the target output voltage is reached, the PWM signal should be halted to avoid overcharging the photoflash capacitor, since the signal at the CHARGE pin overrides the refresh timer.

A simple method to achieve adjustable input current is shown in Figure 8. The PWM signal has a frequency of 1kHz. When ON is logic high, the circuit is enabled and the CHARGE pin is driven by the PWM signal. When

continued on page 11

Small 1.25A Step-Down Regulator Switches at 4MHz for Space-Sensitive Applications

by Damon Lee

Introduction

Cell phones, pagers, PDAs and other portable devices are shrinking, and as they shrink, the demand for smaller components grows. The ubiquitous switching regulator, which solves the problem of creating a constant voltage from inconstant batteries, is not exempt from the demand to become smaller. One way to shrink regulator circuitry is to increase the switching frequency of the regulator, allowing the use of smaller and less costly capacitors and inductors to complete the circuit. Another way is to shrink the switcher itself by putting the switcher and MOSFETs in a small monolithic package. The LTC3411 DC/DC converter does both.

The LTC3411 is a 10-lead MSOP, synchronous, step-down, current mode, DC/DC converter, intended for medium power applications. It operates within a 2.5V to 5.5V input voltage range and switches at up to 4MHz, making it possible to use tiny capacitors and inductors that are under 2mm in height. By using the LTC3411 in a small MS10 package, a complete DC/DC converter can consume less than 0.3 square inches of board real estate, as shown in Figure 1.

The output of the LTC3411 is adjustable from 0.8V to 5V. For

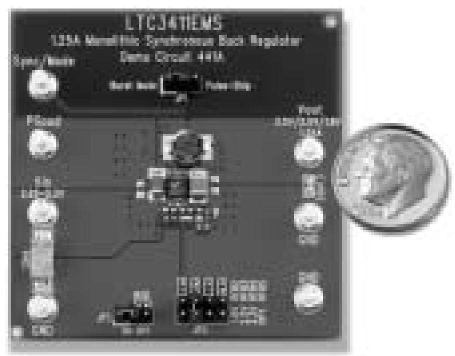


Figure 1. A complete DC/DC converter can take less than 0.3in^2 of board space.

battery-powered applications that have input voltages above *and* below the output, the LTC3411 can also be used in a single inductor, positive Buck-Boost converter configuration. A built-in 0.11Ω switch allows up to 1.25A of output current at high efficiency. OPTI-LOOP® compensation allows the transient response to be optimized over a wide range of loads and output capacitors.

Efficiency takes on grave importance in battery-powered applications, and the LTC3411 keeps efficiency high. Automatic, power saving Burst Mode® operation reduces gate charge losses at low load currents. With no load, the converter draws only 62µA, and in shutdown, it draws less than 1µA, making it ideal for low current applications.

The LTC3411 uses a current mode, constant frequency architecture that benefits noise sensitive applications. Burst Mode operation is an efficient solution for low load current applications, but sometimes noise suppression takes on more importance than efficiency, especially in telecommunication devices. To reduce noise problems, the LTC3411 provides a pulse-skipping mode and a forced-continuous mode. These modes decrease the ripple noise and improve noise filterability. Although not as efficient as Burst Mode operation at low load currents, pulse-skipping mode and forced continuous mode can still provide high efficiency for moderate loads (see Figure 3). In dropout, the internal P-channel MOSFET switch is turned on continuously, thereby maximizing the usable battery life.

A High Efficiency 2.5V Step-Down DC/DC Converter with all Ceramic Capacitors

The low cost and low ESR of ceramic capacitors make them a very attrac-

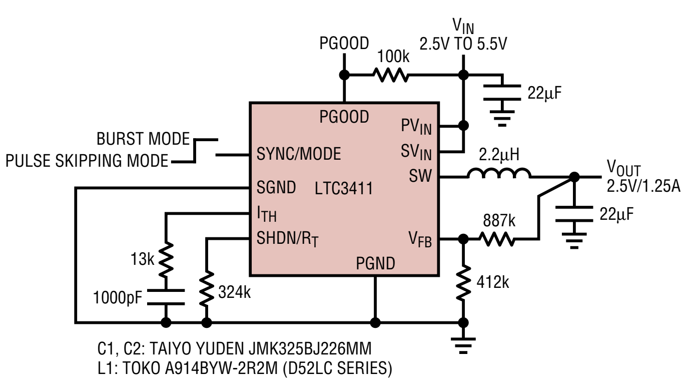


Figure 2. Step-down 2.5V/1.25A regulator

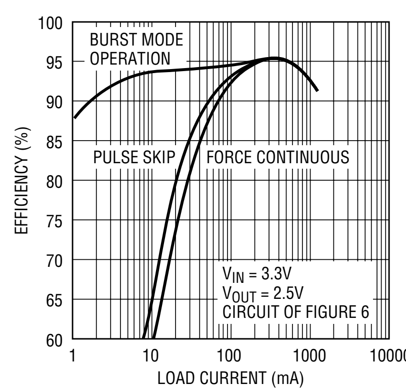


Figure 3. Efficiencies for the circuit shown in Figure 2, under different operating modes

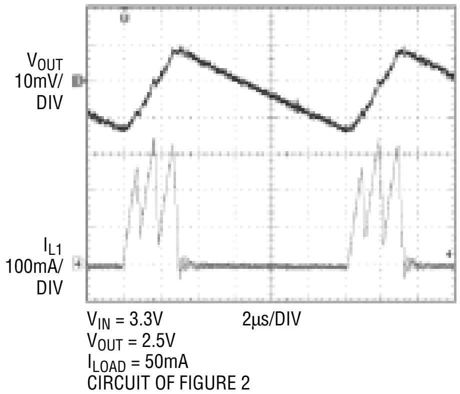


Figure 4. Burst Mode operation

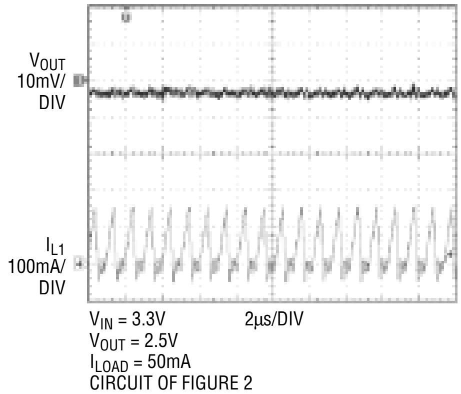


Figure 5. Pulse skipping mode

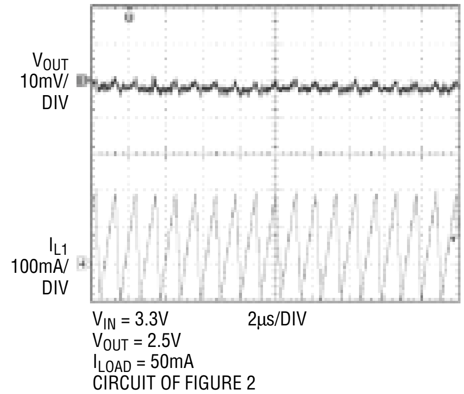


Figure 6. Forced continuous mode

tive choice for use in switching regulators. Unfortunately, the ESR is so low that it can cause loop stability problems. Solid tantalum capacitor ESR generates a loop zero at 5KHz to 50KHz that is instrumental in giving acceptable loop phase margin. Ceramic capacitors remain capacitive to beyond 300KHz and usually resonate with their ESL before ESR becomes effective. Also, ceramic caps are prone to temperature effects, requiring the designer to check loop stability over the operating temperature range. For these reasons, great care must be taken when using only ceramic input and output capacitors. The LTC3411 helps solve loop stability problems with its OPTI-LOOP phase compensation adjustment, allowing the use of ceramic capacitors. For details, and a process for optimizing compensation components, see Linear Technology Application Note 76.

A typical application for the LTC3411 is a 2.5V step-down converter using only ceramic capacitors, as shown in Figure 2. This circuit provides a regulated 2.5V output, at up to 1.25A, from a 2.5V to 5.5V input. Efficiency for the circuit is as high as 95% for a 3.3V input as shown in Figure 3.

Although the LTC3411 is capable of operating at 4MHz, the frequency in this application is set for 1MHz by R4 to improve the efficiency. Also, the availability of capacitors and inductors capable of 4MHz operation is limited.

Figures 3 through 6 show the tradeoff between noise and efficiency for the different modes for the circuit. Figure 3 shows the efficiencies, while Figures 4, 5 and 6 show the output voltage and inductor current for different operating modes.

Burst Mode operation is the most efficient for low current loads, but it is also generates the most complicated noise patterns. Figure 4 shows how Burst Mode operation produces a single pulse or a group of pulses that are repeated periodically. By running cycles in periodic bursts, the switching losses—dominated by the gate charge losses of the power MOSFET—are minimized. Figure 5 shows how in pulse skipping mode, the LTC3411 continues to switch at a constant frequency down to very low currents, minimizing the ripple voltage and ripple current. Finally, Figure 6 shows how in forced continuous mode, the inductor current is continuously cycled, creating a constant ripple at all output currents. Forced continuous mode is particularly useful in noise-sensitive telecom applications since the constant frequency noise is easy to filter. Another advantage of this mode is that the regulator is capable of both sourcing and sinking current into a load. This mode is enabled by forcing the mode pin to half of V_{IN}.

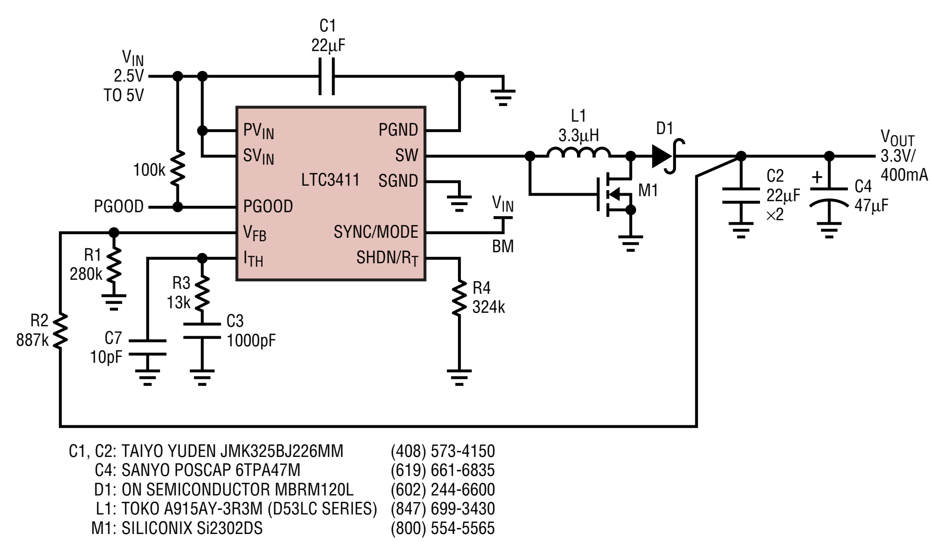


Figure 7a. Single inductor, positive, buck-boost converter

Single Cell Li-Ion to 3.3V DC/DC Converter

Lithium-Ion batteries are popular in many portable applications because of their light weight and high energy density, but the battery voltage ranges from a fully charged 4.2V down to a drained 2.5V. When a device requires a voltage output that falls somewhere

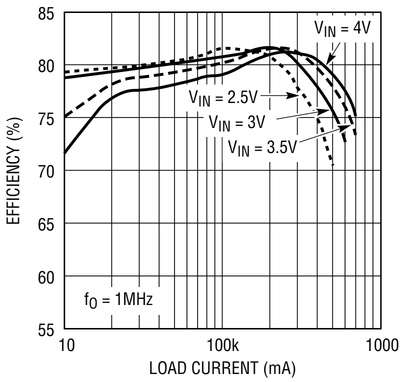


Figure 7b. Efficiency for the circuit in Figure 7a

the middle of the Li-Ion operating range, say 3.3V, a simple buck or boost converter does not work. One solution is a single inductor, positive buck-boost converter, which allows the input voltage to vary above and below the output voltage.

In Figure 7, the LTC3411 is used in a single Inductor, positive buck-boost configuration to supply a constant 3.3V with 400–600mA of load current, depending on the battery voltage. This circuit is well suited to portable applications because none of the components exceed 3mm in height.

The efficiency varies with the input supply, due to resistive losses at high currents and to switching losses at low currents. The typical efficiency across both battery voltage and load current is about 78%.

It's Only 2mm High: 2MHz, Li-Ion to 1.8V Converter

In some applications, minimizing the height of the circuit takes prime importance. One method of lowering the DC/DC converter height is to run the LTC3411 at the 2MHz switching frequency, which allows one to use low-profile inductors and capacitors. Figure 8 shows a circuit built with low profile components to produce a 2mm tall (nominal), 1.8V step-down converter that occupies less than 0.3 square inches. In the spirit of keeping things as small as possible, this circuit uses tantalum capacitors for their relatively small size when compared to equivalent ceramic capacitors.

The downside to running at a higher frequency is that efficiency suffers a little due to higher switching losses. The efficiency for this particular circuit peaks at 93% with V_{IN} = 2.5V.

Conclusion

The LTC3411 is a monolithic, step-down regulator that switches at high frequencies, lowering component costs and board real estate requirements of DC/DC converters. Although the LTC3411 is designed for basic buck applications, its architecture is versatile enough to produce an efficient single inductor, positive buck-boost converter, due in part to its power saving Burst Mode operation and the OPTI-LOOP compensation feature.

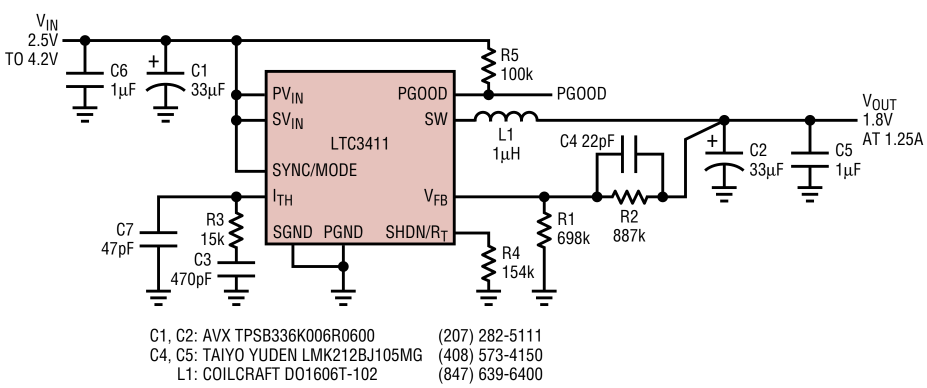


Figure 8a. Tiny 1.8V/1.25A step-down converter uses low profile components

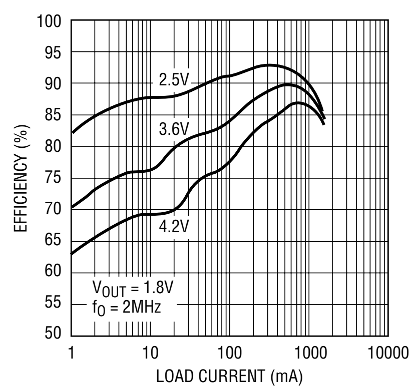


Figure 8b. Efficiency for the circuit in Figure 8a

LT3420, continued from page 8

the target output voltage is reached, DONE goes high while CHARGE is also high. The output of A1 goes high, which forces CHARGE high regardless of the PWM signal. The part is now in the Refresh mode. Once the refresh period is over, the DONE pin goes low, allowing the PWM signal to drive the CHARGE pin once again. This function can be easily implemented in a microcontroller. Figure 9 shows the input current for the circuit of Figure 1 as the duty cycle of the PWM signal is varied.

Conclusion

The LT3420 provides a highly efficient and integrated solution for charging photoflash capacitors. Many important features are incorporated into the device, including automatic refresh, tightly controlled currents and an integrated power switch, thus reducing external parts count. The LT3420 comes in a small, low profile, MSOP-10 package, making for a complete solution that takes significantly less PC board space than more traditional methods. Perhaps most importantly, the LT3420 provides a simple solution to a complicated high voltage problem, freeing camera designers to spend time on other important matters, like increasing the pixel count or adding new camera features.

Dual DC/DC Controller Brings 2-Phase Benefits to Low Input Voltage Applications

by Jason Leonard

Introduction

The LTC3701 is an efficient, low input voltage, dual DC/DC controller that fits into the tight spaces required by the latest portable electronics. It uses 2-phase switching techniques to reduce required input capacitance (saving space and cost) and increase efficiency. The versatile LTC3701 accepts a wide range of input voltages, from 2.5V to 9.8V, making it useful for single lithium-ion cell and many multicell systems. It can provide output voltages as low as 0.8V and output currents as high as 5A. The 100% duty cycle allows low dropout for maximum energy extraction from a battery, and the optional Burst Mode operation enhances efficiency at low load currents. It also includes other popular features, such as a Power Good voltage monitor, a phase-locked loop, and an internal soft start. Its small 16-lead narrow SSOP package and relatively high operating frequency (300kHz–750kHz) allow the use of small, surface mount components, making for a compact overall power supply solution.

Operation

Figure 1 shows the LTC3701 used in a step-down converter with an input of from 2.5V to 9.8V and two outputs of 2.5V at 2A and 1.8V at 2A. Figure 2 shows its efficiency versus load current. The LTC3701 uses a constant frequency, current mode architecture with the two controllers operating 180 degrees out of phase.

The LTC3701 offers the benefits of 2-phase operation, which include lower input filtering requirements, reduced electromagnetic interference (EMI) and increased efficiency.

During normal operation, each external P-channel power MOSFET is turned on every cycle when the oscillator for that controller sets a latch and turned off when the current comparator resets the latch. The peak inductor current at which the current comparator resets the latch is controlled by the voltage on the ITH/RUN pin, which is the output of the error amplifier. The VFB pin receives the output voltage feedback signal, which is compared to the internal 0.8V reference by the error amplifier. When the load current increases, it causes a slight decrease in VFB relative to the reference, which, in turn, causes the ITH/RUN voltage to increase until the average inductor current matches the load current.

2-Phase Operation

The LTC3701 offers the benefits of 2-phase operation, which include lower input filtering requirements, reduced electromagnetic interference (EMI) and increased efficiency.

In a single phase dual switching regulator, both top-side P-channel MOSFETS are turned on at the same time, causing current pulses of up to twice the amplitude of those from a single regulator to be drawn from the input capacitor. These large amplitude pulses increase the total RMS current flowing into the input capacitor, requiring the use of more expensive input capacitors, and increasing both EMI and losses in the input capacitor and input power supply.

With 2-phase operation, the two channels of the LTC3701 are operated 180 degrees out of phase. This effectively interleaves the current

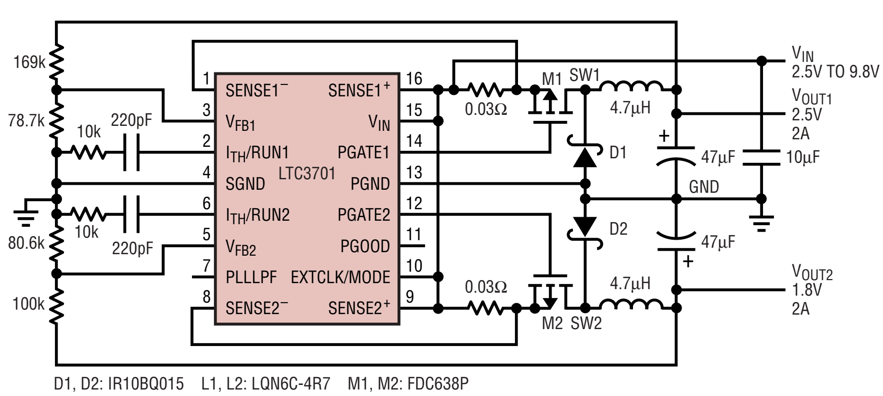


Figure 1. 2-phase step-down converter with an input of 2.5V to 9.8V and two outputs: 2.5V at 2A and 1.8V at 2A

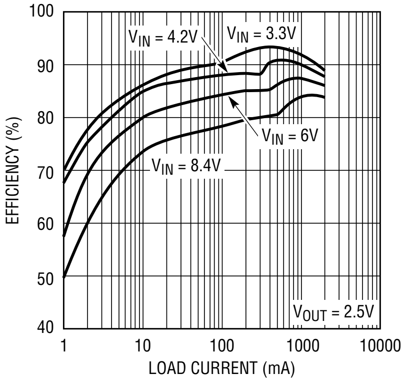


Figure 2. Efficiency vs load current

Figure 3. These input waveforms for the circuit in Figure 1 show how 2-phase operation reduces ripple. Reduced ripple translates directly to less expensive input capacitors, reduced shielding requirements for EMI, and improved efficiency.

