# *Linear Technology* Magazine Circuit Collection, Volume IV
## Power Products

Richard Markell, Editor

## INTRODUCTION

Application Note 84 is the fourth in a series that excerpts useful circuits from *Linear Technology* magazine to preserve them for posterity. This application note highlights "power" circuits from issue VI:1 (February 1996) through issue VIII:4 (November 1998). Another application note will feature data conversion, interface and signal processing circuits from the same era. Like its predecessor, AN 66, this Application Note includes circuits that can power most any system you can imagine, from "server" power supplies that generate in excess of 50 amps to micropower systems for portable and handheld equipment. Also included are power converters that can be voltage programmed using Intel's VID code. Charge pump converters, linear regulators and battery charger circuits are included here, with Li-Ion batteries receiving extra attention. There are, of course, circuits that cannot be so simply categorized. Come browse. I'll get out of the way and let the authors describe their creations.

Note: Article Titles appear in this application note exactly as they originally appeared in *Linear Technology* magazine. This may result in some inconsistency in the usage of terminology.

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Regulators—Switching (Buck)

## NEW LTC1435–LTC1439 DC/DC CONTROLLERS FEATURE VALUE AND PERFORMANCE
*by Randy Flatness, Steve Hobrecht and Milton Wilcox*

### Introduction

The new LTC1435–LTC1439 multiple-output DC/DC controllers bring unprecedented levels of value to supplies for notebook computers and other battery-powered equipment, while eliminating previous performance barriers.

For example, a new Adaptive Power™ output stage allows two previously incompatible parameters, constant frequency operation and good low current efficiency, to coexist in the same power supply. A second breakthrough allows N-channel power MOSFETs to be used exclusively, while maintaining low dropout operation previously available only with P-channel MOSFETs. Other innovations include an auxiliary linear regulator loop, a phase-locked loop (PLL) to synchronize the oscillator to an external source, a self-contained power-on-reset (POR) timer and programmable run delays useful for staging output voltages.
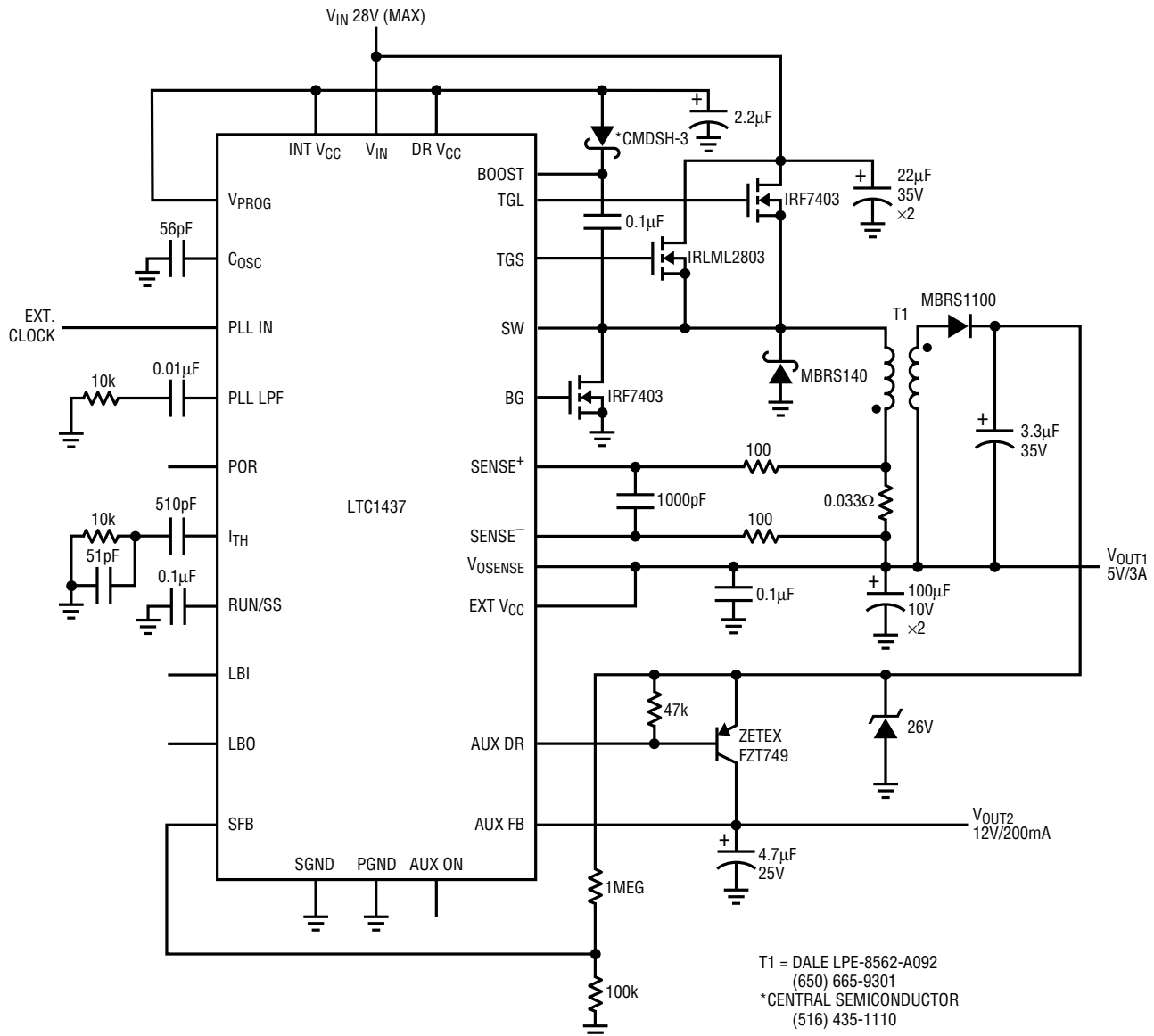


**Figure 1. High Efficiency, Constant Frequency, Dual-Output Supply Delivers 3A at 5V and 250mA at 12V**

## Cost Effective LTC1437 Switcher/Linear Combination with 5V/3A and 12V/200mA Outputs

The main switcher loop, shown in the schematic in Figure 1, is set to 5V by strapping the $V_{PROG}$ pin high. Other output options include 3.3V ($V_{PROG}$ low) and adjustable ($V_{PROG}$ open).

The 12V output in Figure 1's circuit is provided by the auxiliary linear regulator operating in conjunction with a secondary winding feedback loop using the SFB pin. The turns ratio for the transformer is 1:2.2, resulting in a

secondary output voltage of approximately 15V. The secondary resistive divider causes the SFB pin voltage to drop below the internal 1.19V reference if the secondary output is loaded and the 5V output has little or no load. This forces continuous operation as necessary to guarantee sufficient headroom for the linear regulator to maintain 12V regulation independent of the 5V load. The auxiliary output is turned on and off with the AUX ON pin.

The auxiliary regulator can also be used in an adjustable mode, determined by the voltage on the AUX DR pin. When the AUX DR voltage is higher than 9.5V, as is the
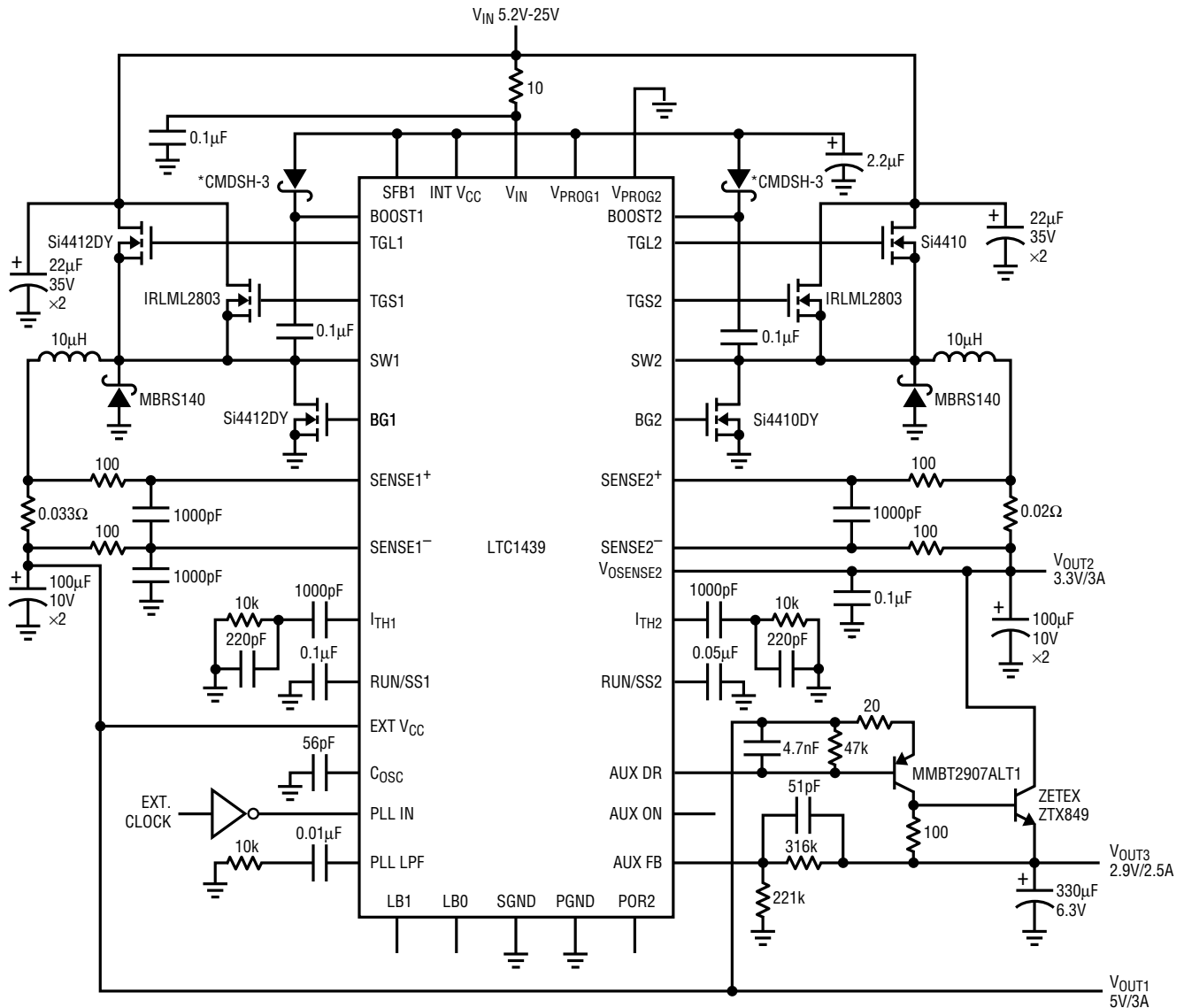


*CENTRAL SEMICONDUCTOR (516) 435-1110

**Figure 2. High Efficiency, Constant-Frequency, Triple-Output Supply Features 200mV Dropout**

case in Figure 1, the regulator automatically configures itself for fixed 12V operation using an internal AUX FB resistive divider. When AUX DR is less than 8.5V, the internal divider is removed and the user can adjust the output voltage via an external divider referenced to 1.19V. The external auxiliary regulator PNP pass transistor is sized for the desired output current; in this case a SOT-223 device is used to deliver up to 200mA.

**Synchronizable, Triple-Output, Low Dropout Supply**

The LTC1439-based supply shown in Figure 2 is an example of how three logic supply voltages, 5V, 3.3V and 2.9V, can be easily derived using only two simple inductors. The two main DC/DC controller loops are used to supply 5V/3A and 3.3V/5.5A. Up to 2.5A of the 3.3V output current is then used to supply a 2.9V output using the adjustable capability of the auxiliary linear regulator.

The 2.9V output also illustrates the use of an external NPN pass transistor with the auxiliary regulator. Because only 0.4V is dropped across the NPN transistor, 2.9V efficiency remains in the 85% range. And thanks to the 99% duty cycle capability of the switcher loops, Figure 2's supply can maintain all three output voltages in regulation down to $V_{IN}$ = 5.2V with a 2A load on the 5V output.

The phase-locked loops built into the LTC1437/LTC1436-PLL and LTC1439 offer a convenient means of synchronization for the applications in Figures 1 and 2. The internal oscillator is actually a voltage-controlled oscillator (VCO) controlled by the voltage on the PLL LPF pin. When no

PLL IN signal is present, the PLL LPF pin goes low, causing the oscillator to run at its minimum frequency ($f_{MIN}$ = 180kHz with $C_{OSC}$ = 56pF). Applying a 3.3V or 5V logic signal of any duty cycle to the PLL IN pin will cause the oscillator frequency to lock to the external frequency and to track it up to a maximum of $f_{MAX} = 2 \cdot f_{MIN}$. A logic signal may also be coupled to PLL LPF to effect a 2:1 frequency shift, provided that the initial frequency has been set to less than 200kHz.

Figure 3 is a photograph showing the 3.3V output staged to start 10ms before the 5V output when power is first applied to Figure 2's circuit.

An internal regulation monitor is continually monitoring the main controller output in the LTC1436/LTC1437, and the controller 2 output (3.3V in Figure 2) in the LTC1438/LTC1439. When out of regulation or in shutdown mode, the POR open drain output pulls low. At start-up, once the output voltage has reached 5% of its final value, an internal timer is started, after which the POR pin is released. The timer is accomplished by counting $2^{16}$ oscillator cycles, yielding a delay-to-release reset of approximately 300ms in a typical application.

The EXT $V_{CC}$ pin is normally connected to the 5V output to allow INT $V_{CC}$ power to be derived from the regulator itself. Quiescent current is then reduced because driver and control currents are scaled by a factor approximately equal to the 5V controller duty cycle. EXT $V_{CC}$ can also be connected to other external high efficiency sources, up to a maximum of 10V.
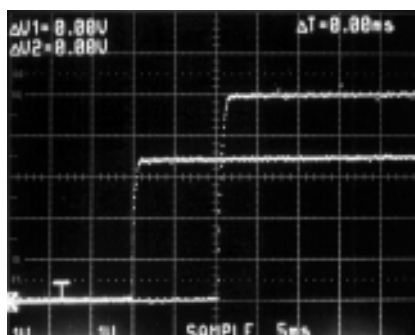


**Figure 3. Start-Up of 3.3V and 5V Supplies is Easily Staged Upon Initial Application of Input Power**

## THE LTC1266 OPERATES FROM ≥12V AND PROVIDES 3.3V OUT AT 12A
*by Craig Varga*

### Circuit Description and Operation

The design in Figure 4 relies on a floating high-side driver that provides enough gate-drive capability to easily switch a large power MOSFET. The LTC1266 is configured to drive a P-channel MOSFET by tying pin 3 (PINV) to ground. This is required because there will be a net inversion by the floating driver. Q4 controls the driver stage and provides gate-discharge capability through D3. When the low-side switches are on, C16 charges to 12V through D1. When the LTC1266 signals Q1 to turn on, Q4 is turned off. R11 provides base current for Q6, which, in conjunction with Q5, acts like an SCR. Once fired, the regenerative behavior of Q5 and Q6 rapidly charges the gate of Q1. Since C16 is referenced to the source of Q1, the top of C16 rises above the 12V supply rail as Q1 turns on, forcing the gate of Q1

to nearly 24V above ground. When the LTC1266 takes pin 1 high, Q4 turns on, pulling charge from the gate capacitance of Q1 through D3. This back biases the base-emitter junction of Q6, forcing the pull-up circuit, and therefore Q1, off.

Since the input voltage is high relative to the output, the nominal duty factor of the high-side switch is small (in this case approximately 31%). As a result, the RMS current through Q1 is relatively low. By contrast, the low-side switches are on nearly 70% of the time, and therefore see a much higher RMS current. This explains why the low-side switch employs two MOSFETs, whereas the high-side switch uses only one. Schottky diode D2 is used to help keep the body diodes of Q2 and Q3 from turning on during the short dead time before switching transitions. These body diodes exhibit relatively long reverse recovery times, contributing to commutation losses. The Schottky diode improves overall efficiency several percent, but the circuit will function correctly without it. Switching losses in the
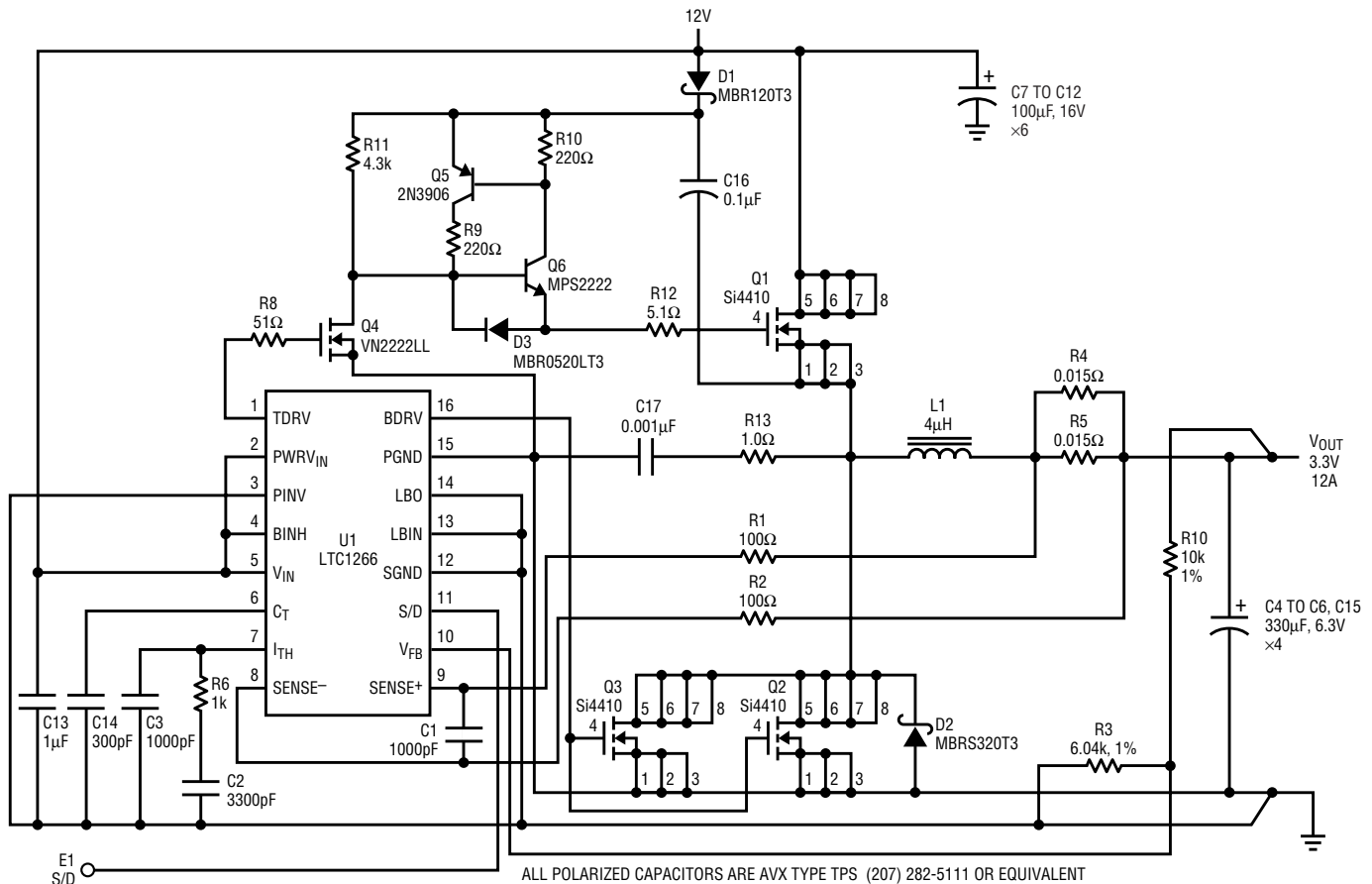


**Figure 4. 12V In, 3.3V/12A Out Supply**

two low-side switches are nearly zero, since these devices are turned on and off into nearly zero volts (the forward drop of the Schottky).

There is no fundamental limitation on how high the maximum input voltage can be with this approach. The drive level shift is limited by the breakdown rating of Q4. Obviously, the power transistors and input capacitors must be rated for the intended input voltage. A low power 12V supply is needed to provide power for the LTC1266 and voltage for the bootstrap supply.

Figure 5 shows a 24V input design. As the input supply voltage is increased, one thing to watch for is the potential for overlap in the high- and low-side turn-on/turn-off transitions. The LTC1266 is designed to prevent shoot-through by actually waiting until the gate voltage of one switch is low before allowing the other switch to be turned on. Using the floating driver defeats this capability, so this condition must be checked for. The high-side drive turn-on time may be reduced by lowering the value of R11. Using a larger device for Q4 will speed up the turn-off transition. The value of C16 may also need to be a bit larger if R11 is reduced to limit drooping of the bootstrap supply voltage.
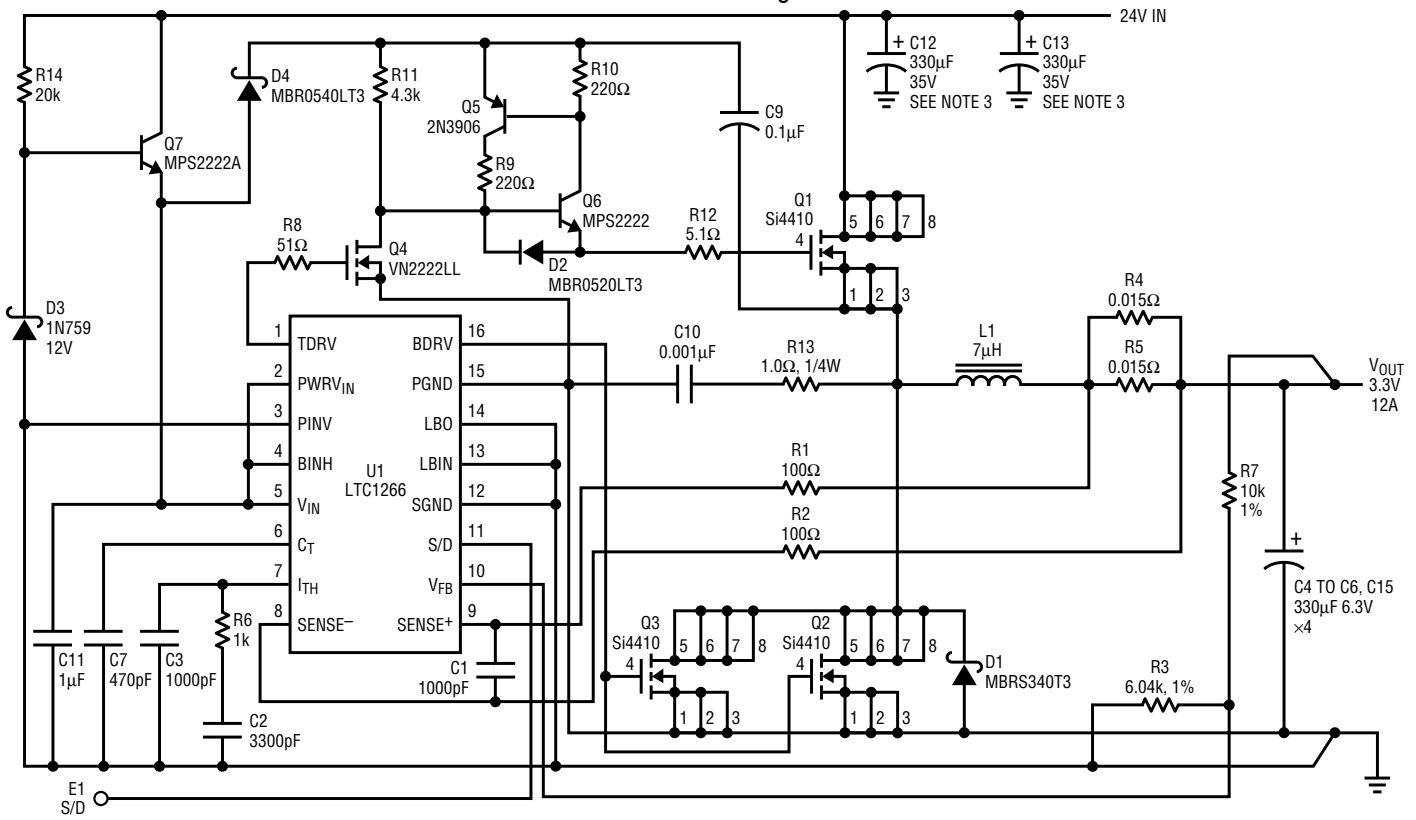


1. ALL POLARIZED CAPACITORS ARE AVX TYPE TPS OR EQUIVALENT UNLESS NOTED OTHERWISE. (207) 282-5111

2. L1 CONSISTS OF 15 TURNS OF #16 AWG ON MAGNETICS, INC. 77848-A7 Kool Mμ CORE (800) 245-3984

3. C12 AND C13 ARE PANASONIC TYPE HF OR EQUIVALENT (201) 348-7522

**Figure 5. 24V In, 3.3V/12A Out Supply**

## THE NEW LTC1435 MAKES A GREAT MICROPROCESSOR CORE VOLTAGE REGULATOR
*by John Seago*

Current microprocessor architectures require different voltages for the core and the I/O ring. For portable computer applications, the microprocessor core voltage is reduced for lower power consumption. Three high current regulated voltages, 5V, 3.3V and 2.9V, are commonly required. Several IC manufacturers offer two-output controllers, like the LTC1438, which are normally used for 5V and 3.3V. Another controller is required to generate the 2.9V. Figure 6 shows a simple circuit using the LTC1435 to provide 2.9V at 2.65 amps for the Intel portable Pentium® processor.

The circuit's 165kHz switching frequency was selected as a compromise between transient response and circuit efficiency. This frequency is determined by the value of C1. Output voltage transient response is shown in Figure 7. The transient response can be adjusted for other applications by changing the values of compensation components R1, C3 and C14. Efficiency curves for different input voltages and load currents up to 3.2 amps are shown in Figure 8.

Another feature of the LTC1435 is the option to maintain constant switching frequency under all load conditions or to select Burst Mode™ operation for the highest efficiency at light loads. Pulling the SFB pin high enables Burst Mode when load current drops to a low value. However, Burst Mode can degrade transient response at low input voltages and should not be used for pulsed load applications where good transient response at low input voltage is required.
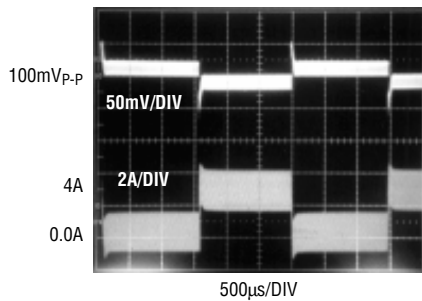
**Figure 7. Output Voltage vs Transient Response**
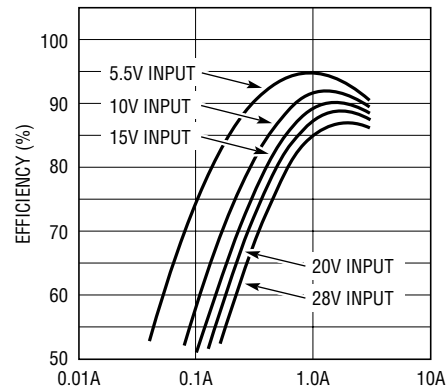
**Figure 8. LTC1435 Efficiency Curves for Different Input Voltages**
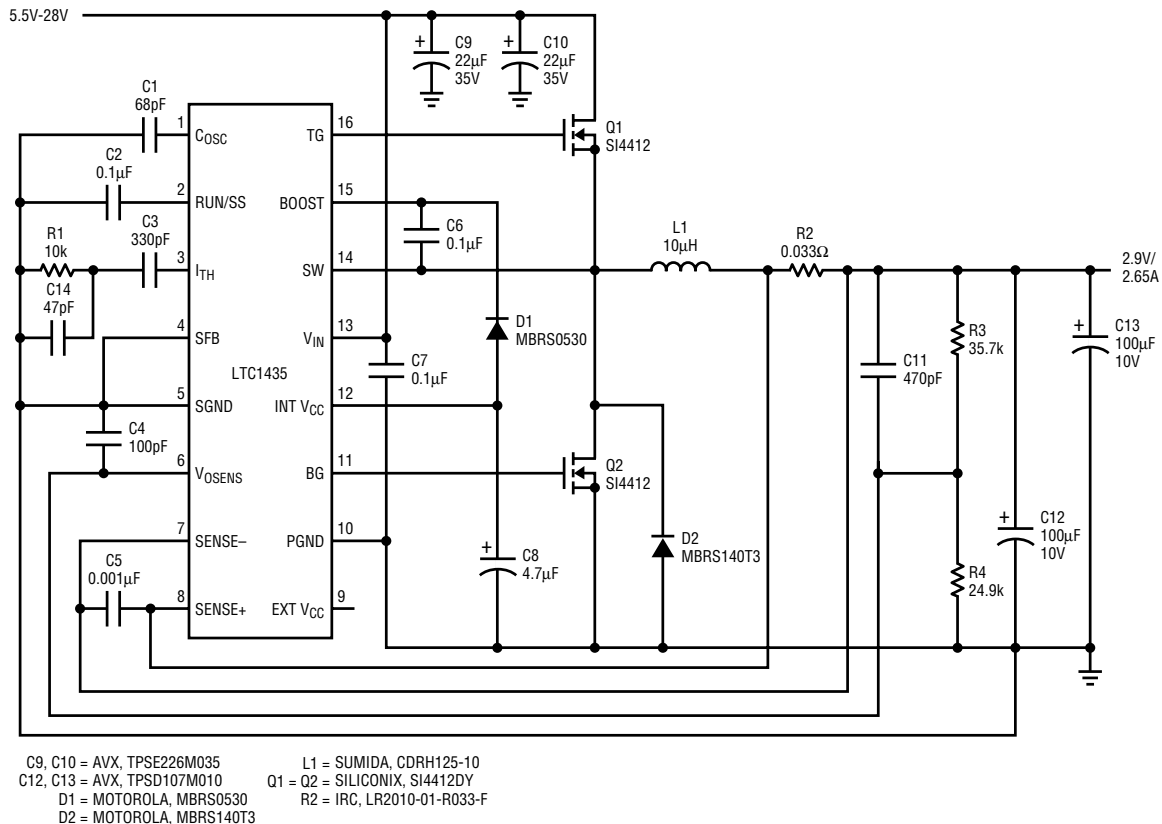
**Figure 6. 2.9V Regulator for Portable Pentium Processor**
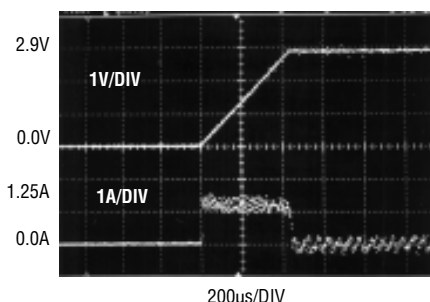
Figure 9. Inductor Input Voltage and Current Waveforms



Figure 10. Soft-Start Output Voltage and Inductor Current

The SFB pin in the circuit of Figure 6 is grounded, which will defeat the Burst Mode and ensure constant frequency operation.

It is sometimes necessary to shut down power to the load. RUN/SS is a dual-function pin on the LTC1435 that provides both output voltage on/off control and output current soft-start capability. When RUN/SS (pin 2) is pulled low by an open collector or open drain device, the output voltage is turned off and the controller shuts down. The soft-start feature takes over when the low is removed from pin 2. Figure 9 shows the output voltage under no-load conditions at turn-on, with the soft-start capacitor C2 equal to 0.1μF. This simulates the start up conditions of a microprocessor held in standby until after the input voltage has stabilized. If the regulator is started under full-load conditions, the output current ramp time will be approximately 0.5s/μF of soft-start capacitance. The output volt-

age during this soft-start period depends on the load impedance. If soft-start is not required, capacitor C2 is not used and the current limit setting of the regulator determines the maximum load current during start-up.

In order to properly enhance the top MOSFET (Q1), INT $V_{CC}$ is level shifted by charge pumping capacitor C6 to INT $V_{CC}$ minus one diode drop. C6 provides the power to turn Q1 on and off. The INT $V_{CC}$ of the LTC1435 is regulated to 5V, but will increase with higher voltage applied to EXT $V_{CC}$, up to a maximum of 10V. For outputs between 5V and 10V, the output should be connected to EXT $V_{CC}$. The power loss of the INT $V_{CC}$ linear regulator will be replaced by the more efficient switcher output and the gate-drive voltage of both MOSFETs will be increased for lower "ON" resistance. Figure 10 shows L1 input voltage and current with a 10 volt input, 2.9 volt output, and 2.65 amp load current.

## LTC1433/LTC1434: HIGH EFFICIENCY, CONSTANT-FREQUENCY MONOLITHIC BUCK CONVERTER
*by San-Hwa Chee*

### Typical Application: Buck Converter Supplies 3.3V at 600mA

Figure 11 shows a practical LTC1433 circuit that can be used for cellular telephone applications. Efficiency curves for this circuit at various input voltages are shown in Figure 12. Note that the efficiency reaches 93% at a supply voltage of 5V and a load current of about 150mA. This high efficiency makes the LTC1433 and LTC1434 attractive for power-sensitive applications. The circuit works all the way down to 3.6V at a load current of 250mA before dropping out and the oscillator frequency is a constant 210kHz down to 20mA load current.

### Typical Application: Positive-to-Negative Converter

Both the LTC1433 and LTC1434 can easily be set up for a negative output voltage. Figure 13 shows the schematic using the LTC1433. The efficiency curve is shown in Figure 14. This circuit is set up so that the output is taken from the device ground. Components that are normally referenced back to the device ground, such as the Run/SS capacitor, oscillator frequency capacitor and the $I_{TH}$ compensation network, are connected to the output instead of to the circuit ground.