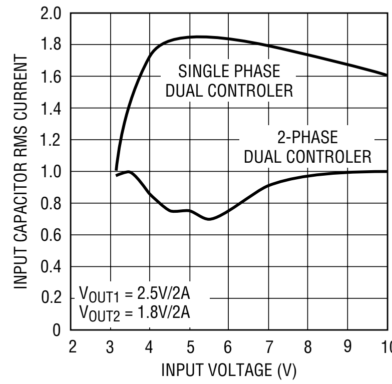


Figure 4. RMS input current comparison

pulses coming from the switches, greatly reducing the amount of time where they overlap and add together. The dead bands in the input current waveform are "filled up," so to speak. The result is a significant reduction in the total RMS input current, which in turn allows for the use of less expensive input capacitors, reduces shielding requirements for EMI, and improves efficiency. Figure 3 shows the input waveforms for the circuit in Figure 1. The RMS input current is significantly reduced by the interleaving current pulses. Of course, the improvement afforded by 2-phase operation is a function of the dual switching regulator's relative duty cycles, which are dependent on the input voltage V_{IN}. Figure 4 shows how the RMS input current varies for single-phase and 2-phase operation for 2.5V and 1.8V regulators over a wide input voltage range.

Burst Mode Operation

The LTC3701 can be enabled to enter Burst Mode operation at low load currents by connecting the EXTCLK/MODE pin to VIN. In this mode, the minimum peak current is set as if VITH/RUN = 1V, even though the voltage at the ITH/RUN pin is at a lower value. If the inductor's average current is greater than the load requirement, the voltage at the ITH/RUN pin will drop as VOUT rises slightly. When the ITH/RUN voltage goes below 0.85V, a sleep signal is generated, turning off the external MOSFET and much of the LTC3701's internal circuitry. The load current is then supported by the output capacitor. When the ITH/RUN voltage goes above 0.925V, the sleep signal goes low and normal operation resumes. For frequency sensitive applications, Burst Mode operation can be inhibited by connecting the EXTCLK/MODE pin to ground. In this case, constant frequency operation is maintained at a lower load current with a lower output voltage ripple. If the load current is low enough, cycle skipping occurs to maintain regulation.

Frequency Selection/Synchronization (Phase-Locked Loop)

The LTC3701 operates at a constant frequency between 300kHz and 750kHz. The frequency can be selected by forcing a voltage at the PLLLPF pin. Grounding the PLLLPF pin selects 300kHz, while tying it to VIN or a voltage greater than 2V selects 750kHz. Floating the PLLLPF pin selects 550kHz operation.

The LTC3701 can also be synchronized to an external clock source (300kHz to 750kHz) using the LTC3701's true phase-locked loop. The clock signal is applied to the EXTCLK/MODE pin and an RC filter is connected between the PLLLPF pin and ground. Burst Mode operation is disabled when synchronized to an external clock.

Run/Soft Start

Either controller can be shutdown by pulling its respective ITH/RUN pin below 0.35V, which turns off most circuits associated with that control-

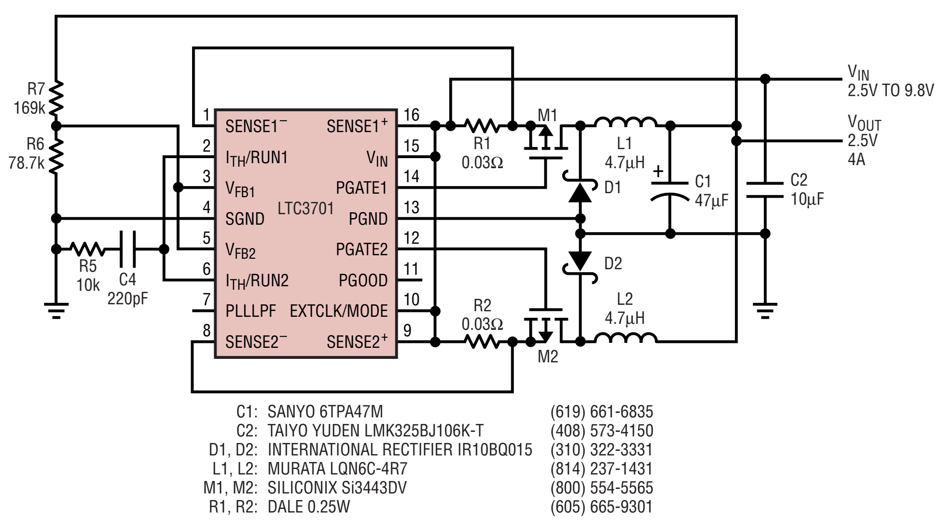
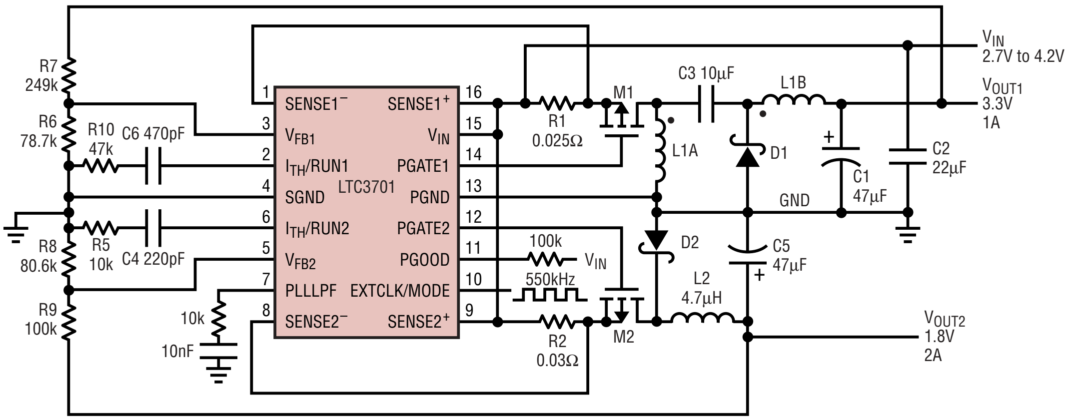


Figure 5. 2.5V–9.8V to 2.5V/4A 2-phase step-down converter operating at 550kHz



C1, C5: SANYO 6TPA47M (619) 661-6835
C2: TAIYO YUDEN JMK325BJ226MM (408) 573-4150
C3: TAIYO YUDEN JMK316BJ106ML
D1, D2: INTERNATIONAL RECTIFIER IR10BQ015 (310) 322-3331
L1A, L1B: COILTRONICS CTX5-2 (561) 752-5000
L2: MURATA LQN6C-4R7 (814) 237-1431
M1, M2: SILICONIX Si3443DV (800) 554-5565
R1, R2: DALE 0.25W (605) 665-9301

Figure 6. Single cell Li-Ion to 3.3V/1A and 1.8V/2A DC/DC converter

ler and holds its external MOSFET off. If both ITH/RUN pins are pulled low, the LTC3701 is shutdown and draws only 9µA.

The LTC3701 has separate internal soft start functions that allow each output to power up gently. The maximum allowed inductor current is stepped up from 0 to 120mV/R_{SENSE} in four equal steps of 30mV/ R_{SENSE}, with each step lasting 512 clock cycles (just under 1ms per step at 550kHz).

Power Good Output Voltage Monitor

A window comparator monitors both output voltages and the open-drain PGOOD output is pulled low when the divided down output voltages are not within ±8% of the reference voltage of 0.8V.

2-Phase 2.5V/2A and 1.8V/2A Step-Down Regulator

Figure 1 shows a typical application of the LTC3701. This circuit supplies a 2A load at 2.5V and a 2A load at 1.8V with an input supply from 2.5V to 9.8V. Due to the reduced input current ripple associated with 2-phase operation, only a single 10µF ceramic input capacitor is required. The 0.03Ω sense resistors ensure that both outputs are capable of supplying 2A with a low input voltage. The circuit operates at the internally set frequency of 550KHz. 4.7µH inductors are chosen so that the inductor currents remain continuous during burst periods at low load current.

2-Phase Single Output 2.5V/4A Step-Down Regulator

In addition to dual output applications, the LTC3701 can also be used in a single output configuration to take advantage of the benefits of 2-phase operation, as shown in Figure 5. This circuit provides a 2.5V output with up to 4A of load current. In this case, 2-phase operation reduces both the input and output current ripple, in turn reducing the required input and output capacitances.

Single Cell Li-Ion to 3.3V/1A (Zeta Converter) and 1.8V/2A

In addition to step-down applications, the LTC3701 can also be used in a zeta converter configuration that will do both step-down and step-up conversions, as shown in Figure 6. This circuit delivers 1A at 3.3V (zeta converter) and 2A at 1.8V (step-down converter) from an input of 2.7V to 4.2V (Li-Ion voltage range). The circuit takes advantage of the LTC3701's true phase-locked loop by synchronizing to an external clock source.

Conclusion

The LTC3701 brings the benefits of 2-phase operation to low-voltage dual power supply systems. It offers flexibility, high efficiency, and many other popular features in a small 16-pin narrow SSOP package.

ThinSOT RF Power Controllers Save Critical Board Space and Power in Portable RF Products

by Ted Henderson and Shuley Nakamura

Introduction

The LTC4400-1 and LTC4401-1 provide RF power controller solutions for the latest cellular telephones. They feature very small footprints, low power consumption and wide frequency ranges while minimizing adjacent channel interference by carefully controlling RF power profiles. The LTC4400-1 and LTC4401-1 are both available in a low profile 6-pin ThinSOT package, and require few external parts. For example, when used with a directional coupler, only two resistors and two capacitors are required (Figure 1a, Figure 1b). Both devices require minimal power to operate, typically 1mA when enabled and 10µA when in shutdown.

The LTC4400-1's 450kHz loop bandwidth is optimized for applications involving fast turn-on (<2µs) and medium gain (200-300dB/V) RF power amplifiers. The LTC4401-1's 250kHz loop bandwidth is optimized for slow turn-on (>2µs) and/or high gain (300-400dB/V) RF power amplifiers. The RF frequency range for both parts is 800MHz to 2.7GHz and the supply voltage range is 2.7V to 6.0V. This wide frequency and voltage range allow these products to be used in a variety of RF power control applications including GSM/GPRS, PCS and TDMA. The LTC4400-1 and LTC4401-1 include an auto zero system that requires periodic updates between single or multiple consecutive bursts. Therefore these power controllers are *not* suitable for continuous time applications.

Figure 2 shows the block diagram of the LTC4400/4401. When the part is in shutdown all circuitry except the reference is turned off and VPCA is held at ground. When the part is enabled, the auto zero system samples both internal and external offsets.

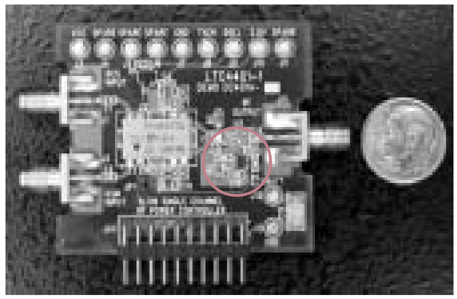


Figure 1a. The DC401A RF demo board. The circled area encloses the LTC4401 power controller (U4) and its required external components (C11, C12, C13 and R2).

After 10µs the auto zero system is disabled; the sampled offset voltage correction factor is held on two internal capacitors. A differential hold scheme is used to convert hold capacitor voltage droop (due to leakage currents) to a common mode voltage droop. This common mode voltage droop is rejected by the auto zero amplifier, resulting in greatly increased auto zero hold time. The auto zero system improves temperature dependent characteristics by removing temperature offset voltage drifts from internal and external sources.

The external power control ramp is applied 12µs after SHDNB is asserted high by the baseband microprocessor. When the ramp is applied, the VPCA voltage begins to rise. The RF power amplifier turns on when VPCA reaches the RF power amplifier's threshold voltage. VPCA actually starts from 450mV. This start voltage reduces the time required to turn on the RF power amplifier and is lower than power amplifier threshold voltages used in mobile radio applications. The power control loop is open until the RF power amplifier turns on and starts supplying an RF output signal. While the loop is open, the VPCA rise time is limited by the LTC4400/4401 bandwidth and the magnitude of the PCTL signal. A portion of the RF output voltage is fed back to the LTC4400/4401 RF pin. This signal is then peak detected by an internal Schottky diode and capacitor. The detected voltage is applied to the negative input of the loop amplifier thereby closing the power control loop. Once the loop has closed, the RF output signal follows the power ramp signal at PCTL.

RF Detector Performance

The LTC4400 and LTC4401 incorporate two features to improve detector dynamic range. An auto zero system eliminates both internal offsets and

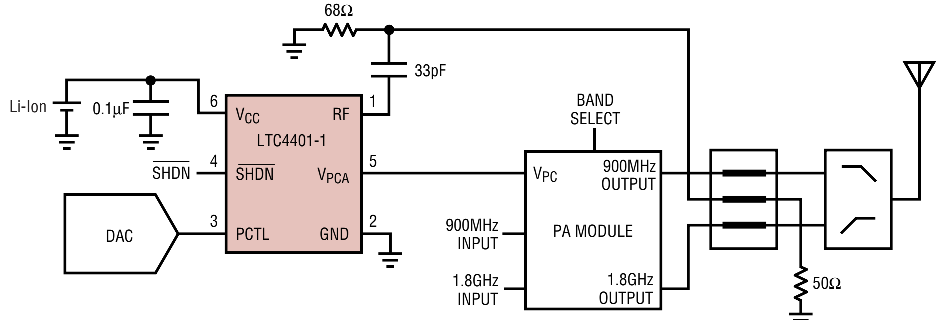


Figure 1b. Typical power control block diagram

external power control DAC offsets. Secondly, a compression circuit allows for higher feedback signals at lower RF power levels to extend the power detector range. The fully integrated detector has a small temperature coefficient as shown in Figure 3.

Measuring RF Power Amplifier Rise Times

To determine which LTC RF power controller fits a particular application, the designer must first understand the RF power amplifier turn-on characteristics. Figure 4 shows a recommended test setup.

A pulse generator is used to drive the RF amplifier power control pin, with its duty cycle set to minimize power dissipation (i.e. 1/8 duty cycle). Terminate the RF power control pin with a 50Ω resistor to match the pulse generator and avoid ringing. With a square wave pulse at various amplitudes, determine the RF output power response. Measure at several output power levels since the rise time may be power level dependent.

Use a high frequency digital scope to measure the RF output voltage shape. Figure 5 shows a typical RF output voltage response. This waveform consists of two regions, delay and ramp. The ramp time is measured from the start of the RF output to 90% of the final amplitude. Generally the LTC4400 is used for amplifiers with total delay and ramp times <2µs; the LTC4401 is used for amplifiers with total times >2µs. Other factors such as power amplifier gains, coupler and antenna switch losses, may also impact this selection. Very high gain power amplifiers may require the LTC4401 independent of the response times.

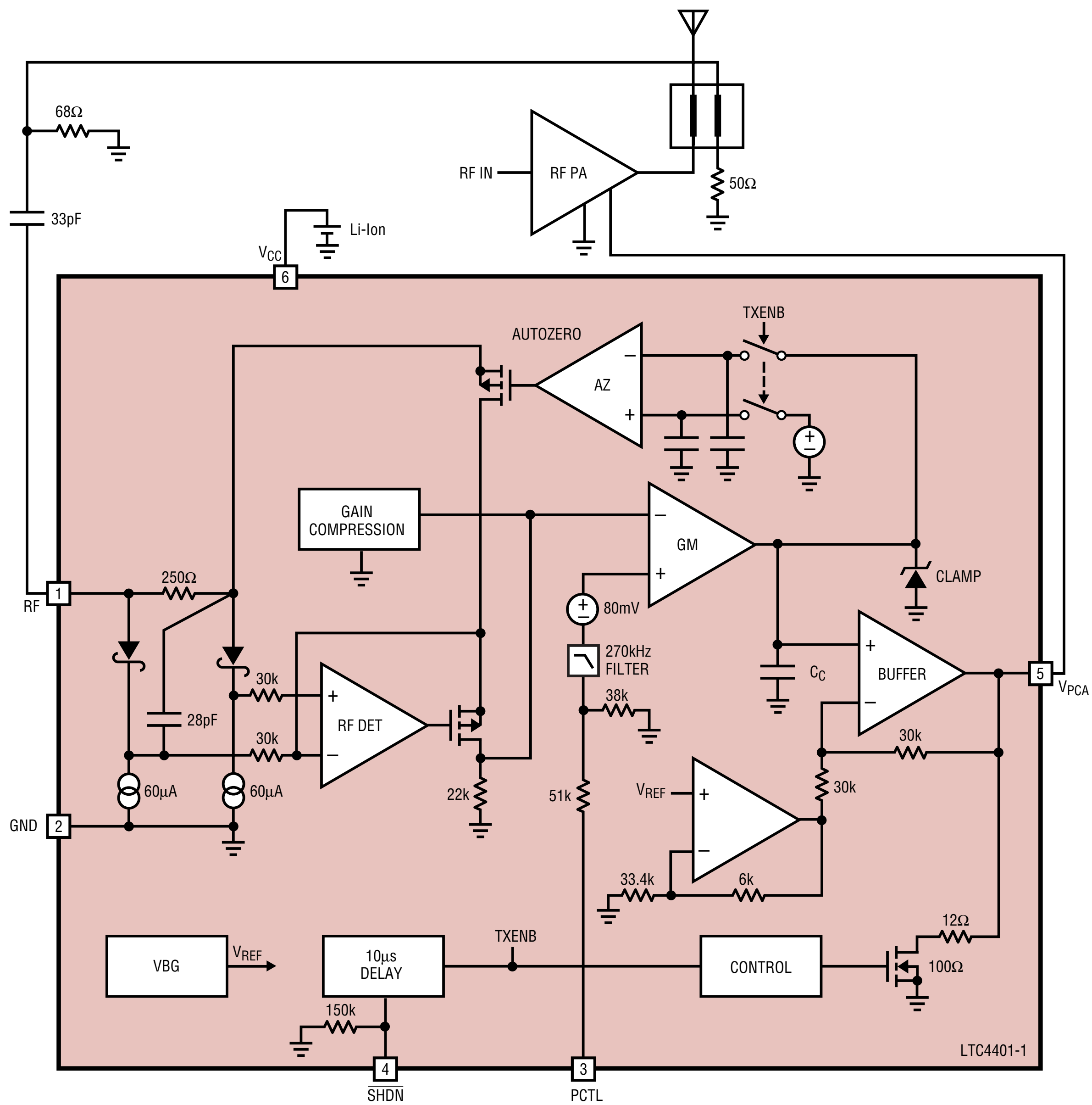


Figure 2. LTC4400/4401 block diagram

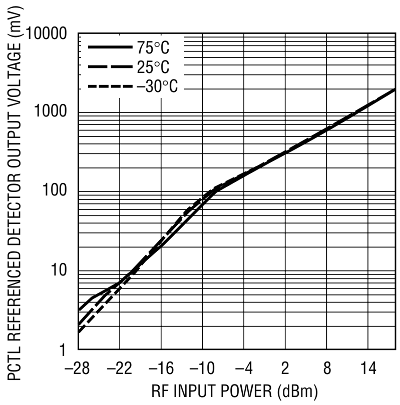


Figure 3a. Detector characteristics at 900MHz

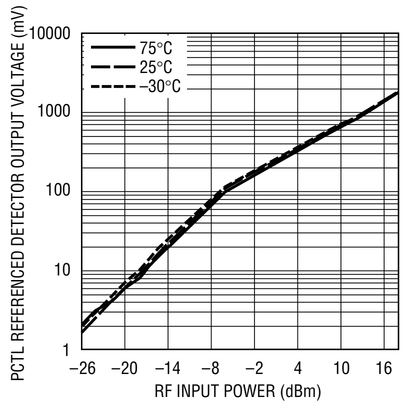


Figure 3b. Detector characteristics at 1800MHz

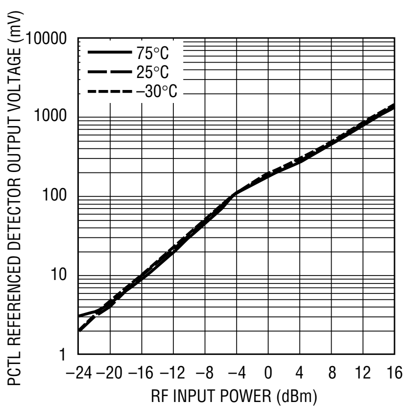


Figure 3c. Detector characteristics at 2400MHz

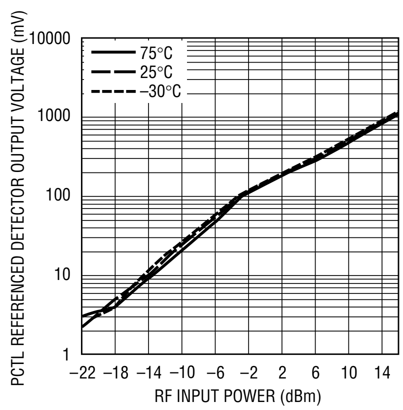


Figure 3d. Detector characteristics at 2700MHz

Figure 3. Typical detector characteristics

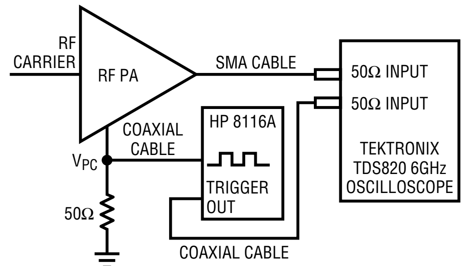


Figure 4. RF power amplifier rise time test setup

Powerful and Easy to Use Development Tools Optimize PA Control

Version 2 of the LT ramp-shaping program (LTRSv2.VXE) is available from Linear Technology. Figure 6 shows the program window of LTRSv2.VXE in ramp-shaping mode. This program lets users generate, reshape, and load ramp profile waveforms onto the DC314A digital demo board. The DC314A digital demo board provides regulated power supplies, control logic and a 10-bit DAC to generate the SHDNB signal and the power control PCTL signal. Flash memory and a serial port interface are also included for updating DAC profiles stored on the DC314A. Eight power control profiles can be stored in the flash memory. A rotary switch (SW1) can be used to select the desired power profile. The DC314A provides signals to the DC401A RF demo board, which contains a GSM/DCS RF channel, LTC4401-1 ThinSOT power controller, and Hitachi PF08107B power amplifier (Figure 1B). The RF test measurement setup is shown in Figure 7.

LTRSv2.VXE creates smooth ramp waveforms based on user inputs. The user controls all aspects of the ramp parameters such as initial DAC offset, step voltage and time, rise and fall times, and maximum voltage amplitude and time. LTRSv2.VXE uses a raised-cosine function to create smooth transitions between areas of varying amplitudes, such as between the step and the maximum amplitude (Figure 8).

LTRSv2.VXE ramp profile parameters are saved in tables as text files. Linear Technology distributes LTRSv2.VXE with HP VEE Runtime,

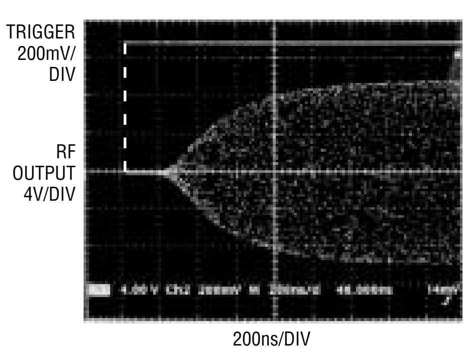


Figure 5. RF output voltage

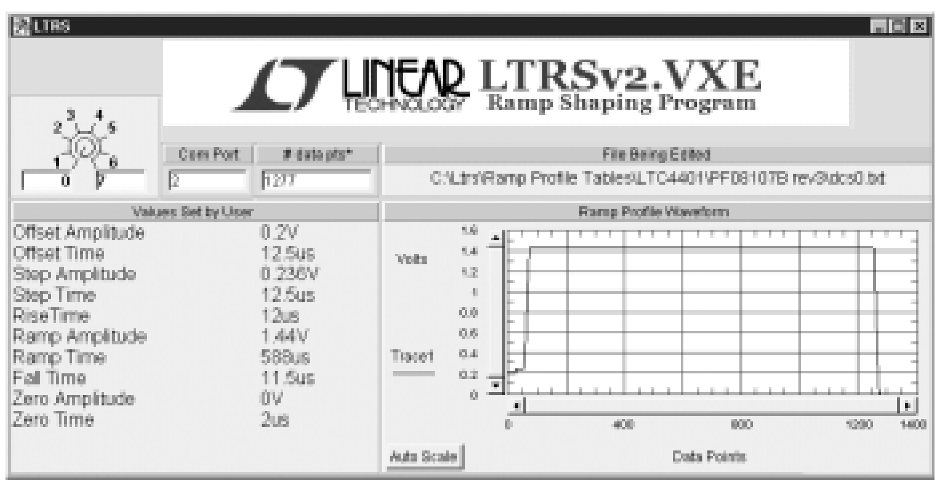


Figure 6. LTRSv2.vxe program window

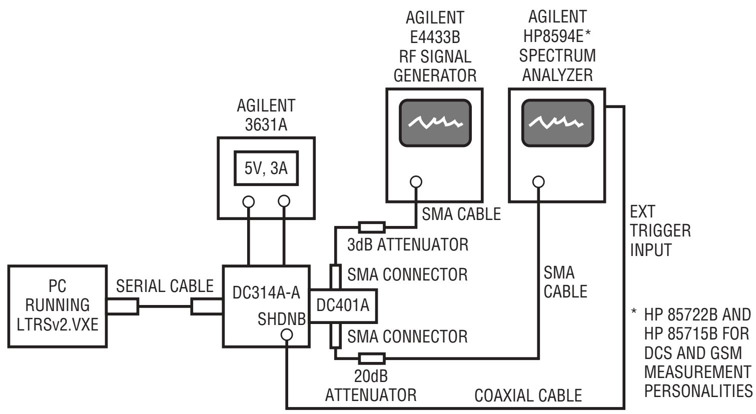


Figure 7. Demo board evaluation setup for GSM/DCS measurements

HP I/O Libraries, and ramp profile table templates for various power amplifiers. Each ramp profile waveform table can be edited or overwritten using LTRSv2.VXE. These ramp profile table templates serve as an excellent starting place for ramp-shaping. Figure 9 is an illustration of a typical ramp profile waveform with the ramp parameters labeled. Figure 10 shows the program window where the user changes ramp profile waveform parameters.

Ramp shapes vary depending on which power controller and power amplifier are being used. For example, power amplifiers that exhibit “slow” turn-on/off times (2µs and greater) require larger step amplitude and time values and a higher DAC offset voltage. Similarly, rise and fall times for slow power amplifiers are longer.

Ramp-shaping is an iterative process. Changes should be made one parameter at a time since each affects different aspects of the output. Placing oscilloscope probes on PCTL and VPCA greatly facilitates the ramp-shaping process.

The ramp waveform begins with the DAC offset voltage. The offset improves ramp down characteristics of the power amplifier. A 100mV offset voltage is sufficient for the LTC4400-1 and fast power amplifiers, while a 200mV offset voltage is sufficient for the LTC4401-1 and slow power amplifiers. The offset time for the ramp is typically 12µs during which auto zeroing occurs. Figure 11 shows the timing relationship between SHDNB, VPCA and PCTL.

The first step in ramp-shaping is determining the correct output power. Increase or decrease the maximum amplitude to effect a corresponding change in the RF output power. Once the output power is set, adjust the initial step amplitude and time.

The initial step values are responsible for closing the voltage loop. VPCA must quickly rise to the RF power amplifier threshold voltage in order to meet power versus time specifications. If the initial step time or amplitude values are too low, the control voltage waveform will resemble VPCA in Figure 12. The resolution of the DAC allows for amplitude changes as small as 2mV. The step voltage can be changed in 1/2 microsecond multiples. There are some tradeoffs to take into consideration when choos-

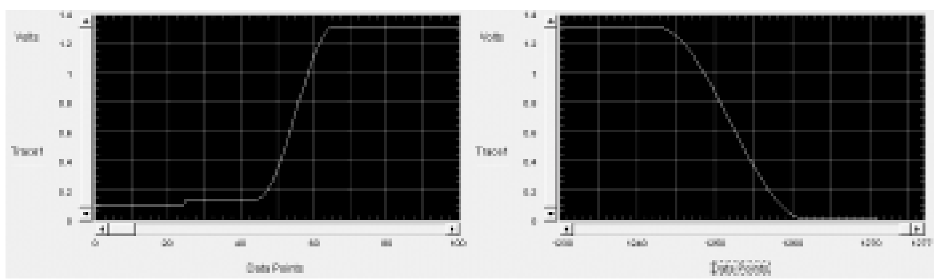


Figure 8. Rise and fall ramp shapes with raised-cosine function

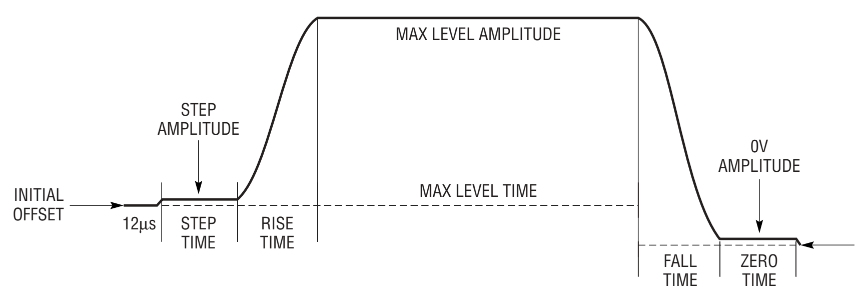


Figure 9. Typical ramp profile waveform

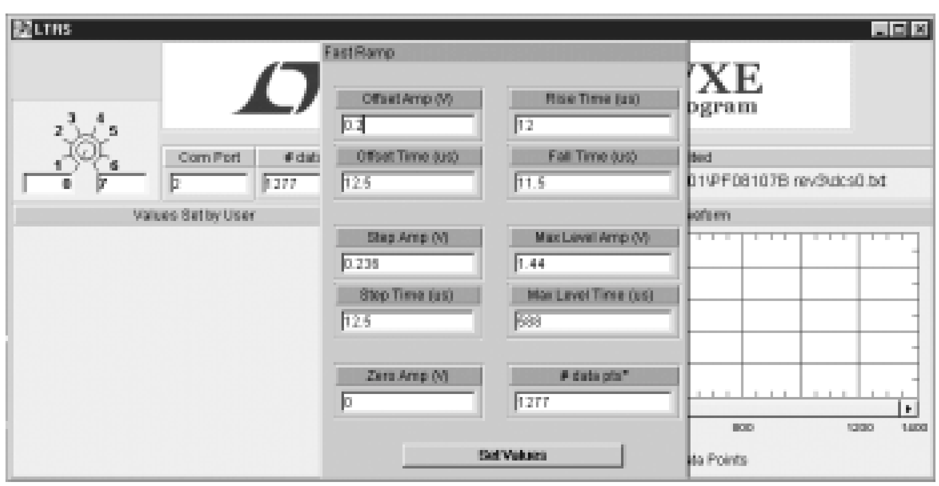
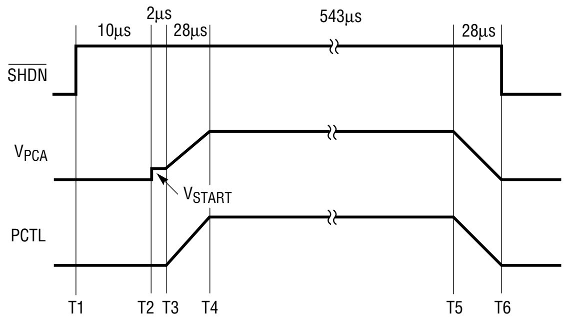


Figure 10. Ramp profile waveform parameter editing window



T1: PART COMES OUT OF SHUTDOWN 12µs PRIOR TO BURST
T2: INTERNAL TIMER COMPLETES AUTOZERO CORRECTION, <10µs
T3: BASEBAND CONTROLLER STARTS RF POWER RAMP UP AT 12µs AFTER SHDN IS ASSERTED HIGH
T4: BASEBAND CONTROLLER COMPLETES RAMP UP
T5: BASEBAND CONTROLLER STARTS RF POWER RAMP DOWN AT END OF BURST
T6: PART RETURNS TO SHUTDOWN MODE BETWEEN BURSTS

Figure 11. LTC4400/4401 timing diagram

ing which parameters to change. For instance, if the step amplitude is too high, the RF output spectrum may exhibit spurs. However, if the step time is too long, as shown in Figure 13, meeting required power versus time is compromised because the time allotted for the burst portion is insufficient. Figure 14 shows the ideal shape for the control voltage. The rise portion of VPCA is smooth and has a constant slope until the maximum amplitude is reached.

Once the step values are set, the rise and fall times should be adjusted. If the rise time is too short for slow amplifiers, an overshoot will occur and will be visible in the power versus time measurement. Lengthening the fall time generally lowers the spurs ±400 kHz from the center frequency. Rise and fall times vary from 8µs–14µs.

The last step is adjusting the width of the maximum amplitude. This is necessary to meet power versus time specifications. Typically, the burst portion of the width is 588µs. The total time of the maximum ramp amplitude must be enough to pass the power versus time measurement and leave suitable time at 0 volts to turn the power amplifier off. Usually, 1µs is required to turn off a power amplifier.

After each parameter is changed, a graph of the waveform created appears in the program window along with the option to load the ramp onto the DC314A demo board.

Ramp-shaping is more challenging with slower power amplifiers because more time is required on the step, rise and fall. If there is not enough time to meet the power versus time mask and turn off the PA, then it is necessary to change the step amplitude and time. A change of 4mV to 6mV accounts for 1µs. Be careful to not let the step amplitude become too high to avoid spurs in the RF output.

Figure 15 shows the control voltage waveform for maximum output power at 1800MHz (DCS0). The waveform has an initial start voltage of 450mV. By starting the output control voltage at 450mV, the time required to reach the power amplifier threshold voltage is reduced. The start voltage is generated by the LTC4400/4401 and not by the program.

Figure 16 is the corresponding output RF spectrum for the control voltage shown in Figure 15. The center frequency is 1710.2MHz and the input power to the power amplifier is 0dBm. Figure 17 shows the power versus time measurement. The on-screen table, shown in Figure 6, represents the values entered to create the ramp waveform. The input step and ramp amplitudes include a 200mV offset amplitude. Therefore, the actual step voltage is 36mV and the ramp amplitude voltage is 1.24V.

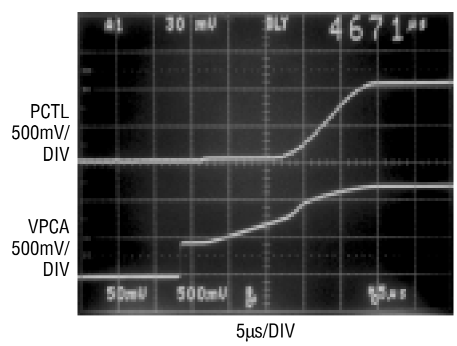


Figure 12. PCTL and VPCA waveforms with low ramp step amplitude and step time

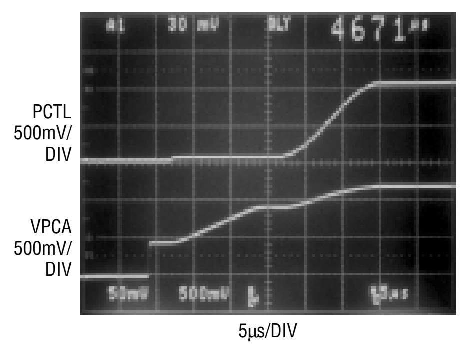


Figure 13. PCTL and VPCA waveforms with high ramp step amplitude and step time

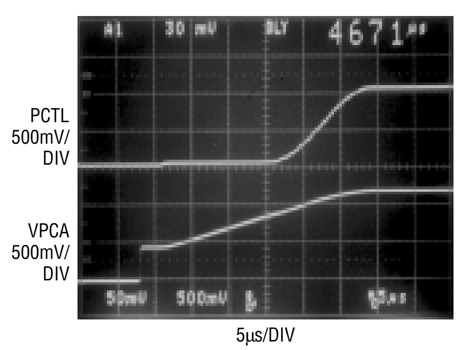


Figure 14. Optimized PCTL input and VPCA output

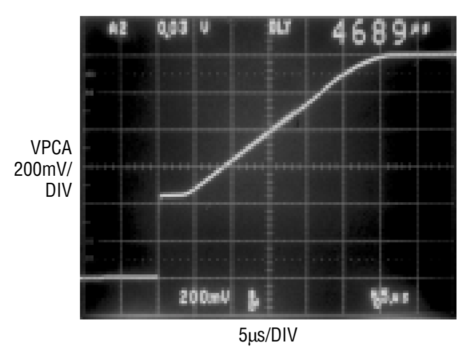
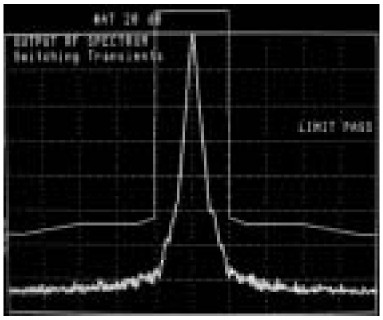
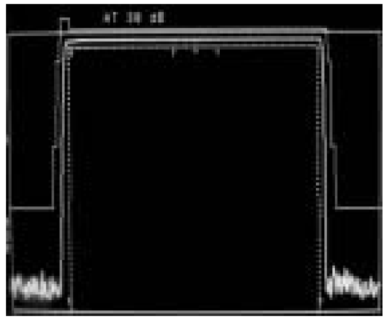


Figure 15. Correct VPCA response and 450mV start voltage



REFERENCE = 30.0dBm
CENTER = 1.710200 GHz SPAN = 4.00MHz
RESOLUTION BANDWIDTH = 30kHz
VIDEO BANDWIDTH = 100kHz
SWEEP = 2.00s

Figure 16. Output RF spectrum switching transients for DCS0



REFERENCE = 32.0dBm
CENTER = 1.710200 GHz SPAN = 0Hz
RESOLUTION BANDWIDTH = 300kHz
VIDEO BANDWIDTH = 300kHz
SWEEP = 800µs

Figure 17. Power versus time measurement for DCS0

Directional Coupler Alternatives

The DC401A board contains the LTC4401-1 power controller and Hitachi PF08107B dual-band power amplifier as well as a Murata dual band directional coupler and Murata diplexer (Figure 18). The directional coupler has a coupling loss of 14±1.5dB for the DCS frequencies and 19±1dB for the GSM frequencies. While the directional coupler is a viable solution, there is a cheaper and smaller solution that is comparable in performance (Figure 19).