* MBRS130LT3
** COILCRAFT DO3316-104
† AVX TPSD107M010R0100
†† AVX TPSE686M020R0150
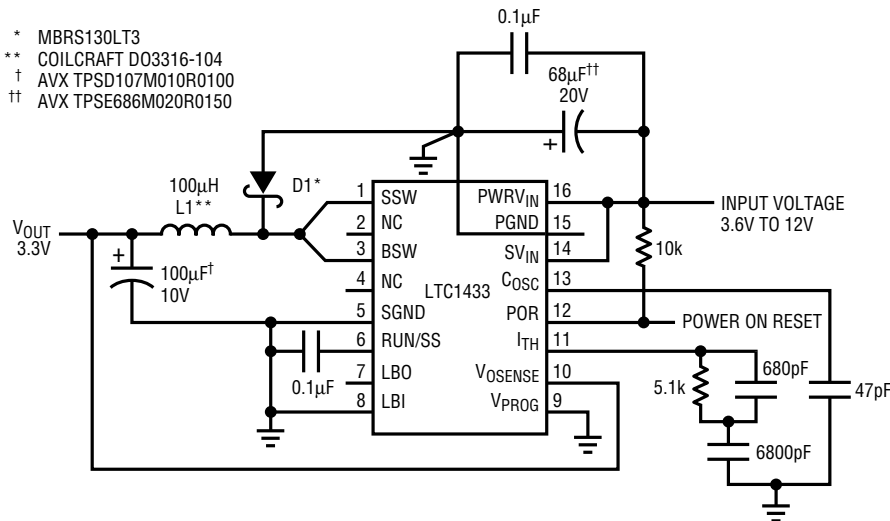
**Figure 11. LTC1433 Typical Application: 3.3V Output at 600mA**
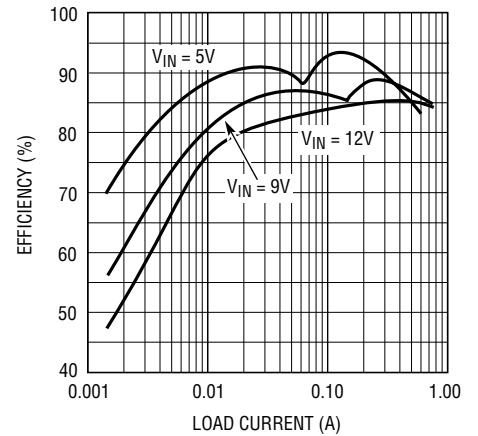
**Figure 12. Efficiency vs Load Current for Figure 11's Circuit**

* MOTOROLA MBRS130LT3
** COILCRAFT DO3316 SERIES
† AVX TPSD107M010R0100
†† AVX TPSE107M016R0100

| $V_{IN}$ (V) | $I_{OUT\ MAX}$ (mA) |
|--------------|---------------------|
| 3.0          | 180                 |
| 4.0          | 240                 |
| 5.0          | 290                 |
| 6.0          | 340                 |
| 7.0          | 410                 |
| 7.5          | 420                 |
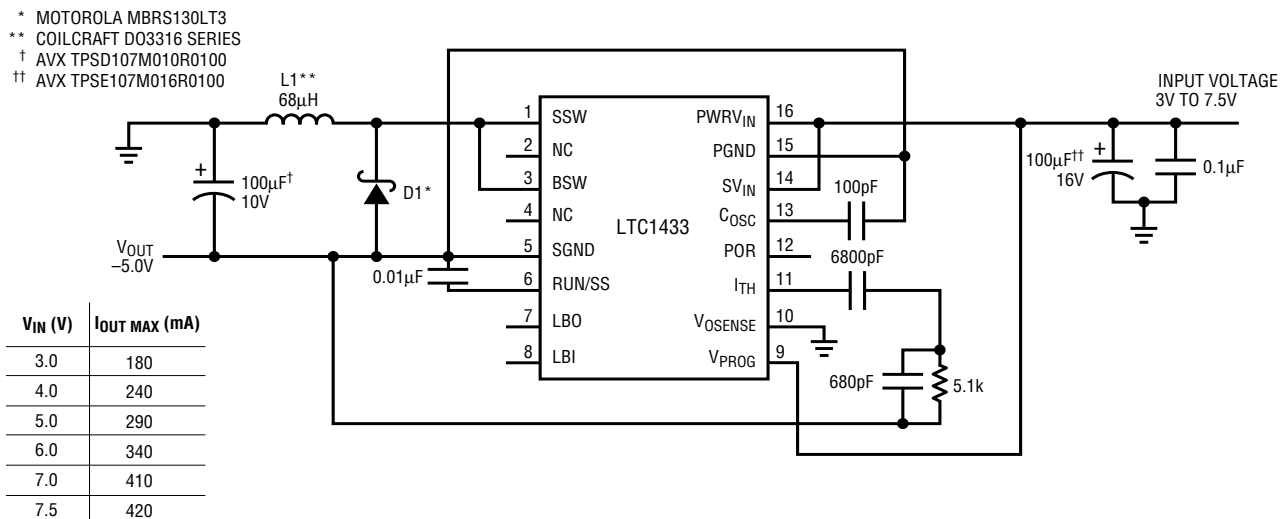
**Figure 13. Positive-to-Negative (–5.0V) Converter**
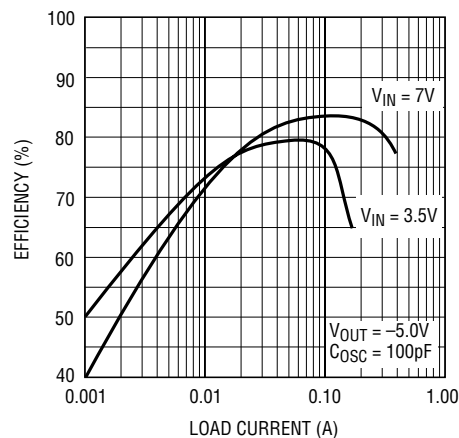
**Figure 14. Efficiency Curves for Figure 13's Positive-to-Negative Converter**

## 24 VOLT TO 14 VOLT CONVERTER PROVIDES 15 AMPS
*by John Seago*

Combining the LTC1435 with a large geometry power MOSFET and good PCB layout allows large currents to be processed easily and efficiently. With the use of a current sense transformer, output voltages greater than 10V can be implemented. The circuit in Figure 15 shows an LTC1435 configured as a conventional buck regulator using a single N-channel MOSFET to control an output voltage greater than 10V with load current exceeding 15 amps. The efficiency of the breadboard measured 94% with a 24V input, 14V output and 15A of load current. If maximum efficiency is required, adding a second power MOSFET for synchronous switching will improve efficiency by about 1%.

This circuit's 100kHz switching frequency was selected to reduce switching losses so that PCB mounted heat sinks could be used without requiring additional air flow. The switching frequency can be set from 50kHz to 400kHz by selecting an appropriate value for C1. The current sense transformer T1 uses a 1:100 turns ratio to scale down the buck inductor input current and develop the voltage across R9, used by the ±SENSE inputs for regulation. Short-circuit protection is provided by Q4 and Q5. When the current transformer secondary voltage developed across
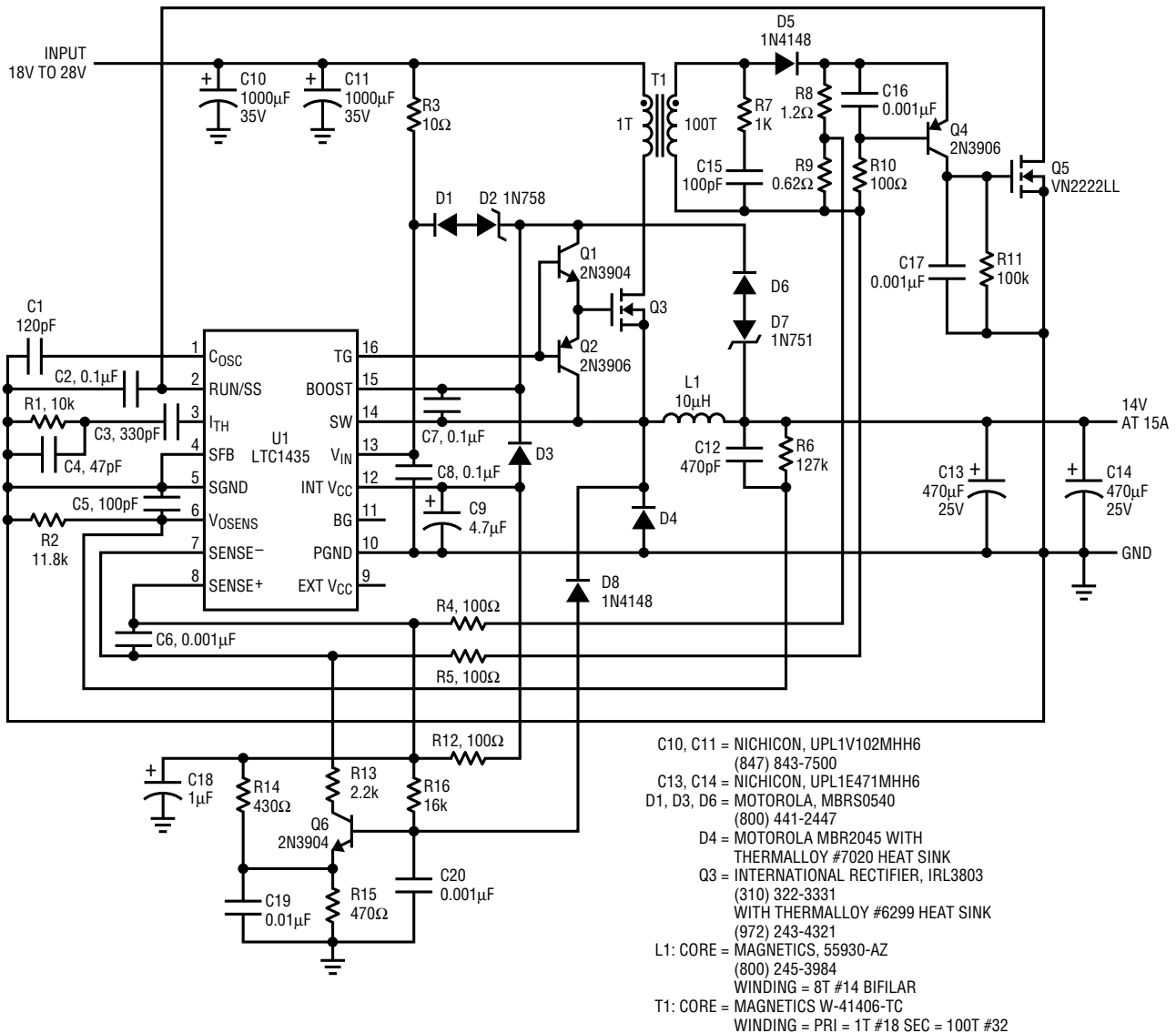


**Figure 15. 14V, 15A Buck Regulator**

A = Q3 SWITCH VOLTAGE
    20V/DIV

B = L1 CURRENT
    10A/DIV

0.0V

0.0A

C = T1 PRIMARY CURRENT
    10A/DIV

0.0A

D = OUTPUT VOLTAGE RIPPLE
    0.2V/DIV          14VDC

2μS/DIV
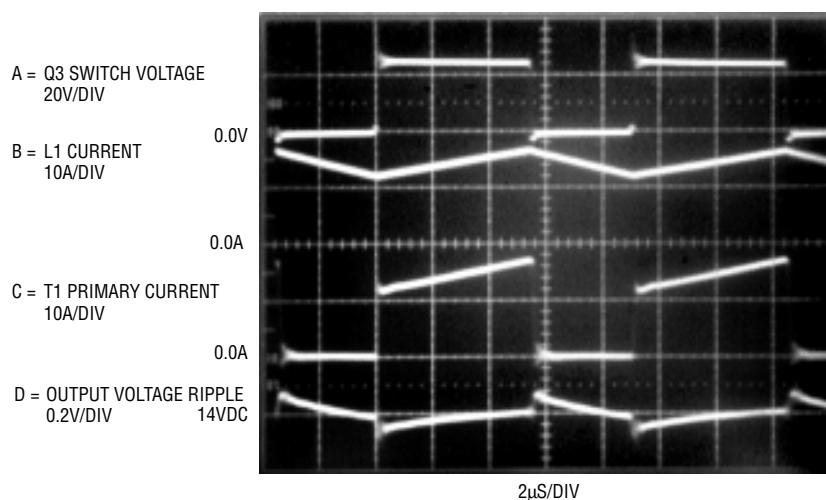
**Figure 16. Buck Regulator Circuit Waveforms**

R8 and R9 is enough to turn on Q4, Q5 temporarily pulls the RUN/SS pin low, turning off the regulator. Output current soft-starts when Q5 releases the RUN/SS pin. This results in frequent attempts to establish output voltage if a short exists, without high current continuously flowing through the power elements. The power elements consist of input capacitors C10 and C11, Current sense transformer T1, buck inductor L1, power MOSFET Q3, commutating diode D4 and output capacitors C13 and C14.

Although the wide 3.6V–36V input voltage range and 99% duty cycle operation of the LTC1435 are ideal for battery/wall adapter input applications, operating above 95% duty cycle causes problems for the current sense transformer. To avoid transformer saturation, the Q6 stage limits duty cycle to approximately 90%. Current through R16 tries to charge C20 to the 3V base voltage of Q6. If the switch cycle terminates at less than a 90% duty cycle, C20 is reset by D8. If the duty cycle exceeds 90%, C20 charges until Q6 turns on, ending the switch cycle.

Switch voltage, inductor current, T1 primary current, and output voltage ripple waveforms are shown in Figure 16. These waveforms were measured with a 24V input, 14V output, and 15A load current. When MOSFET Q3 turns on, the switch voltage (Trace A) goes high, the inductor current (Trace B) increases, as does the T1 primary current (Trace C) and the output ripple voltage (Trace D). When Q3 turns off, the switch voltage goes low, inductor current decreases as its stored energy supplies load current through D4, T1 primary current goes to zero and the output voltage decreases slightly.

## LTC1553 SYNCHRONOUS REGULATOR CONTROLLER POWERS PENTIUM® PRO AND OTHER BIG PROCESSORS
*by Y.L. Teo, S.H. Lim and Craig Varga*

The LTC1553 provides current-limit and short-circuit protection without the use of an external sense resistor. It has excellent (±1%) output regulation over temperature, line voltage and load current variations. To compliment the main voltage-feedback loop, the LTC1553 includes two additional feedback loops to provide good large-signal transient response. The LTC1553 adds additional internal circuits to conform to the Intel Pentium Pro processor power converter requirements while minimizing the number of external components. An on-chip 5-bit digital-to-analog converter (DAC) provides output voltages conforming to Intel's specifications. This allows the LTC1553 to read the code sent by the processor and provide it with the requested voltage. The LTC1553 also provides a power-good indication (PWRGD) to the system. There is also an on-chip overvoltage protection circuit that latches the regulator in an off state if the output voltage ever rises 15% or more above the DAC-requested voltage.

In applications with other processors, the four DAC inputs can be routed to a jumper block, zero ohm resistors or a

DIP switch, or hard wired, to set the desired output voltage. This allows the output voltage to be programmed easily in steps while eliminating the need to stock an assortment of precision resistors. This flexibility in output voltage setting is cheap insurance against last-minute power supply voltage changes by microprocessor manufacturers.

## LTC1553 Overview

The on-chip, 5-bit digital-to-analog converter (DAC) allows the output voltage to be adjusted from 1.80V to 3.5V, as shown in Table 1. Current limiting is maintained by sensing the voltage drop across the $R_{DS(ON)}$ of the high-side MOSFET. The DAC accuracy, initial reference voltage tolerance and internal feedback resistor tolerances result in a maximum initial output voltage error of ±1% of the selected output voltage. The line and load regulation plus temperature drift over the 0°C to 70°C temperature range will contribute another ±1% to the output error budget. This gives a total static operating error of less than ±2%, providing sufficient headroom (3%) for the dynamic response to remain within a ±5% output voltage tolerance, while still requiring a reasonable amount of output capacitance.

## Typical Application

A typical application for LTC1553 is converting 5V to 1.8V–3.5V in a Pentium Pro processor based personal computer. The supply may be in the form of a voltage regulator module (VRM) or may be implemented directly on the motherboard. The output is used to power the Pentium Pro processor and the input is taken from the system's 5V supply. The circuit shown in Figure 17 provides 1.80V–3.5V at 14A while maintaining output regulation within ±1%. The output voltage is determined by connecting the five DAC inputs to the VID pins of the processor. The power MOSFETs are sized to minimize board space and allow operation without the need of a heat sink. With proper airflow, ambient temperature conditions of up to 50° Celsius are acceptable. Typical efficiency is above 90% from 1A to 10A at 3.3V out. (see Figure 18). Achieving higher output currents from LTC1553 based designs is simply a matter of selecting appropriate MOSFETs and passive components.

It pays to look at the regulator design from two perspectives: electrical and thermal. Most processor applications operate at average currents that are approximately 80% or less of the specified peak current. As such, the thermal
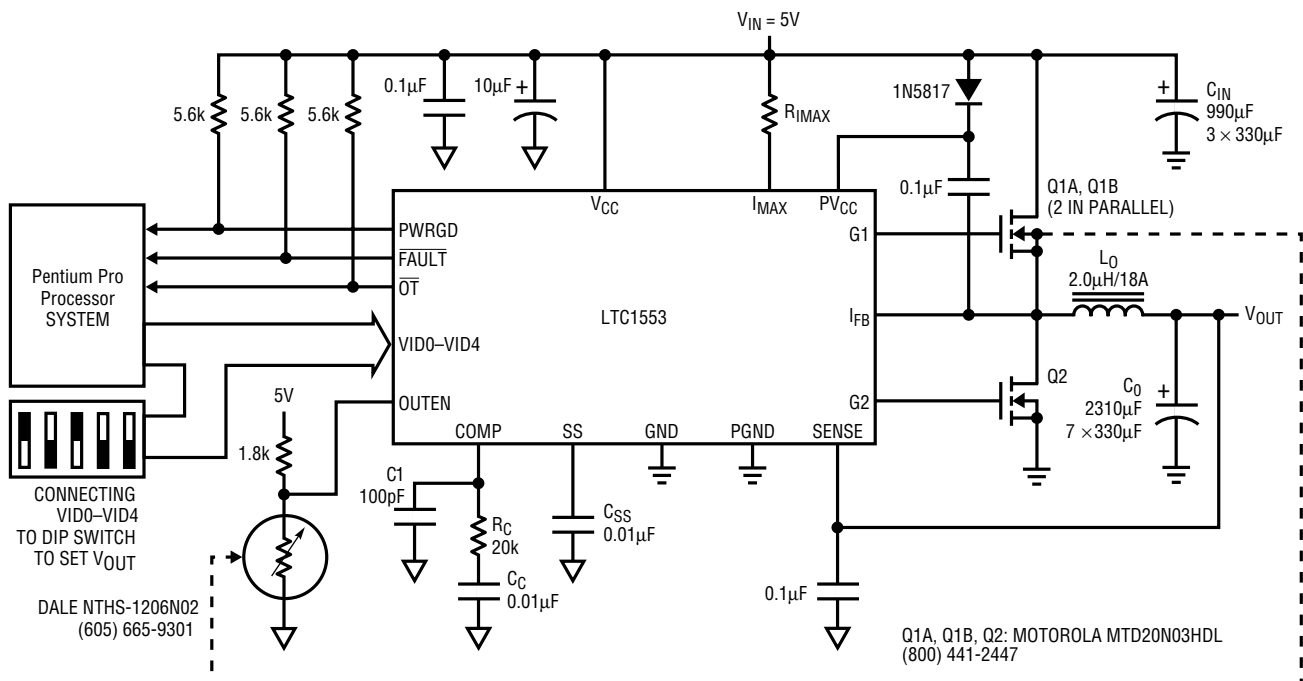


**Figure 17. Typical 5V to 1.8V–3.5V/14A LTC1553 Application**

**Table 1. Output Voltage vs VIDx Code**

| VID4 | VID3 | VID2 | VID1 | VID0 | (VDC) |
|------|------|------|------|------|--------|
| 0 | 1 | 1 | 1 | 1 | * |
| 0 | 1 | 1 | 1 | 0 | * |
| 0 | 1 | 1 | 0 | 1 | * |
| 0 | 1 | 1 | 0 | 0 | * |
| 0 | 1 | 0 | 1 | 1 | * |
| 0 | 1 | 0 | 1 | 0 | * |
| 0 | 1 | 0 | 0 | 1 | * |
| 0 | 1 | 0 | 0 | 0 | * |
| 0 | 0 | 1 | 1 | 1 | * |
| 0 | 0 | 1 | 1 | 0 | * |
| 0 | 0 | 1 | 0 | 1 | 1.80 |
| 0 | 0 | 1 | 0 | 0 | 1.85 |
| 0 | 0 | 0 | 1 | 1 | 1.90 |
| 0 | 0 | 0 | 1 | 0 | 1.95 |
| 0 | 0 | 0 | 0 | 1 | 2.00 |
| 0 | 0 | 0 | 0 | 0 | 2.05 |
| 1 | 1 | 1 | 1 | 1 | No CPU |
| 1 | 1 | 1 | 1 | 0 | 2.1 |
| 1 | 1 | 1 | 0 | 1 | 2.2 |
| 1 | 1 | 1 | 0 | 0 | 2.3 |
| 1 | 1 | 0 | 1 | 1 | 2.4 |
| 1 | 1 | 0 | 1 | 0 | 2.5 |
| 1 | 1 | 0 | 0 | 1 | 2.6 |
| 1 | 1 | 0 | 0 | 0 | 2.7 |
| 1 | 0 | 1 | 1 | 1 | 2.8 |
| 1 | 0 | 1 | 1 | 0 | 2.9 |
| 1 | 0 | 1 | 0 | 1 | 3.0 |
| 1 | 0 | 1 | 0 | 0 | 3.1 |
| 1 | 0 | 0 | 1 | 1 | 3.2 |
| 1 | 0 | 0 | 1 | 0 | 3.3 |
| 1 | 0 | 0 | 0 | 1 | 3.4 |
| 1 | 0 | 0 | 0 | 0 | 3.5 |

*Reserved for future expansion



**Figure 18. Efficiency Plot for Figure 17's Circuit**

mal load results in a huge, and most likely unnecessary, design margin. A good understanding of your system requirements can result in substantial savings in the size and cost for the power supply.

$R_{IMAX}$ sets current limit to the desired level. Add one-half of the inductor ripple current to the maximum load current to determine the peak switch current. Multiply this current by the maximum on-resistance of the selected MOSFET switch to determine the minimum current limit threshold voltage. It's a good idea to add at least a 10% margin to this limit. Also, be sure to use the hot on-resistance of the MOSFET. A multiplier of about 1.4 times the room temperature $R_{DS(ON)}$ should be used to determine the hot resistance. In the case of two parallel MTD20N03HDLs (Q1A and Q1B), the cold resistance is approximately 0.035Ω each; therefore, assume the hot resistance to be approximately 0.050Ω. Divide this by two because the FETs are in parallel. The threshold voltage is programmed by multiplying the $I_{MAX}$ pin's sink current by the value of $R_{IMAX}$. Since we now can determine the required threshold, we need to calculate the value of $R_{IMAX}$. Use the specified minimum sink current, 150μA, to calculate the resistor value.

The soft-start time is programmed by the 0.01μF cap connected to the SS pin. The larger the value of this capacitor, the slower the turn-on ramp.

Inductor $L_O$ is sized to handle the full load current, up to the onset of current limit, without saturating. A value of between 2μH and 3μH is adequate for most processor supply designs. Be careful not to overspecify the inductor.

design can be based on the lower current level. Higher currents, while present, are typically not of sufficient duration to significantly heat the power devices. The design does, however, need to be capable of delivering the peak current without entering current limit or resulting in device failures. Keep in mind that the power dissipation in a resistive element, such as a MOSFET, varies as the square of load current. As such, raising the load current from 80% to 100% translates to approximately 56% more power dissipation ($1/0.8^2$). Designing for this higher ther-

The inductor need not retain its no-load inductance up to the current-limit threshold. If the inductor still retains on the order of 25% to 30% of its initial inductance under worst-case short-circuit current conditions, the circuit should prove reliable. However, you do want to ensure that approximately 60% to 75% of the initial inductance is retained at nominal full load. Excessive inductance roll-off will result in higher than expected output ripple voltage at high loads, along with increased dissipation in the power FETs and the inductor itself.

Proper loop compensation is critical for obtaining optimum transient response while ensuring good stability margins. The compensation network shown here gives good response when used with the inductor and the output capacitors values shown in Figure 17. Several low ESR capacitors are placed in parallel to reduce the total output ESR, resulting in lower output ripple and improved transient performance. Generally speaking, low ESR, high value output capacitors should be chosen to optimize the use of board space. However, if the ESR value is too low for a given capacitor value, loop stability problems can occur. The feedback loop depends on the frequency of the ESR "zero" being well below the loop crossover frequency. There is 45° of positive phase shift at the frequency where the capacitive reactance equals the ESR of the capacitor. Without this phase shift, the loop would be impossible to stabilize. Low ESR, AVX TPS-series tantalum capacitors are a very good compromise between ESR, capacitance value and physical size.

Input capacitors are included to suppress the input switching noise and to keep the input 5V supply variation to a minimum during the Q1 ON/OFF cycle. Excessive conducted emissions are usually traced back to inadequate input capacitance or poor layout of the power-path traces. The crucial parameter for the input capacitors is ripple current rating. A reasonable rule of thumb says that the input capacitor ripple current is going to be approximately 50% of the load current. Therefore, in a typical Pentium Pro processor application, the input capacitors should be rated for close to $7A_{RMS}$. An excellent choice for the input capacitors are Sanyo OS-CONs or the equivalent. They have extremely high ripple current ratings for their size and have demonstrated excellent reliability in this type of application. Low ESR aluminium electrolytic capacitors are a viable option from both input and output. Although lower in cost than OS-CONs or tantalum capacitors, their long-term reliability is not as good. Using 105°C capacitors and keeping operating temperatures low will help to obtain reasonable capacitor life.

The combination of the Dale NTHS-1206N02 thermistor and the 1.8k resistor are for overtemperature monitoring. The OT flag trips if the ambient temperature at Q1 reaches about 90°C; at 100°C the G1 and G2 drivers stop operating. If the system monitors the OT flag, there should be ample time to take precautions, saving data and system configuration information prior to an overtemperature shutdown. Alternatively, CPU activity could be reduced, lowering power supply current and allowing the supply to cool down.

The PWRGD pin gives the CPU rail-voltage OK indication. If, for any reason, the output regulation falls out of the ±5% limit (including an overtemperature shutdown), PWRGD will provide a logic low signal to the system monitor.

## SYNCHRONIZING LTC1430s FOR REDUCED RIPPLE
*by Craig Varga*

The recent move to split-plane microprocessors by several CPU makers has led to the inclusion of multiple switching regulators on many motherboard designs. These regulators typically provide 3.3V for system logic and a separate supply for the processor core. Current requirements of 5A–10A or more per supply are not unusual. The LTC1430 synchronous buck regulator is commonly used to provide these tightly regulated supplies. By nature, the input current waveform in the buck topology is discontinuous, resulting in large input ripple current. By synchronizing a pair of supplies out of phase, it is possible to achieve a degree of ripple current cancellation. This results in less stress on the input capacitors (the number of input capacitors could be reduced) and lower EMI. The ripple is easier to filter since the frequency is effectively doubled and the peak-to-peak current is reduced.

It is extremely simple to synchronize a pair of LTC1430s in an appropriate phase relationship. Simply connect a resistor divider from the low gate drive of a "master"
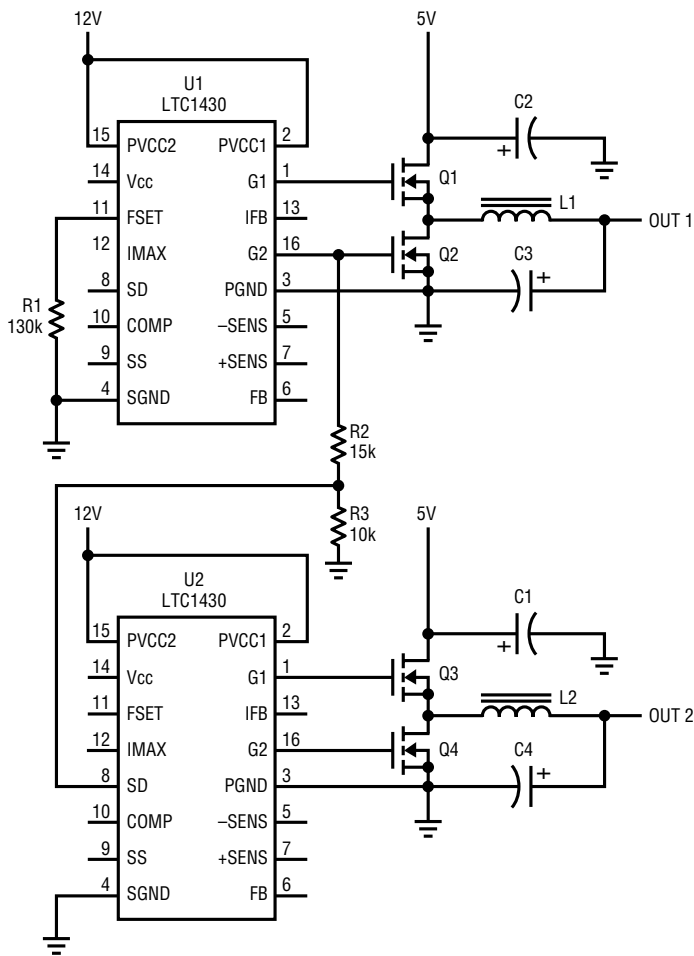
Figure 19. Simplified Schematic Diagram
of Synchronization Circuitry



**Figure 20. Phase Relations Between the Switching Nodes
of the Two Regulators**

regulator to the sync pin of a "slave" regulator. The resistors should divide the gate-drive voltage down to something slightly less than the $V_{CC}$ supply of the slave regulator, typically from 12V down to approximately 4.5V. Total divider resistance of 20k to 30k is adequate. Also, the slave regulator must be set up to free run slower than the master regulator. If, for example, the master is configured

to run at approximately 300kHz (a 130k resistor from FSET to ground) the slave can be left to run at its natural frequency of 200kHz. The slave frequency will be forced up to that of the master.

The sync function on the LTC1430 works as follows: when the shutdown pin is pulled low, the high-side switch turns off; normal duty factor control determines when the high-side switch will turn back on. As long as the shutdown pin is held low for less than approximately 40μs, the chip will not shut down.

The simplified schematic (Figure 19) shows the synchronization circuitry. For a detailed description of LTC1430-based regulator designs, see the LTC1430 data sheet. The scope photo (Figure 20) shows the voltage at the common connection of the two FETs of each regulator.

## COMBINE A SWITCHING REGULATOR AND AN ULTRAFAST LINEAR REGULATOR FOR A HIGH PERFORMANCE 3.3V SUPPLY
*by Craig Varga*

### Introduction

It is becoming increasingly necessary to provide low voltage power to microprocessor loads at very high current levels. Many processors also exhibit high speed load transients. The Pentium® Pro processor from Intel exhibits both of these requirements. This processor requires 3.3V ±5% at approximately 14A peak (9A average) and is capable of making the transition from a low

power state to full load in several clock cycles. Generally, switching regulators are used to supply such high power devices, because of the unacceptable power losses associated with linear regulators. Unfortunately, switching regulators exhibit much slower transient response than linear regulators. This greatly increases the output capacitor requirements for switchers.

### Circuit Operation

The circuit shown in Figure 21 takes advantage of a new, ultrahigh speed linear regulator combined with a switching regulator to get the best of both worlds. An LTC1435 synchronous buck regulator is combined with an LT1575
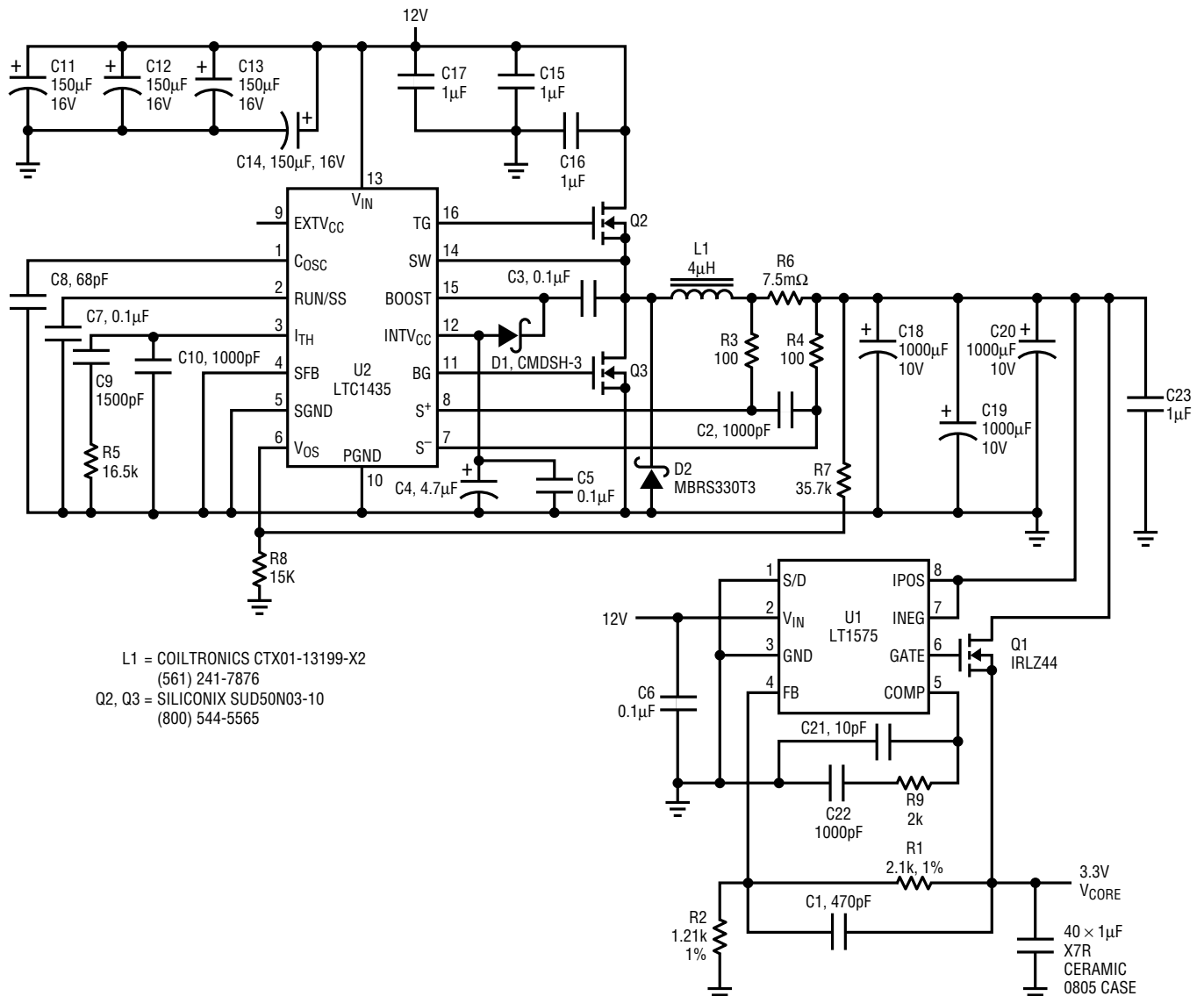


**Figure 21. 12V to 3.3V/9A (14A Peak) Hybrid Regulator**
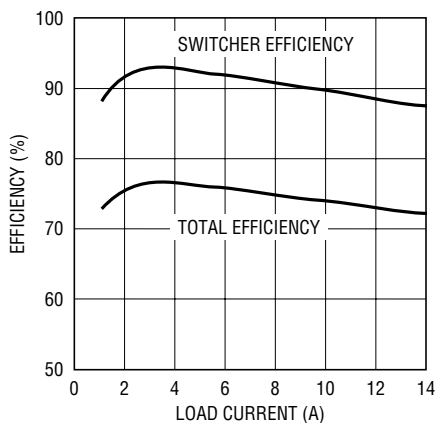
Figure 22. Efficiency of Figure 21's Circuit



Figure 23. Transient Response of Figure 21's Circuit to a 10A Load Step

linear regulator to generate a 3.3V output from a 12V input with an overall conversion efficiency of approximately 72%. The output is capable of current slew rates of approximately 20A per microsecond.

The LT1575 uses an IRLZ44 MOSFET as the pass transistor, allowing the dropout voltage to be less than 550mV. Setting the switching supply's output to only 700mV above the output of the linear regulator ensures output regulation. The switcher is therefore set up to deliver 4.0V at 14A from the 12V supply. Conversion efficiency of the switcher is around 90% (depending on load), whereas the LT1575's efficiency is 82.5% (see Figure 22). The 12V input current is only about 5.5A. At an average current of 9A, the power dissipation in the linear pass transistor is only 6.3W. A small stamped aluminum heat sink is adequate.

Figure 23 shows the transient response to a 10A load step with a rise time of approximately 50ns. The only output capacitance is 40, 1μF ceramic capacitors. No additional bulk capacitance is required at the processor. The circuit eliminates approximately a dozen low ESR tantalum capacitors at the load, which would be required without the linear postregulator. The switching supply's output is decoupled with three aluminum electrolytic capacitors. Because the transient response at this point is much less critical than at the load, the long-term degradation of the aluminum capacitors will not be as detrimental to the circuit's performance as it would be if they were used for load decoupling.

## THE LTC1624: A VERSATILE, HIGH EFFICIENCY, SO-8 N-CHANNEL SWITCHING REGULATOR CONTROLLER
*by Randy G. Flatness*

### Introduction

The LTC1624 is a current mode switching regulator controller operating at an internally set frequency of 200kHz. This versatile 8-pin controller uses the same constant frequency current mode architecture and Burst Mode operation as the LTC1435–LTC1439 controllers, but without the synchronous switch. The LTC1624, like the other members of the family, drives a cost-effective, external N-channel MOSFET for the topside switch and maintains low dropout operation previously available only with P-channel MOSFETs.

The LTC1624 can be configured to operate in all standard switching configurations, including boost, step-down, inverting, SEPIC and flyback, without a limitation on the output voltage. A wide input voltage range of 3.5V to 36V allows operation from a variety of power sources, from as few as four NiCd cells up though high voltage wall adapters. Tight load regulation, coupled with a reference voltage trimmed to 1%, provides very accurate output voltage control.

### Application Circuits

The LTC1624 can be used in a wide variety of switching regulator applications, the most common being the step-down converter. Other switching regulator architectures discussed here include step-up and SEPIC converters.
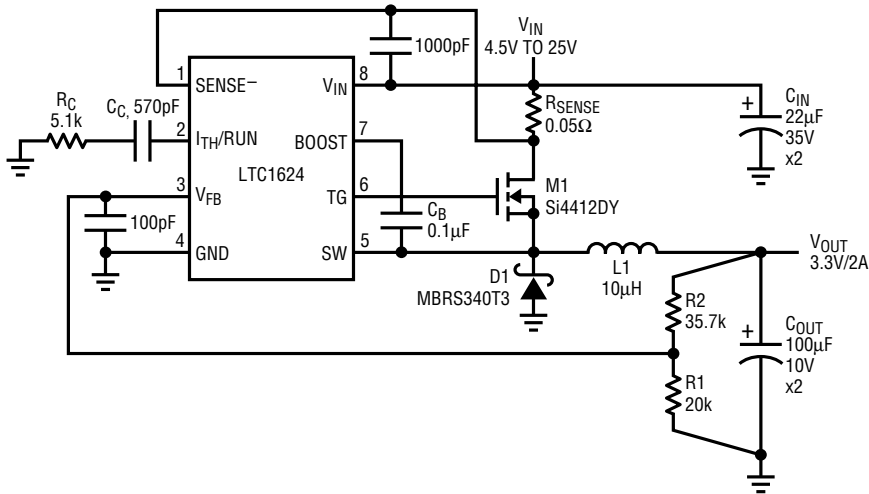
**Figure 24. High Performance 3.3V/2A Step-Down DC/DC Converter**



**Figure 25. Efficiency Plot
of Figure 24's Circuit**

The basic step-down converter is shown in Figure 24. This application shows a 3.3V/2A converter operating from an input voltage range of 4.5V to 25V. The efficiency for this circuit is shown in Figure 25.

Step-up and SEPIC applications require a low-side switch pulling the inductor to ground (see Figures 26 and 28). Since the source of the MOSFET must be grounded, the switch pin (SW) on the LTC1624 is also grounded in order for the driver to supply a gate-to-source signal to control the MOSFET. In these applications, the voltage on the boost pin is a constant 5V, resulting in a 0V–5V gate-drive level. A capacitor from boost to switch is still required, since this capacitor supplies the gate-charge currents.

The basic step-up converter is shown in Figure 26. The LTC1624 is used to create 12V/1A from a 5V source with the efficiency shown in Figure 27. Efficiency is above 90% from 20mA up to close to full load, dropping only to 89% at 1A.

In order to allow input voltages both above and below the output voltage, a SEPIC converter can be used. An example of the LTC1624 used as a 12V/0.5A SEPIC converter operating from an input range of 5V to 20V is shown in Figure 28.
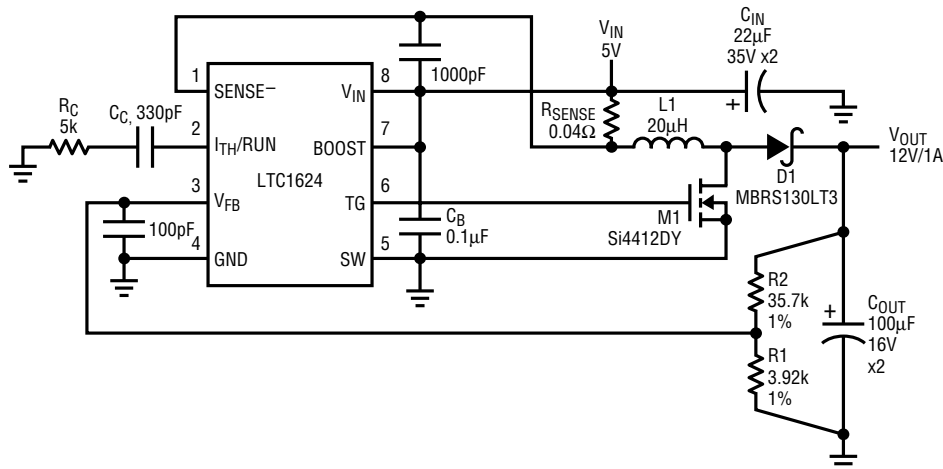


**Figure 26. 12V/1A Step-Up Converter**
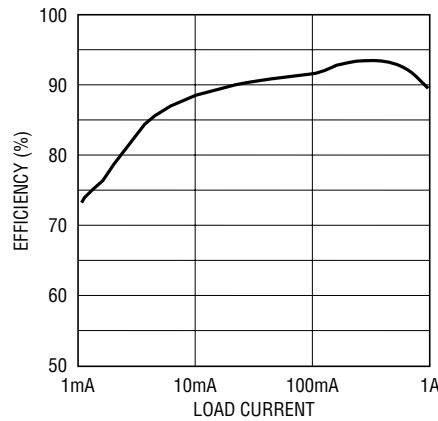
**Figure 27. Efficiency Plot for Figure 26's Circuit**
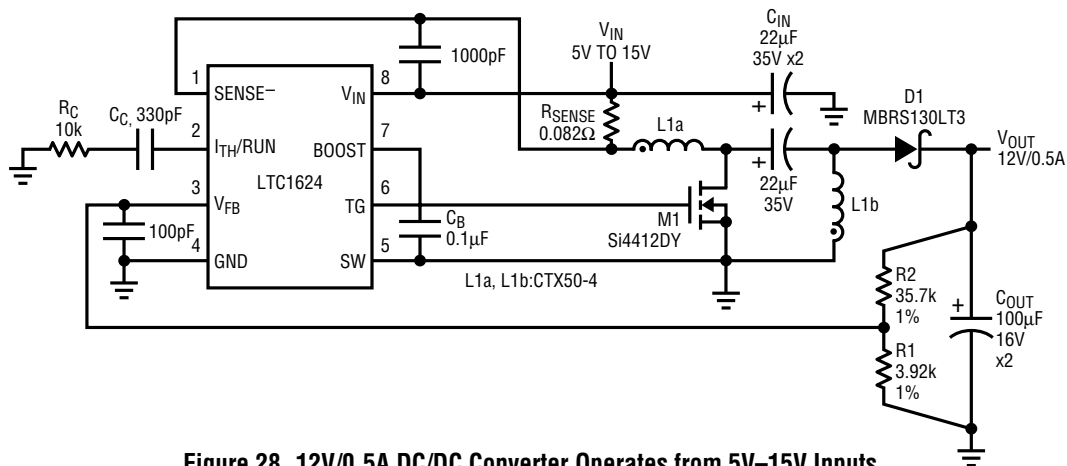


**Figure 28. 12V/0.5A DC/DC Converter Operates from 5V–15V Inputs**

## LOW COST 3.3V TO 1.XV 6 AMP POWER SUPPLY
*by Sam Nork*

As voltage requirements for microprocessors drop, the need for high power DC/DC conversion from a 3.xV supply to a lower voltage keeps growing. The LTC1430 is a very attractive choice for such DC/DC applications, due to its low cost, high efficiency and high output power capability. However, there are two problems: first, 3.xV does not provide enough gate drive to ensure low $R_{DS(ON)}$ using external logic-level FETs; and second, the LTC1430 has a 4V minimum input requirement. These obstacles are both overcome by using an LTC1517-5 regulated charge pump to generate the input voltage for the LTC1430.

The circuit shown in Figure 29 uses the LTC1430 to produce a synchronous 3.3V to 1.9V step-down DC/DC converter. The circuit achieves 90.5% efficiency at 3 amps of output current and has a 6 amp maximum output capability. (Refer to the LTC1430 data sheet for detailed description of LTC1430-based designs). Power for the LTC1430 is derived from the output of the LTC1517-5.

The LTC1517-5 is a switched capacitor charge pump available in a tiny, 5-pin SOT-23 package. The part uses Burst Mode operation to generate a 5V output from a 2.7V to 5V input. The regulated 5V supply powers the internal circuitry of the LTC1430 and ensures that the LTC1430 can

provide adequate gate drive to the external N-channel FETs. With insufficient gate drive, output power and efficiency will be significantly reduced due to high $R_{DS(ON)}$ of the FETs. In this circuit, typical supply current drawn by the LTC1430 is between 25mA and 30mA, the vast majority of which is needed to charge and discharge the external FETs. Because the LTC1517-5 has a maximum effective output impedance of 50Ω, this current can be comfortably supplied from a 3.3V input. If the input voltage drops to 3V or lower, the LTC1517-5 output may also drop. However, with the FETs shown in Figure 29, the LTC1517-5 will provide a 4.5V minimum supply to the LTC1430 at input voltages down to 3V. The circuit's efficiency is shown in Figure 30.

Pulling the $\overline{\text{SHDN}}$ pin on the LTC1430 low will shut down the power supply. Q1 and Q2 will be forced off and the LTC1430 quiescent current will drop to 1μA. Although the LTC1517-5 does not have a shutdown feature, the no-load operating current is an extremely low 6μA. This keeps the

overall shutdown current below 10μA plus external FET leakage. (For further reductions in shutdown current, an 8-pin LTC1522 may be used in place of the LTC1517-5; the LTC1522 is the same as an LTC1517-5 with shutdown.) The additional LTC1517-5 circuitry will not take up much board space. The entire circuit consumes only 0.045 in².
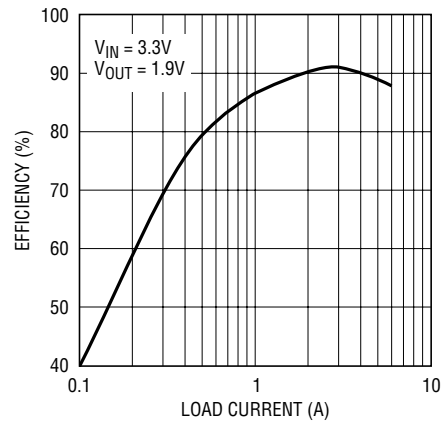


**Figure 30. Efficiency of Figure 29's Circuit**
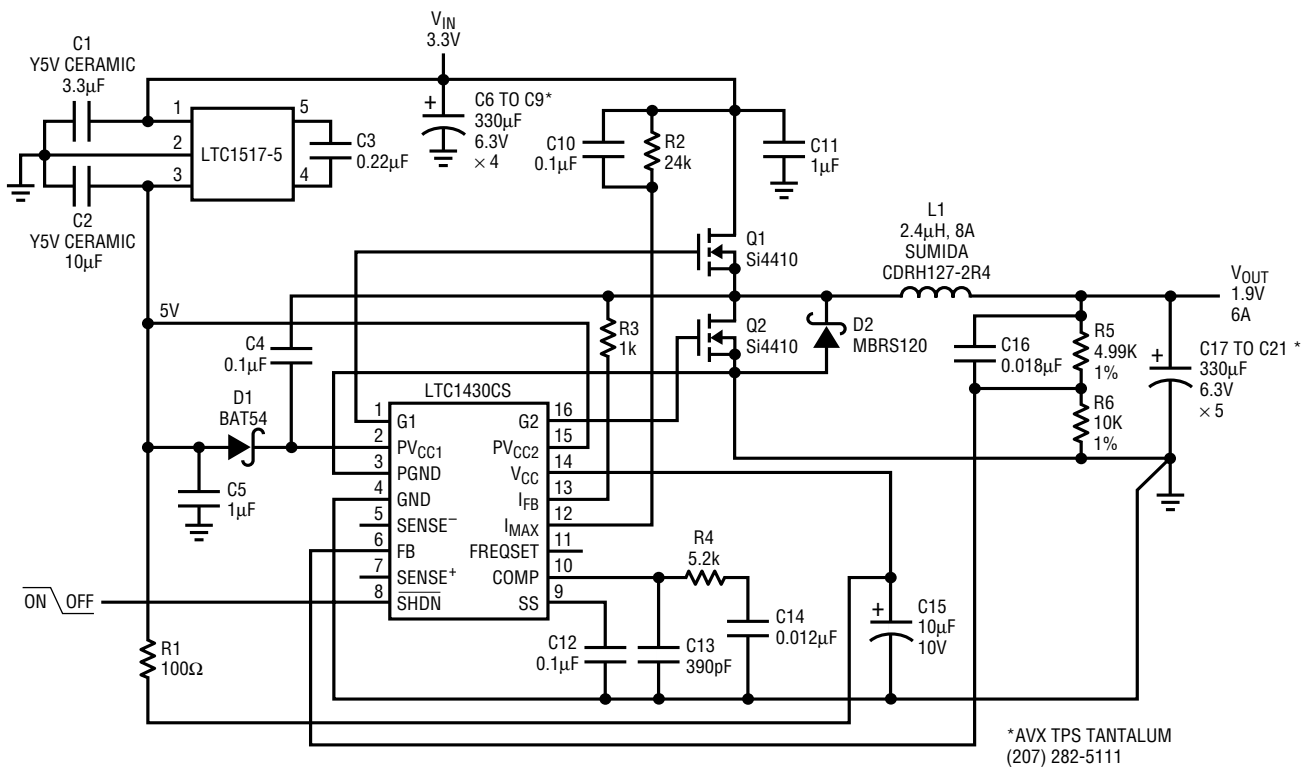


**Figure 29. 3.3V to 1.9V/6A Power Supply**

## THE LT1374: NEW 500kHz, 4.5A MONOLITHIC BUCK CONVERTER
*by Karl Edwards*

### Introduction

The LT1374 is a 4.5A buck converter using an on-chip 80mΩ switch. With its 500kHz operating frequency and integral switch, only a few external, surface mount components are required to produce a complete switching regulator. The LT1374's features include current mode control, external synchronization and a low current (typically 20μA) shutdown mode. Improvements have been made to reduce start-up headroom and switching noise. A novel power device layout makes it possible to fit a high speed, bipolar, 80mΩ switch into a surface mount SO-8 package. The LT1374 is also available in DD and TO-220 packages for higher power applications.

### Application: 5V/4.25A Buck Converter

With its 25V input and 4.5A minimum switch current, the LT1374 will fit into a wide range of applications. Figure 31 shows a typical buck converter with a 6V to 25V input range, a 5V output and 4.25A of output current capability. Due to the low on-resistance of the switch, efficiency remains high over a wide range of currents, as shown in Figure 32. To reduce power dissipation, both the BIAS pin and boost circuit are supplied from the 5V output.

Several factors, including maximum current, core and copper losses, size and cost, affect the choice of inductor, L1. A high value, high current inductor gives the highest output current with the lowest ripple, at the expense of a large physical size and cost. Lower inductance values tend to be physically smaller, have higher current ratings and are cheaper, but output ripple current, and hence ripple voltage, increases.

The input capacitor, C3, experiences very high ripple currents, up to $I_{OUT}/2$, so low ESR tantalum capacitors are needed. At 4.25A output current, two capacitors in parallel are required to meet the ripple current requirement. The ripple current in the output capacitor, C1, is lower, but its ESR still needs to be low to limit output voltage ripple. The voltage drop across the catch diode, D1, has a significant effect on overall converter efficiency, especially at higher input voltages when the switch duty cycle is low. Its ability to survive short-circuit conditions may increase its power rating. For good electrical performance, D1 must be placed close to the LT1374. The power dissipated in D1 will raise the PC board's temperature around the LT1374. This must be taken into account when modeling or taking bench measurements of die temperature.
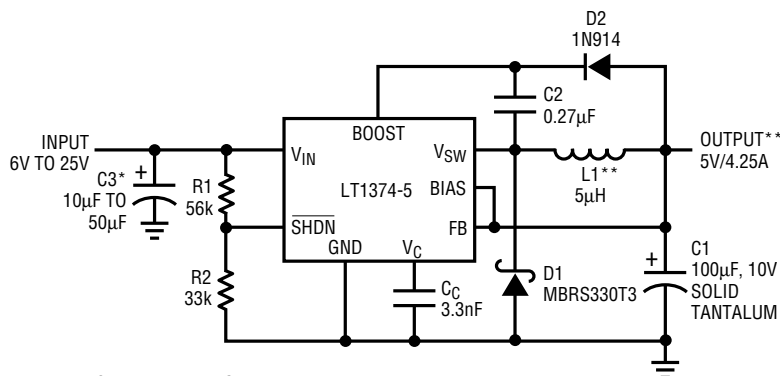


* RIPPLE CURRENT RATING > $I_{OUT}/2$
** L1 = COILTRONICS UP2-4R7; (561) 241-7876
  INCREASE L1 TO 10μH FOR LOAD CURRENTS ABOVE 3.5A AND TO 20μH ABOVE 4A

**Figure 31. 5V Buck Converter**
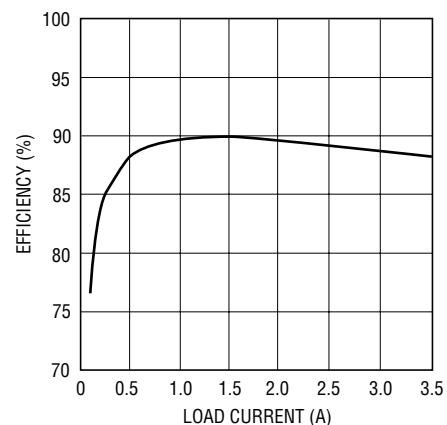


**Figure 32. Efficiency of Figure 31's Circuit: 10V In, 5V Out**

The loop compensation capacitor, $C_C$, produces a pole in the frequency response at 240Hz. Unity-gain phase margin can be further improved with the addition of a resistor, typically 2k, in series with $C_C$, adding a zero to the frequency response. This, however, can cause a large-signal subharmonic problem in the loop. The output ripple voltage feeds back through the error amplifier to the $V_C$ pin, changing the current trip point of the next cycle. This changes the voltage ripple at the output, and the loop is closed. Adding a second capacitor directly from the $V_C$ pin to ground to form a pole at one-fifth the switching frequency solves the problem.

## LTC1504: FLEXIBLE, EFFICIENT SYNCHRONOUS SWITCHING REGULATOR CAN SOURCE OR SINK 500mA
*by Dave Dwelley*

### Introduction

The LTC1504 is an 8-pin step-down switching regulator. It consists of a 200kHz fixed frequency, voltage-feedback, buck-mode switching regulator controller and a pair of 1.5Ω power switches in an 8-pin SO package. The LTC1504 also includes a synchronous rectifier on-chip, maximizing efficiency and minimizing external parts count while allowing the output to both sink and source current: it can source or sink up to 500mA with input voltages from 3.3V to 10V and output voltages as low as 1.26V. The LT1504 can achieve 100% duty cycle at the output switch, maximizing dropout performance with low input-to-output voltage differentials. The LTC1504 includes an onboard precision reference and user-programmable current-limit and soft-start circuits, allowing implementation of full-featured power conversion circuits with a minimum of external components.

The LTC1504 architecture is optimized for maximum efficiency at loads above 50mA and does not include a light-load Burst Mode™ circuit. This penalizes efficiency at very light loads but allows the device to seamlessly shift between sourcing and sinking current, opening up a whole new class of applications. A micropower shutdown mode

## PCB Layout

All high current, high speed circuits require careful layout to obtain optimum performance. When laying out the PCB, keep the trace length around the high frequency switching components as short as possible. This minimizes the EMI and RFI radiation from the loop created by this path. These traces have a parasitic inductance of approximately 20nH/inch, which can cause an additional problem at higher operating voltages. At switch-off, the current flowing in the trace inductance causes a voltage spike. This is in addition to the input voltage across the switch transistor. At higher currents, the additional voltage can potentially cause the output switching transistor to exceed its absolute maximum voltage rating.

is included. The diminutive SO-8 package minimizes the amount of space the LTC1504 fills while allowing adequate thermal dissipation for 500mA load current levels. The LTC1504 allows previously impossible (or at least awkward) tasks to be completed with ease.

### Minimum Component-Count Circuits

Figure 33 shows a fully functional LTC1504 5V to 3.3V regulator, including current limit and soft-start, using the fixed-output LTC1504-3.3 and only six external components. Efficiency is above 90% with load currents between
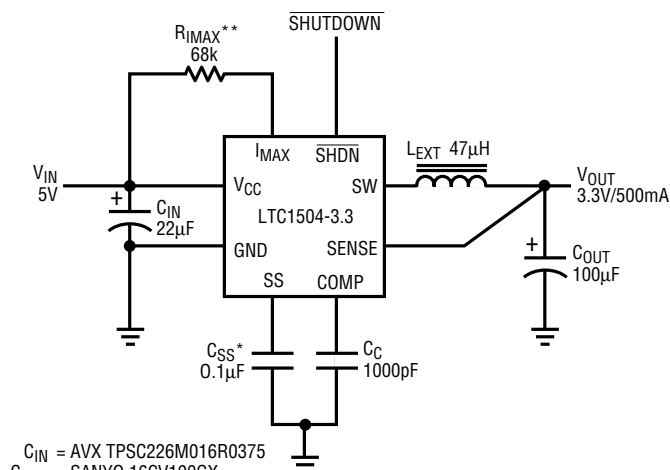
$C_{IN}$ = AVX TPSC226M016R0375
$C_{OUT}$ = SANYO 16CV100GX
$L_{EXT}$ = SUMIDA CD54-470
\* OPTIONAL: DELETE TO DISABLE SOFT START
\*\* OPTIONAL: DELETE TO DISABLE CURRENT LIMIT

**Figure 33. Minimum Parts-Count 5V–3.3V Converter**
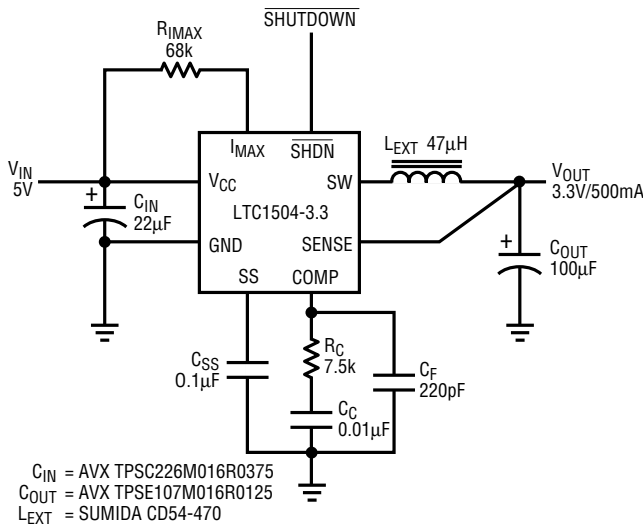
Figure 34. Improved Transient Response



Figure 35. SCSI-2 Active Terminator

50mA and 200mA, peaking at 92% at 100mA and remaining above 82% all the way to the maximum 500mA load. Current limit is set at 500mA in this example; it can be reduced by lowering the value of $R_{IMAX}$. $C_{SS}$ sets the start-up time at approximately 25ms.

The circuit in Figure 33 relies on the ESR of the output capacitor to maintain loop stability with just a single capacitor at the COMP pin. Figure 33 uses a surface mount electrolytic capacitor with about 400mΩ ESR. A low ESR tantalum output capacitor can improve the transient response at the output but requires a more complex compensation network at the COMP pin (Figure 34). There is a tradeoff to be made here: the minimum component count solution is the simplest and uses the least expensive components but pays a penalty in transient response. The low ESR circuit in Figure 34 has improved transient response and actually uses less board space: the tantalum output capacitor is smaller than the electrolytic device used in Figure 33 and the additional compensation components are tiny 0603 surface mount devices.

Note that the input bypass capacitor in both Figures 33 and 34 is an AVX TPS type, a relatively costly surge-tested tantalum capacitor. This is a small, surface mount device that has a surge current rating adequate to support the 500mA maximum load current of the LTC1504. Buck regulators (like the LTC1504) inherently draw large RMS currents from the input bypass capacitor, and the capacitor type chosen must be capable of withstanding this current

without overheating. As with all switching regulator circuits, layout is critical to obtaining maximum performance; if in doubt, contact the LTC Applications Department for component selection and layout advice.

**Sink/Source Capability Improves SCSI Terminators and Supply Splitters**

Figure 35 shows an adjustable-output LTC1504 connected as a 2.85V regulator for use as a SCSI terminator. The ability of the LTC1504 circuit to sink current makes it ideal for use in terminator applications, where the load is just as likely to be putting current into the regulator as taking it out. The synchronous-buck architecture of the LTC1504 allows it to shift cleanly between sourcing and sinking current, making it ideal for such applications. The small number of tiny external components required minimizes the space used by the terminator circuit. A low ESR output capacitor is used along with an optimized compensation network to improve output transient response and maintain maximum data fidelity.
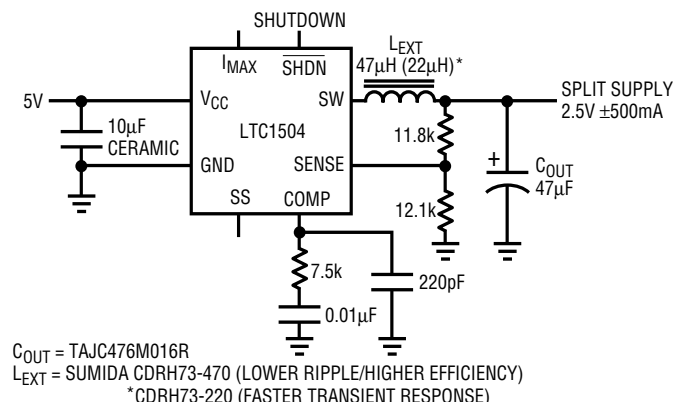


Figure 36. 5V Supply Splitter

Substituting a different set of feedback resistors (Figure 35) creates a 5V supply splitter, which creates a 2.5V "ground" to allow analog circuitry to operate from split supplies. Op amp circuits and data converters like to operate from dual supplies, and the sink/source capability of the LTC1504 allows load currents to be returned directly to the 2.5V "ground" supply.

## HIGH EFFICIENCY DISTRIBUTED POWER CONVERTER FEATURES SYNCHRONOUS RECTIFICATION
*by Dale Eagar*

### Introducing the LT1339

The LT1339 is the buck/boost converter that needs no steroids. As a full-featured switching controller, the LT1339 incorporates the features needed for system-level solutions. The LT1339 has an innovative slope-compensation function that allows the circuit designer freedom in controlling both the slope and offset of the slope-compensation ramp. Additionally, the LT1339 has an average current limit loop that yields a constant output current limit, regardless of input and/or output voltage. The LT1339's RUN pin is actually the input to a precision comparator, giving the designer freedom to select an undervoltage lockout point and hysteresis appropriate for the design. The SYNC and SS (soft-start) pins allow simple solutions to system-level design considerations. Like all Linear Technology controllers, the LT1339 has anti-shoot-through circuitry that ensures the robustness that is demanded in real-world applications for medium and high power conversion.

For input voltages ranging from 12V to 48V and output voltages ranging from 1.3V to 36V, the LT1339 is a simple, robust solution to your power-conversion problems. The LT1339 is ideal for power levels ranging from tens of watts to tens of kilowatts. The LT1339 is straightforward and remarkably easy to use. This is one power converter that's not afraid of 20A, 50A or even 150A of load current.

### Distributed Power

Figure 36 details a typical low voltage buck converter. This circuit has a $V_{IN}$ range of 10V to 18V with configurable output current and voltage. This simple circuit delivers 250W of load power into a 5V load while maintaining efficiencies in the mid-nineties.