This new scheme completely eliminates the directional coupler, 50Ω termination resistor, and 68Ω shunt resistor. Instead, the RF signal is fed directly to the diplexer from the power amplifier. The RF signal is coupled back to the LTC4401-1 via a capacitor and a series resistor. The component count is reduced by two.

The series capacitor should be in the range of 0.3pF to 0.4pF and have a tolerance of ±0.05pF or less. The tolerance is important because it directly affects how much RF signal is coupled back to the RF pin on the LTC4400/4401. ATC has ultra-low ESR, high Q microwave capacitors with the tolerances desired. The ATC 600S0R3AW250XT and ATC 600S0R4AW250XT are 0.3pF and 0.4pF capacitors with 0.05pF tolerance. These capacitors come in a small 0603 package. The series resistor is 49.9Ω with 2% tolerance as shown in Figure 19. There are several factors to consider when using this technique, such as board layout and loading in the main line. For example, parasitic effects can significantly alter the feedback network characteristics.

Conclusion

Linear Technology has introduced two new controllers to its RF power controller family. The LTC4400-1 and LTC4401-1 represent small, low power solutions for RF power control. The integration of the RF detector, auto zero system and compensated loop amplifier have produced a temperature stable RF power control solution. External and internal voltage offset changes due to temperature or power supply are cancelled whenever the part cycles through shutdown. These products are available in a small, low profile ThinSOT package and operate over a frequency range of 800MHz to 2700MHz. The demo boards discussed here and ramp-shaping software are available upon request. Demo boards featuring power amplifiers made by Anadigics, Conexant, Hitachi and RFMD are also available.

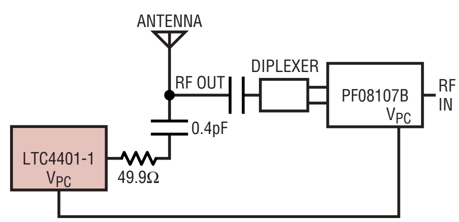


Figure 19. Block diagram of directional coupler alternative

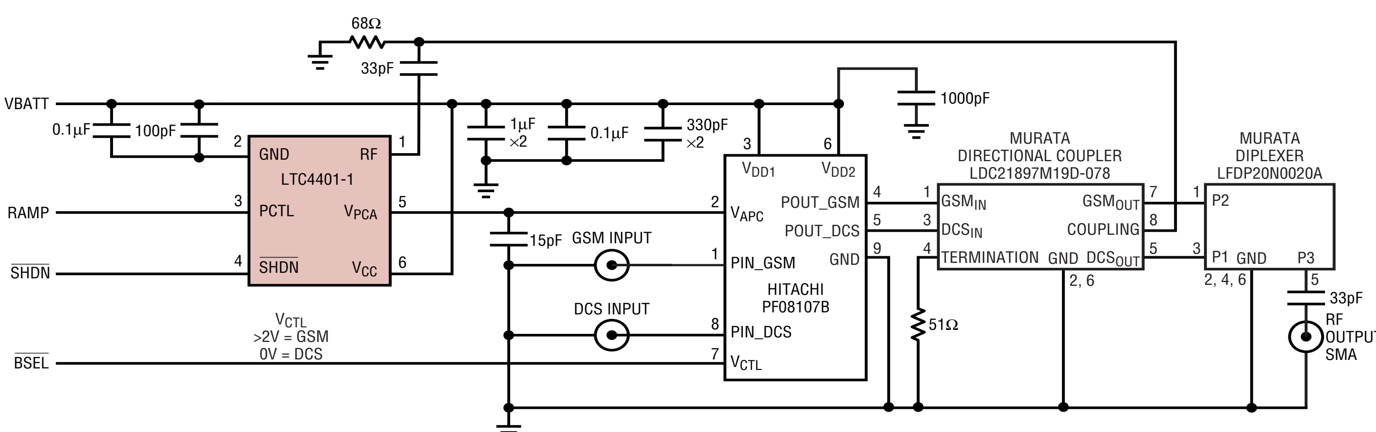


Figure 18. DC401A RF demo board schematic

Versatile LTC3830 and LTC3832 Deliver High Efficiency for Step-Down, Step-Up and Inverting Power Conversions

by Wei Chen and Charlie Zhao

Introduction

The LTC3830 and the LTC3832 are pin-to-pin compatible upgrades to the LTC1430—a popular IC for low voltage step-down applications due to its simplicity and high efficiency. The LTC3830 and the LTC3832 remove the LTC1430's frequency foldback at startup, thus eliminating inrush current and resulting output overshoot. Other improvements over the LTC1430 include tighter g_m distribution of the error amplifier and tighter current limiting. The LTC3832 is identical to the LTC3830, except that it incorporates a 0.6V reference for the output feedback, a larger g_m and a default frequency of 300kHz (instead of the 200kHz for the LTC3830), making it good match for very low output applications. The higher frequency of the LTC3832 also allows the use of smaller inductors and capacitors, making for a smaller overall solution.

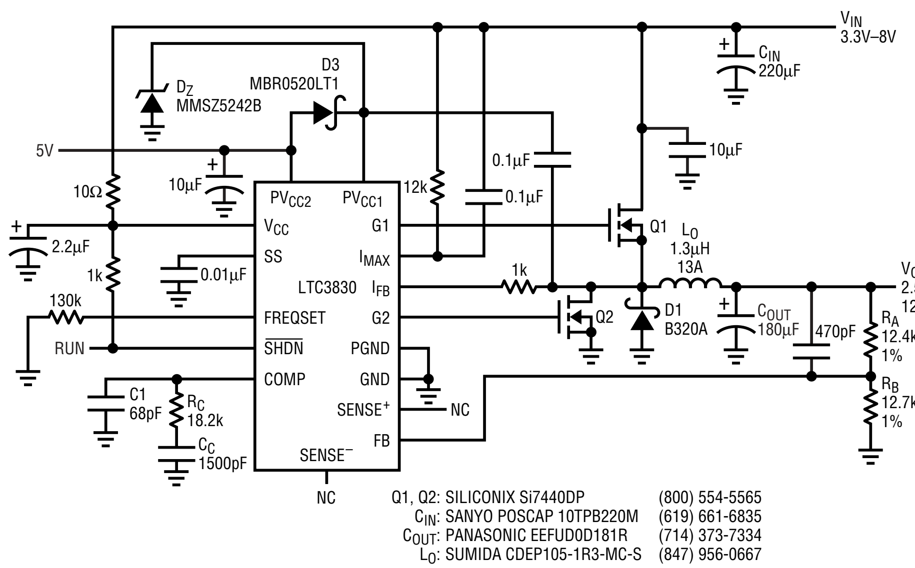


Figure 1. Schematic diagram of 2.5V/12A synchronous step-down power supply

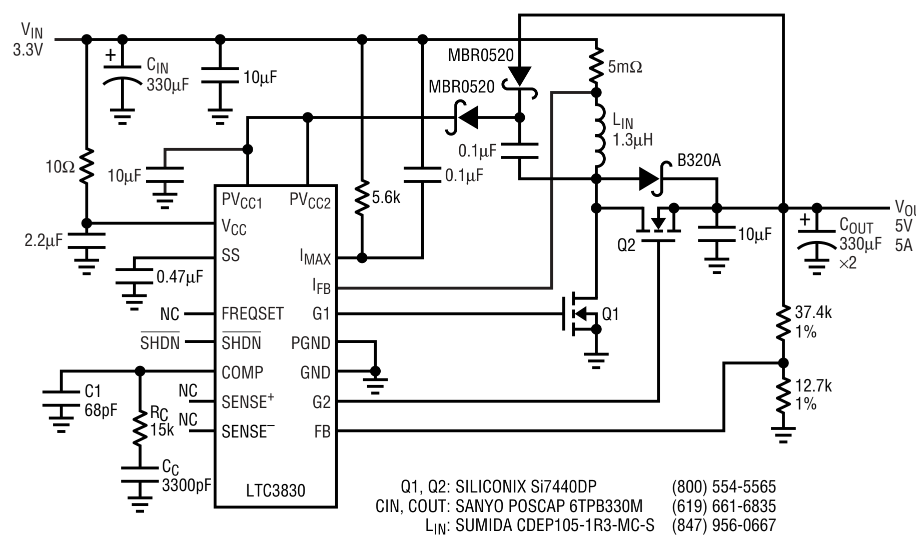


Figure 2a. Schematic diagram of 3.3V to 5V synchronous boost converter

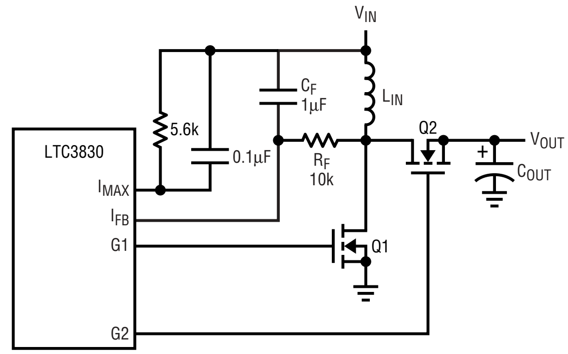


Figure 2b. How to use the DC resistance of the boost inductor to control current limiting

This article shows several designs using the LTC3830 for step down, step up and inverting applications. The LTC3832 can be used in place of the LTC3830 in any of these designs. All that is required are some minor adjustments to the feedback resistor divider and the compensation RC component values.

12A High Efficiency Step Down Power Supply Converts 3.3V–8V Input to a 2.5V Output

LTC3830/3832 are voltage mode synchronous buck controllers with two powerful MOSFET drivers for both the main MOSFET and a synchronous MOSFET. The $R_{DS(ON)}$ of the main MOSFET is used to establish the current limit, thus eliminating the sense resistor and its associated power loss. The current limit and switching frequency can be programmed easily through external resistors.

Figure 1 shows the schematic diagram of a 12A step down design based on LTC3830. The input is 3.3V to 8V and the output is 2.5V. To obtain different output voltages, vary the ratio of R_A/R_B. With only two tiny PowerPak SO8 MOSFETs and 300kHz switching frequency, this design achieves close to 90% efficiency with 5V input and 2.5V output. The overall footprint of this design is less than 1"×1.2", with all of the components placed on the same side of the board. For higher output currents, simply parallel more MOSFETs and use an inductor with a higher current rating.

5A Step Up Power Supply Converts 3.3V to 5V

Although intended for synchronous buck applications, LTC3830 and LTC3832 can also be used in other circuit topologies. Figure 2a shows a synchronous boost design using LTC3830 converting 3.3V to 5V. Compared to a conventional boost converter, this design uses a low $R_{DS(ON)}$ N-channel MOSFET to implement the synchronous rectification, therefore improving efficiency by 5% to 10%. The maximum output current is 8A with only two PowerPak SO8 MOSFETs. A current sense resistor is used for more accurate current limiting than can be achieved by sensing $R_{DS(ON)}$ of the MOSFET. One may also use the DCR of the inductor to implement the current limit function, as shown in Figure 2b. R_F and C_F filters out the AC voltage components of the inductor voltage to obtain the DC voltage drop on the DC resistance of the inductor. This scheme eliminates the sense resistor and its associated power loss, but the response to overcurrent conditions is slower than a topology that uses a sense resistor. The delay time is determined by the product of $R_F \bullet C_F$.

5A Inverter Converts 3.3V to –5V

The LTC3830 and LTC 3832 can also be used in inverting applications. Figure 3 shows a synchronous buck-boost power supply which converts 3.3V into –5V. The total V_{CC} supply voltage in this design is the sum of the absolute values of input and output voltages, which is about 8.3V; and the PV_{CC1} voltage is the V_{CC} voltage plus 5V, which is 13.3V. Since these voltage stresses are very close to the maximum voltage ratings for the LTC3830 and the LTC3832 ($V_{CC(MAX)}$ = 9V and $PV_{CC1(MAX)}$ = 14V), Zener diodes should be placed on V_{CC} and PV_{CC1} pins to provide overvoltage protection.

Conclusion

The LTC3830 and LTC3832 are versatile voltage mode controllers that can be used in variety of applications including step up, step down and voltage inversion. Their integrated high current MOSFET drivers and programmable frequencies allow users to minimize power loss and total solution size.

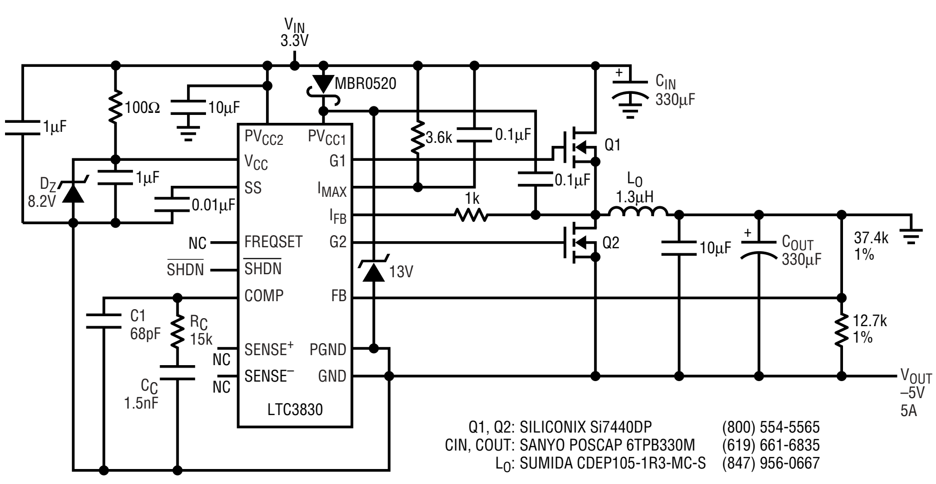


Figure 3. Schematic diagram of 3.3V to –5V inverting converter

How to Use the LTC6900 Low Power SOT-23 Oscillator as a VCO

by Nello Sevastopoulos

Introduction

The LTC6900 is a precision low power oscillator that is extremely easy to use and occupies very little PC board space. It is a lower power version of the LTC1799, which was featured in the February 2001 issue of this magazine.

The output frequency, f_{OSC}, of the LTC6900 can range from 1kHz to 20MHz—programmed via an external resistor, R_{SET}, and a 3-state frequency divider pin, as shown in Figure 1.

$$R_{SET} = 20k \bullet \left(\frac{10MHz}{N \bullet f_{OSC}}\right),\ N = \begin{cases}100\\10\\1\end{cases} \qquad (1)$$

A proprietary feedback loop linearizes the relationship between R_{SET} and the output frequency so the frequency accuracy is already included in the expression above. Unlike other discrete RC oscillators, the LTC6900 does not need correction tables to adjust the formula for determining the output frequency.

Figure 2 shows a simplified block diagram of the LTC6900. The LTC6900 master oscillator is controlled by the ratio of the voltage between V^+ and the SET pin and the current, I_{RES}, entering the SET pin. *As long as I_{RES} is precisely the current through resistor R_{SET}*, the ratio of $(V^+ - V_{SET}) / I_{RES}$ equals R_{SET} and the frequency of the LTC6900 depends solely on the value of R_{SET}. This technique ensures accuracy, typically ±0.5% at ambient temperature.

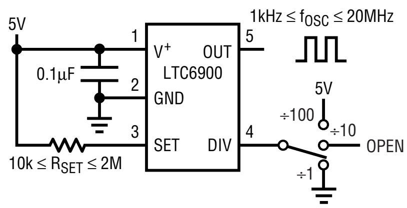


Figure 1. Basic connection diagram

As shown in Figure 2, the voltage of the SET pin is controlled by an internal bias, and by the gate to source voltage of a PMOS transistor. The voltage of the SET pin (V_{SET}) is typically 1.1V below V^+.

Programming the Output Frequency

The output frequency of the LTC6900 can be programmed by altering the value of R_{SET} as shown in Figure 1 and the accuracy of the oscillator will not be affected. The frequency can also be programmed by steering current in or out of the SET pin, as conceptually shown in Figure 3. This technique can degrade accuracy as the ratio of $(V^+ - V_{SET}) / I_{RES}$ is no longer uniquely dependent on the value of R_{SET}, as shown in Figure 2. This loss of accuracy will become noticeable when the magnitude of I_{PROG} is comparable to I_{RES}. The frequency variation of the LTC6900 is still monotonic.

Figure 4 shows how to implement the concept shown in Figure 3 by connecting a second resistor, R_{IN}, between the SET pin and a ground referenced voltage source V_{IN}.

For a given power supply voltage in Figure 4, the output frequency of the LTC6900 is a function of V_{IN}, R_{IN}, R_{SET}, and $(V^+ - V_{SET}) = V_{RES}$:

$$f_{OSC} = \frac{10MHz}{N} \bullet \frac{20k}{R_{IN} \| R_{SET}} \bullet \left[1 + \frac{(V_{IN} - V^+)}{V_{RES}} \bullet \left(\frac{1}{1 + \frac{R_{IN}}{R_{SET}}}\right)\right] \qquad (2)$$

When $V_{IN} = V^+$ the output frequency of the LTC6900 assumes the highest value and it is set by the parallel combination of R_{IN} and R_{SET}. Also note, the output frequency, f_{OSC}, is independent of the value of $V_{RES} = (V^+ - V_{SET})$ so, the accuracy of f_{OSC} is within the datasheet limits.

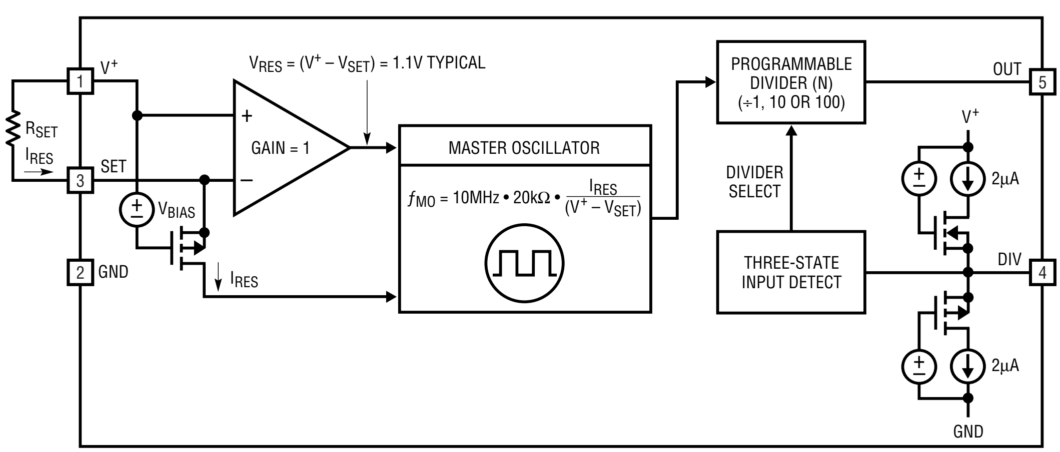


Figure 2. Simplified block diagram

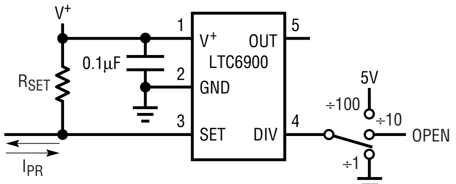


Figure 3. Concept for programming via current steering

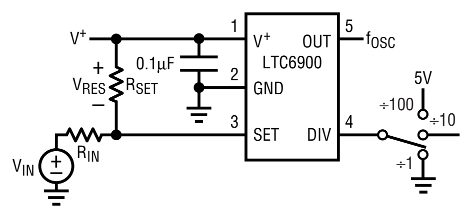


Figure 4. Implementation of the concept shown in Figure 3

When V_{IN} is less than V^+, and especially when V_{IN} approaches the ground potential, the oscillator frequency, f_{OSC}, assumes its lowest value and its accuracy is affected by the change of $V_{RES} = (V^+ - V_{SET})$. At 25°C V_{RES} varies by ±8%, assuming the variation of V^+ is ±5%. The temperature coefficient of V_{RES} is 0.02%/°C.