### Higher Input Voltages

The circuit shown in Figure 37 is limited to 20V because of the maximum rating (Abs Max) of the LT1339 $V_{IN}$ pin. The input voltage can be extended above 20V by inserting a 10V Zener diode where the asterisk (*) is shown in Figure 37. This will extend the input voltage of Figure 37's circuit up to 30V (the Abs Max rating of the MOSFETs).



| $R_{SENSE}$ | $I_{LIMIT}$ |
|---|---|
| 0.01Ω | 10A |
| 0.005Ω | 20A |
| 0.002Ω | 50A |

| $R_{FB}$ | $V_{OUT}$ |
|---|---|
| 3K | 5V |
| 1.66K | 3.3V |
| 1.25K | 2.8V |
| 450Ω | 1.8V |
| 40Ω | 1.3V |

*SEE TEXT
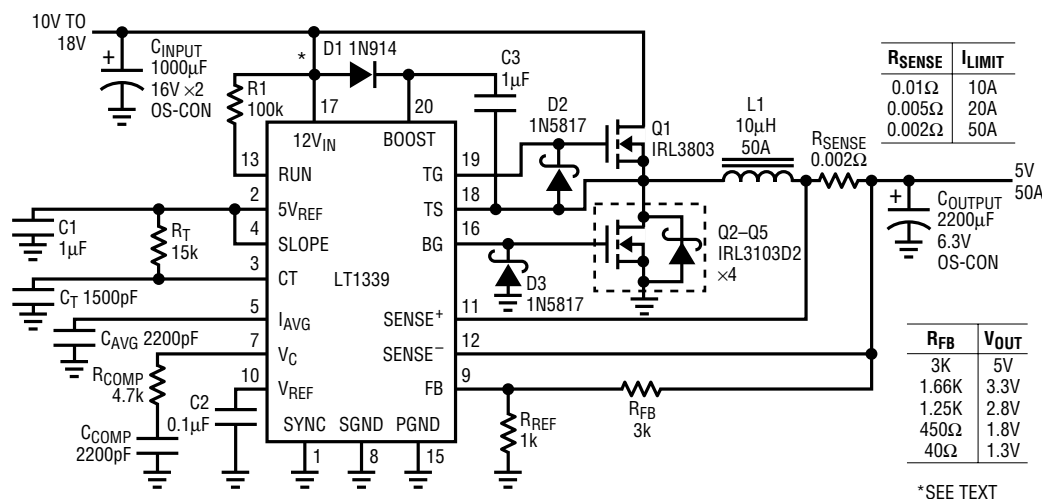
**Figure 37. 10V–18V In, 5V/50A Out Buck Converter**
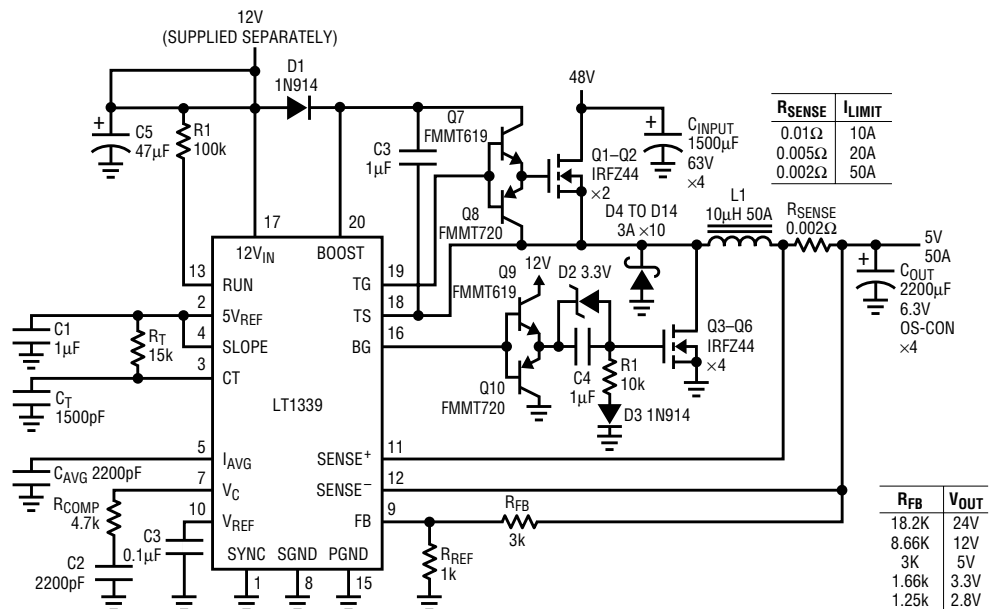
**Figure 38. 48V In, 5V/50A Out, High Power Buck Converter**

| $R_{SENSE}$ | $I_{LIMIT}$ |
|---|---|
| 0.01Ω | 10A |
| 0.005Ω | 20A |
| 0.002Ω | 50A |

| $R_{FB}$ | $V_{OUT}$ |
|---|---|
| 18.2K | 24V |
| 8.66K | 12V |
| 3K | 5V |
| 1.66k | 3.3V |
| 1.25k | 2.8V |
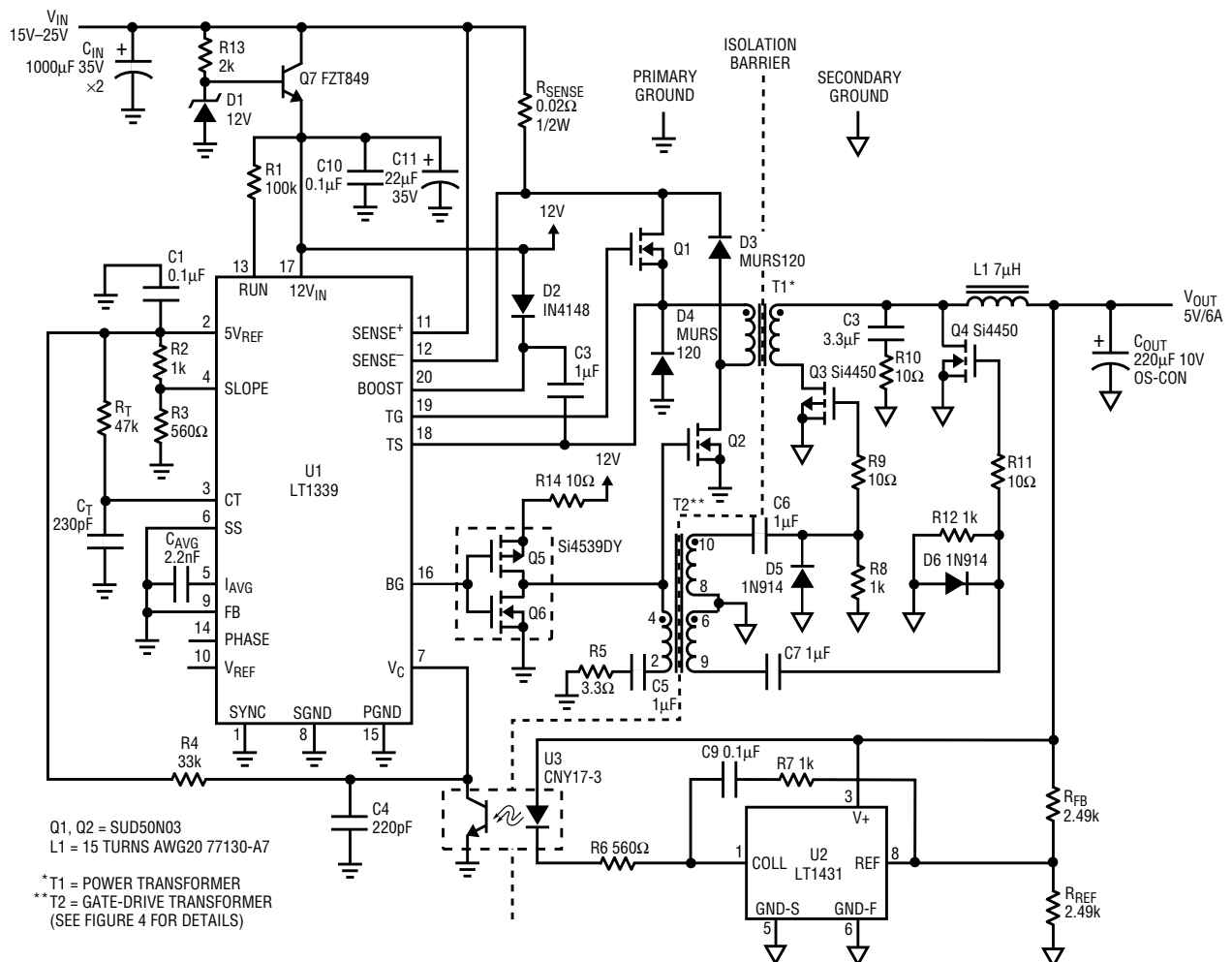


**Figure 39. Galvanically Isolated Synchronous Forward Converter (see Figure 40 for Details of T1 and T2)**

T1: PHILIPS EFD20-3F3 CORE
Lp = 93µH, Al = 1150nH/T² (NO GAP)

SECONDARY, 9 TURNS
TRIFILAR 26AWG
PRIMARY, 9 TURNS
TRIFILAR 26AWG

2MIL
POLYESTER
FILM

1500VDC ISOLATION
TUCK TAPE ENDS

T2: COILTRONICS VP1-1400
(500V ISOLATION)
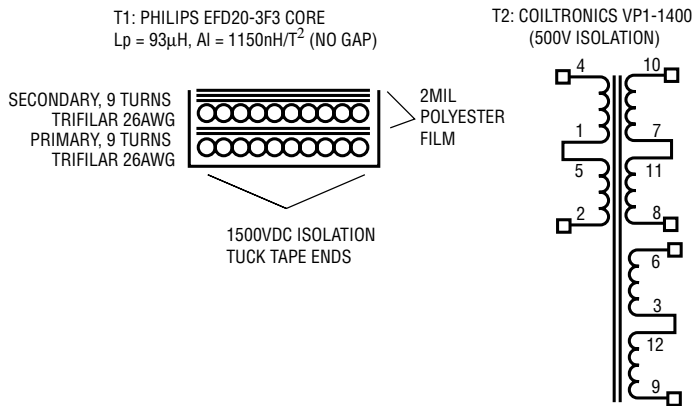
**Figure 40. Transformer Details of Figure 39's Circuit**

## Blame it on the Physicists

As the input voltage approaches 30V, the bottom MOS-FETs will begin to exhibit "phantom turn-on." This phenomenon is driven by the instantaneous voltage step on the drain, the ratio of $C_{MILLER}$ to $C_{INPUT}$, and yields localized gate voltages above $V_T$, the threshold voltage of the bottom MOSFET. To defeat the physicists, we add 3.3V of negative offset to the bottom gate drive, effectively making the threshold of the bottom MOSFETs 3.3V harder to reach (see Figure 38). This offset is provided by the 3.3V Zener, 1µF capacitor, 10k resistor and the 1N914 diode preceding the gate of the bottom MOSFETs.

## The Synchronous Forward Converter

Figure 39 details a Galvanically isolated LT1339 synchronous forward converter. Operating at its rated load of 6V at 5A, this circuit achieves 87% efficiency with a 15V input and 85% efficiency with a 24V input. Figure 40 shows details of the transformers used in Figure 39's circuit.

## The Synchronous Boost Converter

The LT1339 becomes a synchronous boost controller when the PHASE pin is grounded. Figure 41 details a 250W boost converter that outputs 28V at 9A from a 5V supply.
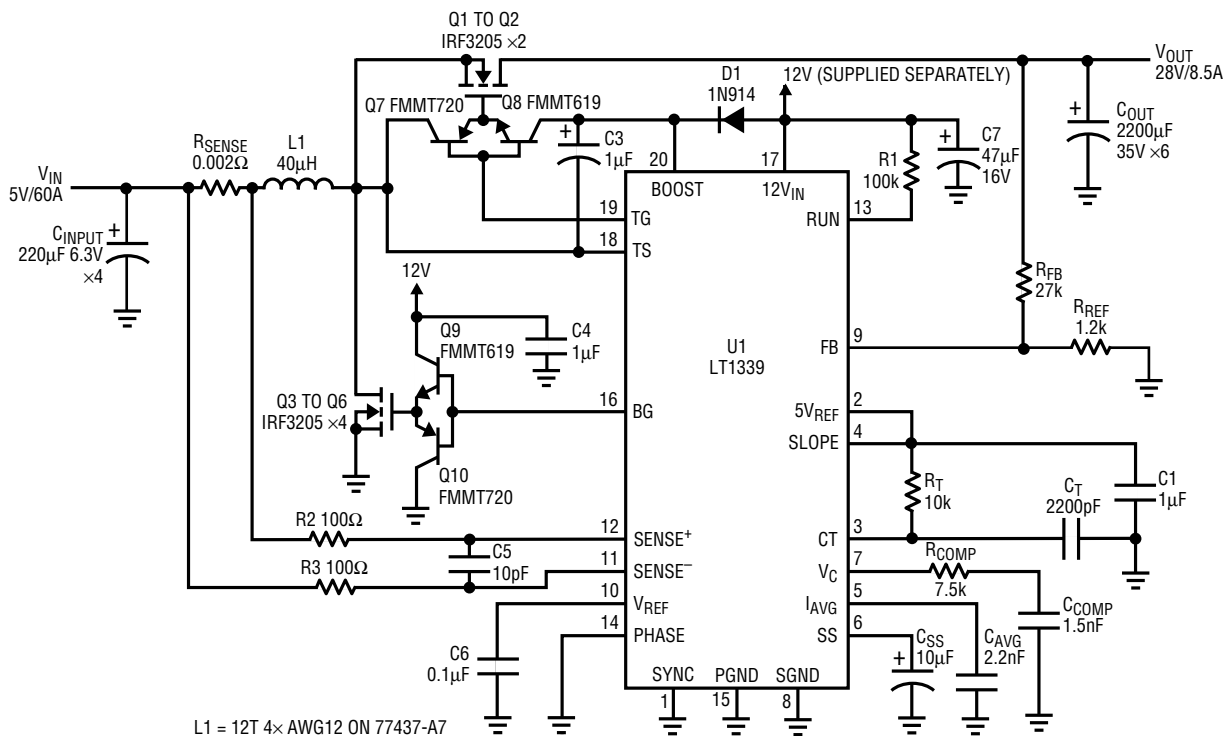


**Figure 41. This 5V to 28V Synchronous Boost Converter Limits Input Current at 60A (DC)**

**FIXED FREQUENCY, 500kHz, 4.5A STEP-DOWN CONVERTER IN AN SO-8 OPERATES FROM A 5V INPUT**
*by Karl Edwards*

**Introduction**

The LT1506 is a 500kHz monolithic buck mode switching regulator, functionally identical to the LT1374 but optimized for lower input voltage applications. Its high 4.5A switch rating makes this device suitable for use as the primary regulator in small to medium power systems. The small SO-8 footprint and input operating range of 4V to 15V is ideal for local onboard regulators operating from 5V or 12V system supplies. The 4.5A switch is included on the die, along with the necessary oscillator, control and logic circuitry to simplify design. The part's high switching frequency allows a considerable reduction in the size of external components, providing a compact overall solution.

The LT1506 is available in standard 7-pin DD and fused-lead SO-8 packages. It maintains high efficiency over a wide output current range by keeping quiescent supply current to 4mA and by using a supply-boost capacitor to saturate the power switch. The topology is current mode for fast transient response and good loop stability. Full cycle-by-cycle short-circuit protection and thermal shutdown are provided. Both fixed 3.3V and adjustable output voltage parts are available.

**5V to 3.3V Buck Converter**

The circuit in Figure 42 is a step-down converter suitable for use as a local regulator to supply 3.3V logic from a 5V power bus. The high efficiency, shown in Figure 43, removes the need for bulky heat sinks or separate power devices, allowing the circuit to be placed in confined locations. Since the boost circuit only needs 3V to operate, the boost diode can still be connected to the output, improving efficiency. Figure 42's circuit shows the shutdown pin option. If this pin is pulled to a logic low, the output is disabled and the part goes into shutdown mode, reducing supply current to 20µA. An internal pull-up ensures correct operation when the pin is left open. The SYNC pin, an option for the DD package, can be used to synchronize the internal oscillator to a system clock. A logic-level clock signal applied to the SYNC pin can synchronize the switching frequency in the range of 580kHz to 1MHz.

**Current Sharing Multiphase Supply**

The circuit in Figure 44 uses multiple LT1506s to produce a 5V, 12A power supply. There are several advantages to using a multiple switcher approach compared to a single larger switcher. The inductor size is considerably reduced. Inductor size is proportional to the energy that needs to be stored in the core. Three 4A inductors store less energy ($1/2Li^2$) than a single 12A coil, so they are much smaller. In addition, synchronizing three converters 120° out of phase with each other reduces input and output ripple currents. This reduces the ripple rating, size and cost of the filter capacitors.



Figure 42. 5V to 3.3V Step-Down Converter
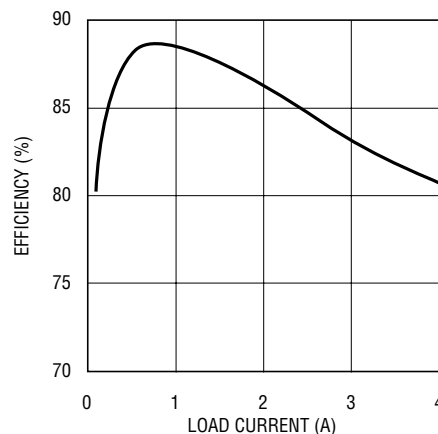


Figure 43. Efficiency vs Load Current for Figure 42's Circuit

C1, C3: MARCON THCS50E1E106Z
D1: ROHM RB051L-40
D2: 1N914
L1: DO3316P-682
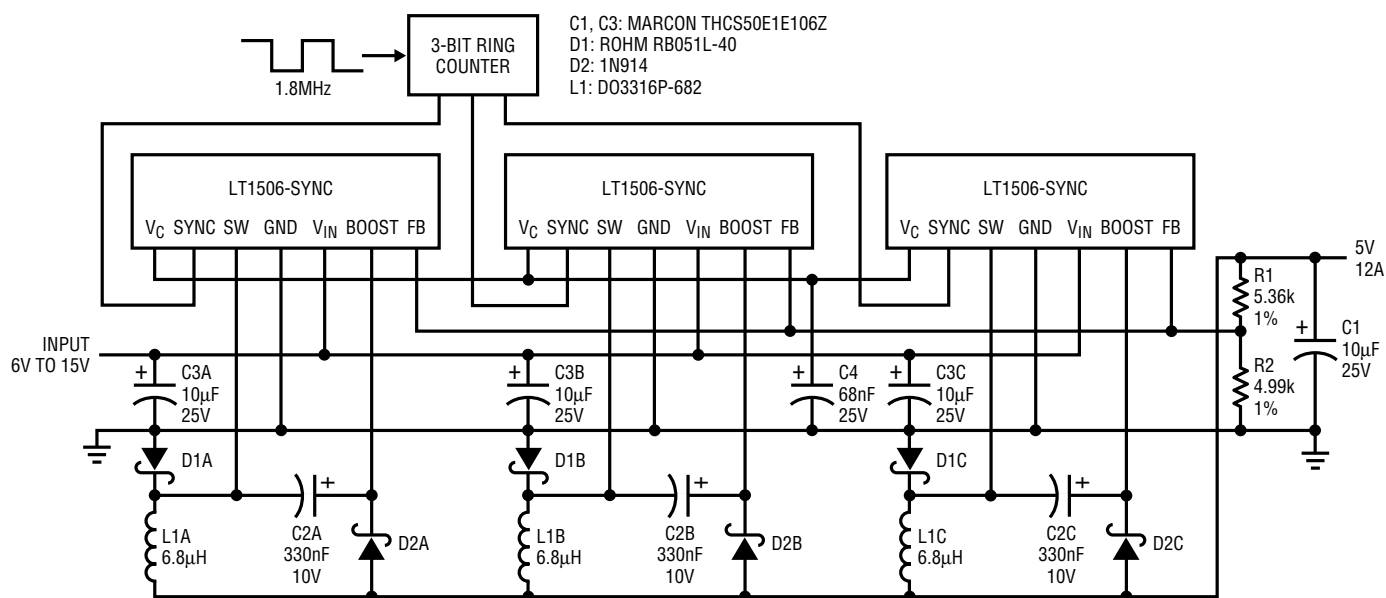
**Figure 44. Current-Sharing 5V/12A Supply**

## Current Sharing/Split Input Supplies

Current sharing is accomplished by connecting the $V_C$ pins to a common compensation capacitor. The output of the error amplifier is a $g_m$ stage, so any number of devices can be connected together. The effective $g_m$ of the composite error amplifier is the product of the individual devices. In Figure 44, the compensation capacitor, C4, has been increased by 3×. Tolerances in the reference voltages cause small offset currents to flow between the $V_C$ pins. The overall effect is that the loop regulates the output at a voltage somewhere between the minimum and maximum references of the devices used. Switch-current matching between devices will be typically better than 300mA over the full current range. The negative temperature coefficient of the $V_C$-to-switch-current transconductance prevents current hogging.

A common $V_C$ voltage forces each LT1506 to operate at the same switch current, not at the same duty cycle. Each device operates at the duty cycle defined by its input voltage. This is a useful feature in a distributed power system. The input voltage to each device could vary due to drops across the backplane, copper losses, connectors and so on. The common $V_C$ signal ensures that loading is still shared between the devices.

## Synchronized Ripple Currents

A ring counter generates three synchronization signals at 600kHz, 33% duty cycle, phased 120° apart. The sync input will operate over a wide range of duty cycles, so no further pulse conditioning is needed. At full load, each device's input ripple current is a 4A trapezoidal wave at 600kHz, as shown in Figure 45. Summing these waveforms gives the effective input ripple for the complete system. The resultant waveform, shown at the bottom of Figure 45, remains at 4A but its frequency has increased to 1.8MHz. The higher frequency eases the requirements on the value of input filter without the 3× increase in ripple current rating that would normally occur. Although only a single input capacitor is required, practical layout restrictions usually dictate an individual capacitor at each device. Figure 46 shows the output ripple current waveforms. The resultant 1.8MHz triangular waveform has a maximum amplitude of 350mA at an input voltage of 10V. This is significantly lower than would be expected for a 12A output. Interestingly, at inputs of 7.6V and 15V, the theoretical summed output ripple current cancels completely. To reduce board space and ripple voltage, C1 and C3 are ceramic capacitors. Loop compensation capacitor C4 must be adjusted when using ceramic output capacitors, due to the lack of effective series resistance (ESR).

**Redundant Operation**

The typical tantalum compensation value of 1.5nF is increased to 22nF ($\times$3) for the ceramic output capacitor. If synchronization is not used and the internal oscillators free run, the circuit will operate correctly, but ripple cancellation will not occur. Input and output capacitors must be ripple rated for the individual output currents.

The circuit shown in Figure 44 is fault tolerant when operating at less than 8A of output current. If one power stage fails open circuit, the output will remain in regulation. The feedback loop will compensate by raising the voltage on the $V_C$ pin, increasing the switch current of the two remaining devices.
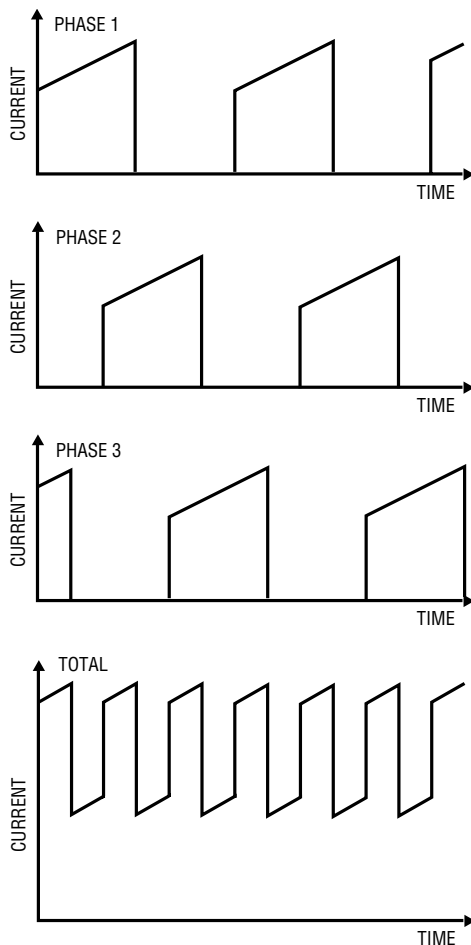


Figure 45. Input Current
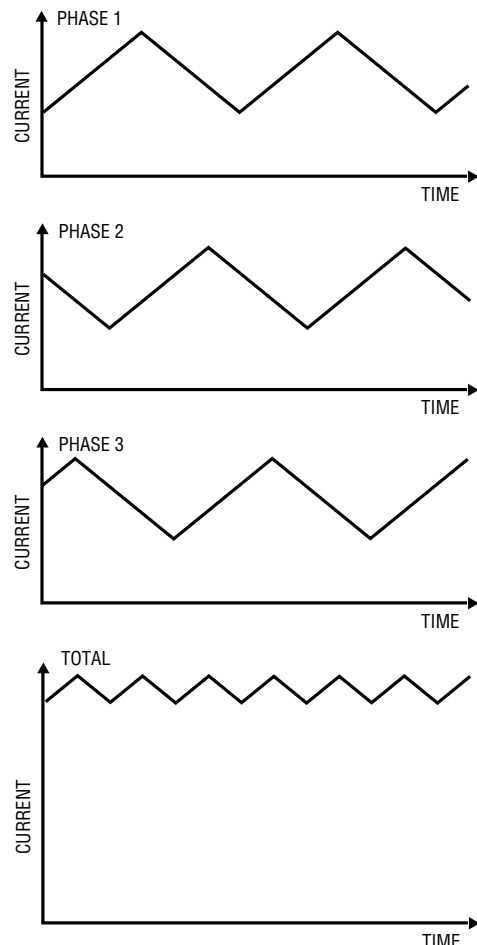


Figure 46. Output Current

## VID VOLTAGE PROGRAMMER
## FOR INTEL MOBILE PROCESSORS
*by Peter Guan*

Figure 47 shows a VID-programmed DC/DC converter for an Intel mobile processor that uses the LTC1435A and LTC1706-19 to deliver 7A of output current with a programmable $V_{OUT}$ of 1.3V to 2.0V from a $V_{IN}$ of 4.5V to 22V. Simply connecting the LTC1706-19's FB and SENSE pins to the LTC1435A's $V_{OSENSE}$ and SENSE$^-$ pins, respectively, closes the loop between the output voltage sense and the feedback inputs of the LTC1435A regulator with the appropriate resistive divider network, which is controlled by the LTC1706-19's four VID input pins.

Each VID pin must be grounded or driven low to produce a digital low input, whereas a digital high input can be generated by either floating the VID pin or connecting it to $V_{CC}$. The LTC1706-19 is fully TTL compatible and operational over a VID input voltage range that is much higher than $V_{CC}$.

Table 2 shows the VID inputs and their corresponding output voltages. VID3 is the most significant bit (MSB) and VID0 is the least significant bit (LSB). When all four inputs are low, the LTC1706-19 sets the regulator output voltage to 2.00V. Each increasing binary count is equivalent to decreasing the output voltage by 50mV. Therefore, to
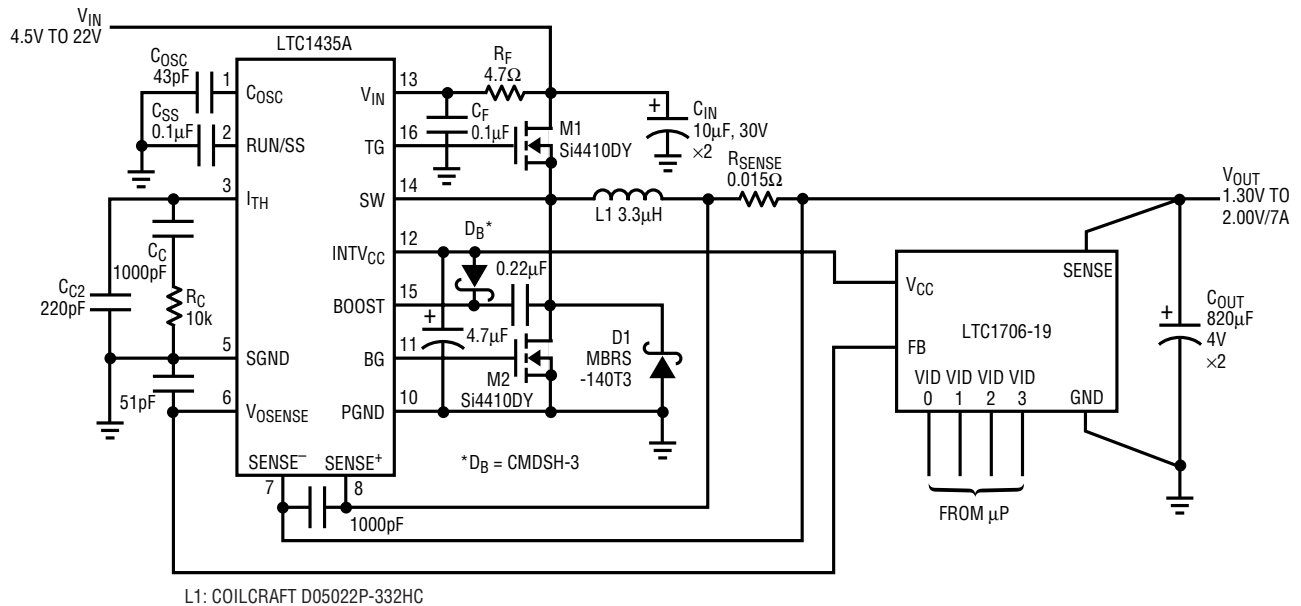


**Figure 47. Intel Mobil Pentium II VID Power Converter**
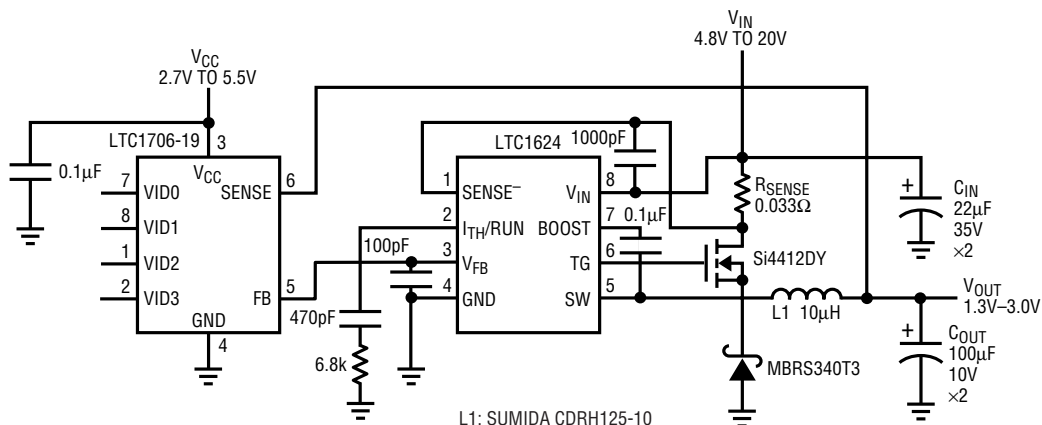


**Figure 48. High Efficiency SO-8, N-Channel 3A Switching Regulator with Programmable Output**

**Table 2. VID Inputs and Coresponding Output Voltages**

| Code | VID3 | VID2 | VID1 | VID0 | Output |
|------|------|------|------|------|--------|
| 0000 | GND | GND | GND | GND | 2.00V |
| 0001 | GND | GND | GND | Float | 1.95V |
| 0010 | GND | GND | Float | GND | 1.90V |
| 0011 | GND | GND | Float | Float | 1.85V |
| 0100 | GND | Float | GND | GND | 1.80V |
| 0101 | GND | Float | GND | Float | 1.75V |
| 0110 | GND | Float | Float | GND | 1.70V |
| 0111 | GND | Float | Float | Float | 1.65V |
| 1000 | Float | GND | GND | GND | 1.60V |
| 1001 | Float | GND | GND | Float | 1.55V |
| 1010 | Float | GND | Float | GND | 1.50V |
| 1011 | Float | GND | Float | Float | 1.45V |
| 1100 | Float | Float | GND | GND | 1.40V |
| 1101 | Float | Float | GND | Float | 1.35V |
| 1110 | Float | Float | Float | GND | 1.30V |

obtain a 1.30V output, the three MSBs are left floating while only VID0 is grounded. In cases where all four VID inputs are tied high or left floating, such as when no processor is present in the system, a regulated 1.25V output is generated at $V_{SENSE}$.

Figure 48 shows a combination of the LTC1624 and the LTC1706-19 configured as a high efficiency step-down switching regulator with a programmable output of 1.3V to 2.0V from an input of 4.8V to 20V. Using only one N-channel power MOSFET, the two SO-8 packaged LTC parts offer an extremely versatile, efficient, compact regulated power supply.

Figure 49 shows the LTC1436A-PLL and the LTC1706-19, a combination that yields a high efficiency low noise synchronous step-down switching regulator with programmable 1.3V to 2V outputs and external frequency synchronization capability.

Besides the LTC family of 1.19V-referenced DC/DC converters, the LTC1706-19 can also be used to program the output voltages of regulators with different onboard references. Figure 50 shows the LTC1706-19 programming the output of the LT1575, an UltraFast™ transient response, low dropout regulator that is ideal for today's power-hungry desktop microprocessors. However, since the LT1575 has a 1.21V reference instead of a 1.19V reference, the output will range from 1.27V to 2.03V in steps of 50.8mV.
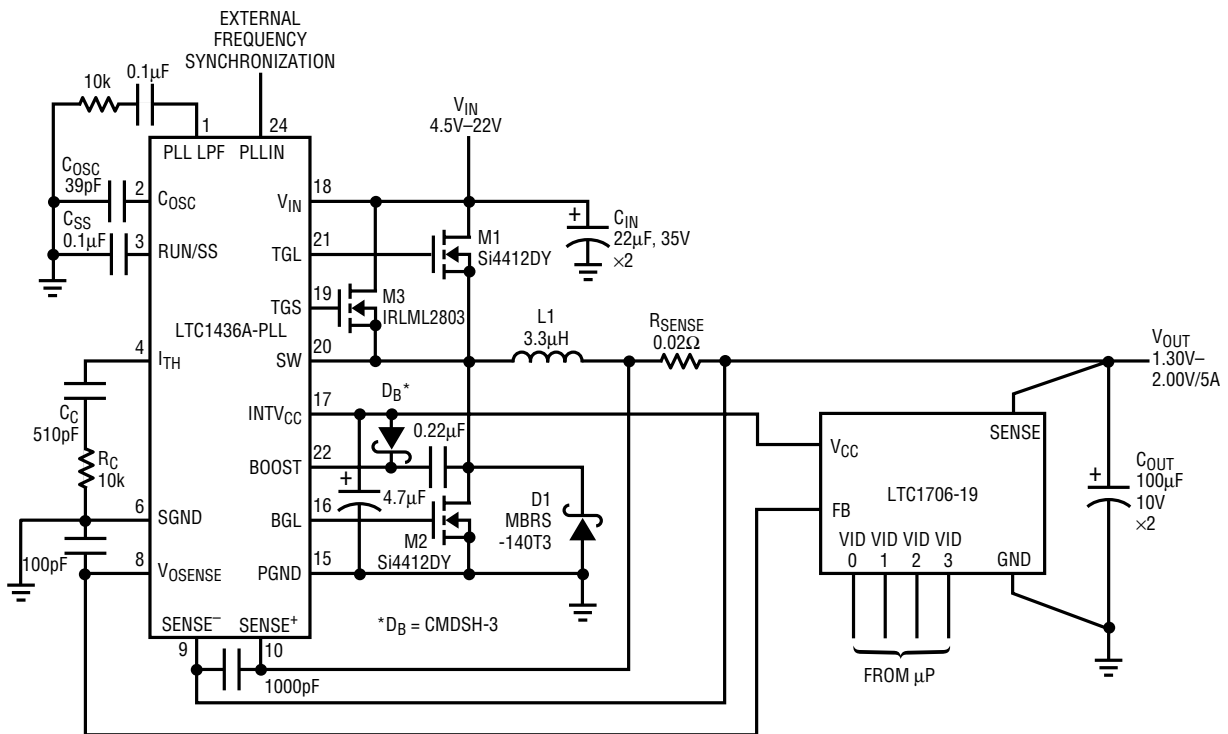


**Figure 49. High Efficiency, Low Noise, Synchronous Step-Down Switching Regulator with Adjustable Output Voltage**
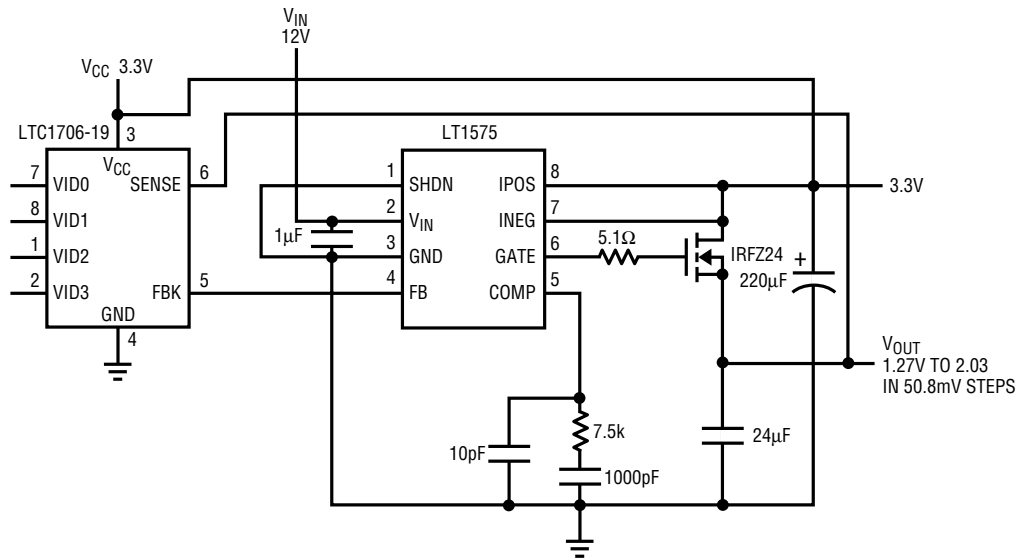
**Figure 50. UltraFast Transient Response, Low Dropout Regulator with Adjustable Output Voltage**

## NEW DC/DC CONTROLLER ENABLES
## HIGH STEP-DOWN RATIOS
*by Greg Dittmer*

### Capabilities of the LTC1435

The LTC1435 high efficiency synchronous DC/DC controller has been extremely popular for notebook computers and other battery-powered equipment due to its low noise, constant-frequency operation and its dual N-channel drive for outstanding high current efficiency without sacrificing low dropout operation. However, its 400ns to 500ns minimum on-time requires lower operating frequencies (<150kHz) to regulate output voltages below 2.0V if $V_{IN}$ is

high. This occurs because $t_{ON} = V_{OUT}/(V_{IN} \cdot f)$; thus, at low duty ratios, frequency must be decreased to keep $t_{ON} > t_{ON(MIN)}$. Lowering the operating frequency is usually not desirable because it increases noise and component size.