By manipulating the algebraic relation for f_{OSC} above, a simple algorithm can be derived to set the values of external resistors R_{SET} and R_{IN}, as shown in Figure 4:

1. Choose the desired value of the maximum oscillator frequency, $f_{OSC(MAX)}$, occurring at maximum input voltage $V_{IN(MAX)} \le V^+$.
2. Set the desired value of the minimum oscillator frequency, $f_{OSC(MIN)}$, occurring at minimum input voltage $V_{IN(MIN)} \ge 0$.
3. Choose $V_{RES} = 1.1$ and calculate the ratio of R_{IN}/R_{SET} from the following:

$$\frac{R_{IN}}{R_{SET}} = \frac{\left(V_{IN(MAX)} - V^+\right) - \left(\frac{f_{OSC(MAX)}}{f_{OSC(MIN)}}\right)\left(V_{IN(MIN)} - V^+\right)}{V_{RES}\left[\left(\frac{f_{OSC(MAX)}}{f_{OSC(MIN)}}\right) - 1\right]} - 1 \qquad (3)$$

Once R_{IN}/R_{SET} is known, calculate R_{SET} from:

$$R_{SET} = \frac{10MHz}{N} \bullet \frac{20k}{f_{OSC(MAX)}} \bullet \left[\frac{\left(V_{IN(MAX)} - V^+\right) + V_{RES}\left(1 + \frac{R_{IN}}{R_{SET}}\right)}{V_{RES}\left(\frac{R_{IN}}{R_{SET}}\right)}\right] \qquad (4)$$

Example 1: In this example, the oscillator output frequency has small excursions. This is useful where the frequency of a system should be tuned around some nominal value.

Let $V^+ = 3V$, $f_{OSC(MAX)} = 2MHz$ for $V_{IN(MAX)} = 3V$ and $f_{OSC(MIN)} = 1.5MHz$ for $V_{IN}=0V$. Solve for R_{IN}/R_{SET} by equation (3), yielding $R_{IN}/R_{SET} = 9.9/1$. $R_{SET} = 110.1k\Omega$ by equation (4). $R_{IN} = 9.9R_{SET} = 1.089M\Omega$. For standard resistor values, use $R_{SET} = 110k\Omega$ (1%) and $R_{IN} = 1.1M\Omega$ (1%). Figure 5 shows the measured f_{OSC} vs V_{IN}. The 1.5MHz to 2MHz frequency excursion is quite limited, so the curve f_{OSC} vs V_{IN} is linear.

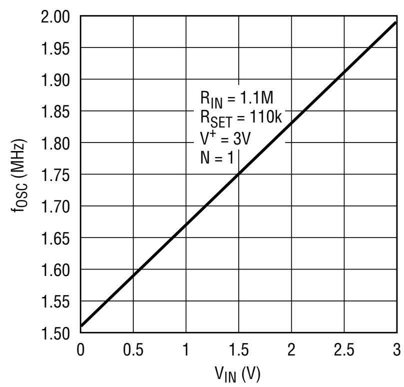


Figure 5. Output frequency vs input voltage

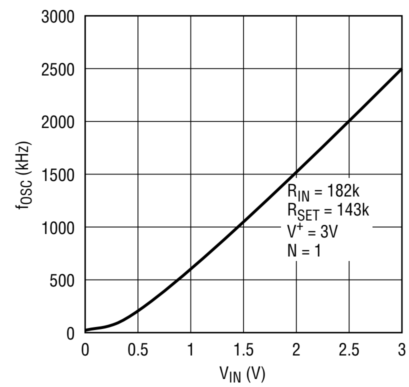


Figure 6. Output frequency vs input voltage

Example 2: Vary the oscillator frequency by one octave per volt. Assume $f_{OSC(MIN)} = 1MHz$ and $f_{OSC(MAX)} = 2MHz$, when the input voltage varies by 1V. The minimum input voltage is half supply, that is $V_{IN(MIN)} = 1.5V$, $V_{IN(MAX)} = 2.5V$ and $V^+ = 3V$.

Equation (3) yields $R_{IN}/R_{SET} = 1.273$ and equation (4) yields $R_{SET} = 142.8k\Omega$. $R_{IN} = 1.273R_{SET} = 181.8k\Omega$. For standard resistor values, use $R_{SET} = 143k\Omega$ (1%) and $R_{IN} = 182k\Omega$ (1%).

Figure 6 shows the measured f_{OSC} vs V_{IN}. For V_{IN} higher than 1.5V the VCO is quite linear; nonlinearities occur when V_{IN} becomes smaller than 1V, although the VCO remains monotonic.

The VCO modulation bandwidth is 25kHz that is, the LTC6900 will respond to changes in the frequency programming voltage, V_{IN}, ranging from DC to 25kHz.

Note:

All of the calculations above assume $V_{RES} = 1.1V$, although $V_{RES} \approx 1.1V$. For completeness, Table 1 shows the variation of V_{RES} against various parallel combinations of R_{IN} and R_{SET} ($V_{IN} = V^+$). Calculate first with $V_{RES} \approx 1.1V$, then use Table 1 to get a better approximation of V_{RES}, then recalculate the resistor values using the new value for V_{RES}.

Table 1: Variation of V_{RES} for various values of $R_{IN} || R_{SET}$

$R_{IN} \| R_{SET}$ ($V_{IN} = V^+$)	V_{RES}, $V^+ = 3V$	V_{RES}, $V^+ = 5V$
20k	0.98V	1.03V
40k	1.03V	1.08V
80k	1.07V	1.12V
160k	1.1V	1.15V
320k	1.12V	1.17V

V_{RES} = Voltage across R_{SET}

Save Space and Expense by Extracting Two Lowpass Filters Out of a Single LTC1563

by Doug La Porte

Introduction

Lowpass filters are required in systems for a variety of reasons: to limit the noise bandwidth, smooth out transition edges or remove unwanted signals. To make it easy for designers to use lowpass filters, Linear Technology Corporation developed the LTC1563-2 and LTC1563-3, for which a simple formula and a single resistor value set the cutoff frequency. The LTC1563 features two 2nd order building block sections, which can be cascaded to form a 4th order filter. Some applications, though, do not require the higher order filtering, but they do require more filters. For these applications, the LTC1563 building block sections can be used separately to produce a dual 2nd or 3rd order filter, thus saving the space and expense of additional ICs.

The FilterCAD™ filter design program from Linear Technology Corporation also helps designers create custom lowpass filters using Linear Technology Corporation products. FilterCAD does not directly support dual filters, but it can be tricked into putting one together. This article shows how to use FilterCAD and the LTC1563 to create a single IC dual lowpass filter.

About the LTC1563

The LTC1563 is designed to be an easy-to-use 4th order lowpass filter. The LTC1563-2 provides a Butterworth transfer function while the LTC1563-3 provides a Bessel response when applied with six equally valued resistors. The LTC1563 family is not limited to these transfer functions though. One can generate nearly any arbitrary fourth order transfer function with the LTC1563 by using varied resistor values. For custom filtering, use FilterCAD to analyze the frequency response and step response. Otherwise, using equally valued resistors, setting the cutoff frequency is simply a matter of choosing the appropriate resistor value:

$R = 10k \bullet (256kHz/f_C)$
where f_C = Cutoff Frequency

Figure 1 shows the LTC1563 circuit topology. As mentioned above, the 4th order filter is obtained by cascading two 2nd order section building blocks. The sections are similar, but not identical—their capacitor values are different. Figure 1 shows the LTC1563 with the two sections connected separately, instead of cascaded, to form two 2nd order filters, or with the addition of two capacitors (one for each filter), two 3rd order filters. The rest of this article shows how to design this and similar dual lowpass filters with the LTC1563.

Using FilterCAD to Design a Dual Filter with the LTC1563

The following procedure shows how to design a dual lowpass filter using FilterCAD. The accompanying illustrations show the design of a dual 3rd order filter: one filter is a 3rd order Butterworth with a cutoff frequency of 50kHz, and the other is a 3rd order Bessel with a cutoff of 100kHz. The values can be modified to fit other applications.

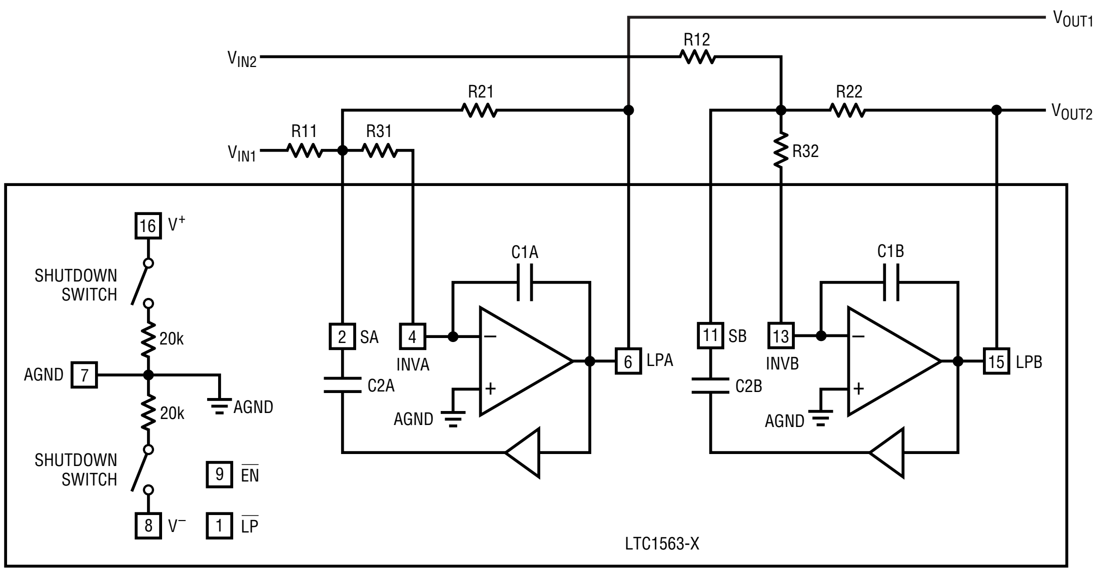


Figure 1. This block diagram shows how the LTC1563's two filter sections can be hooked up separately to yield a dual filter from a single-IC.

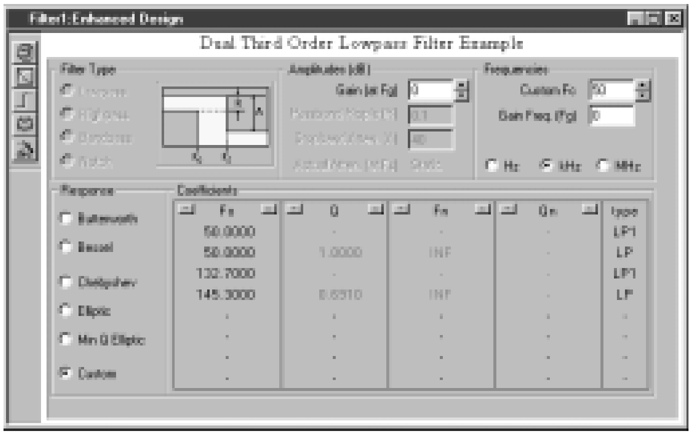


Figure 2. Enhanced Design window

The first order of business is to identify the filter order and transfer function. This is determined by the usual parameters of passband bandwidth, attenuation requirement and step response, though transfer function selection is a classical engineering trade-off problem. The "ideal brick wall" filter has outstanding attenuation just beyond the passband but suffers from a step response with large overshoot, substantial ringing and a long settling time. At the other end of the spectrum, filters with ideal step responses tend to have poor attenuation just beyond the passband. Choosing the best transfer function for any specific application ultimately requires a compromise. FilterCAD can help you decide, but you will need the values in Table 1 and a little trial and error.

Table 1 lists the coefficients for most of the popular 2nd and 3rd order lowpass filters. In the table, find the coefficients for the filters that best match your application needs. Then, enter the coefficients into FilterCAD to see the frequency and step responses of the filters. Here's how:

1. Launch FilterCAD.
2. Select the Enhanced Design option.
3. Click Next. The Enhanced Design window appears (Figure 2).
4. In the Enhanced Design Window, click Custom (for the Response item).
5. Enter 0 for the Gain Frequency (Fg), indicating a lowpass filter.
6. Enter the filter coefficients from Table 1 into the Coefficients table in FilterCAD.
7. Enter the cutoff frequency in the Custom Fc box. Note that the f_O entered in step 6 is now multiplied by the Custom Fc value.

FilterCAD can only evaluate one filter at a time, so you will need to enter the coefficients for one filter, evaluate it, and then replace those coefficients for the other filter to evaluate it. If you put in the coefficients for both filters, FilterCAD will assume you want the results of a composite filter, which is not what we are interested in here.

If you are designing a dual 2nd order filter, you only need to enter one row of coefficients for each filter. For a 3rd order, you need two rows: one 1st order (corresponding to an external RC) and one 2nd order (corresponding to a 2nd order section of the LTC1563). That's why in Table 1 there are two rows of coefficients for the 3rd order filters and only one row for the 2nd order filters.

For each 3rd order filter in this example, enter the first row of coefficients, and choose LP1 (1st order) as the coefficient type, corresponding to the 1st order external RC lowpass.

Enter the second row of coefficients, and choose LP (2nd order) as the coefficient type, corresponding to the 2nd order filter built into LTC1563.

8. Evaluate the filter by clicking the Frequency Response and Step Response buttons in the Enhanced Design window.
9. Adjust the coefficients to get the performance you require.
10. Repeat for the second filter.

Table 1: Coefficients for popular 2nd and 3rd order lowpass filters (f_O normalized for a 1Hz cutoff frequency)

Filter Type	Bessel		12dB Transitional Gaussian		6dB Transitional Gaussian		Butterworth		0.01dB Ripple Chebyshev		0.1dB Ripple Chebyshev		0.5dB Ripple Chebyshev	
Charactersitics														
Passband Gain	Flat, no ripple		Flat, no ripple		Flat, no ripple		Flat, no ripple		0.01dB ripple		0.1dB ripple		0.5dB ripple	
Attenuation Slope	Poor, unselective ⟷ Best, most selective													
Step Response	Best, no overshoot ⟷ Poor, most overshoot													
Coefficients	f_O	Q	f_O	Q	f_O	Q	f_O	Q	f_O	Q	f_O	Q	f_O	Q
2nd Order	1.2736	0.5773	–	–	–	–	1.0000	0.7071	0.9774	0.7247	0.9368	0.7673	0.8860	0.8638
3rd Order	1.4530	0.6910	1.5352	0.8201	1.5549	0.8080	1.0000	1.0000	0.9642	1.1389	1.2999	1.3409	1.0689	1.7062
	1.3270	–	0.9630	–	0.9776	–	1.0000	–	0.8467	–	0.9694	–	0.6289	–

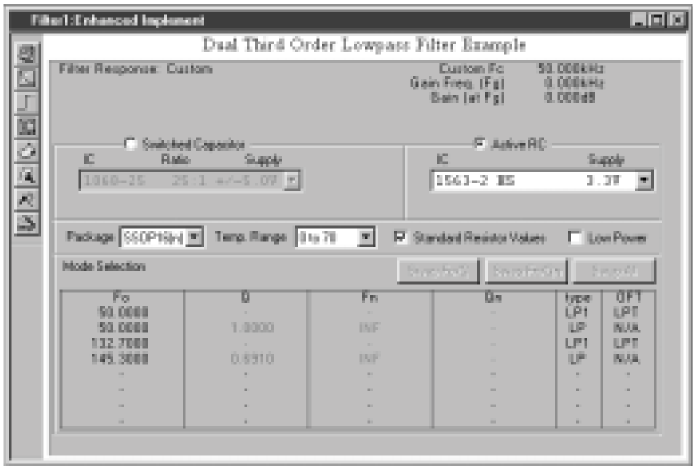


Figure 3. Enhanced Implement window

11. Once you have determined the coefficients of the two filters, enter them all in the Enhanced Design window custom response coefficient table.

Figure 2 shows this window for the 50kHz Butterworth and 100kHz Bessel example. The individual filters that make up the Butterworth and Bessel filters must be entered in a specific order. That is, enter one filter with the LP1 section listed first. Then, enter the second filter with its LP1 section first. At this point do not bother to look at the frequency or step response results, unless you are somehow interested the combined 6th order filter. This is where some trickery comes in. FilterCAD does not directly support dual filters, but it can still be used to design the dual filter that we want.

The next step is to choose the part you would like FilterCAD to use, in this case the LTC1563-2.

12. In the Enhanced Design window, click the Implement button. The Enhanced Implement window appears (Figure 3) with the coefficient table you entered in Enhanced Design.

13. Click the Active RC button.

14. Select LTC1563-2 from the list of available parts.

Why not use the LTC1563-3? The reason is that both the LTC1563-2 and the LTC1563-3 have f_O-Q limitations for the first building block section. The limitation is greater with the LTC1563-3, so use the LT1563-2.

Check the order of the sections to make sure that it hasn't changed from how you entered them in the Enhanced Design window. FilterCAD usually leaves the order alone, but sometimes it shuffles things a little. To change the order of any offending rows, click one of the rows to select it, and while pressing the Control key, click on another row to select it too. With both rows selected, click the Swap All button to swap the two rows. Figure 3 shows the Enhanced Implement window.

The final step is to generate the schematic for the dual filter.

15. Click the Schematic button in the Enhanced Implement window. The Schematic window appears showing a single 6th order lowpass filter. This is not exactly what we want, but it's easy to fix.

16. Print the schematic.

17. Fix the schematic.

Break out some Liquid Paper® and a pencil, and look at Figure 4. The connection from the first section to the second section must be broken. A dab or two of Liquid Paper should do the trick. Also, the inputs and outputs must be labeled. In Figure 4, V_{IN1} and V_{OUT1} correspond to the 50kHz Butterworth; V_{IN2} and V_{OUT2} correspond to the 100kHz Bessel.

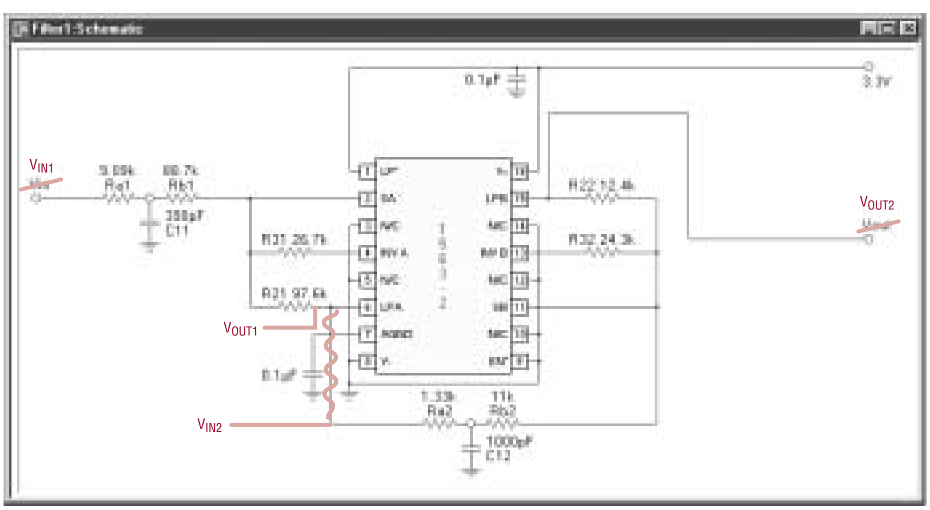


Figure 4. FilterCAD does not produce the exact schematic you want, but all you need are a few simple modifications to FilterCAD's design.