What happens if minimum on-time is violated in the LTC1435? If $V_{IN}$ is increased so that the on-time falls below $t_{ON(MIN)}$, the LTC1435 will begin to skip cycles to remain in regulation. During this "cycle-skipping" mode, the output remains in regulation but the operating frequency decreases, causing the inductor ripple current and output ripple voltage to increase.
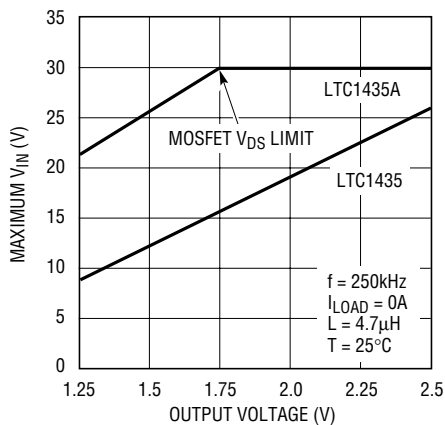


**Figure 51. LTC1435/LTC1435A Maximum $V_{IN}$ Comparison**
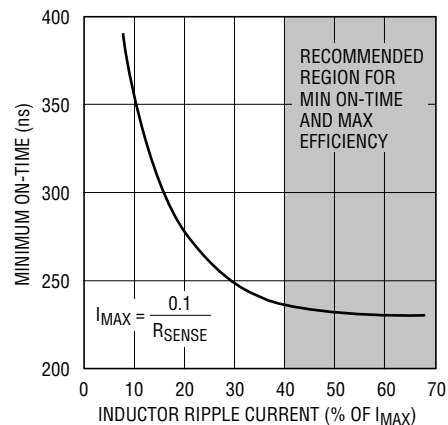


**Figure 52. LTC1435A Minimum On-Time vs Inductor Ripple Current**
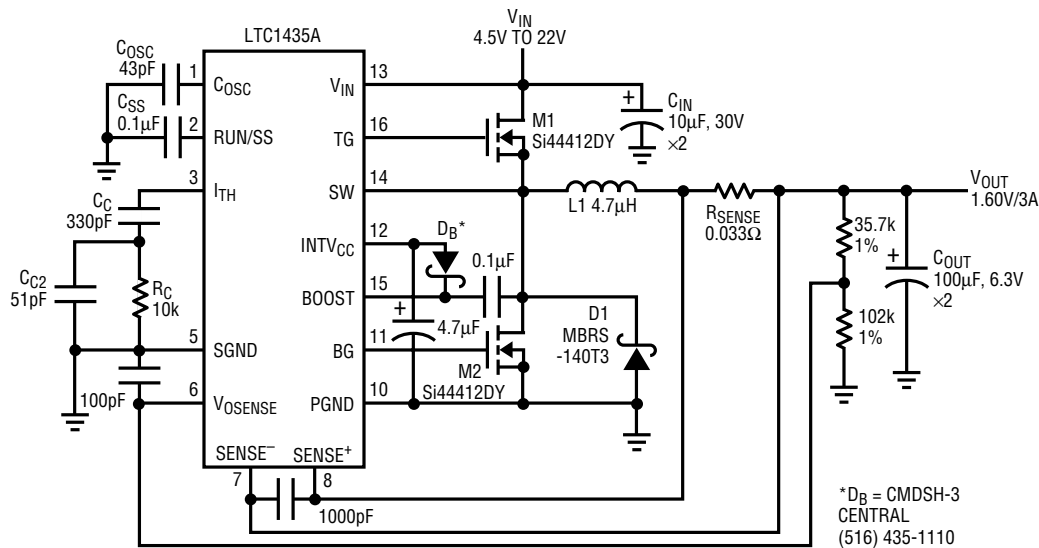
**Figure 53. LTC1435A 22V to 1.6V/3A Converter (f = 250kHz)**

## Enter the LTC1435A

The operating envelope has been substantially expanded with the introduction of the new LTC1435A DC/DC controller, which has all the outstanding features of the LTC1435 with a reduced minimum on-time of 300ns or less and improved noise immunity at low output voltages. With these improvements, high performance at output voltages down to 1.3V can be achieved with operating frequencies in excess of 250kHz from input supply voltages above 22V. Figure 51 shows the resulting improvement of maximum $V_{IN}$ vs output voltage as a result of the reduced minimum on-time.

The LTC1435A's minimum on-time is dependent on the speed of the internal current comparator, which in turn is dependent on the amplitude of the signal the comparator is monitoring: inductor ripple current. Thus, the higher the ripple current, the lower the minimum on-time. Figure 52 shows how minimum on-time varies as a function of the inductor ripple amplitude. At higher amplitudes, $t_{ON(MIN)}$ is less than 250ns; at low amplitudes it can be 350ns or more. This means that for low duty cycle applications where the on-time is approaching $t_{ON(MIN)}$, there may be a minimum ripple current amplitude, and hence, a maximum inductance necessary to prevent cycle skipping. Or, expressed differently, the lower the inductance, the higher the maximum $V_{IN}$ that can be achieved before the minimum on-time is violated and cycle skipping occurs. For most applications, 40% ripple not only reduces the minimum on-time but also optimizes efficiency.

## 22V to 1.6V Converter at 250kHz

Figure 53 shows the LTC1435A configured in an all N-channel synchronous buck topology as a 22V to 1.6V/3A converter running at 250kHz. The 43pF $C_{OSC}$ capacitor sets the internal oscillator frequency at 250kHz and the 33mΩ sense resistor sets the maximum load current at 3A. For a 22V to 1.6V converter, the on-time required is:

$$t_{ON} = 1.6/(22 \times 250kHz) = 291ns$$

Can the LTC1435A do this? At maximum $V_{IN}$ the inductor ripple is

$$\Delta I_L = \frac{V_{OUT} \cdot (1 - V_{OUT}/V_{IN})}{F \cdot L}$$

$$= \frac{1.6 \cdot (1 - 1.6/22)}{250kHz \cdot 4.7\mu H} = 1.3A$$

which is 43% of the 3A maximum load. From Figure 52, 43% ripple gives a minimum on-time of 235ns, which is well below the 291ns required by this application, so no cycle skipping will occur. If a 10μH inductor is used, the ripple amplitude drops to 0.6A or 20% and the minimum on-time increases to 280ns. This does not provide much margin below the 291ns on-time required, and thus the 4.7μH inductor is a better choice.

## Intel Mobile Processor VID Power Converter

Figure 54 shows the LTC1435A used with an LTC1706-19 to implement an Intel Mobile Pentium® II Processor VID power converter. This DC/DC converter provides digitally selectable output voltages over the range of 1.3V to 2.0V in 50mV increments at 250kHz and a 7A load current. The selectable output voltage is implemented by replacing the conventional feedback resistor network with the LTC1706-19, which provides the appropriate feedback resistor ratios internally. The proper ratio is selected with the 4-bit digital input pins.
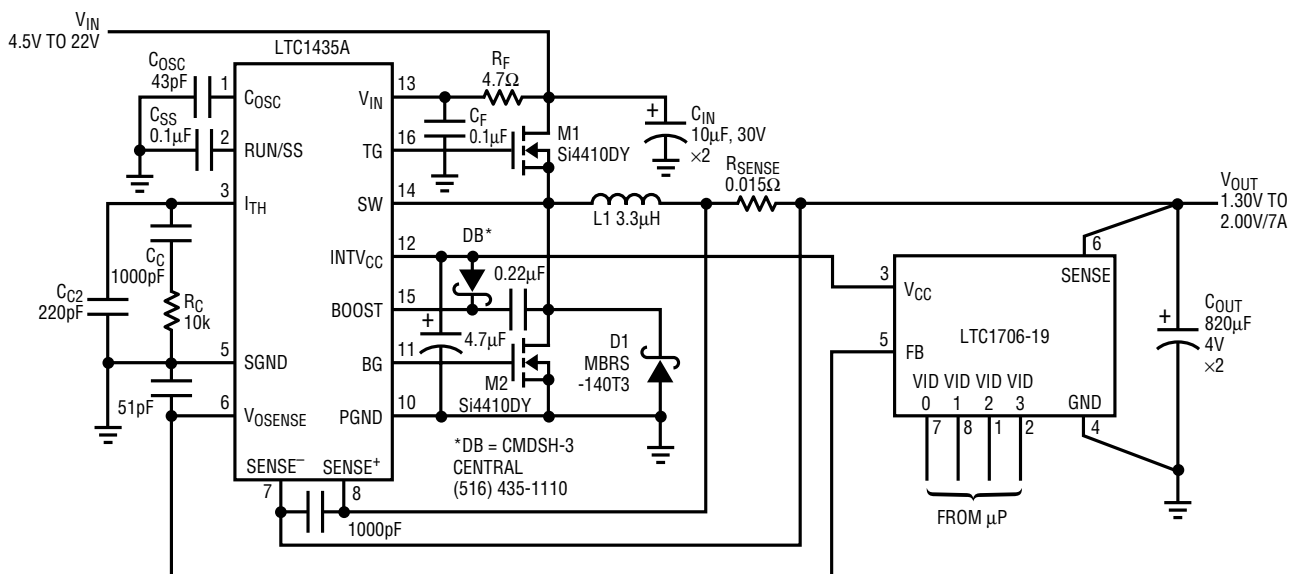


**FIgure 54. Intel Mobil Pentium II VID Power Converter**

## LTC1627 MONOLITHIC SYNCHRONOUS STEP-DOWN REGULATOR MAXIMIZES SINGLE OR DUAL LI-ION BATTERY LIFE
*by Jaime Tseng*

### Introduction

The LTC1627 is a new addition to a growing family of power management products optimized for Li-Ion batteries. Li-Ion batteries, with their high energy density, are becoming the chemistry of choice for many handheld products. As the demand for longer battery operating time continues to increase and the operating voltages of submicron DSPs and microcontrollers decreases, more demands are placed on DC/DC conversion. The LTC1627 monolithic, current mode synchronous buck regulator was specifically designed to meet these demands.

The LTC1627, with its operating supply range of 2.65V to 8.5V, can operate from one or two Li-Ion batteries as well as 3- to 6-cell NiCd and NiMH battery packs.

The LTC1627 incorporates power saving Burst Mode operation and 100% duty cycle for low dropout to maximize the battery operating time. In Burst Mode operation, both power MOSFETs are turned off for increasing intervals as the load current drops. Along with the gate-charge savings, unused circuitry is shut down between burst intervals, reducing the quiescent current to 200µA. This extends operating efficiencies exceeding 90% to over two decades of output load range.

### Typical Applications

The LTC1627, with its synchronous switching and attendant circuitry, provides the means of easily constructing a secondary flyback regulator, as shown in Figure 55. This flyback regulator is regulated by the secondary feedback resistive divider tied to the SYNC/FCB pin. This pin forces continuous operation whenever it drops below its ground-referenced threshold of 0.8V. Power can then be drawn from the secondary flyback regulator whether the main output is loaded or not.
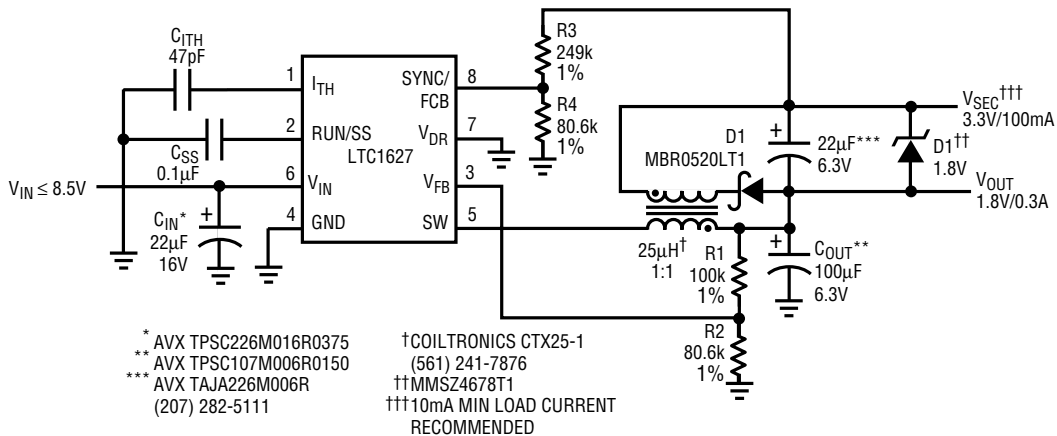
**Figure 55. Dual-Output 1.8V/0.3A and 3.3V/100mA Application**

## 1 or 2 Li-Ion Step-Down Converter

Figure 56 is a schematic diagram showing the LTC1627 being powered by one or two Li-Ion batteries. All the components shown in this schematic are surface mount and have been selected to minimize the board space and height. The output voltage is set at 3.3V, but is easily programmed to other voltages.

## Single Li-Ion Step-Down Converter

The circuit in Figure 57 is intended for input voltages below 4.5V, making it ideal for single Li-Ion battery applications. Diodes D1 and D2 and capacitors C1 and C2 comprise the bootstrapped charge pump to realize a negative supply at the $V_{DR}$ pin, the return pin for the top P-channel MOSFET driver. This allows Figure 57's circuit to maintain low switch $R_{DS(ON)}$ all the way down to the UVLO trip voltage.
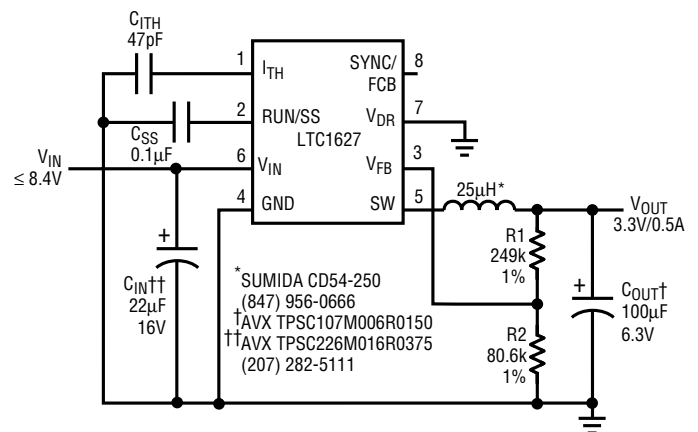


**Figure 56. Lithium-Ion to 3.3V/0.5A regulator**
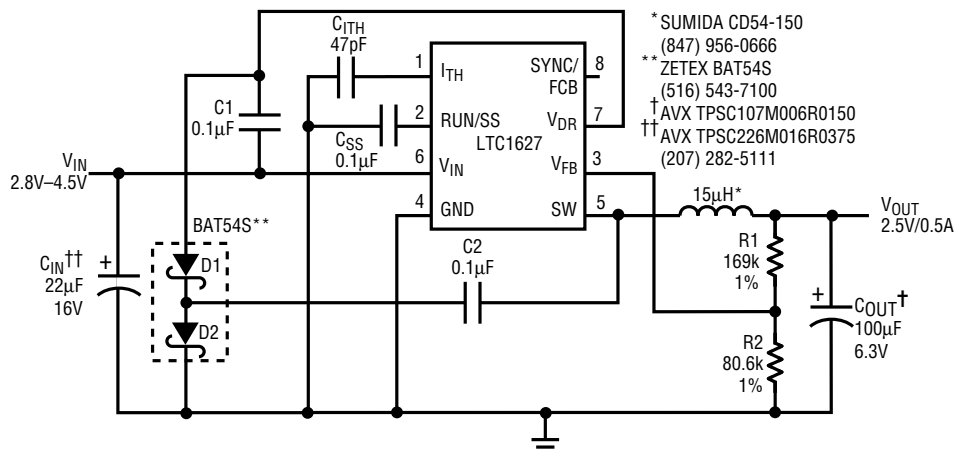


**Figure 57. Single Lithium-Ion to 2.5V/0.5A Regulator**

## THE LTC1625 CURRENT MODE DC/DC CONTROLLER ELIMINATES THE SENSE RESISTOR

*by Christopher B. Umminger*

### Introduction

Power supply designers have a new tool in their quest for ever higher efficiencies. In the past, when designing a step-down DC/DC converter, one had to choose between the high efficiency of voltage mode control and the many benefits of current mode control. Although voltage mode control offers high efficiency and a simple topology, it is difficult to compensate, has poor rejection of input-voltage transients and does not inherently limit output current under fault conditions, such as an output short circuit. Current mode control overcomes these problems by adding a control loop to regulate the inductor current in addition to the output voltage. Unfortunately, a sense resistor is required to measure this current, which adds cost and complexity while reducing converter efficiency. However, with the new LTC1625 No R$_{SENSE}$™ controller, one can enjoy all of the benefits of current mode control without the penalties of using a sense resistor.

The LTC1625 is a step-down DC/DC switching regulator controller that is capable of a wide range of operation with inputs from 3.7V to 36V. Fixed output voltages of 5V and 3.3V can be selected or an external resistive divider can be used to obtain output voltages from 1.19V up to nearly the
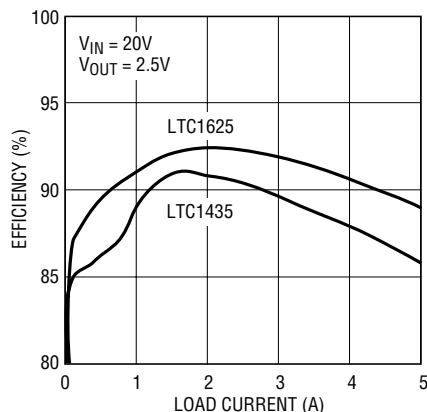


**Figure 59. Efficiency vs Load Current**

full input voltage. The controller provides synchronous drive for N-channel power MOSFETs and retains the advantage of low dropout operation typically associated with P-channel MOSFETs. Burst Mode™ operation maintains efficiency at low load currents, but can be overridden to assist secondary-winding regulation by forcing continuous operation. In addition to eliminating the sense resistor, the LTC1625 further reduces the external parts count by incorporating the oscillator timing capacitor. The oscillator frequency can be set to 150kHz, 225kHz, or can be injection locked to any frequency between these points.

### Design Examples

Figure 58 shows the LTC1625 in an application supplying a 2.5V output using an external feedback divider. Si4410DY MOSFETs from Siliconix allow this converter to deliver up
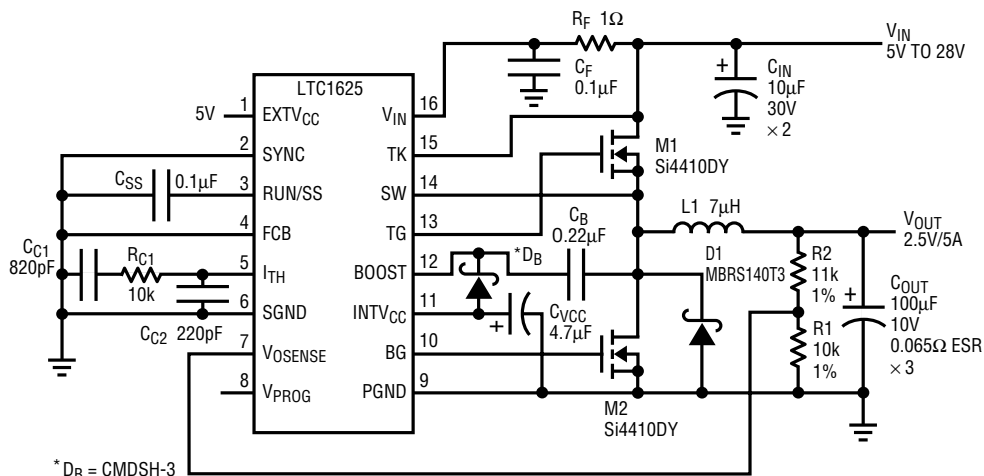


**Figure 58. 2.5V/5A Adjustable-Output Supply**
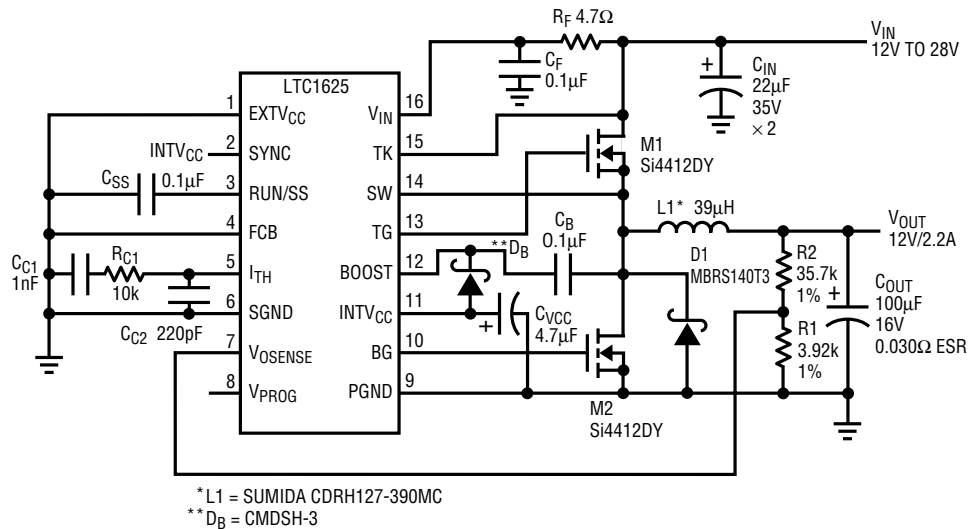
**Figure 60. 12V/2.2A Adjustable-Output Supply**

to 5A of load current. Ripple current is 1.8A (36% of full load) and current limit occurs around 6A. Note also that the EXTV$_{CC}$ pin is connected to an external 5V supply. This increases efficiency by drawing the roughly 7mA gate charge current from a supply lower than V$_{IN}$.

An efficiency plot of this circuit is shown in Figure 59. An LTC1435 with identical components in the power path is also plotted for comparison. At lower output voltages such as this, the sense resistor is responsible for an increasing share of the total power loss. By eliminating this source of loss, the LTC1625 is easily able to deliver an efficiency

greater than 90% at high load current. The benefit of reduced I$^2$R loss is readily apparent at the highest loads. The controller makes a transition to Burst Mode operation below around 1.1 A which keeps the efficiency high at moderate loads.

A circuit demonstrating the wide output range of the LTC1625 is shown in Figure 60. This application uses Si4412DY MOSFETs to deliver a 12V output at up to 2.2A. Note that the SYNC pin is tied high for 225kHz operation in order to reduce the inductor size and ripple current.

# PolyPhase SWITCHING REGULATORS OFFER HIGH EFFICIENCY IN LOW VOLTAGE, HIGH CURRENT APPLICATIONS
*by Craig Varga*

## Introduction

In recent years, there has been a tendency in the digital world toward smaller device geometries and higher gate counts. This has led to requirements for lower voltages and higher currents for logic supplies. As this trend continues, to levels under 2V and over 30A, the conventional buck regulator approach ceases to be viable. Switch currents are too high for a single device to handle, inductor energy storage exceeds what is available in surface mount technology and ripple current requirements on input capacitors dictate the use of many capacitors in parallel. Although all this may seem like enough of a challenge, the

transient response requirements also become much more severe. The question that arises is: "is there a topology that can solve all of these problems simultaneously? " The answer is "PolyPhase™."

## What is PolyPhase, Anyway?

Since it is apparent that multiple FETs need to be paralleled to handle the current requirements, the question is whether there is a way to drive them intelligently, rather than by brute force. The solution is to stagger the turn-on times so that the dead bands in the input current waveform are "filled up," so to speak. In the simplest implementation, there are essentially two independent synchronous buck regulators operating 180° out of phase. The net effect of this is that the input and output ripple currents of the two channels tend to cancel during steady-state operation. This results in significant reductions in both input and
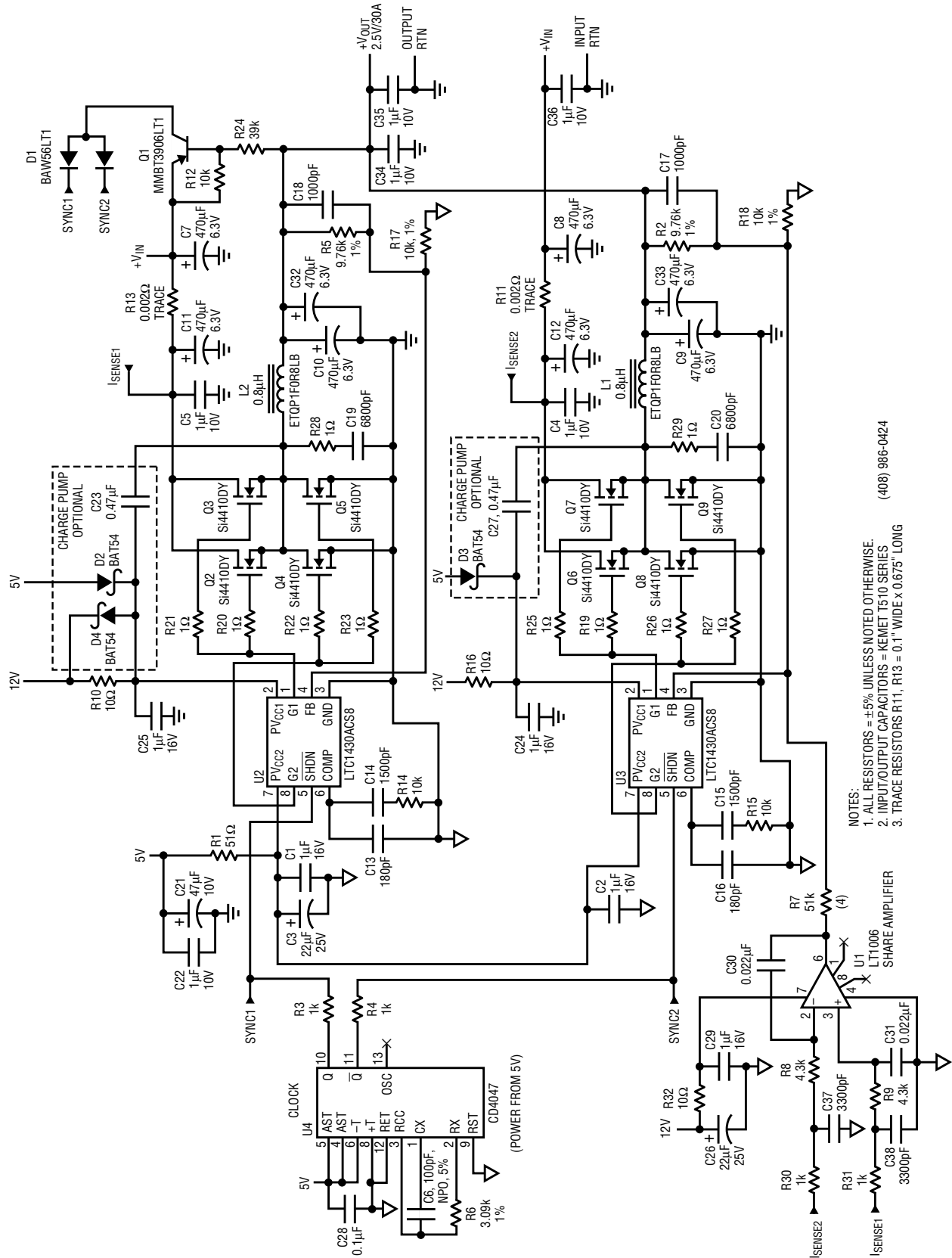
**Figure 61. 2-Phase Synchronous Buck Regulator**
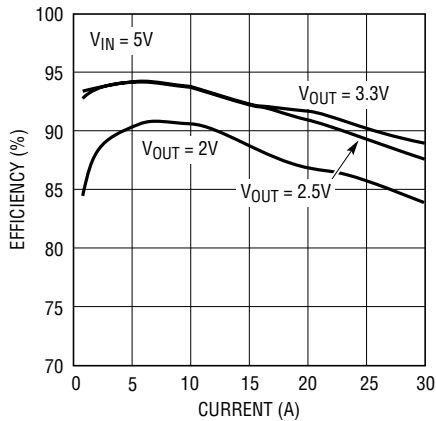
Figure 62. Efficiency of Figure 61's Circuit, $V_{IN}$ = 5V



Figure 64. Transient Response with
10A Load Step (100ns Rise Time)

output capacitor requirements. There is also a fourfold reduction in the total inductor energy storage requirement, which means much smaller inductors and vastly improved transient dynamics. During a large load step, the two channels operate at maximum duty factor in an attempt to maintain the desired output voltage. Both inductor currents slew rapidly and are now additive, since they are going in the same direction. Hence, the slew rate is double what a single channel could do for equal inductor values. However, due to the ripple current cancellation during steady-state conditions, the two inductors can be reduced to approximately one-half the value that a single channel design would require for equal ripple currents. Since during slew they appear to be operating in parallel, the actual slew rate is four times that of a single channel design with equal steady-state output ripple current. Both input and output ripple frequencies are double those of a single-channel design, further simplifying filtering requirements.

## Why Stop at Two?

If two channels are good, aren't more channels better? In a word, yes. In principle, there is no limit to the number of parallel channels that can be added. As the number of channels, $n$, increases, the ripple frequency increases to $n$ times the single-channel frequency. Input and output RMS ripple currents continue to decrease. Diminishing returns are reached as $n$ rises above three. At three stages, the ripple reductions are very substantial and dynamic performance is excellent. Adding more channels produces slight improvements but the dramatic gains will have been realized by $n = 3$. The only real penalty is added complexity.

The bottom line is that PolyPhase designs offer a considerable reduction in the cost and volume of the power devices at the expense of a little added complexity in the control circuitry.



$V_O$ = 2.5V    2µs/DIV
$V_{IN}$ = 5V

Figure 63. Output Ripple with 30A Load



CHANNEL A
5A/DIV

CHANNEL A + B =
TOTAL INPUT
RIPPLE CURRENT,
UNFILTERED

CHANNEL B
5A/DIV

$I_O$ = 15A
f = 306kHz

2µs/DIV
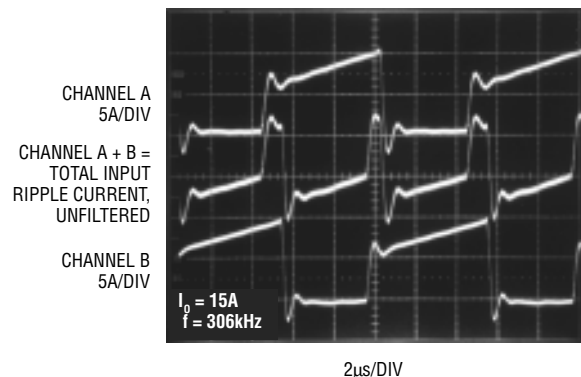
Figure 65. Ripple Cancellation—Input

A + B

5A/DIV

CHANNEL A
CHANNEL B

$I_0 = 25A$

2μs/DIV
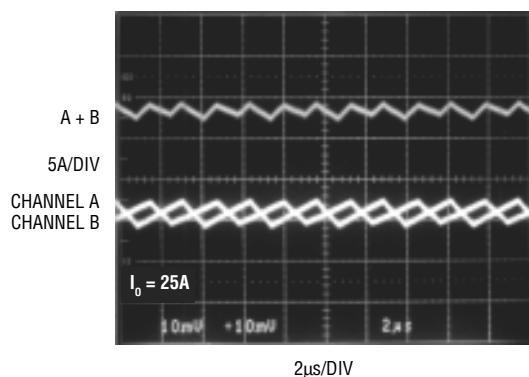
**Figure 66. Ripple Cancellation—Output**

## 2-Phase Design Example

The circuit shown in Figure 61 is a 2-phase, voltage mode–control, synchronous buck regulator designed for a 5V input and output voltages below 3.3V. It is intended to power large memory arrays, ASICs, FPGAs and the like in server and workstation applications. The output is capable of more than 30 amps continuous at outputs of 2.5V and below, with peak current capability of greater than 40 amps. The design is entirely surface mount and the maximum height above the board is 5.5mm. Overall board area is only 4.24 in². Efficiency is excellent, as can be seen in the curve in Figure 62. Output ripple voltage is shown in Figure 63. The circuit's dynamic response to a 10 amp load step is shown in Figure 64. The response is dominated by the output capacitor's ESR and shows the output voltage recovered to the original level in under 10μs. Figures 65 and 66 show how the input and output ripple currents cancel.

## Circuit Operation

The basic design consists of two LTC1430CS8-based synchronous buck regulators connected in parallel and operated 180° out of phase. U4, the CD4047 oscillator, is used to generate the required clock signals and synchro-

nize the two LTC1430s. Unfortunately, simply connecting two regulators in parallel is a recipe for instant disaster. The output voltages of the two regulators will be slightly different due to normal component tolerances. Therefore, the higher output voltage channel will attempt to supply the full load current, while the lower voltage output will sink current from the output in a desperate attempt to reduce the output voltage to where it thinks it should be. The result is like a dog chasing its tail, with large currents running around in a circle and going nowhere.

Op amp U1 solves this problem. Because the two channels are identical, if the output currents are the same, the input currents will be also. Low value sense resistors are included in the input power path to allow the circuit to measure input current. U1 then forces the input current of channel two to match the input current of channel one by making small adjustments in channel two's output voltage. It does this by adding or subtracting a small amount of current from channel two's feedback divider. The two sense resistors are short lengths of PCB trace and only need to be ratiometrically accurate. The absolute value of these resistors is not important (see Linear Technology Application Note 69, Appendix A, for a discussion on how to design trace resistors).

The only remaining trick in the circuit is the role of Q1 and its associated circuitry. At start-up, the LTC1430's clock frequency is slowed down to approximately 10kHz until the output voltage rises to approximately 50% of the desired level. If, during this start-up phase, an attempt is made to synchronize the controller to a very high frequency, the oscillator ramp amplitude never rises to a level sufficiently high to trip the PWM comparator and enable the FET drivers. Therefore, the output gets stuck on ground. Q1 fixes this by forcing the sync signals high during the turn-on transient. Once the output voltage nears its final level, the clock signals are allowed to synchronize the two PWM controllers.