Conclusion

FilterCAD does not directly support single part dual filter design, but it can still help you design a dual filter with the LTC1563. The example in this article illustrates that the procedure is a little tricky, but the end result is a simple, compact and cost effective solution.

Liquid Paper is a registered trademark of the Gillette Company

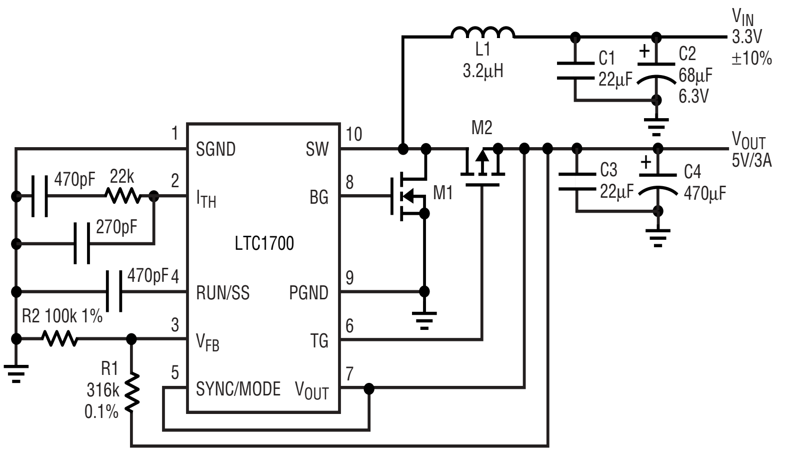
Tiny and Efficient Boost Converter Generates 5V at 3A from 3.3V Bus

by Dongyan Zhou

Introduction

Circuits that require 5V remain popular despite the fact that modern systems commonly supply a 3.3V power bus, not 5V. The tiny LTC1700 is optimized to deliver 5V from the 3.3V bus at very high efficiency, though it can also efficiently boost other voltages. The small MSOP package and 530kHz operation promote small surface mount circuits requiring minimal board space, perfect for the latest portable devices. By taking advantage of the synchronous rectifier driver, the LTC1700 provides up to 95% efficiency. To keep light load efficiency high in portable applications, the LTC1700 draws only 180µA in sleep mode. The LTC1700 features a start-up voltage as low as 0.9V, adding to its versatility.

The LTC1700 uses a constant frequency, current mode PWM control scheme. Its No R_{SENSE}™ feature means the current is sensed at the main MOSFET, eliminating the need for a sense resistor. This saves cost, space and improves efficiency at heavy loads. For noise-sensitive applications, Burst Mode operation can be disabled when the SYNC/MODE pin is pulled low or driven by an external clock. The LTC1700 can be synchronized to an external clock ranging from 400kHz to 750kHz.



C1, C3: TAIYO YUDEN CERAMIC JMK325BJ226M	(408) 573-4150
C2: AVX TAJB686K006R	(207) 282-5111
C4: SANYO POSCAP 6TPB470M	(619) 661-6853
L1: SUMIDA CEP1233R2	(847) 956-0667
M1: INTERNATIONAL RECTIFIER IR7811W	(310) 322-3331
M2: SILICONIX Si9803	(800) 554-5565

Figure 1. 3.3V to 5V, 3A boost regulator

3.3V Input, 5V/3A Output Boost Regulator

Figure 1 shows a 3.3V input to 5V output boost regulator which can supply up to 3A load current. Figure 2 shows that the efficiency is greater than 90% for a load current range of 200mA to 3A and stays above 80% all the way down to a 3mA load.

C2 is a tantalum capacitor providing bulk capacitance to compensate for possible long wire connections to the input supply. In applications where the regulator's input is con-

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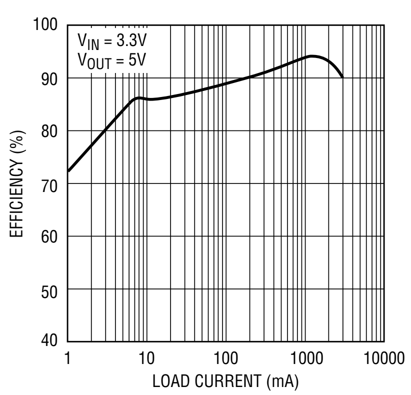
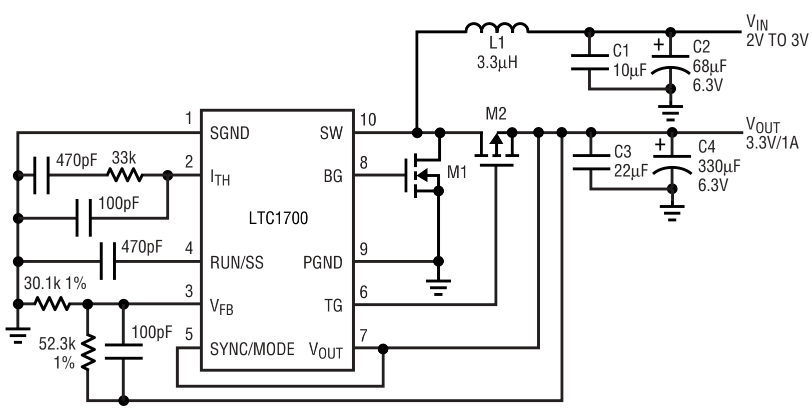


Figure 2. Efficiency of the circuit in Figure 1



C1: TAIYO YUDEN CERAMIC JMK316BJ106ML	(408) 573-4150
C2: AVX TAJB686K006R	(207) 282-5111
C3: TAIYO YUDEN CERAMIC JMK325BJ226M	
C4: SANYO POSCAP 6TPB330M	(619) 661-6853
L1: MURATA LQN6C	(814) 237-1431
M1: SILICONIX Si9804	(800) 554-5565
M2: SILICONIX Si9803	

Figure 3. 2-cell to 3.3V, 1A boost regulator

Small, Portable Altimeter Operates from a Single Cell

by Todd Owen

Some sports enthusiasts want to know altitude changes from an initial elevation. A small, lightweight, portable altimeter is easy to design using modern micromachined pressure transducers. Inverting barometric pressure and compensating for nonlinearities in air-pressure changes with respect to altitude produces a reasonably accurate altimeter.

Figure 1 shows a small, handheld altimeter based on a micromachined pressure transducer. The circuit takes advantage of the inverse relationship between air pressure and altitude. The aim of this circuit is to be small, lightweight, and portable. Accuracy is not paramount; errors as high as 3%, such as a 300ft error at 10,000ft altitude, are acceptable. The speed of the circuit is also not critical: Extreme changes in altitude in milliseconds may prove fatal to whoever is attempting to read the output.

The heart of the altimeter is an NPC-1220-015-A-3L pressure transducer. This 5k bridge provides 0mV to 50mV of output voltage for a 0psi to 15psi pressure range. To power the transducer and signal-conditioning circuitry, LT1307 (IC1), generates 5V from a single AA battery, and a charge pump generates a –5V supply. The pressure transducer is driven by IC3B (LT1490), which uses a reference voltage and a setting resistor on the transducer to generate appropriate drive current.

The output of the transducer drives an LT1167 instrumentation amplifier (IC2) which provides an initial gain of 21. A nonlinear gain stage, composed of IC3A and associated components, then inverts the output of the instrumentation to provide a voltage that is inversely proportional to air pressure. D4 and R1 introduce the nonlinear gain, and the final output is directly proportional to altitude.

R2 performs gain calibration in the signal-conditioning circuitry. This potentiometer calibrates out any normal variations in part tolerances and sets the altimeter for a 100mV change in output for every 1000ft of altitude. The circuit has some initial offset, as well as an offset that is determined by barometric pressure variations. You can use R3 to R5 to null this offset, giving a 0V to 1V output for 0ft to 10,000ft of altitude.

Altimeter testing was performed using a DeHavilland DHC-6 Twin Otter for an ascent to 13,000ft, followed by free descent—limited by the engineer's parasitic drag—to 3000ft. Subsequent deployment of an aerodynamic decelerator (Precision Aerodynamics Icarus Omega 190) prevented engineer injury or circuit damage. Aircraft rental for testing is available at many local airports. Extensive instruction in free descent and the use of aerodynamic decelerators are highly recommended before undertaking testing of this nature. Contact USPA at (703) 836-3495 for further information.

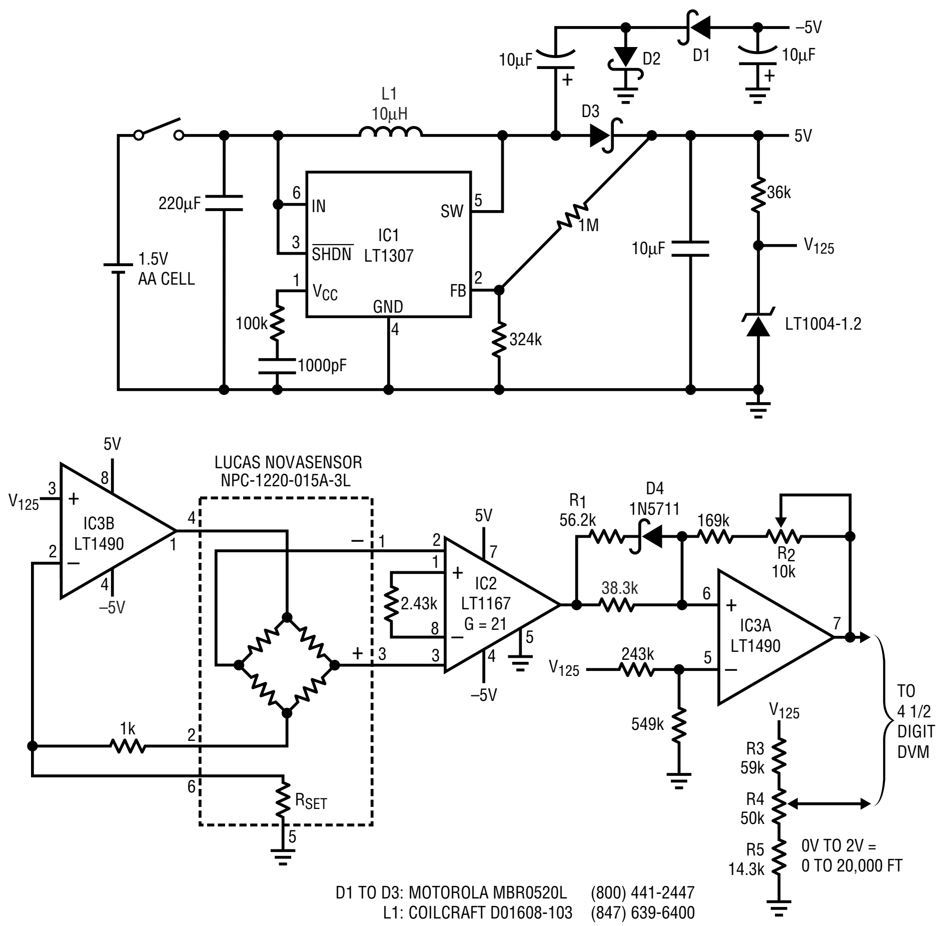


Figure 1. To produce a reasonably accurate altimeter, conditioning circuitry inverts the barometric pressure of a micromachined pressure transducer and compensates for nonlinearities in air-pressure changes with respect to altitude.

Simple Isolated Telecom Flyback Circuit Provides Regulation Without Optocoupler

by John Shannon

Introduction

The LT1725 uses a proprietary technique to regulate an isolated output voltage *without* an optocoupler, thus greatly simplifying flyback converter design and reducing the component count. The result is reduced design time, smaller space requirements, lower cost, and improved performance.

Traditional isolated flyback converters employ a secondary side voltage reference and error amplifier that drive an optocoupler, which sends the control signals back to the primary side. In addition to being parts intensive, this approach places an optocoupler in the feedback loop, which introduces a host of design problems. Optocouplers are poorly defined components—their gain is variable and subject to degradation over time. They are also relatively slow. Optocoupler shortcomings add considerably to the total converter design time and ultimately limit performance.

Consider instead the schematic of Figure 1. This is a flyback converter based on the LT1725. There are extremely few components and yet a high level of functionality. This design is short circuit proof and includes an input undervoltage lockout for increased reliability. The performance of this converter is shown in Figure 2. Output voltage is regulated to within 1% over a 2:1 input voltage range with 10% or greater load. No load regulation is within 2% over a 2:1 input voltage range. This is well within the typical requirement of 5% regulation.

Circuit Operation

The LT1725 flyback controller is a current mode control IC. Current mode operation provides for inherent line transient rejection and simple loop compensation. Current mode controllers have an "inner" fast current control loop and a slower "outer" voltage control loop. The inner current loop has immediate pulse-by-pulse control of the switching MOSFET M1. A normal switching cycle is as follows. The MOSFET M1 is turned on to begin the cycle. Once M1 is turned on, the current in the primary winding of the flyback transformer ramps up. When the primary current reaches a level determined by the value of the voltage on the V_C pin, M1 is turned off. The voltage on the V_C pin is set by the LT1725's output voltage control loop—the outer loop. Once M1 turns

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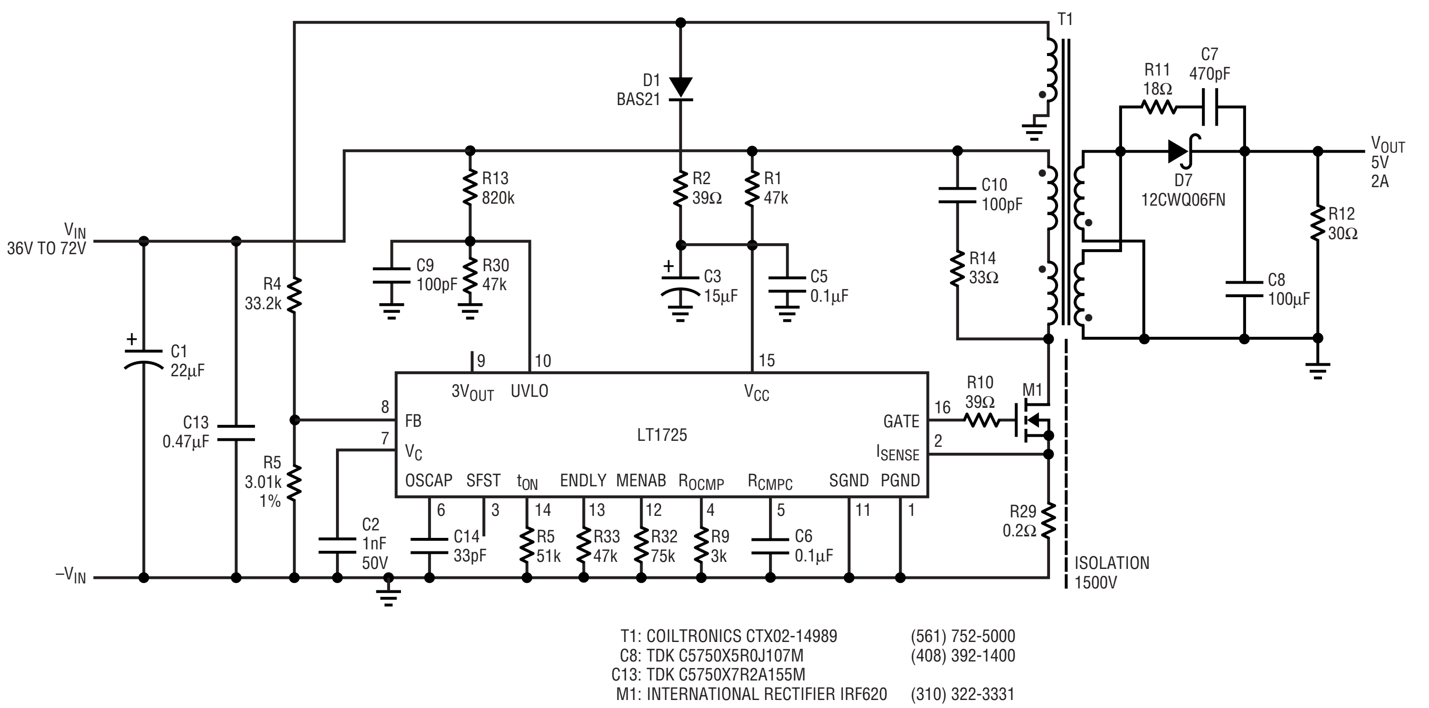

Figure 1. –48V to 5V 2A isolated flyback converter

Space Saving Dual Output ±5V High Current Power Supply Requires Only One 1.25MHz Switcher and One Magnetic Component

by Keith Szolusha

ADSL modems, disc drives, notebook computers, and other data acquisition circuits require high current, ±5V power supplies with switching frequencies greater than 1.1MHz to avoid interfering with noise sensitive circuitry. Figure 1 shows a very simple, compact and efficient solution that uses a single 1.25MHz LT1765EFE monolithic step-down switching regulator and only one magnetic component. This circuit can provide ±5V supplies from a 12V source with greater than 1A capabilities on both rails. The LT1765EFE's internal 3A power switch saves space by eliminating the requirement for an external MOSFET and its traces. Typical efficiency is 84%, as shown in Figure 2. An alternative option is to use two ICs, which means paying a heavy toll in board space, overall cost, and complexity.

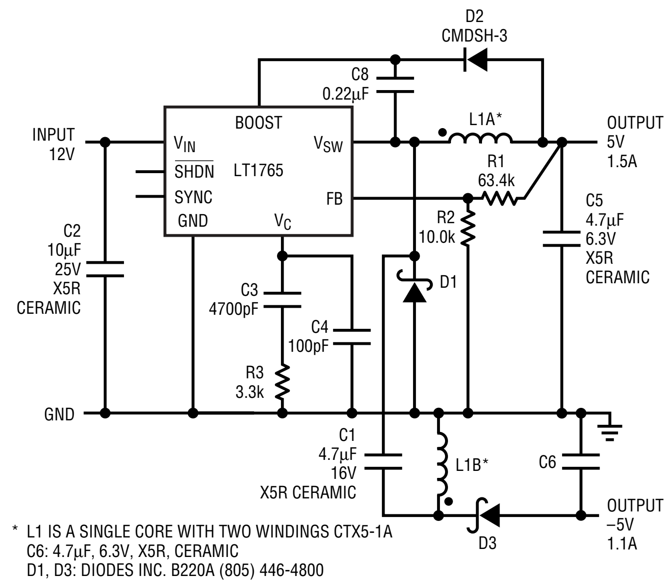

Figure 1. This high-current dual output power supply conserves space by consolidating the magnetics into a single component (L1) and by using ceramic capacitors.

The LT1765EFE uses current-mode control to regulate the positive output with its step-down converter topology. The off-the-shelf CTX5-1A(L1) transformer, which has a greater than 3A current rating and a 1:1 turns ratio, induces the same voltage across the secondary winding as the primary winding and maintains a –5V output. A high current-density ceramic coupling capacitor creates a low-impedance path for current to run between the IC and the negative output, maintaining excellent cross-regulation, as shown in Figure 3. The 3A minimum switch-current limit of the LT1765EFE and the thermally

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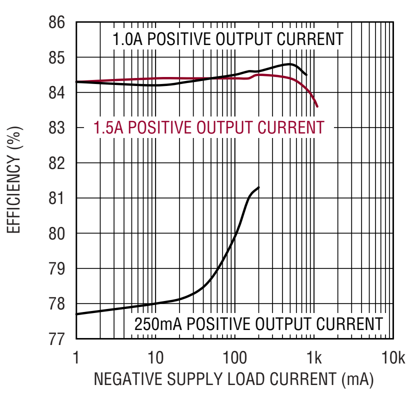


Figure 2. The efficiency of the circuit in Figure 1 is typically greater than 80%, and as high as 85% for varying output currents.

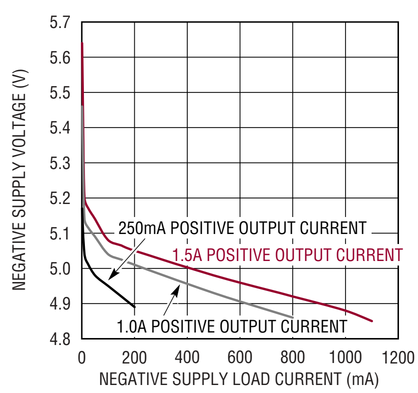


Figure 3. The negative (–5V) supply maintains excellent regulation (±3%) over a wide range of output currents without the use of a negative supply feedback network.

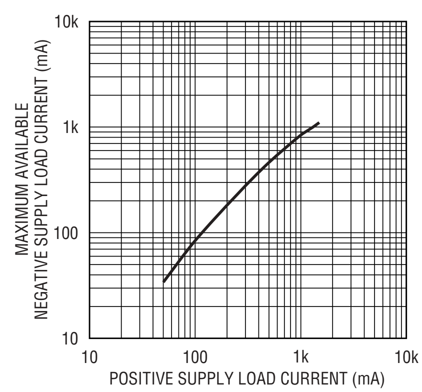


Figure 4. The available negative output current (±3% voltage regulation on –5V output) increases as positive supply (5V) current increases until switch current or thermal limitation are reached.

Efficient DC/DC Converter Provides Two 15A Outputs from a 3.3V Backplane

by David Chen

Introduction

The 3.3V DC bus has become popular for broadband networking systems, where it is tapped for a variety of lower voltages to power DSPs, ASICs and FPGAs. These lower voltages range from 1V to 2.5V and often require high load currents. To maintain high conversion efficiency, power MOSFET conduction losses from the step-down converters must be minimized. The problem is that the 3.3V bus also brings with it frequent use of sub-logic level MOSFETs. Such MOSFETs have a relatively high $R_{DS(ON)}$, limiting the full-load efficiency of a converter to around 85%. A more efficient solution is to use logic-level MOSFETs, which have very low $R_{DS(ON)}$ but require a 5V supply. The LTC1876 allows the use of logic-level MOSFETs by combining a 1.2MHz boost regulator, which produces a 5V bias supply from a 3.3V input, with two step-down controllers, which provide the low voltage outputs. By integrating all three regulators in a single IC, the LTC1876 makes for efficient power supplies that can be small and inexpensive.

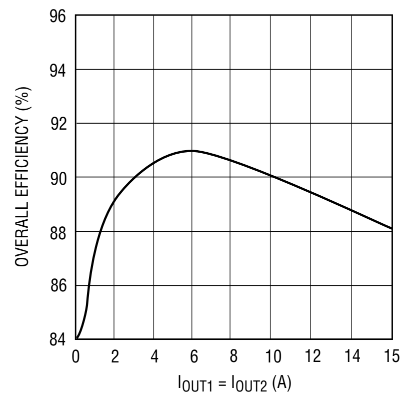


Figure 2. High efficiency of the design in Figure 1

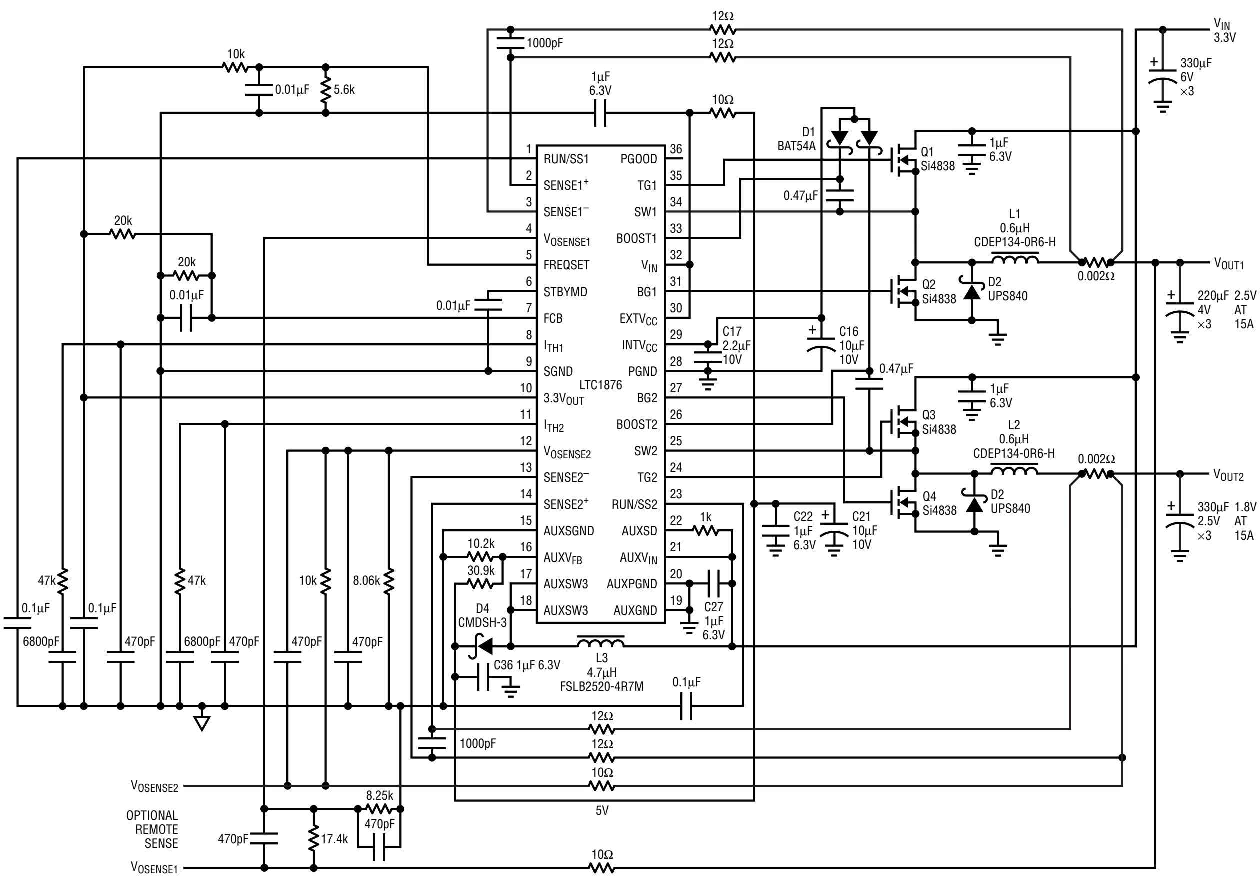


Figure 1. An LTC1876 design converts 3.3V to 2.5V at 15A and 1.8V at 15A

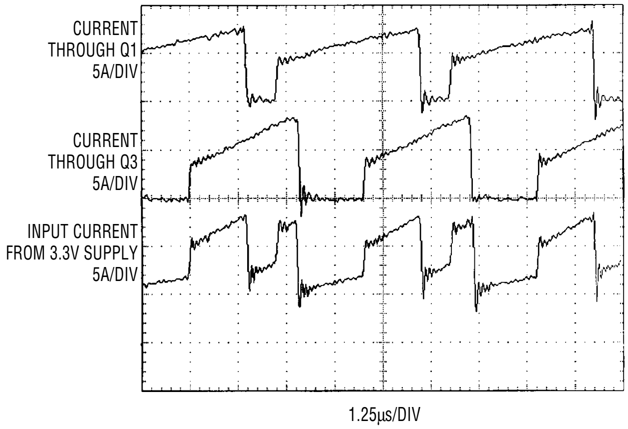


Figure 3. Each switcher has 5A peak current, but the total ripple at the input is still only 5A, minimizing C_{IN} requirements.

Design Example

Figure 1 shows a design that provides 2.5V/15A and 1.8V/15A from a 3.3V input. Because the LTC1876 provides a 5V bias for MOSFET gate drive, a very low $R_{DS(ON)}$ MOSFET Si4838 (2.4mΩ typical) can be used to achieve high efficiency. Figure 2 shows that the overall efficiency is above 90% over a wide range of loads.

Figure 2 also shows that the light load efficiency of this design is more than 84%. This is a direct benefit of the Burst Mode operation of the LTC1876. Further efficiency improvements come from operating the two step-down channels out-of-phase. The top MOSFET of the first channel is fired 180° out of phase from that of the second channel, thus minimizing the RMS current through the input capacitors. This significantly reduces the power loss associated with the ESR of input capacitors. Figure 3 shows detailed current waveforms of this operation.

Conclusion

The LTC1876 uses three techniques to efficiently power low voltage DSPs, ASICs and FPGAs from a low input voltage. The first technique uses an internal boost regulator to provide a separate 5V for the MOSFET gate drive. Secondly, its Burst Mode operation achieves high efficiency at light loads. Lastly is the out-of-phase technique which minimizes input RMS losses and reduces input noise. Complete regulator circuits are kept small and inexpensive, because all three switchers (one step-up regulator and two step-down controllers) are integrated into a single IC. For systems where a separate 5V is available or the input supply is greater than 5V, the internal boost regulator can be used to provide a third step-up output with up to 1A switch current.

LT1725, continued from page 30

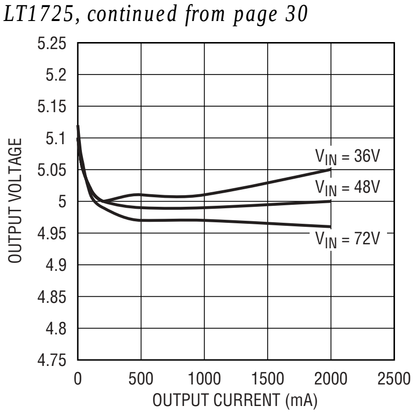


Figure 2. LT1725 regulation

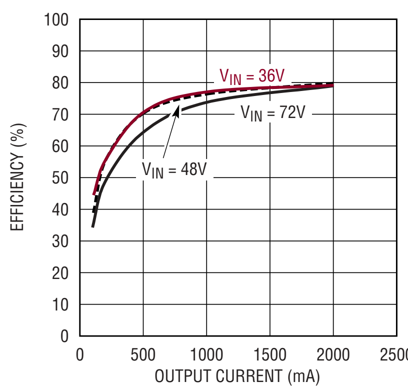


Figure 3. Efficiency vs output current for the circuit in Figure 1

off, the current that had been flowing in the primary of the transformer begins to flow in the secondary. The voltage on the drain of M1 rises to a level determined by the transformer turns ratio and the output voltage. Similarly, the voltage on the feedback winding rises to a level set by the output voltage. The LT1725 reads the voltage on the feedback winding during the flyback pulse using a proprietary sampling technique. This sampled voltage is then compared a precision internal reference and current is added to or subtracted from the capacitor on the V_C pin. This has the effect of modifying the M1 turn-off current in such a way as to regulate the output voltage. An important benefit of this sampling technique is that output voltage information arrives at the controller about a microsecond after the switching cycle is terminated. In a conventional optocoupler-based design. Delays of tens to hundreds of microseconds occur in the optocoupler alone, severely limiting the converters transient response. Additionally the LT1725 features internal slope compensation. This suppresses sub-harmonic oscillations that can occur with less sophisticated current mode controllers. Sub-harmonic oscillations increase output voltage ripple and increase switching stress.

Conclusion

The LT1725 isolated flyback controller greatly simplifies the design of isolated flyback converters. Compared to traditional opto-isolated designs, an LT1725 based circuit has far fewer components, superior transient response and is easier to stabilize.

Design Low Noise Differential Circuits Using the LT1567 Dual Amplifier Building Block

by Philip Karantzalis

Introduction

Many communications systems use differential, low level (400mV – 1V peak-to-peak), analog baseband signals, where the baseband circuitry operates from with a single low voltage power supply (5V to 3V). Any differential amplifier circuit used for baseband signal conditioning must have very low noise, and an output voltage swing that includes most of the power supply range for maximum signal dynamic range. The LT1567, a low noise operational amplifier ($1.4nV/\sqrt{Hz}$ voltage noise density) and a unity-gain inverter, is an excellent analog building block (see Figure 1) for designing low noise differential circuits. The typical gain bandwidth of the LT1567 amplifier is 180MHz and op amp slew rate is sufficient for signal frequencies up to 5MHz. The LT1567 operates from 2.7V to 12V total power supply. The output voltage swing is guaranteed to be 4.4V and 2.6V peak-to-peak, at 1k load with a single 5V and 3V power supply respectively. The LT1567 is available in a tiny MS8 surface mount package.

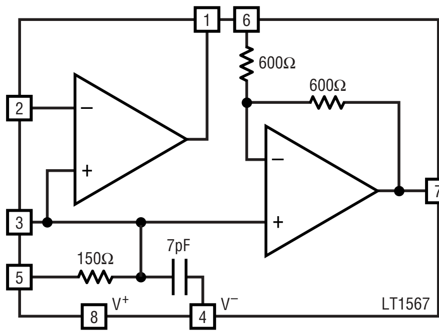


Figure 1. LT1567 analog building block

A Single-Ended To Differential Amplifier

Figure 2 shows a circuit for generating a differential signal from a single-ended input. The differential output noise is a function of the noise of the amplifiers, the noise of resistors R1 and R2 and the noise bandwidth. For example, if R1 and R2 are each 200Ω, the differential voltage noise density is $9.5nV/\sqrt{Hz}$ and in a 4MHz noise bandwidth the total differential noise is $19\mu V_{RMS}$ (with a low level $0.2V_{RMS}$ differential signal, the signal-to-noise ratio is an excellent 80.4dB). The voltage on Pin 5 (V_{REF}) provides flexible DC bias for the circuit and can be set by a voltage divider or a reference voltage source (with a single 3V power supply, the V_{REF} range is $0.9V \le V_{REF} \le 1.9V$). In a single supply circuit, if the input signal is DC coupled, then an input DC voltage (V_{INDC}) is required to bias the input within the circuit's linear region. If V_{INDC} is within the V_{REF} range, then V_{REF} can be equal to V_{INDC} and the output DC common mode voltage (V_{OUTCM}) at V_{O1} and V_{O2} is equal to V_{REF}. To maximize the unclipped LT1567 output swing however, the DC common mode output voltage must be set at $V^+/2$. In addition, the input signal can be AC coupled to the circuit's input resistor R1 and V_{REF} set to the DC common mode voltage required by any following circuitry (for example the input of an I and Q modulator).

A Differential Buffer/Driver

Figure 3 shows an LT1567 connected as a differential buffer. The differential output voltage noise density is $7.7nV/\sqrt{Hz}$. The differential buffer circuit of Figure 3, translates the input common mode DC voltage (V_{INCM}) to an output common mode DC voltage (V_{OUTCM}) set by the V_{REF} voltage ($V_{OUTCM} = 2 \bullet V_{REF} - V_{INCM}$). For example, in a single 5V power supply circuit, if V_{INCM} is 0.5V and V_{REF} is 1.5V then V_{OUTCM} is 2.5V.

A Differential to Single-Ended Amplifier

Figure 4 shows a circuit for converting a differential input to a single-ended output. For a gain equal

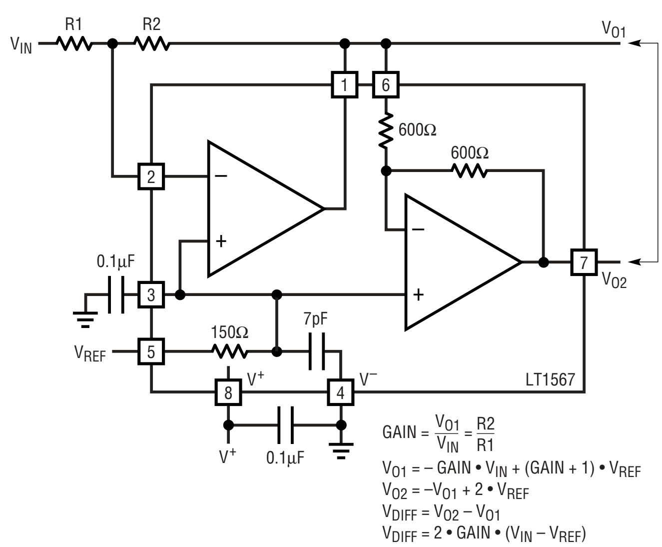


Figure 2. A single-ended input to differential output amplifier

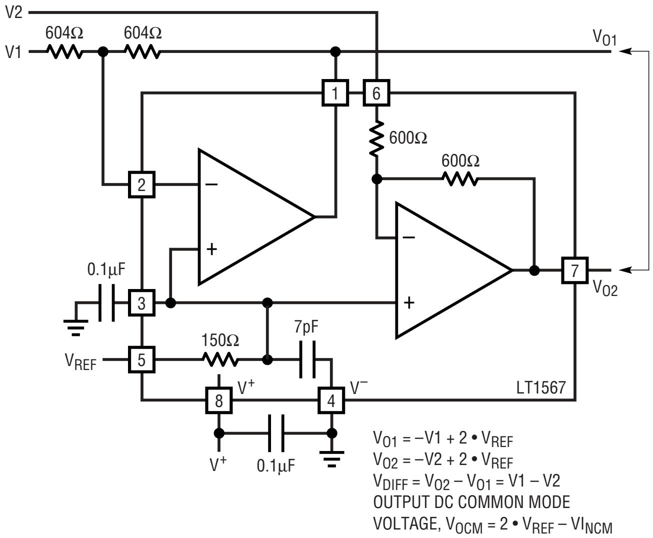
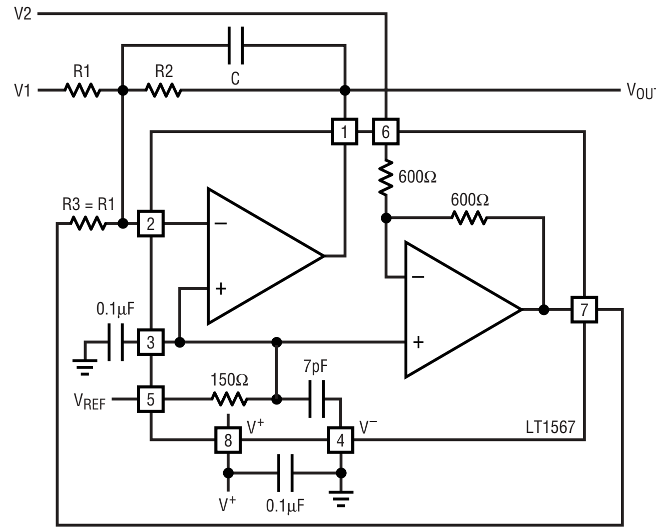


Figure 3. A differential input and output buffer/driver



$\text{GAIN} = \frac{R2}{R1}$, R3 = R1

$V_O = \text{GAIN}\,(V2 – V1) + V_{REF}$

$f_{–3dB}$ BANDWIDTH AT $V_{OUT} = \frac{1}{2 • \pi • R2 • C} \leq 5\text{MHz}$

IF R1 = R3 = 604Ω, THEN

R2	V_η	GAIN
604Ω	9.0	1
1.21k	8.4	2
2.43k	8.1	4

NOISE AT $V_{OUT} = \text{GAIN} • V_\eta • \sqrt{f_{\eta BW}}$ $\quad f_{\eta BW} = 1.57 • f_{–3dB}$

V_η IS THE INPUT REFERRED DIFFERENTIAL VOLTAGE NOISE DENSITY IN $nV/\sqrt{Hz}$

Figure 4. A differential input-to-single-ended output amplifier

to one (R1 = R2 = 604Ω and V_{OUT} = V2 – V1) the input referred differential voltage noise density is $9nV/\sqrt{Hz}$ and differential input signal-to-noise ratio is 80.9dB with $0.1V_{RMS}$ input signal in a 4MHz noise bandwidth. The input AC common mode rejection depends on the matching of resistors R1 and R3 and the LT1567 inverter gain tolerance (common mode rejection is at least 40dB up to 1MHz with one percent resistors and two percent inverter typical gain tolerance). If the differential input is DC coupled, then V_{REF} must be set equal to input common mode voltage (V_{INCM}) (if V_{REF} is greater than V_{inCM} then a peak voltage on Pin 7 may exceed the output voltage swing limit). The DC voltage at the amplifier's output (V_{OUT}, Pin 1) is V_{REF}.