# LTC1622: LOW INPUT VOLTAGE, CURRENT MODE PWM BUCK CONVERTER

*by San-Hwa Chee*

## Introduction

The 8-pin LTC1622 step-down DC/DC controller is designed to help system designers harness all of the available energy from lithium-ion batteries in several ways. Its wide operating input-voltage range (2.0V to an absolute maximum of 10V) and 100% duty cycle allows low dropout for maximum energy extraction from the battery. The part's low quiescent current, 400μA, with a shutdown current of 15μA, extends battery life. Its user-selectable Burst Mode operation enhances efficiency at low load current.

For portable applications where board space is a premium, the LTC1622 operates at a constant frequency of 550kHz and can be synchronized to frequencies of up to 750kHz. High frequency operation allows the use of small inductors, making this part ideal for communications products. The LTC1622 comes in a tiny 8-lead MSOP package, providing a complete power solution while occupying only a small area.

## 2.5V/1.5A Step-Down Regulator

A typical application circuit using the LTC1622 is shown in Figure 67. This circuit supplies a 1.5A load at 2.5V with an input supply between 2.7V and 8.5V. The 0.03Ω sense resistor is selected to ensure that the circuit is capable of
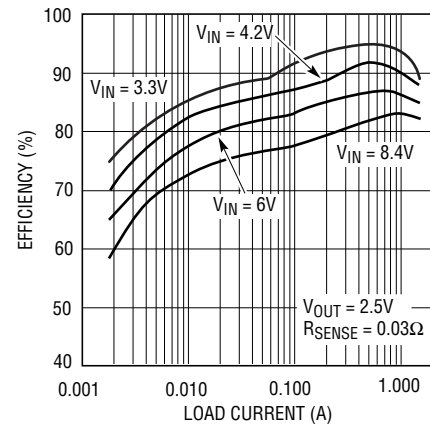
**Figure 68. Efficiency vs Load Current for Figure 67's Circuit (Burst Mode Operation Enabled)**

supplying 1.5A at a low input voltage. In addition, a sublogic threshold MOSFET is used, since the circuit operates at input voltages as low as 2.7V. The circuit operates at the internally set frequency of 550kHz. A 4.7μH inductor is chosen so that the inductor's current remains continuous during burst periods at low load current. For low output voltage ripple, a low ESR capacitor (100mΩ) is used.

## Efficiency Considerations

The efficiency curves for Figure 67's circuit are shown in Figures 68 and 69. Figure 68 shows the efficiency with Burst Mode enabled, whereas Figure 69 has Burst Mode defeated. (Burst Mode is defeated by connecting the SYNC/Mode pin to ground.) Note that, at low load currents, the efficiency is higher with Burst Mode operation. However, constant frequency operation is still achievable

C1: MURATA CERAMIC GRM235Y5V106Z
   (814) 236-1431
C2: SANYO POSCAP 6TPA47M
   (619) 661-6835
L1: MURATA LQN6C-4R7M04
   (814) 237-1431

D1: IR10BQ015
   (310) 322-3331
R2: DALE, 0.25W
   (605) 665-9301
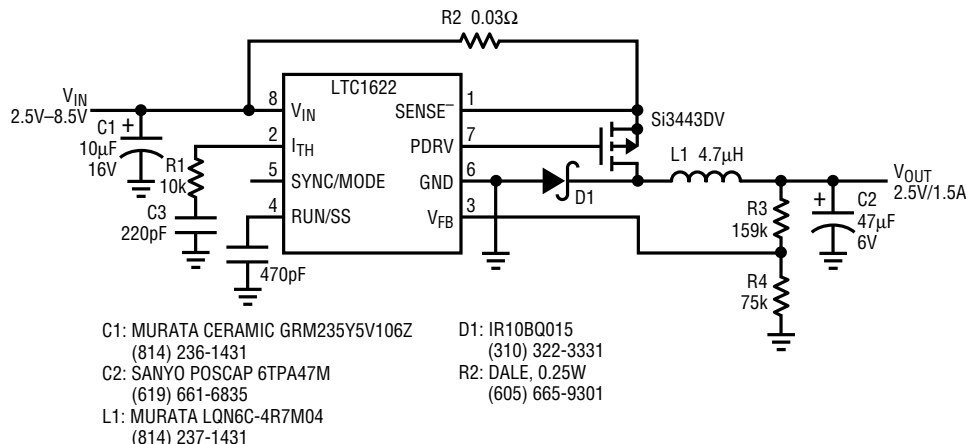
**Figure 67. LTC1622 Typical Application: 2.5V/1.5A Converter**
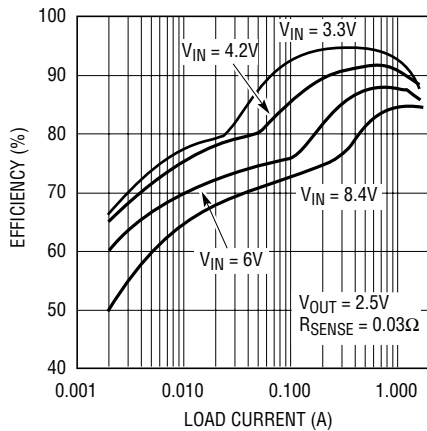
Figure 69. Efficiency vs Load Current for Figure 66's Circuit (Burst Mode Operation Disabled)

at a lower load currents with Burst Mode operation defeated. The kinks in the efficiency curves indicate the transition out of Burst Mode operation.

The components of Figure 67 have been carefully chosen to provide the amount of output power using a minimum of board space. Efficiency is also a prime consideration in selecting the components, as illustrated in Figures 68 and 69. Figures 70 and 71 show the transient response of $V_{OUT}$ with a load step from 50mA to 1.2A. Figure 70 has Burst Mode enabled, while Figure 71 has it defeated. Note that the output voltage ripple (in the middle portion of the photographs) is higher for Burst Mode operation than with Burst Mode disabled at 50mA load current.

Applications that require better transient response can use the circuit in Figure 72, whose components are selected specifically for this requirement. Figures 73 and 74 show the response with and without Burst Mode operation, respectively. Note that the transient response has been



0.1ms/DIV
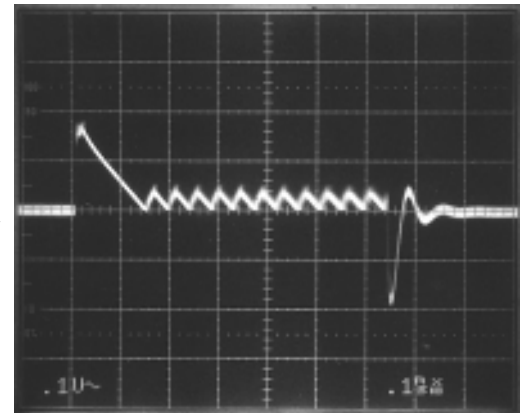
Figure 70. Transient Response with Burst Mode Operation Enabled; Load Step = 50mA to 1.2A

enhanced significantly. However, this comes at the expense of slightly reduced efficiency at low load currents, as indicated by the efficiency curves of Figures 75 and 76.



0.1ms/DIV
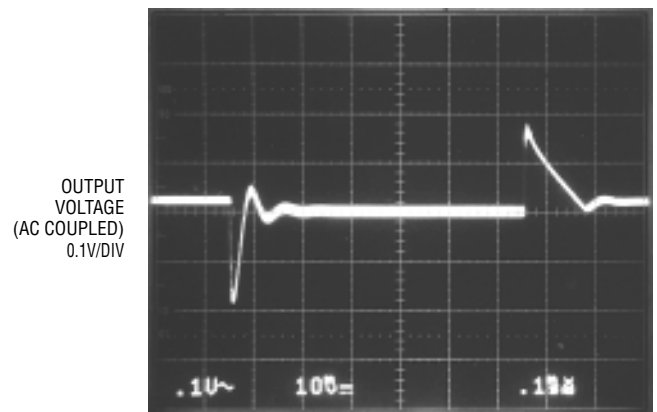
Figure 71. Transient Response with Burst Mode Operation Inhibited; Load Step = 50mA to 1.2A
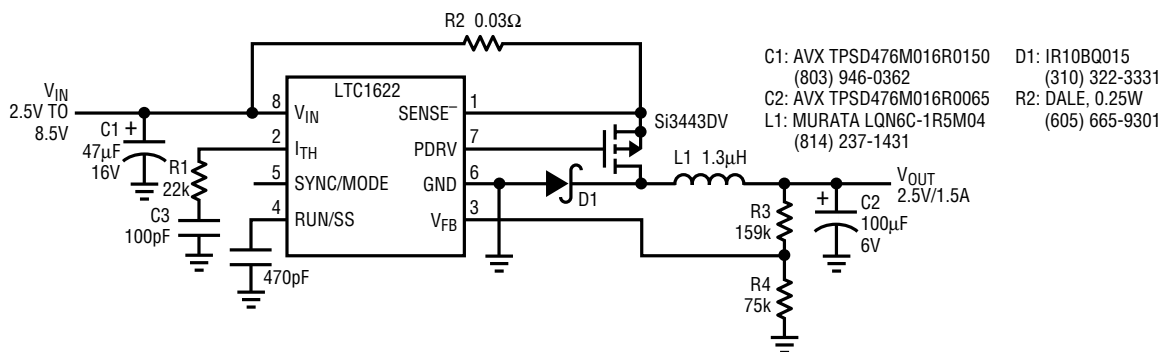


Figure 72. 2.5V/1.5A Converter with Improved Transient Response

OUTPUT
VOLTAGE
(AC COUPLED)
0.1V/DIV

0.1ms/DIV
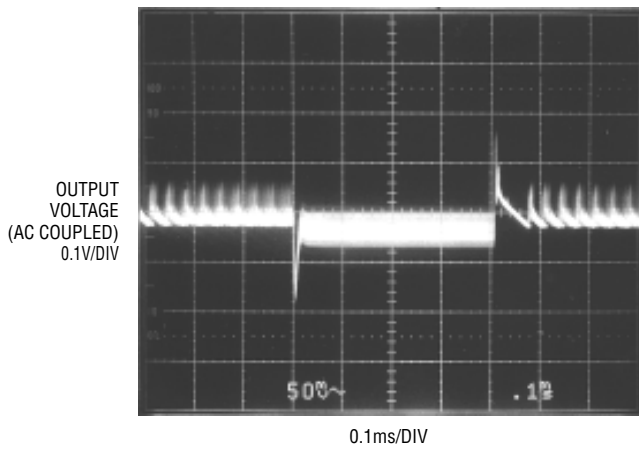
**Figure 73. Transient Response with Burst Mode Operation Enabled; Load Step = 50mA to 1.2A**



OUTPUT
VOLTAGE
(AC COUPLED)
0.1V/DIV

0.1ms/DIV
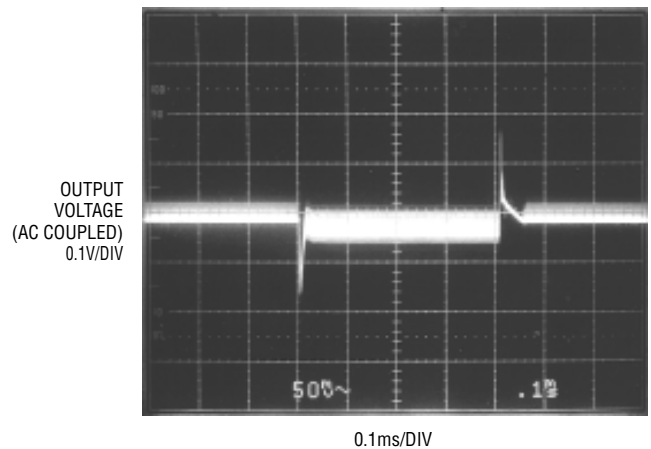
**Figure 74. Transient Response with Burst Mode Operation Inhibited; Load Step = 50mA to 1.2A**
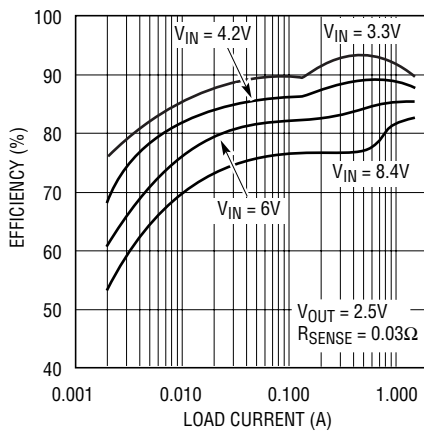


**Figure 75. Efficiency vs Load Current for Figure 72's Circuit (Burst Mode Operation Enabled)**
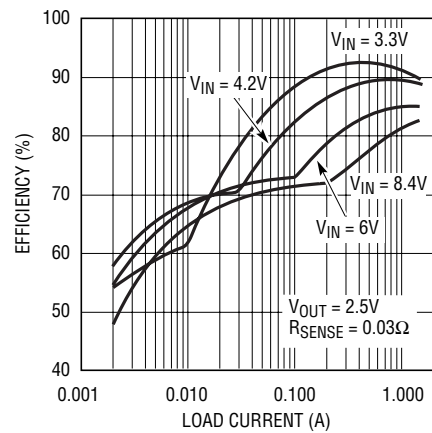


**Figure 76. Efficiency vs Load Current for Figure 72's Circuit (Burst Mode Operation Disabled)**

## WIDE INPUT RANGE, HIGH EFFICIENCY STEP-DOWN SWITCHING REGULATORS
*by Jeff Schenkel*

### Introduction

The LT1676/LT1776 are current mode switching regulator ICs optimized for high efficiency operation in high input voltage, low output voltage buck topologies. These two parts are pin-for-pin compatible and virtually identical in operation, the only difference being their internal oscillator frequencies—100kHz for the LT1676 vs 200kHz for the LT1776. They operate in a fixed frequency mode (as opposed to constant off-time or on-time, for instance) and can be externally synchronized to a higher switching frequency.

The internal output switch is rated at a nominal peak current of 700mA, which typically accommodates DC output currents of up to 500mA. The input voltage range is 7.4V to 60V. Maintaining acceptable efficiency in the upper half of this input voltage range requires very fast output-switch edge rates. The LT1676/LT1776 contain specialized output circuitry to deliver this performance. Additionally, they contain circuitry to monitor output load level and reduce leading-edge switch rate (turn-on) when the output load is light. This arrangement helps avoid pulse skipping at light load, with its consequent subharmonic behavior.

True current mode operation is supported, with all its well known advantages for switching regulator operation. The shutdown pin implements a pair of functions. Pulling it down to near ground turns off the part almost completely and reduces the quiescent current to a few tens of micro-amperes. The second shutdown pin function acts at a threshold of roughly 1.25V. Below this level, the part operates normally, except that output switching action is inhibited. This allows the implementation of an undervoltage lockout function set by, for instance, an external resistor divider. The LT1676/LT1776 are available in both 8-pin SO and PDIP packages.

### Applications

#### *Minimum Component-Count Application*

Figure 77 shows a basic "minimum component count" application using the LT1676. The circuit produces 5.0V at up to 500mA $I_{OUT}$ with input voltages in the range of 12V to 48V. The typical $P_{OUT}/P_{IN}$ efficiency is shown in Figure 78. No pulse skipping is observed down to zero external load. (The several milliamperes drawn by the $V_{CC}$ pin acts as a sufficient preload.) As shown, the SHDN and SYNC pins are unused, however either (or both) can be optionally driven by external signals as desired.



C1: PANASONIC HFQ
(201) 348-7522
C2: AVX D CASE TPSD107M010R0080
(803) 946-0362
C4, C5: X7R OR COG/NPO
D1: MOTOROLA 100V, 1A, SMD SCHOTTKY
(800) 441-2447
L1: COILCRAFT DO3316P-224
(847) 936-6400

FOR 3.3V $V_{OUT}$ VERSION:
R1: 24.3k, R2: 14.7k
L1: 150µH, DO3316P-154
$I_{OUT}$: 0mA TO 500mA
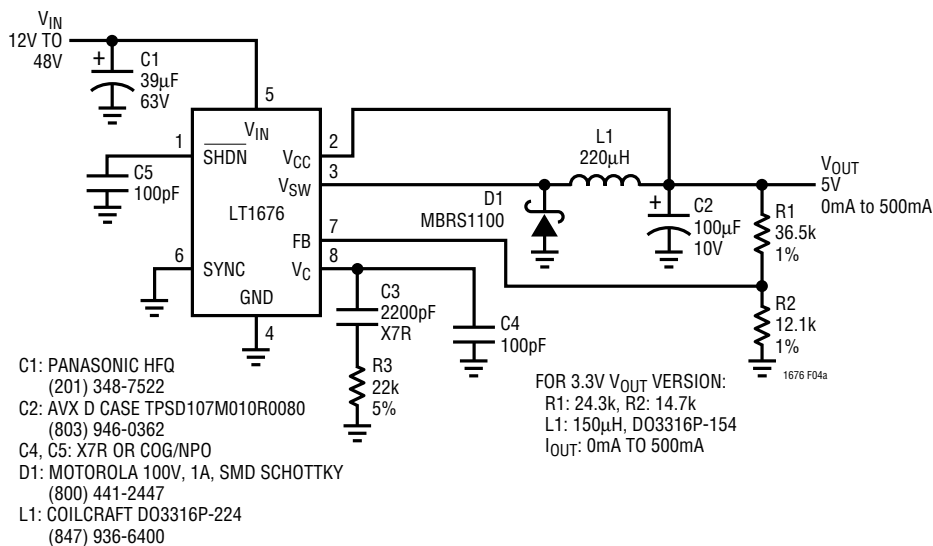
**Figure 77. Minimum Component-Count Application**
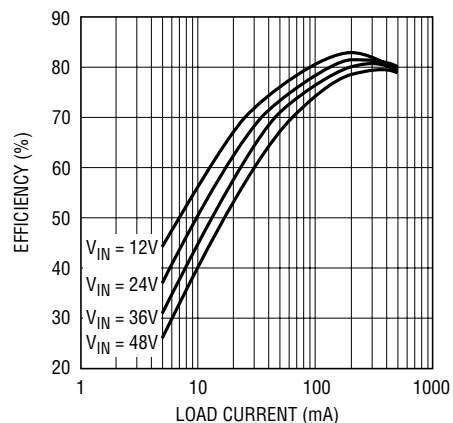


**Figure 78. Efficiency of Figure 77's Circuit**
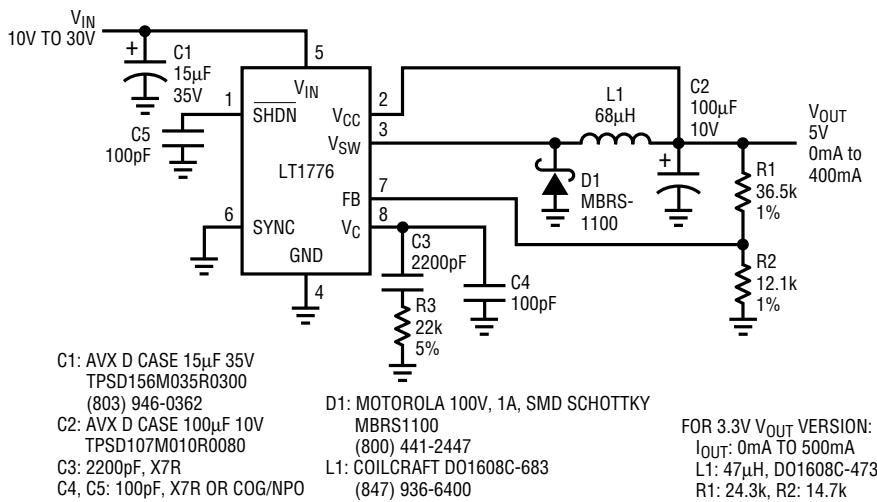
Figure 79. Minimum PC Board Application
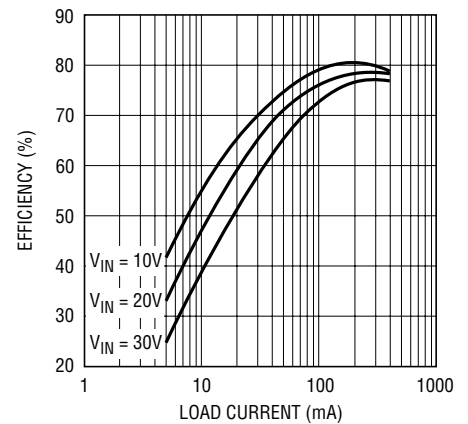


Figure 80. Efficiency of Figure 79's Circuit

### Minimum PC Board Area Application

The previous application example used the LT1676 to demonstrate simultaneously the maximum input voltage and output current capability. As such, the input bypass capacitor choice was a high frequency aluminum electrolytic type, rated to 63V. Also, the 100kHz switching rate of the LT1676 requires an inductor of about 220μH. The DO3316 device size was chosen to support the output current requirements. However, both of these components are physically large.

The application example in Figure 79 shows a circuit that is much smaller physically than the previous minimum component count application. The nominal 200kHz switching frequency of the LT1776 allows the use of a physically smaller 68μH inductor—a Coilcraft DO1608C-683. This inductor will support output current to 400mA at 5V. However, the part is incapable of withstanding an indefinite short circuit to ground. (Momentary shorts of a few seconds or less can still be tolerated.) Additionally, the bulky aluminum electrolytic capacitor previously on $V_{IN}$ has been replaced by a compact 35V-rated tantalum type. The result is a postage-stamp-sized circuit with efficiency as shown in Figure 80.

### Burst Mode Application

The minimum component count application demonstrates that power supply efficiency degrades with lower output

load current. This is not surprising, as the LT1676 itself represents a fixed power overhead. A possible way to improve light load efficiency is to use Burst Mode™ operation.

Figure 81 shows the LT1676 configured for Burst Mode operation. Output voltage regulation is now provided in a "bang-bang" digital manner, via comparator U2, an LTC1440. Resistor divider R4/R5 provides a scaled version of the output voltage, which is compared against U2's internal reference. Intentional hysteresis is set by the R6/R7 divider. As the output voltage falls below the regulation range, the LT1676 is turned on. The output voltage rises and, as it climbs above the regulation range, the LT1676 is turned off. Efficiency is maximized as the LT1676 is only powered up while it is providing heavy output current. Figure 82 shows that efficiency is typically maintained at 75% or better down to a load current of 10mA. Even at a load current of 2mA, efficiency is still a respectable 65% to 75% (depending on $V_{IN}$).

Resistor divider R1/R2 is still present, but does not directly influence output voltage. It is chosen to ensure that the LT1676 delivers high output current throughout the voltage regulation range. Its presence is also required to maintain proper short-circuit protection. Transistors Q1 and Q2 and resistor R7 form a high $V_{IN}$, low quiescent current voltage regulator to power U2.
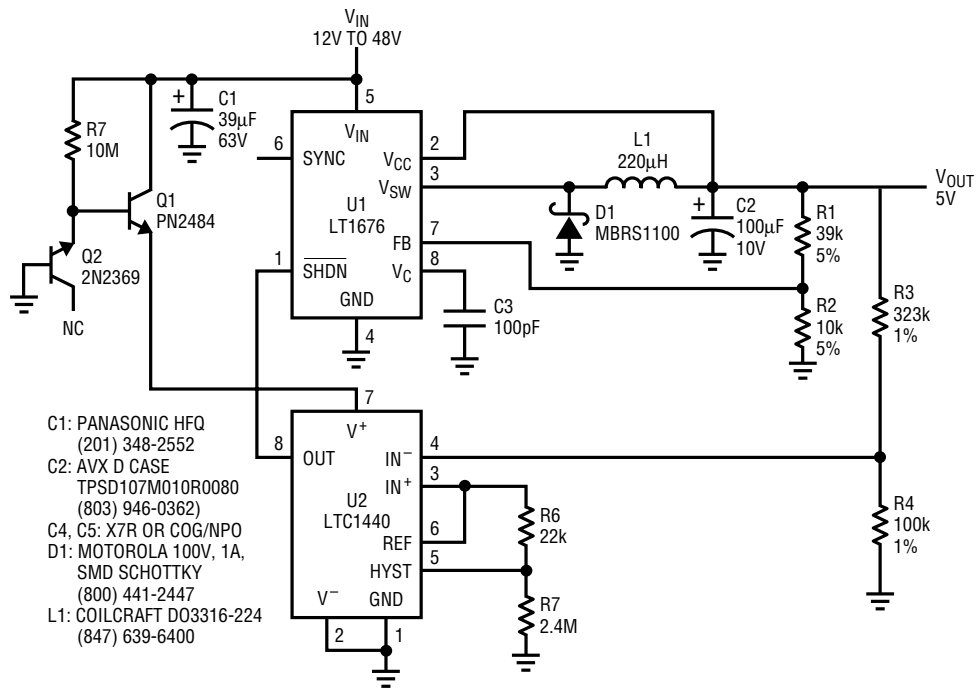
**Figure 81. Burst Mode Operation Configuration**

C1: PANASONIC HFQ
(201) 348-2552
C2: AVX D CASE
TPSD107M010R0080
(803) 946-0362)
C4, C5: X7R OR COG/NPO
D1: MOTOROLA 100V, 1A,
SMD SCHOTTKY
(800) 441-2447
L1: COILCRAFT DO3316-224
(847) 639-6400

## Battery Charger Application

Figure 83 shows the LT1776 configured as a constant-current/constant-voltage battery charger. An LT1620 rail-to-rail current sense amplifier (U2) monitors the differential voltage across current sense resistor R4. As this equals and exceeds the voltage across resistor R5 in the R5/R6 divider, the LT1620 responds by sinking current at its $I_{OUT}$ pin. This is connected to the $V_C$ control node of the LT1776 and therefore acts to reduce the amount of power delivered to the load. The overall constant-current/constant-voltage behavior can be seen in Figure 84.
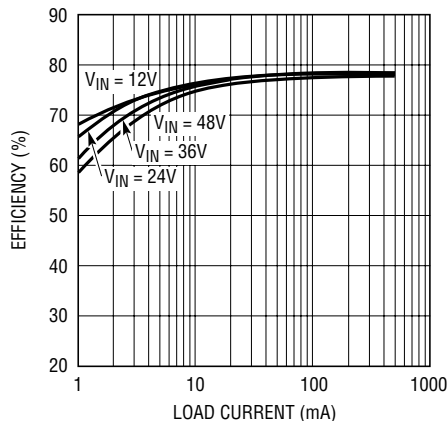


**Figure 82. Efficiency of Figure 81's Circuit**

Target voltage and current limits are independently programmable. The output voltage of 7.2V, which corresponds to the charging voltage of a 3-cell lead-acid battery, is set by the R1/R2 divider and the internal reference of the LT1776. Output current, presently 200mA, is set by current sense resistor R4 and the R5/R6 divider. (A 16-pin version of the LT1620 that implements end-of-cycle detection is also available. This is useful for implementing lead-acid battery "top-off" charger behavior or the like. See the LT1620 data sheet for further information.)

The circuit as shown accommodates an input voltage range of 11V to 30V. The upper input voltage limit of 30V is determined not by the LT1776, but by the LT1121-5 regulator (U3). (A regulated 5V is required by the LT1620.) This regulator was chosen for its micropower behavior, which helps maintain good overall efficiency. However, the basic catalog part is only rated to 30V. Substitution of the industry standard LM317, for example, extends the allowable input voltage to 40V (or more with the HV version), but its greater quiescent current drain degrades efficiency from that shown.
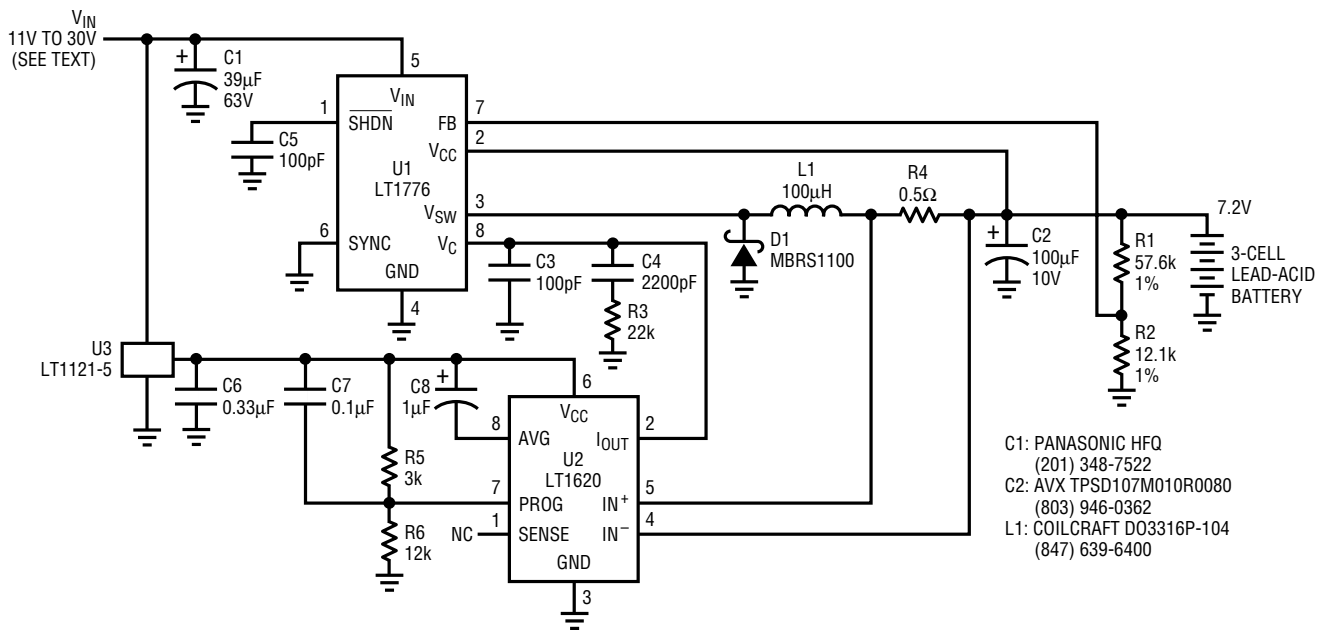
**Figure 83. Wide V$_{IN}$ Range, High Efficiency Battery Charger**

### Dual Output SEPIC Converter

All of the previous applications provide a single positive output voltage. Real world situations often require dual supply voltages. The SEPIC topology (single-ended primary inductance converter) offers a cost-effective way to simultaneously generate a negative voltage with a single piece of magnetics. The circuit in Figure 85 uses an LT1776 to generate both positive and negative 5V. The two inductors shown are actually just two windings on a standard Coiltronics inductor. Capacitor C3 creates the SEPIC topology, which improves regulation and reduces ripple current in L1.

For the best negative supply voltage regulation, this output should have a preload of at least 1% of the maximum positive load. Total available current from both outputs is limited to 500mA. Maximum negative supply current is limited by the positive 5V load. A typical limit is one-half of the positive current, but a more exact calculation includes the input voltage. For this and further details of this topology, see Linear Technology Design Note 100.

### Positive-to-Negative Converter

The previous example used a dual inductor to create a pair of output voltages, one positive and the other negative. The positive-to-negative converter topology illustrated in Figure 86 generates a single negative output voltage from a positive input voltage, using just an ordinary inductor. The topology is somewhat similar to the original step-down arrangement, but the inductor is grounded and the LT1776 ground is now referred to the negative output voltage. Note that the integrated circuit must now be rated
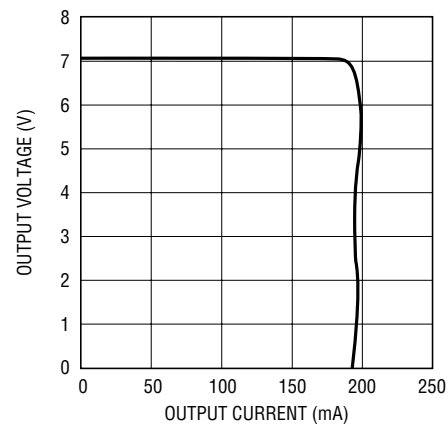


**Figure 84. Battery Charger Output Voltage vs Output Current for Figure 83's Circuit**

for the worst case sum of the input voltage plus the absolute value of the output voltage. The relatively high input voltage rating of the LT1676/LT1776 parts along with their good efficiency under such conditions make them an excellent choice for implementing this topology. The circuit as shown converts an input voltage in the range of 10V to 28V to a –5V output. Available output current is 300mA at the worst case $V_{IN}$ of 10V.