Conclusion

With one LT1567 and two or three resistors, it is easy to design low noise, differential circuits for signals up to 5MHz. The LT1567 can also be used to make of low noise second and third order lowpass filters and second order bandpass filters with differential outputs. See www.linear.com for a spreadsheet-based design tool for just this purpose.

LTC1700, continued from page 28

nected very close to a low impedance supply, this capacitor is not needed.

2-Cell Input, 3.3V/1A Output Regulator

In digital cameras and other battery-powered devices, the LTC1700 makes for a high efficiency boost regulator in a small package. Figure 3 shows a 2-alkaline cell to 3.3V output circuit. This circuit can supply 1A maximum output current. Figure 4 shows the efficiency at different battery voltages. Efficiency of this circuit peaks at 93%. If a lower $R_{DS(ON)}$ MOSFET (such as Si6466) is used for M1, the maximum output current can be increased to 1.4A with about a 2% reduction in efficiency due to the increase in gate capacitance. MOSFETs with lower than 2.5V gate threshold voltages are recommended. The LTC1700 is also an ideal device for single cell Li-Ion battery to 5V applications.

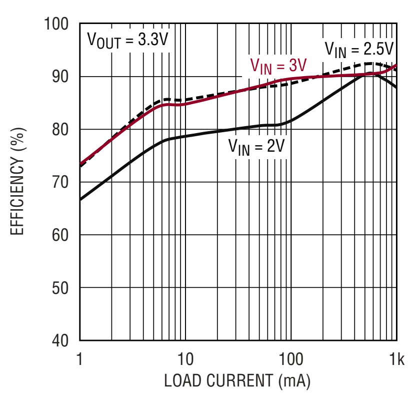


Figure 4. Efficiency of the circuit in Figure 3

Conclusion

The LTC1700 boost controller brings high efficiency and small size to low voltage applications. Its features are ideally suited to both battery-powered and line-powered applications.

New Device Cameos

LTC1706-85 Meets the Requirements of Intel VRM 8.5 Specification for Laptops

Linear Technology announces the release of the LTC1706-85 VID Voltage Programmer to address the requirements of the new Intel® Voltage Regulator Module 8.5 specification. The LTC1706-85 is a precision, digitally-programmed resistive divider which adjusts the output of any 0.8V referenced regulator. The LTC1706-85 adheres to Intel's VRM 8.5 specification and can provide voltages from 1.05V to 1.825V in 25mV increments based on the state of the VID inputs. With an exceptionally tight output accuracy of ±0.25%, the LTC1706-85 in turn relaxes the accuracy requirements on the associated DC/DC converter while still allowing the VRM to easily meet the Intel output specification.

The LTC1706-85 has been designed to work over a wide range of input voltages. Each VID input contains an internal 40kΩ pullup resistor with an integrated blocking diode. The inputs assume a high state if left unconnected, and it is acceptable to drive the VID inputs with as much as 5V while powering the LTC1706-85 from as little as 2.7V, increasing design flexibility.

The LTC1706-85 has a versatile architecture that allows it to be teamed with a wide variety of Linear Technology DC/DC converters. This flexibility makes a large selection of regulator solutions available to power Intel and other microprocessors. For instance, the LTC1706-85 and the LTC1778 can be used together to create a high-efficiency power supply that can tolerate the high input voltage requirement of laptop computers. For extremely high current applications, such as servers and workstations, one can use a single LTC1706-85 to program up to six LTC1629 polyphase high efficiency step-down DC/DC converters. The end result is an extremely compact, powerful, and programmable power supply that relies only on surface-mount components.

Intel is a registered trademark of Intel Corporation

The LTC1778 and LTC1629 are two of the many 0.8V reference DC/DC converters that are served by the LTC1706-85. It also works well with the LTC1622, LTC1628, LTC1702, LTC1735, LTC1772, LTC1773, LTC1929 or LTC3728. In addition, Linear Technology Corp. offers the LTC1709-85 as a single-chip solution to the Intel VRM 8.5 specification.

The LTC1706-85 comes in a miniature 10-lead MSOP package and is specified to operate from –40°C to 85°C.

Versatile Op Amp Family Brings Low Noise and High Speed to Low Voltage Applications

The LT1722, LT1723, and LT1724 are single, dual, and quad operational amplifiers that offer a unique combination of high performance features and low supply current. Performance is fully specified for operation from +5V or ±5V supplies. Each amplifier typically draws a mere 3.7mA, yet offers 200MHz gain-bandwidth-product and 70V/µs slew rate. The amplifiers are unity-gain stable, even while driving capacitive loads up to 100pF. The low input noise densities of $3.8\text{nV}/\sqrt{\text{Hz}}$ and $1.2\text{pA}/\sqrt{\text{Hz}}$ allow designers to build sensitive, high-speed preamps previously requiring amplifiers with 12V–15V split supplies.

The DC characteristics are as impressive as the AC properties, including guaranteed sub-millivolt offset and low residual bias-cancelled input current under 350nA. The bias-cancellation also eliminates the need for matched source resistances, reducing the resistor count in many designs.

The LT1722 single op amp is available in the compact SOT-23 5-lead, or a SO style 8-lead surface-mount package. The LT1723 dual is available in the SO 8-lead package. The LT1724 quad comes in the 14-lead SO package. Each device is available in both commercial and industrial temperature range versions.

With the versatility of having low-noise and offset, along with high gain-bandwidth and packaging options, the LT1722, LT1723, and LT1724 can provide optimal solutions for many sensor preamp, line receiver, line driver, and other applications where high speed, precision, and signal fidelity are key requirements.

Multiprotocol Transceiver Family Works from Single 3.3V Supply

The LTC2844, LTC2845 and LTC2846 are a new family of multiprotocol transceivers designed to operate from a single 3.3V supply and interface with 3.3V logic. These devices are the 3.3V counterparts to the 5V LTC1544, LTC1545 and LTC1546. When combined with the LTC2844 or LTC2845, the LTC2846 forms a complete software configurable DTE or DCE interface that supports RS232, RS449, EIA530, EIA530-A, V.35, V.36 and X.21 protocols. Unlike 3-chip solutions from other manufacturers, cable termination resistors are provided on-chip. The chip set supports V.11 data rates of up to 10Mb/s and V.28 modes of 128kbps and the receivers feature failsafe operation in all modes.

The desired protocol is selected via three mode pins, M0, M1 and M2, which can be driven by a microprocessor, or they can be hardwired in the connector (allowing the protocol to be selected by simply plugging in the appropriate cable). The DCE/DTE pin allows the microprocessor to configure a port as a DCE or DTE port. This pin may also be hardwired to fix the port as a DCE or DTE or to make the selection when the appropriate cable is plugged in.

The LTC2846 consists of a boost switching regulator, a charge pump, three configurable drivers, three configurable receivers and precision resistor termination networks. It serves to handle data and clock signals in the DTE or DCE interface and provides the necessary termination for each protocol. It also provides power to the LTC2844 or LTC2845 companion chip and generates the 5V and ±8V voltage levels needed for the various protocols. The LTC2844 or LTC2845 handle the control signals in the DTE or DCE interface. The LTC2844 has four drivers and four receivers and provides for an optional local loop-back test signal. The LTC2845 has five drivers and five receivers and allows users to add remote loop-back as well as test mode signals.

The LTC2844 is available in a 28-lead SSOP package, while the LTC2845 and LTC2846 are packaged in 36-lead SSOP packages. Both industrial and commercial grades are offered. The LTC2844/LTC2846 and LTC2845/LTC2846 chip sets are in the process of being certified for NET1, NET2 and TBR2 compliance.

Accurate, Low Power and Fast 80MHz Amplifiers Provide Best Solution for Low Voltage Signal Conditioning

The LT1801 and LT1802 are dual and quad, low power, high speed rail-to-rail input and output operational amplifiers. The LT1801 and LT1802 amplifiers consume a mere 2mA max supply current per amp and still provide 80MHz gain-bandwidth product and DC accuracy required by low voltage signal conditioning and data acquisition systems.

The DC performance is exceptional with a maximum input offset voltage of 350µV and input bias current of 250nA. These results come from internal trimming of the input offset voltage and employing Linear Technology Corporation's patent pending technique of the input bias current cancellation.

The LT1801 and LT1802 have the characteristics that are essential in precision systems: common mode rejection of 105dB, power supply rejection of 97dB and an open loop gain of 85V/mV combine to maintain precision performance over a wide common mode input voltage, independent of power supply fluctuation, with minimum gain error.

These amplifiers can operate from supplies as low as 2.3V over industrial temperature ranges; have an input voltage range that includes both power supply rails; and have an output that swings within 20mV of either supply rail to maximize the signal dynamic range in low voltage applications. The rail-to-rail input and output characteristics of the amplifiers can simplify designs by eliminating a negative supply. The LT1801 and LT1802 also possess an 80MHz gain bandwidth product, a 25V/µs slew rate and a 50mA output current that make them suitable for high frequency signal conditioning. In servo loop applications, where avoiding phase reversal is critical, the inputs of these amplifiers can be driven beyond supplies without phase reversal of the output.

The LT1801 is housed in an SO-8 package; the LT1802 in an SO-14—both with standard op amp pin outs.

Tiny TSOT-23 Buck Regulators Are Optimized to Work with Ceramic Output Capacitors for Very Low Output Ripple

The LTC3405A, LTC3405A-1.5 and LTC3405A-1.8 are high efficiency monolithic synchronous buck regulators specifically designed to work with ceramic input and output capacitors. Unlike the LTC3405, the internal loop compensation of these new devices does not rely on the output capacitor ESR for stable operation. Ceramic output capacitors can be used freely for very low output ripple and small circuit size. Housed in tiny 6-lead TSOT-23 packages, the LTC3405A, LTC3405A-1.5 and LTC3405A-1.8 can supply 300mA of output current. Their high switching frequency (1.5MHz) allows the use of very small inductors and capacitors. The internal synchronous switch increases efficiency and eliminates the need for an external Schottky diode. The LTC3405A provides adjustable output voltage, while the LTC3405A-1.5 and LTC3405A-1.8 are fixed at outputs of 1.5V and 1.8V respectively. The fixed output voltage versions eliminate the need for the output voltage setting resistors, further reducing the number of external components and saving space. A complete switching regulator solution can occupy less than 0.06in^2 of board space and require only three external components: an input capacitor, an output capacitor and an inductor.

The LTC3405A, LTC3405A-1.5 and LTC3405A-1.8 all use a constant frequency, current mode architecture to provide excellent transient response and line regulation. The supply voltage ranges from 2.5V to 5.5V making them ideally suited for single Li-Ion battery-powered applications. The supply current during operation is only 20µA while maintaining the output voltage with no load (using Burst Mode operation) and < 1µA in shutdown. This enables the regulators to maintain better than 90% efficiency over three decades of output load current. For noise-sensitive applications, Burst Mode operation can be disabled by connecting a MODE pin to V_{IN} or driving it with a logic high signal. This enables constant-frequency operation, which is maintained at lower load currents together with lower output ripple. If the load current is low enough, cycle skipping occurs to maintain regulation. In constant frequency mode, the efficiency is lower than Burst Mode operation at light loads, but it is comparable to Burst Mode operation when the output load current exceeds 25mA.

All three devices can deliver 300mA in a tiny low profile 1mm height TSOT-23 package.

Rugged CAN Transceiver Survives Loss of Ground and Shorts to ±60V

The LT1796 is a rugged transceiver for Controller Area Network (CAN) bus applications. The LT1796 can withstand ±15kV ESD strikes and faults up to ±60V. This makes it ideal for harsh environments, such as industrial controls with 24V supplies, or heavy-duty truck applications. In these applications, the loss of ground connection or cross-wiring faults can force DC voltages in excess of 24V in either polarity onto the bus pins. The LT1796 can survive such faults without the need for external protection circuitry.

The LT1796 matches the industry standard footprint in the SO-8 package, including a combined slew rate control/standby pin. In standby, the supply current is reduced to 800µA. The slew rate control allows a maximum data rate of 500kbps, or it can be programmed for slower rates to minimize EMI and reduce reflections due to long stubs or improper termination.

LTC4211 Hot Swap Controller Provides Overcurrent Protection and Inrush Current Limiting Without an External Gate Capacitor

The LTC4211 Hot Swap™ controller features dual level overcurrent protection and inrush current limiting without the need for an external gate capacitor. The dual level overcurrent protection is implemented using two comparators, each with different thresholds and response times, to monitor the voltage across an external R_{SENSE} resistor connected between the V_{CC} and SENSE pins. The slow comparator helps to filter out noise by tripping when the voltage across the resistor exceeds 50mV for more than 20µs. The fast comparator protects the board against potentially damaging high energy voltage spikes and severe overloads—it trips when the voltage exceeds 150mV for more than 300ns. During power-up, an internal soft-start circuit gradually ramps up the GATE pin of an external N-channel MOSFET to limit inrush current to $50mV/R_{SENSE}$. The soft-start circuit does not require an external gate capacitor, which means faster turn-off times when an overload occurs, and significant cost and space savings.

The operation of the LTC4211 is sequenced in two timing cycles, with the duration of both cycles set by an external capacitor connected from the TIMER pin to ground. The first timing cycle—the plug-in cycle—provides enough time for a solid connection to be made to the backplane and for the power supply to settle. The second timing cycle determines the duration of the soft-start cycle. The ON pin of the LTC4211 must be taken high for the first timing cycle to be activated. As the GATE voltage ramps up during the second timing cycle, the FB pin monitors the load side voltage via an external resistor divider and forces the $\overline{\text{RESET}}$ open drain pin low until the voltage rises above some nominal value.

The LTC4211 operates from 2.5V to 16.5V and is available in the SO-8, MSOP-8 and MSOP-10 packages. It is specified for both commercial and industrial temperature ranges. The 8-pin versions are pin compatible with the LTC1422, allowing for easy upgrades. The 10-pin versions add FILTER and $\overline{\text{FAULT}}$ pins. The 20µs response time of the slow comparator can be increased by connecting an external capacitor from the FILTER pin to ground. The FILTER pin can also used in conjunction with an external Zener diode to add overvoltage protection. The $\overline{\text{FAULT}}$ pin is an open drain output which is normally pulled high by an external pull-up resistor. It serves as a status indication and pulls low when the LTC4211 is latched off by an overcurrent fault. It can also be tied to the ON pin for auto-retry applications in which the LTC4211 tries to reconnect the load automatically after being latched off.

LT1765, continued from page 31

enhanced TSSOP16 exposed leadframe package provide high-power in a more compact solution than is possible with either dual controllers—at a much higher cost—or a single controller and separately chosen MOSFET—a more complex design using extra board space and design and assembly time. The B220A Schottky diodes have a low forward voltage rating for high efficiency and a small case size to further minimize board space. The ceramic input and output capacitors provide a tiny, low-cost solution with minimal output ripple.

The current-mode topology of the regulator provides stable response to load transients on both outputs—requiring only ceramic output capacitors and a simple RC network located on the V_C pin of the LT1765EFE. This is a space and cost saving advantage over a voltage-mode controller topology, which would require additional compensation components to optimize load transient response. Also, voltage-mode controllers typically require electrolytic or tantalum output capacitors, rather than extremely low ESR ceramic capacitors, to stabilize the control loop and maintain good high frequency response. Given the same RMS current-handling requirement, electrolytic and tantalum capacitors take much more space and create much more output voltage ripple than the equivalent ceramic. Overall, a current-mode step-down regulator with ceramic capacitors is simpler, smaller, and less expensive than a voltage mode solution.

The switch current of the LT1765EFE, which has a minimum rating of 3A, limits the maximum output current of the negative line and positive line. In this topology, the negative output current must be less than (and cannot equal) the positive output current, or the output voltage will drop out, so care must be taken when considering all possible load-transient conditions. The typical maximum negative output current with respect to the positive output current is shown in Figure 4. If cross-regulation is an issue with +5V output current greater than 1.0A and –5V negative output current less than 5mA, a 1k preload resistor on the –5V output can improve regulation.

DESIGN TOOLS

Databooks and Applications Handbooks

1990 Linear Databook, Vol I—This 1440 page collection of data sheets covers op amps, voltage regulators, references, comparators, filters, PWMs, data conversion and interface products (bipolar and CMOS), in both commercial and military grades. The catalog features well over 300 devices. $10.00

1992 Linear Databook, Vol II — This 1248 page supplement to the 1990 Linear Databook includes all products introduced in 1991 and 1992. $10.00

1994 Linear Databook, Vol III—This 1826 page supplement to the 1990 and 1992 Linear Databooks includes all products introduced since 1992. $10.00

1995 Linear Databook, Vol IV—This 1152 page supplement to the 1990, 1992 and 1994 Linear Databooks includes all products introduced since 1994. $10.00

1996 Linear Databook, Vol V—This 1152 page supplement to the 1990, 1992, 1994 and 1995 Linear Databooks includes all products introduced since 1995. $10.00

1997 Linear Databook, Vol VI—This 1360 page supplement to the 1990, 1992, 1994, 1995 and 1996 Linear Databooks includes all products introduced since 1996. $10.00

1999 Linear Databook, Vol VII— This 1968 page supplement to the 1990, 1992, 1994, 1995, 1996 and 1997 Linear Databooks includes all products introduced since 1997. $10.00

1990 Linear Applications Handbook, Volume I— 928 pages full 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. $20.00

1993 Linear Applications Handbook, Volume II— Continues the stream of "real world" linear circuitry initiated by the 1990 Handbook. Similar in scope to the 1990 edition, the new 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. $20.00

1997 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 LTC data sheets, design notes, application notes and *Linear Technology* magazines. $20.00

Brochures and Software

Power Management Solutions Brochure — This 96 page collection of circuits contains real-life solutions for common power supply design problems. There are over 70 circuits, including descriptions, graphs and performance specifications. Topics covered include battery chargers, desktop PC power supplies, notebook PC power supplies, portable electronics power supplies, distributed power supplies, telecommunications and isolated power supplies, off-line power supplies and power management circuits. Selection guides are provided for each section and a variety of helpful design tools are also listed for quick reference. Available at no charge

Data Conversion Solutions Brochure — This 88 page collection of data conversion circuits, products and selection guides serves as excellent reference for the data acquisition system designer. Over 40 products are showcased, solving problems in low power, small size and high performance data conversion applications— with performance graphs and specifications. Topics covered include delta-sigma ADCs, low power and high speed ADCs and low power and high speed DACs. A complete glossary defines data conversion specifications; a list of selected application and design notes is also included. Available at no charge

Telecommunications Solutions Brochure —This 76 page collection of application circuits and selection guides covers a wide variety of products targeted for telecommunications. Circuits solve real life problems for central office switching, cellular phones, high speed modems, basestation, plus special sections covering –48V and Hot Swap™ applications. Many applications highlight new products such as Hot Swap controllers, power products, high speed amplifiers, A/D converters, interface transceivers and filters. Includes a telecommunications glossary, serial interface standards, protocol information and a complete list of key application notes and design notes. Available at no charge

SwitcherCAD™ III — 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. Filter CAD 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 Disk — This IBM-PC (or compatible) high density diskette contains the library of LTC op amp SPICE macromodels. The models can be used with any version of SPICE for general analog circuit simulations. The diskette also contains working circuit examples using the models and a demonstration copy of PSPICE™ by MicroSim. Available at no charge

Noise Disk — This IBM-PC (or compatible) 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. Available at no charge

www.linear.com and the Linear Direct Online Store

LTC Web Site — Customers can quickly and conveniently find and retrieve the latest technical information covering the company's products on LTC's web site. 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 web site 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 general corporate information.

Linear Direct Online Store — As of May 1, 2002 the Linear Direct Online Store will be temporarily under reconstruction for approximately four to six weeks. To purchase LTC products during this time, please contact your local sales office or distributor.

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