The user should exercise caution in modifying this circuit for other applications. The positive-to-negative topology is not as straightforward as the step-down topology. It is actually more like a flyback topology, in that current is delivered to the output in discrete pulses. The output capacitor must supply the entire load current for at least a portion of the switching cycle, so output capacitor ripple current rating and ESR may be an issue. Maximum available output current will usually be a strong function of input voltage. Supporting low $V_{IN}$-to-$V_{OUT}$ ratios may require additional components for maintaining control-loop stability. A detailed theoretical analysis of this topology and its behavior can be found in Linear Technology Application Note 44.
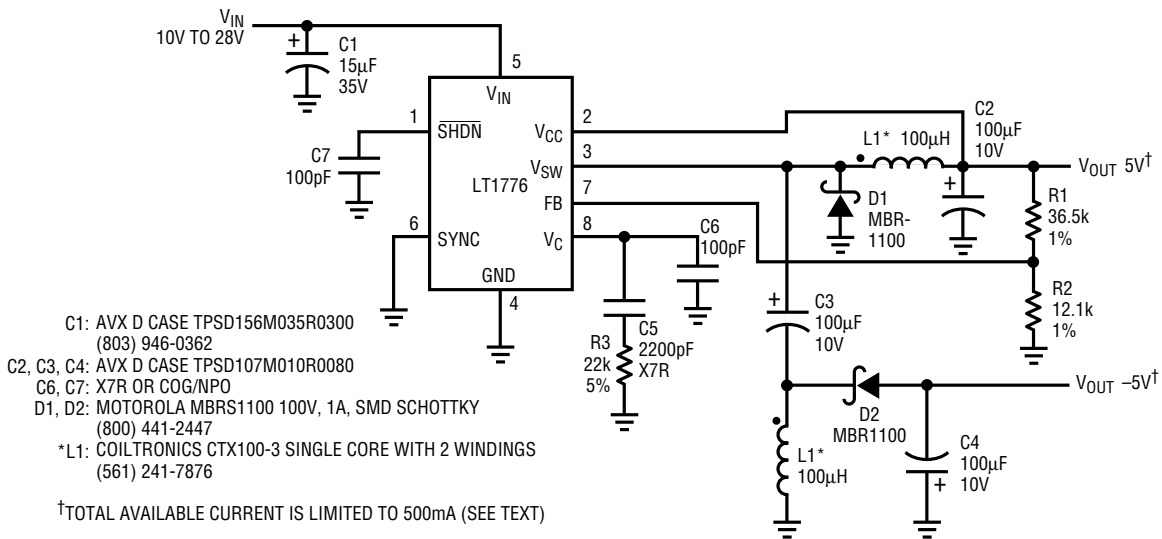


**Figure 85. Dual-Output SEPIC Converter**
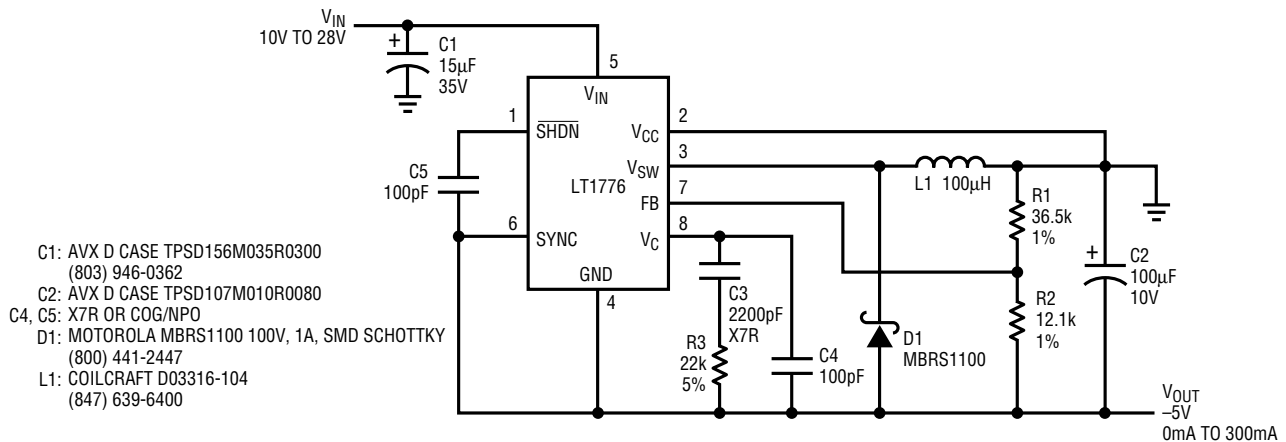


**Figure 86. Positive-to-Negative Converter**

## Regulators—Switching (Boost)

### ±12 VOLT OUTPUT FROM THE LT1377
*by John Seago*

Many applications use positive and negative voltages, with only one voltage requiring tight regulation. Often, cost and board space are more important than regulation of the second output. An equal output of opposite polarity can be added to a boost configuration by means of a negative charge pump. This two-output configuration is shown in Figure 87. The 1MHz switching frequency of the LT1377 decreases required board space, and the availability of both positive and negative feedback amplifiers allows regulation of either positive or negative output.

In the circuit of Figure 87, the LT1377 with L1, D1, D2 and C6 make up a positive boost circuit. As the internal power switch in the IC turns on, the voltage at pin 8 goes low and energy is stored in inductor L1. When the power switch turns off, L1 transfers energy through diodes D1 and D2 to capacitor C6 and the positive output load. C6 supplies load current when the power switch is on. Resistors R2 and R3 provide feedback from the positive output. R1, C3 and C4 provide loop compensation. C1 is the input capacitor and C2 provides local decoupling for the IC.

The charge pump consists of two capacitors, two diodes and a small inductor. When the power switch turns off, L1 also replenishes the charge on C5, forward biasing D3. When the power switch turns on, the charge on C5 reverse biases D3, forward biases D4 and supplies energy to C7 and the negative output load. L2 attenuates capacitive current spikes. D2 was added so that the voltage drop across both D1 and D2 would be approximately equal to the sum of the voltage drops of D3, D4 and the saturation voltage of the power switch in the LT1377. This makes both output voltages approximately equal but opposite in polarity. D1 and D2 can be replaced with a single Schottky diode if equal outputs are not required.

Voltage and current waveforms of the internal power switch are shown in Figure 88. These measurements were taken at pin 8 of the LT1377 with the circuit powered from a 5V supply. Figure 89 shows the ripple voltage from each output. The high frequency spikes can be attenuated with a small LC filter if necessary.

The circuit of Figure 87 was intended to operate from a 5V supply and provide ±12V outputs at 100mA each. It operates over an input range of 4V to 10V and load current variations from 15mA to 100mA. The regulated positive output voltage remains constant for changes in the input



C1 = C5 = 1E106ZY5U-C205-M, TOKIN (408) 432-8020
C6 = C7 = 1E225ZY5U-C203. TOKIN (408) 432-8020
L1 = CTX10-1P, COILTRONICS (407) 241-7876
L2 = PM20-R047M, GARRETT (805) 922-0594
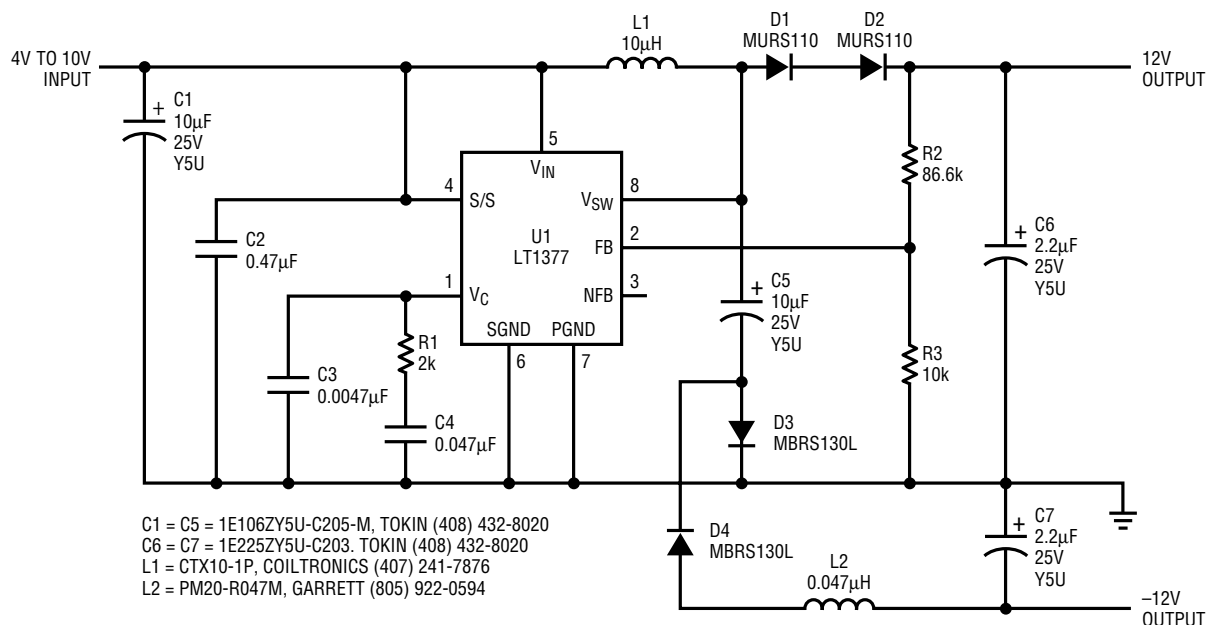
**Figure 87. Positive Output Regulated Supply**

0.5μs/DIV
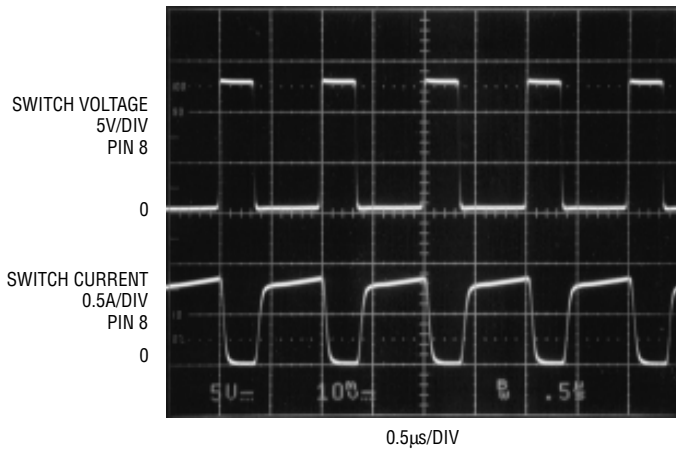
**Figure 88. Switch Voltage and Current Waveforms**



0.5μs/DIV

**Figure 89. Output Ripple Voltage**



**Figure 90. Unregulated Negative Output
Voltage with Positive Output Voltage Regulated**
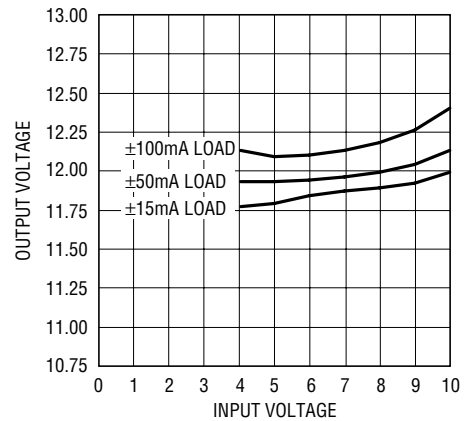


**Figure 91. Unregulated Positive Output Voltage
with Negative Output Voltage Regulated**



C1 = C5 = 1E106ZY5U-C205-M, TOKIN (408) 432-8020
C6 = C7 = 1E225ZY5U-C203, TOKIN (408) 432-8020
L1 = CTX10-1P, COILTRONICS (407) 241-7876
L2 = PM20-R047M, GARRETT (805) 922-0594
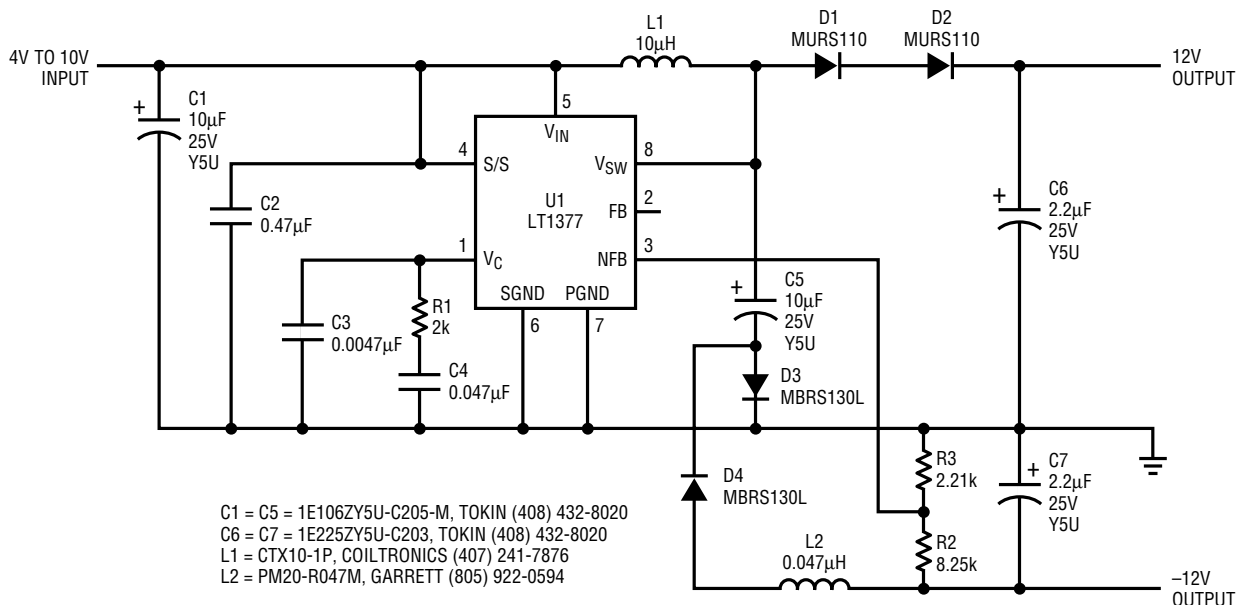
**Figure 92. Negative Output Regulated Dual Supply**

voltage and load current, while the voltage of the unregulated negative output changes as shown in Figure 90. Line and load regulation of the unregulated output will improve with smaller changes of input voltage or load current.

A common requirement is for the positive output to regulate the majority of power while the negative output supplies a much smaller, unregulated bias current. Measurements taken on the test circuit of Figure 87 showed the unregulated –12V output had less than ±1% variation for a fixed 15mA load while the input voltage changed from 4V to 10V with a load current change of 15mA to 200mA on the regulated positive output.

Occasionally, it is more important to regulate the negative output than the positive output. The circuit in Figure 92 is the same as that shown in Figure 87, except feedback resistors R2 and R3 have different values and provide feedback from the negative output to the negative feedback amplifier of the LT1377. Figure 91 shows the variation in unregulated positive output for input voltage and load current variations.

## THE LT1370: NEW 500kHz, 6A MONOLITHIC BOOST CONVERTER
*by Karl Edwards*

### Introduction

The LT1370 is a 500kHz, 6A boost converter. At 65mΩ on-resistance, 42V maximum switch voltage and 500kHz switching frequency, the LT1370 can be used in a wide range of output voltage and current applications.

The high efficiency switch is included on the die, along with the oscillator, control and protection circuitry necessary for a complete switching regulator. This part combines the convenience and low parts count of a monolithic solution with the switching capabilities of a discrete power device and controller. The LT1370, features cur-

rent mode operation, external synchronization and low current shutdown mode (12µA typical). Only a few surface mount components are needed to complete a small, high efficiency DC/DC converter. The LT1370 will operate in all the standard switching configurations, including boost, buck, flyback, forward, inverting and SEPIC.

### 5V to 12V Boost Converter

Figure 93 shows a typical 5V to 12V boost application. The feedback divider network has been selected to give the desired output voltage. As long as R2 is less than 7k, FB input bias current can be ignored. The inductor needs to be chosen carefully to meet both peak and average current values. The output capacitor can see high ripple currents—often, as in this application, higher than the ripple rating of a single capacitor. This requires the use of two surface mount tantalums in parallel; both capacitors should be of the same value and manufacturer. The input capacitor does not have to endure such high ripple currents and

| L1 | I$_{OUT}$ |
|---|---|
| 4.7µH | 1.8A |
| 10µH | 2.0A |
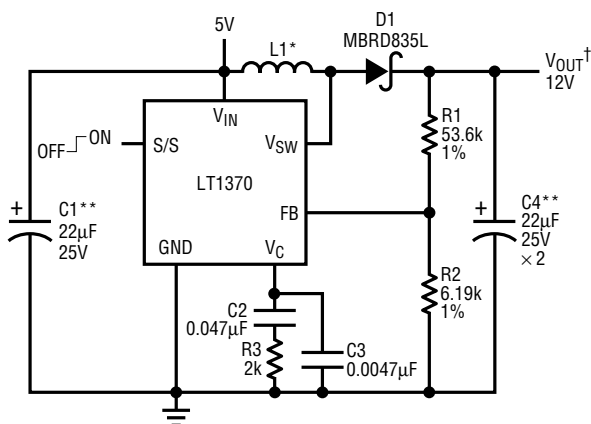
*COILTRONICS
(561) 241-7876
UP2-4R7 (4.7µH)
UP4-100 (10µH)
**AVX TPSD226M025R0200

†MAX I$_{OUT}$

**Figure 93. 5V to 12V Boost Converter**
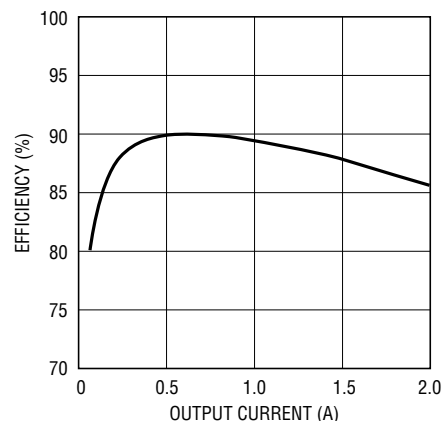
**Figure 94. 12V Output Efficiency**
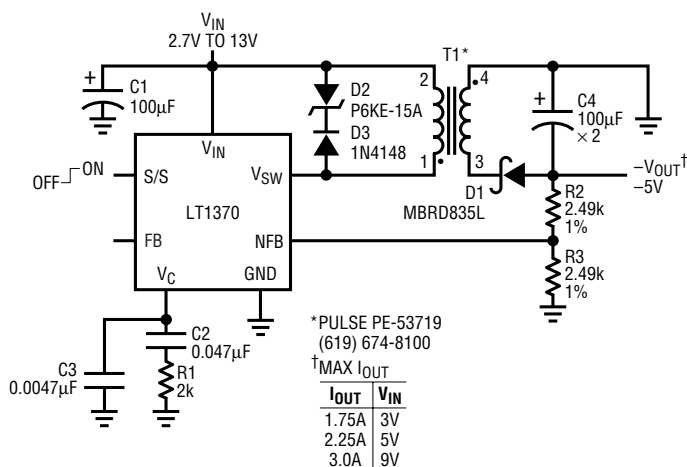
**Figure 95. Positive-to-Negative Converter with Direct Feedback**

V_IN 2.7V TO 13V

C1 100µF

OFF ON S/S

V_IN

LT1370

FB

V_C GND

V_SW

NFB

D2 P6KE-15A
D3 1N4148

T1*

MBRD835L
D1

C4 100µF ×2

−V_OUT† −5V

R2 2.49k 1%

R3 2.49k 1%

C2 0.047µF

C3 0.0047µF

R1 2k

*PULSE PE-53719
(619) 674-8100
†MAX I_OUT

| I_OUT | V_IN |
|-------|------|
| 1.75A | 3V |
| 2.25A | 5V |
| 3.0A | 9V |

a single capacitor will normally suffice. The catch diode, D1, must be rated for the output voltage and average output current. The compensation capacitor, C2, normally forms a pole with the internal $g_m$ of the part in the 2Hz to 20Hz range. It also creates a zero in conjunction with series resistor R3, at 1kHz to 5kHz.

A second capacitor, C3, is sometimes required to prevent erratic switching. Ripple current in the output capacitor's ESR causes voltage ripple. This feeds back through the error amp to the $V_C$ pin, changing the current-trip threshold cycle-to-cycle. The problem appears as subharmonic oscillation. Adding C3, typically one-tenth the value of the main compensation capacitor, reduces the loop gain at the switching frequency, preventing the oscillation.

The ground return from the compensation network must be separate from the high current switch ground. If drops in the ground trace due to switch current cause the $V_C$ pin to dip, premature switch-off will occur. This effect appears as poor load regulation. A solution to this is to return the compensation network to the FB pin. The S/S pin in this example is driven by a logical on/off signal, a low input forcing the LT1370 into its 12µA shutdown mode. Figure 94 shows the overall converter efficiency. Note that peak efficiency is over 90%; efficiency stays above 86% at the device's maximum operating current.

## Positive-to-Negative Converter

The NFB (negative feedback) pin allows negative output regulators to be designed with direct feedback. In the circuit shown in Figure 95, a 2.7V to 13V input, −5V output converter, the output is monitored by the NFB pin and a simple divider network. No complex level shifting or unusual grounding techniques are required. The regular FB pin is left open circuit and the divider network, R2, R3, is calculated based on the −2.49V NFB reference voltage and 30µA of input current. The switch-clamp diodes, D2 and D3, prevent the leakage spike from the transformer, T1, from exceeding the switch's absolute maximum voltage rating. The Zener voltage of D2 must be higher than the output voltage, but low enough that the sum of input voltage and clamp voltage does not exceed the switch voltage rating.

## 5V SEPIC Converter

Figure 96 shows a SEPIC converter. One of the advantages of the SEPIC topology is that the input voltage can range from below to above the output voltage. In Figure 96, the input voltage range is from 4V to 9V, with a 5V output. The magnetic coupling of inductors L1A and L1B is not critical for operation, but generally they are wound on the same core. C2 couples the inductors together and eliminates the

V_IN 4V TO 9V

OFF ON S/S

V_IN

LT1370

GND V_C

V_SW

FB

L1A* 6.8µH

C2 4.7µF

D1 MBRD835L

R2 18.7k 1%

V_OUT† 5V

C3 100µF 10V ×2

L1B* 6.8µH

R3 6.19k 1%

C1 33µF 20V

R1 2k

C4 0.047µF

C5 0.0047µF

C1 = AVX TPSD 336M020R0200
C2 = TOKIN 1E475ZY5U-C304
C3 = AVX TPSD107M010R0100
* BH ELECTRONICS 501-0726
(612) 894-9590
** INPUT VOLTAGE MAY BE GREATER OR
LESS THAN OUTPUT VOLTAGE

†MAX I_OUT

| I_OUT | V_IN |
|-------|------|
| 1.8A | 4V |
| 2A | 5V |
| 2.6A | 7V |
| 2.9A | 9V |
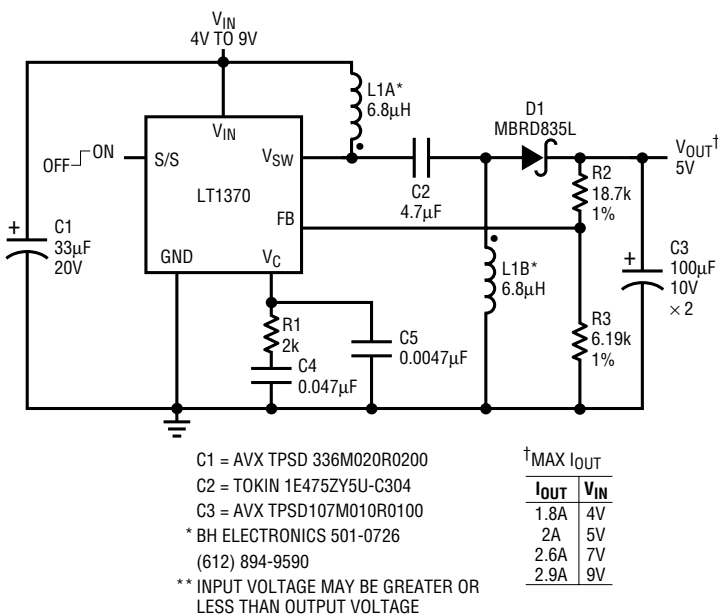
**Figure 96. Two Li-Ion Cells to SEPIC Converter**

need for a switch snubber network. C2 must have a very low ESR, because the ripple current is equal to $I_{SW}/2$. Its capacitance value is not critical and has no significant effect on loop stability. The voltage across C2 is equal to

the input. A 4.7µF, 50V ceramic will work in most SEPIC applications. The S/S pin is used as a logical on/off signal. In the off state, there is no leakage to the output, and only 12µA leakage from the input.

## BOOTSTRAPPED SYNCHRONOUS BOOST CONVERTER OPERATES AT 1.8V INPUT
*by Tom Gross*

Some applications, such as those powered by batteries or solar cells, see their input voltage decrease as they operate. Many regulators that could operate with high input voltages cease to function as the input voltage decreases. The circuit in Figure 97 maintains the maximum load current as the input voltage drops. The regulator boosts a 2.5V–4.2V input to 5V at a maximum load current of 2A (10W of output power).

The circuit is a bootstrapped synchronous boost regulator using an LTC1266 synchronous regulator controller. Diodes D1 through D4 allow the circuit to start-up using the (low) input voltage and then to be powered during normal operation by the higher output voltage. The crucial elements in this circuit are the switches: two IRF7401 N-channel MOSFETs. These MOSFETs are fully enhanced at very low gate-to-source voltages (at 2V of $V_{GS}$, the peak drain current is rated at 15A). The low enhancement



**Figue 98. Efficiency of Figure 97's Circuit**

voltages allow the circuit to start up at low input voltages (crucial for low series-cell-count, battery-powered applications). Diodes D3 and D4, along with capacitor C2, form a charge-pump circuit, which the controller uses for the MOSFETs' gate drive. The switches are driven by an LTC1266 synchronous regulator controller.

Because the circuit is powered from the 5V output, it will still operate if the input supply voltage drops below the minimum input voltage of the IC. This bootstrapping



**Figure 97. Bootstrapped Synchronous Boost Converter**

allows the circuit to start up even when the input voltage is below the minimum input voltage of the IC (3.5V). With a 1A load, the regulator operates down to 1.8V.

Figure 98 shows the efficiency of the regulator versus the input voltage at three different load currents. At 2A of load current, efficiency drops as the input voltage is decreased due to the higher power losses in the inductor. A larger inductor will increase efficiency and/or allow for larger load currents. The efficiency with the indicated inductor is good, averaging above 83% overall. Higher efficiency will help to increase the run time of battery-powered applications.

## Regulators (Switching)—Buck-Boost

### 500kHz BUCK-BOOST CONVERTER NEEDS NO HEAT SINK
*by Mitchell Lee*

Thanks to an efficient 0.25Ω switch, the LT1371 SEPIC converter shown in Figure 99 operates at full power with no heatsink. Up to 9W at 5V output is available, and the circuit works over a wide range of input voltages extending from the LT1371's 2.7V minimum to 20V, limited by the rating of the capacitors.

A 1:1 bifilar-wound toroid is used as the magnetic element. A careful analysis showed that, in spite of the 500kHz operating frequency, a high permeability ($m_r$ = 125) Magnetics Inc. Kool Mμ® core exhibited the best efficiency when compared to powdered iron materials. Copper loss is minimized by the use of the high-perm Kool Mμ material, with only a slight core-loss penalty.



Figure 100. Maximum Avialable Output Current

Maximum available output current varies with input voltage, and is shown (for 3A peak switch current) in Figure 100. Efficiencies for several input voltages are shown in Figure 101. At a 2.7V input, most of the loss is tied up in the LT1371 switch, whereas the output diode is the dominant source of loss with high inputs. Because these losses are small, surface mount construction provides adequate dissipation, eliminating the need for heat sinks.

In this application, the synchronization feature of the LT1371 is not used. When driven with an external clock at the shutdown/sync pin (S/S), the chip can be synchronized to any frequency between 600kHz and 800kHz.



**Figure 99. 5V, 9W Converter Operates Over a Wide Input Range with Good Efficiency**
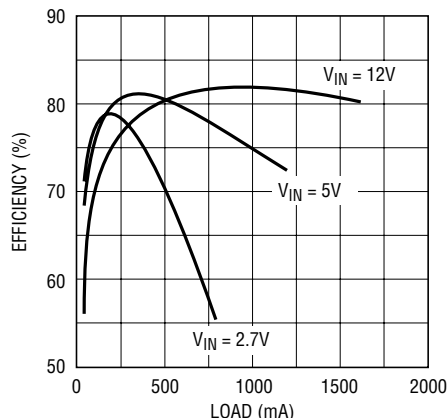


**Figure 101. Efficiency of Figure 99's Circuit**

## BATTERY-POWERED BUCK-BOOST CONVERTER REQUIRES NO MAGNETICS
*by John Seago*

One of the problems that designers of portable equipment face is generating a regulated voltage that is between the charged and discharged voltage of a battery pack. As an example, when generating a 3.3V output from a 3-cell battery pack, the regulator input voltage changes from about 4.5V at full charge to about 2.7V when discharged. At full charge, the regulator must step down the input voltage, and when the battery voltage drops below 3.3V, the regulator must step up the voltage. The same problem occurs when a 5V output is required from a 4-cell input voltage that varies from about 3.6V to 6V. Ordinarily, a flyback or SEPIC configuration is required to solve this problem.

The LTC1515 switched capacitor DC/DC converter, can provide this buck-boost function for load currents up to 50mA with only three external capacitors. The circuit shown in Figure 102 will provide a regulated 3.3V output from a 3-cell input or a 5V output from a 4-cell input. Connecting the 5/3 pin to $V_{IN}$ will program the output to
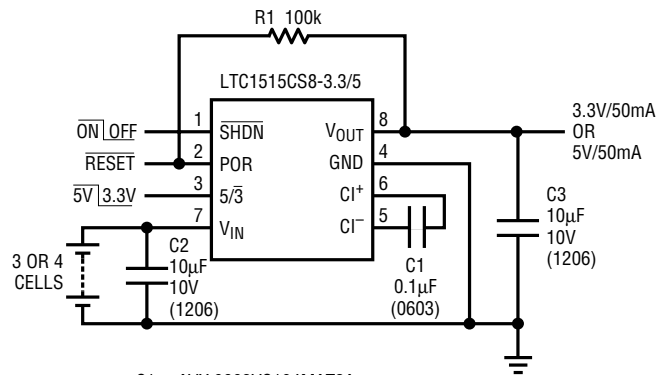


C1 = AVX 0603YC104MAT2A
C2, C3 = TAIYO YUDEN LMK316F106ZL

**Figure 102. Battery-Powered Buck-Boost Converter**

5V, whereas grounding the 5/3 pin programs the output to 3.3V.

The absence of bulky magnetics provides another benefit: this circuit requires only 0.07 square inches of board space in those applications where components can be mounted on both sides of the board. The addition of R1 provides a power-on-reset flag that goes high 200ms after the output reaches 93.5% of its programmed value. The SHDN pin allows the output to be turned on or off with a 3V logic signal.

Regulators—Switching (Inverting)

## MAKING –5V 14-BIT QUIET
*by Kevin R. Hoskins*

Many high performance data acquisition systems reap multiple benefits when using ±5V supplies rather than a single 5V supply. These benefits include the ability to handle larger signal magnitudes than is possible with a single 5V supply. This increases a system's dynamic range and helps improve the signal-to-noise ratio. Operating on ±5V also increases headroom, which is important for signal conditioning. Compared to operating on 5V, conditioning circuitry operating on ±5V has twice the headroom, allowing it to easily handle ±2.5V signals without clipping. Additionally, the greater headroom avoids the limitations of rail-to-rail operation and widens the selection of high performance operational amplifiers and analog-to-digital converters, such as the LTC1419.

Although a switching or charge-pump power supply is an efficient way to create a –5V supply from a single 5V

supply, they are not generally recommended for use with ADCs. Typical ADCs have inadequate PSRR, which decreases with increasing frequency. This poor PSRR performance cannot sufficiently attenuate the noise created by switching or charge-pump supplies. However, LTC's
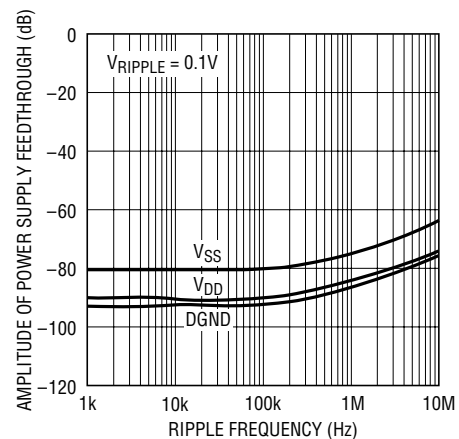


**Figure 103. The LTC1419's Positive Supply PSRR of 90dB to 200kHz is a Significant Contributor to this ADC's Wideband Conversion Performance and 80dB SINAD**

new family of ADCs, here represented by the LTC1419, has excellent PSRR. This family makes it easy to achieve high performance data conversion, even at 14 bits, using a switch-mode regulator for a –5V supply.

The LTC1419's high PSRR is shown in Figure 103. It shows that when operating on ±5V, the negative and positive PSRR are typically 80dB and 90dB, respectively, up to 200kHz for a 100mV ripple voltage. Combined with proper layout, the LTC1419's high PSRR allows it to convert signals without signal degradation while using switching regulators and charge pumps to generate its –5V supply. Applications including high speed communications, high resolution signal processing and wideband multiplexing benefit from the LTC1419's advantages—its 20MHz S/H bandwidth, 800ksps conversion rate and 14-bit resolution. This article shows two supply designs that are quiet enough to use with the LTC1419.

### Low Noise Inverting Converter

The LT1373 switching regulator shown in Figure 104 is configured as an inverting converter, creating –5V from 5V. This configuration has the advantage of a small triangular switching-current waveform through the sec-

ondary inductor. This current waveform is continuous, producing much less harmonic content than is created by a typical positive-to-negative voltage converter, with its rectangular switching current waveform. With the components shown, the LT1373 operates continuously with load currents above 10mA. Because the LTC1419 typically draws 18mA of negative supply current, the LT1373 will always operate in the quiet continuous mode.

### Regulated Charge Pump Converter

The LTC1419's negative PSRR also allows the use of charge pumps to create –5V. The circuit shown in Figure 105 uses the LT1054 regulated charge pump. This circuit has the advantage of reduced board space, since it lacks an inductor and requires fewer passive components. However, the LT1373 circuit can supply more current (150mA) than the LT1054 circuit (100mA).

### Performance Results

What is the effect of using either of these switch-based supplies on the LTC1419's conversion performance? The FFTs in Figures 106–108 show the excellent results. Figure
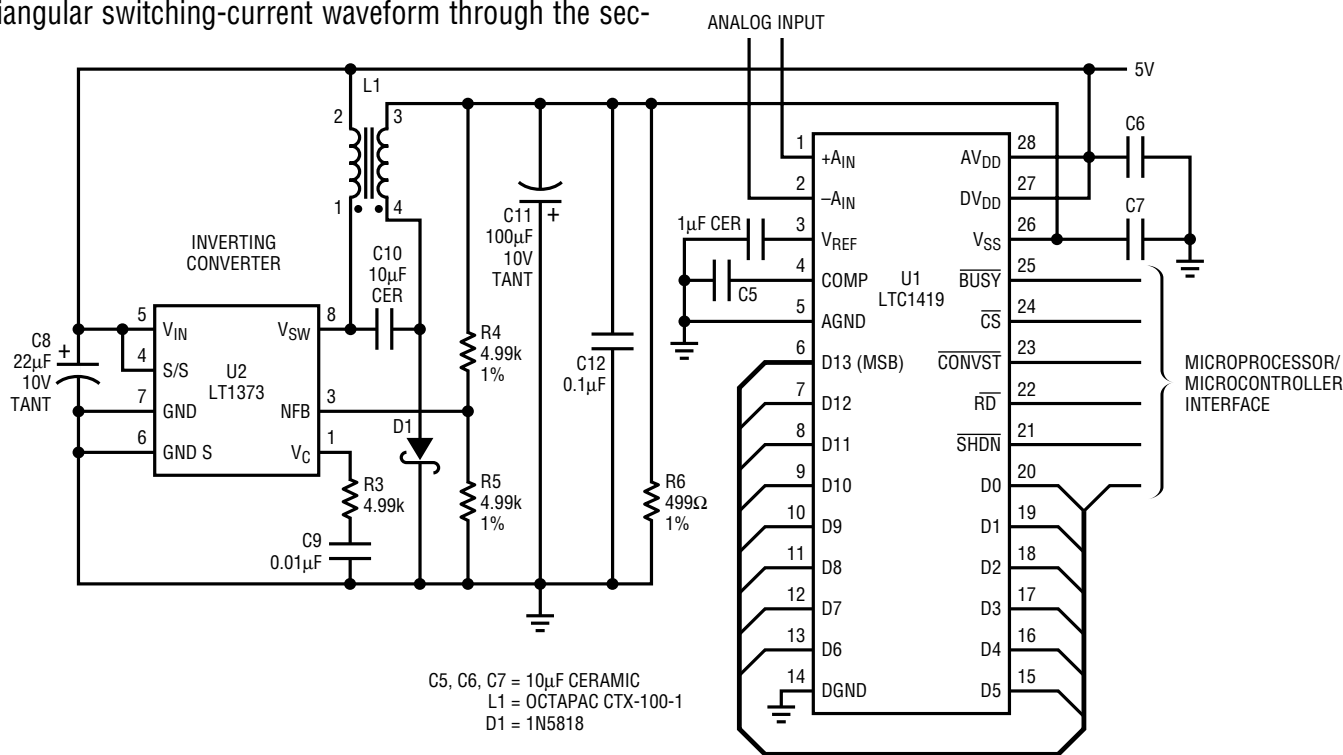


**Figure 104. The LTC1419's 80dB PSRR Allows the LTC1373 to Generate the –5V and Power the ADC without Signal-Coversion Degradation**
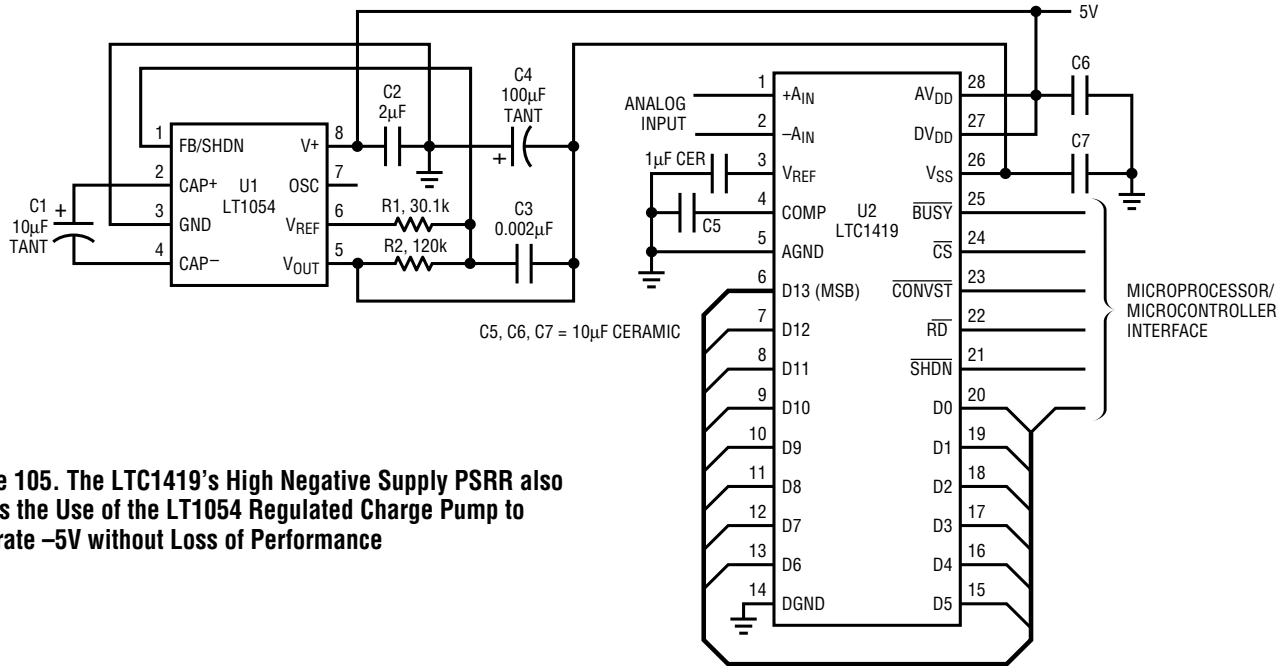
**Figure 105. The LTC1419's High Negative Supply PSRR also Allows the Use of the LT1054 Regulated Charge Pump to Generate –5V without Loss of Performance**

106 is an FFT of a typical LTC1419 operating on ±5V from a lab supply and converting a full-scale 91kHz sine wave at 800ksps. The noise floor is approximately 114dB below full scale, the second harmonic's amplitude is approximately 90dB below full scale and the SINAD is 80.5dB. Figure 107 shows the FFT of the same LTC1419 operating on a 5V lab supply and –5V from the LT1373 circuit. The noise floor and the second harmonic's amplitude remain the same relative to full scale and the SINAD remains the same at 80.5dB. Figure 108 shows the LTC1419's response when its –5V is generated by the LT1054 circuit. As with the LT1373 circuit, the noise floor and the amplitude of the harmonics remain the same and the SINAD is 80.8dB.
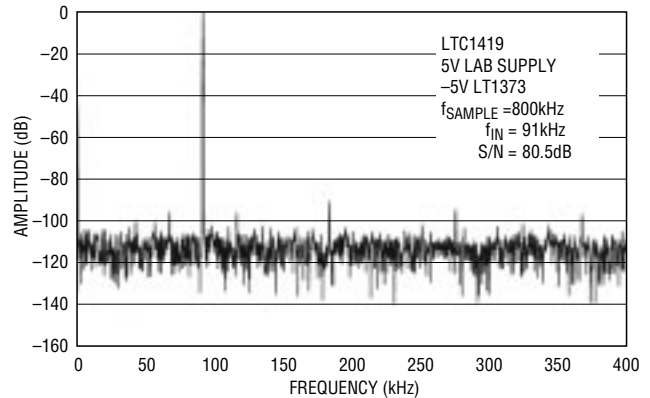


**Figure 107. When the –5V Supply is Generated by an LT1373 Switching Regulator, the SINAD, the Noise Floor, and the 91kHz Fundamental's Harmonics Remain Essentially the Same as in Figure 106**
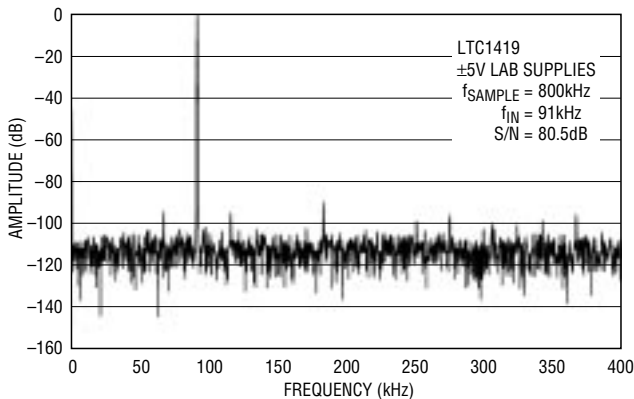


**Figure 106. This FFT of an LTC1419 Powered by a ±5V Lab Supply Shows a SINAD of 80.5dB for a 91kHz Input sampled at 800ksps**
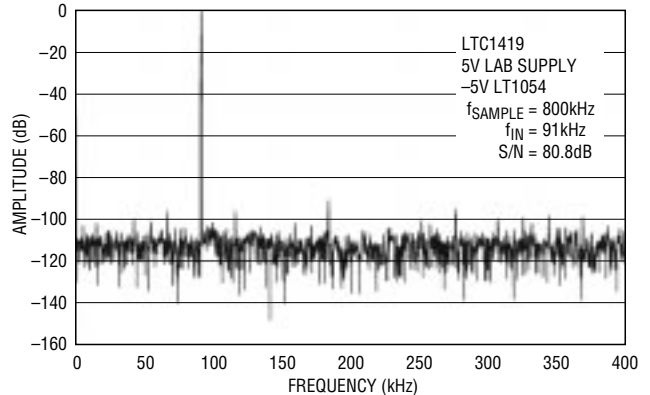


**Figure 108. When the –5V Supply is Generated by an LT11054 Inverter, the SINAD, the Noise Floor, and the 91kHz Fundamental's Harmonics Are Again Unchanged from Those in in Figure 106**

## NEGATIVE-TO-POSITIVE TELECOMMUNICATION SUPPLY
*by Kurk Mathews*

Many telecommunication circuits require a positive supply voltage derived from a –48V input. The traditional approach to negative-to-positive conversion has been to use a buck-boost converter (see Figure 109). Unfortunately, this topology suffers drawbacks as the power level and input-to-output voltage difference increases.

A more appropriate solution for –48V to 5V conversion is shown in Figure 110. The LT1680 is used to implement a forward converter with its output referenced to the input common. Compared to the buck-boost converter, switch current is reduced by a factor of two and output capacitor ripple is reduced by a factor of five.

The LT1680 is referenced to –48V and requires a 12V bias supply. The 12V is generated by using the RUN/SHDN and
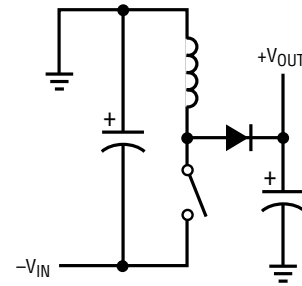


**Figure 109. Buck-Boost Converter**

a bootstrap winding on the output inductor, L1. When input voltage is first applied, R1 begins charging C1. As C1 charges, Q1 is held on by R2, shorting R3. R4 and R5 form a voltage divider that holds the RUN/SHDN pin below its 1.25V threshold until the $12V_{IN}$ pin reaches approximately 14V. Once out of standby, Q1 is turned off by Q2, reducing the run threshold to approximately 9V and allowing C1 time to discharge slightly before the overwinding on L1 takes over. The only remaining issue is feedback. Q3 translates the output voltage to a current, which flows to the $V_{FB}$ pin.
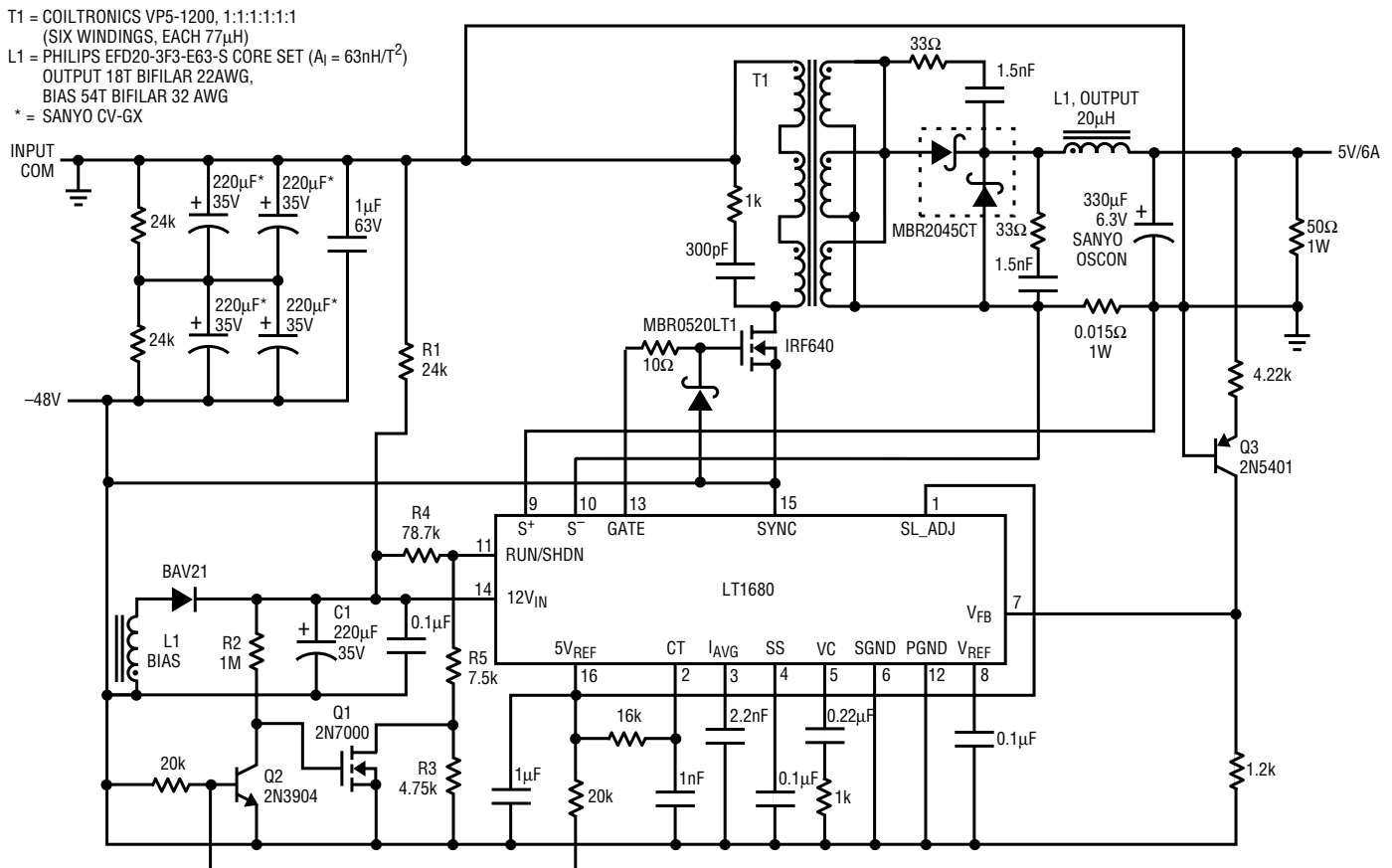


**Figure 110. –48V to 5V Telecommunications Supply**

The LT1680's unique differential current sense amplifier has an input common mode range of –0.3V to 60V. If $V_{IN}$ is expected to exceed 60V, the sense resistor could be relocated in the main FET's source and the input capaci-

tors' voltage increased. Because the forward converter is fundamentally an isolated topology, an optocoupler and reference could be added to provide isolation between the input and output of the supply.

## POSITIVE-TO-NEGATIVE CONVERTER POWERS –48V TELECOM CIRCUITS
*by Mitchell Lee*

If you're designing a system that interfaces to telecom equipment, chances are you'll need a –48V supply. The circuit in Figure 111 supplies up to 6W at –48V and scales to more than 12W with higher power components. Based on the inverting topology, the converter exhibits excellent efficiency over a wide range of loading conditions (see Figure 112).

The LT1171's error amplifier is designed for positive-boost applications, and hence its gain and reference are of the wrong phase and polarity for sensing an inverted output. In this application, the error amplifier is simply bypassed and feedback is applied at the compensation ($V_C$) pin. Zener diode D2 senses the output, pulling down on Q1 and the $V_C$ pin, in response to small increases in output voltage. Pulling down on the $V_C$ pin reduces peak switch current, and constitutes negative feedback. If the output is a little low, the Zener's diminished feedback signal is overcome by an internal 200µA current source at the $V_C$ pin, thereby increasing peak switch current and restoring the output voltage.

The combination of the LT1171 and the VP-2 series VERSA-PAC™ coil (CTX02-13836) are suited for 120mA output current as shown. For lighter loads of up to 60mA, use the LT1172 and a VP-1-series equivalent to the coil shown. For up to 15W, use the LT1171 and a VP-5 equivalent. High voltage versions of the LT1170 family (-HV) allow inputs of up to 20V without exceeding the peak switch-voltage rating.

This converter starts working at 2.7V and will regulate –48V at reduced power. You can add undervoltage lockout by inserting a Zener diode ($V_Z = V_{LOCKOUT} - 2.7V$) between the input supply and the LT1172's $V_{IN}$ pin.



**Figure 111. 12V to –48V Converter Features Good Efficiency Over a Wide Range of Loads**
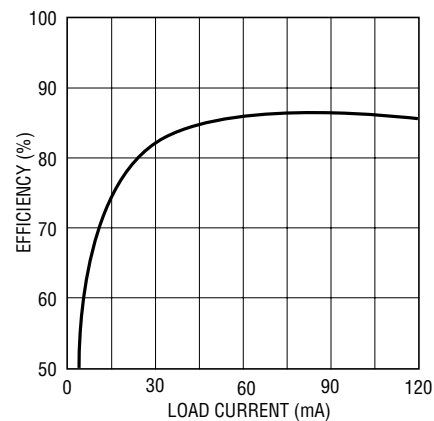


**Figure 112. Converter Efficiency Rises to 80% at Only 20mA Load**

## LOW NOISE LT1614 DC/DC CONVERTER DELIVERS –5V AT 200mA FROM 5V INPUT

*by Steve Pietkiewicz*

The inverting DC/DC converter function is traditionally realized with a capacitor-based charge pump. Although simple, the output impedances of the best charge pump solutions are in the 5Ω to 10Ω range, resulting in significant regulation issues when the load current increases beyond a few tens of milliamperes. The LT1614 inductor-based inverting DC/DC converter uses closed-loop regulation to obtain an output impedance of 0.1Ω, eliminating output voltage droop under load.

Figure 113 details the 5V to –5V converter circuit. The LT1614 contains an internal 0.6Ω switch rated at 30V, allowing up to 28V differential between input and output. Quiescent current is 1mA and the device contains a low-battery detector with a 200mV reference voltage. The device switches at 600kHz, allowing the use of small, inexpensive external inductors and capacitors. In fact, the total cost of the components specified in Figure 113 (excluding the LT1614) is approximately $0.70 in 10,000-piece quantities.

The LT1614 operates by driving its NFB pin to a voltage of –1.24V, allowing direct regulation of the negative output. This converter topology, which consists of inductors in series with both input and output, results in low output noise and also in low reflected noise on the 5V input supply. The output and switch nodes are shown in Figure 114. Output ripple voltage of 40mV is due to the ESR of the tantalum output capacitor C2. Ripple voltage can be reduced substantially by replacing output capacitor C2 with a 10µF ceramic unit, as pictured in Figure 115.
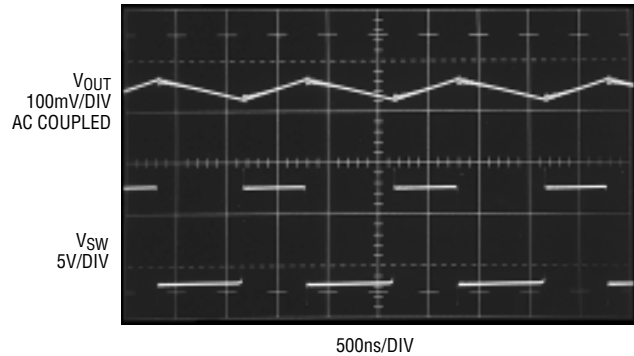


**Figure 114. LT1614 Output and Switch Node with a 33µF Tantalum Capacitor and 200mA Load Current**
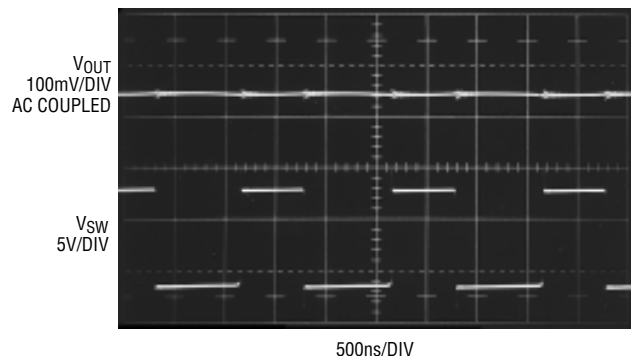


**Figure 115. LT1614 Output and Switch Node with a 10µF Ceramic Output Capacitor and 200mA Load Current**
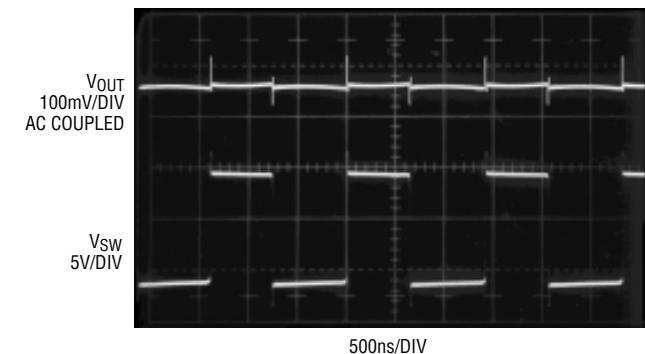


**Figure 116. Improper Placement of D1's Cathode Results in 60mV Switching Spikes at Output, Even with a 10µF Ceramic Output Capacitor**



D1: MBR0520          (800) 441-2447
L1, L2: MURATA LQH3C220   (814) 237-1431
C1, C2: AVX TAJB336M010   (803) 946-0362
C3: AVX1206YC105KAT (CERAMIC, X7R)

**Figure 113. 5V to –5V DC/DC Converter Uses an Inverting Topology and Delivers 200mA.**

In layout, be sure to tie D1's cathode directly to the LT1614's GND pin, as shown in Figure 113. This keeps the switching current loops tight and prevents the introduction of high frequency spikes on the output. The low noise that can be achieved with a ceramic capacitor may be corrupted by noise spikes if proper layout practice is not followed. To illustrate this point, output and switch waveforms from Figure 113's circuit, with a 10μF ceramic output capacitor and 200mA load, but with D1's cathode arbitrarily connected to the ground plane, are shown in Figure 115. 60mV switching spikes ruin an otherwise clean output.

Efficiency of the circuit is detailed in Figure 117. Efficiency reaches 73% at a 50mA load, and is above 70% at a 200mA load. Larger inductors with less copper resistance
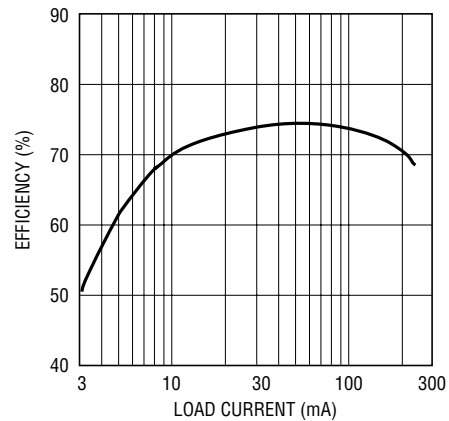


**Figure 117. 5V to –5V Converter Efficiency Reaches 73%**

can be used to increase efficiency, although such inductors are more expensive than the Murata units specified.

## –48V TO 5V DC/DC CONVERTER OPERATES FROM THE TELEPHONE LINE
*by Gary Shockey*

DC/DC converters for use inside the telephone handset require operation from the high source-impedance phone line. Additionally, the CCITT specifications call for on-hook power consumption of 25mW maximum. The DC/DC converter circuit presented here is 70% efficient at an input power of 25mW, providing 5V at 3.4mA. Controlled, low peak switch current ensures that the –48V input line does not experience excessive voltage drops during switching.
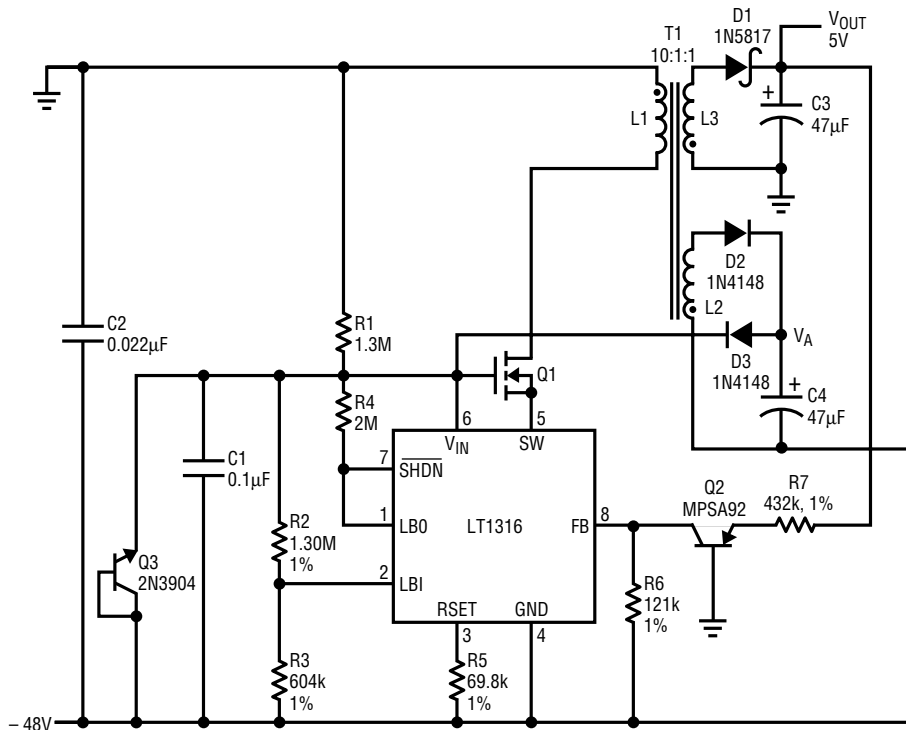


T1 =DALE LPE-4841-A313, L$_{PRI}$ = 2mH  (605) 665-1627  
Q1 =ZETEX ZVN 4424A  (516) 543-7100  

R6, Q2 AND R7 MUST BE PLACED NEXT TO THE FB PIN

**Figure 118. –48V to 5V Flyback Converter**

The circuit shown in Figure 118 operates as a flyback regulator with an auxiliary winding to provide power for the LT1316. To understand the operation of this circuit, examine Figure 118. When power is first applied, the LBI pin is low, causing the $\overline{SHDN}$ pin to be grounded through LBO. This places the part in shutdown mode and only the low-battery comparator remains active. During this state, $V_{IN}$ rises at a rate determined by R1 and C1. The LT1316 draws only 6μA in shutdown mode; R1 needs to supply only this current, the current through R2 and R4, and C1's charging current. When LBI reaches 1.17V ($V_{IN} \approx 3.7V$) the LBO pin lets go of $\overline{SHDN}$ and the part enters the active mode. Once this state is reached, switching action begins and the output voltage begins to increase. As the device switches, the LT1316 $V_{IN}$ pin draws current out of C1; $V_{IN}$ then decreases sufficiently to trip the low-battery detector, stopping the switching. Start-up proceeds in this irregular fashion until, eventually, the voltage at $V_A$ increases to 5V. ($V_A$ is the same as $V_{OUT}$, because L2 and L3 have the same number of turns.) After start-up, current is supplied to the LT1316 from $V_A$ rather than from the –48V rail,

increasing efficiency. $V_{OUT}$ must not be loaded until it reaches 5V or the circuit will not start.

During each switch cycle, current in the transformer primary ramps up until current limit is reached (See Figures 119 and 120). This peak switch current can be set by adjusting R5. The circuit shown uses a 69.8kΩ resistor to give a peak switch current of 50mA. Increasing R5 decreases the current limit. Secondary peak current will be approximately equal to the primary peak current multiplied by the transformer turns ratio. The FB pin has a sense voltage of 1.23V and $V_{OUT}$ can be set by the following formula:

$$V_{OUT} = 1.23(R7/R6) + 0.6V.$$

Efficiency versus load current is detailed in Figure 121. Note that for the range of 4mA to 80mA, 70% efficiency or greater is achieved. Figure 122 shows input current versus output power. Less than 80μA quiescent current flows when the converter supplies 0.5mW over the 36V–72V range.
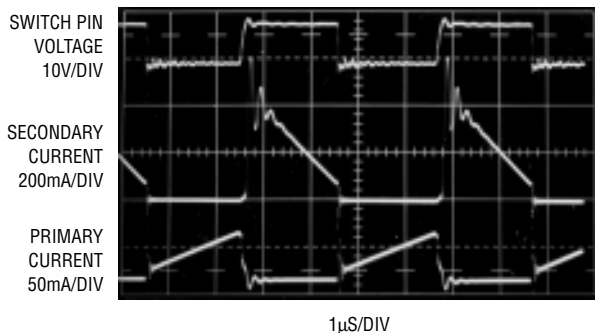


Figure 119. Switch Voltage and Current Waveforms
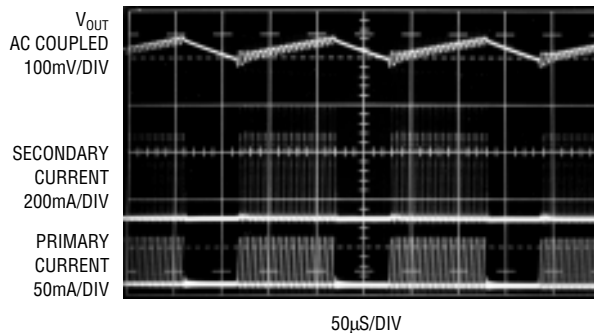


Figure 120. Output Ripple Voltage and Current Waveforms
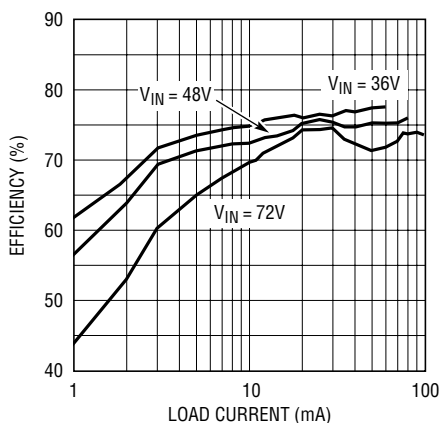


Figure 121. Efficiency vs Load Current
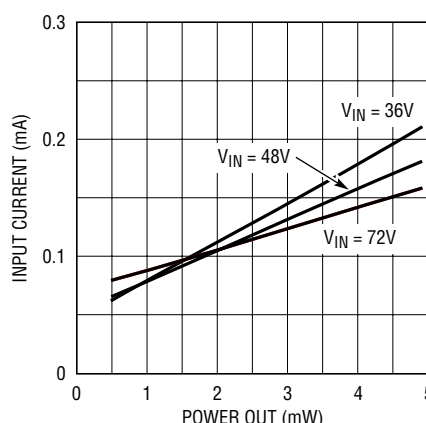


Figure 122. Input Current vs Power Out

Regulators—Switching (Flyback)

## THE LT1425 ISOLATED FLYBACK CONTROLLER
*by Kurk Mathews*

### Introduction

Low voltage circuitry, such as local area networks (LAN), isolation amplifiers and telephone interfaces, frequently requires isolated power supplies. The flyback converter is often the choice for these low power supplies because of its simplicity, size and low parts count. Unfortunately, designers are forced to add optocouplers and references in order to achieve the desired output regulation and transient response.

The LT1425 provides a one-chip solution for these and other applications. The LT1425 is a 275kHz current mode controller with an integral 1.25A switch designed primarily to provide well regulated, isolated voltages from 3V–20V sources. The LT1425 is available in a 16-pin SO. Features include a new error amplifier and load compensation circuitry that eliminate the need for optocouplers while maintaining output regulation typically within a few percent.

### Typical Applications

Figure 123 shows a typical flyback LAN supply using the LT1425. Figure 123 also includes details on an alternate transformer for a complete PCMCIA type II height solution. The output voltage is within 1% of –9V for load currents of 0mA–250mA. Input current is limited to 0.35 amps in the event the output is short circuited. The output voltage droops only 300mV during a 50mA to 250mA load transient (see Figure 124). The off-the-shelf transformers provide 500$V_{AC}$ of isolation. The high switching frequency allows the use of small case size, low cost, high value ceramic capacitors on the input and output of the supply.

Figure 126 shows a ±15V supply with 1.5kV of isolation. Output regulation remains within ±3% over the entire 5V to 15V input voltage and ±60mA output current range,
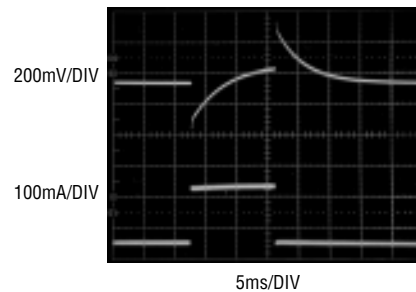


**Figure 124. Transient Response of LT1425 5V to –9V Converter**
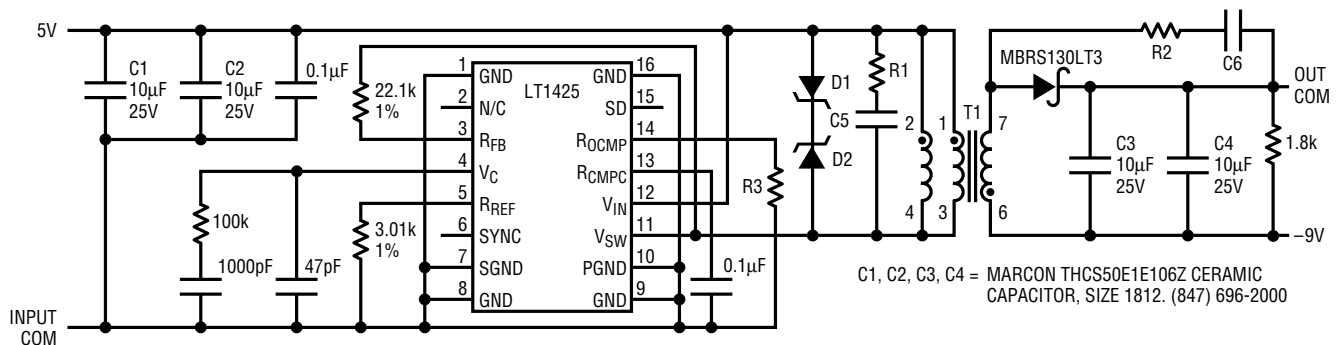


C1, C2, C3, C4 = MARCON THCS50E1E106Z CERAMIC CAPACITOR, SIZE 1812. (847) 696-2000

### TRANSFORMER T1

| | L$_{PRI}$ | TURNS RATIO | ISOLATION | SIZE (L × W × H) | I$_{OUT}$ | EFFICIENCY | D1 | D2 | R1, R2 | C5, C6 | R3 |
|---|---|---|---|---|---|---|---|---|---|---|---|
| DALE LPE-4841-A307 | 36µH | 1:1:1 | 500VAC | 10.7 × 11.5 × 6.3mm | 250mA | 76% | NOT USED | NOT USED | 47Ω | 330pF | 13.3k |
| COILTRONICS CTX02-13483 | 27µH | 1:1 | 500VAC | 14 × 14 × 2.2mm | 200mA | 70% | 1N5248 | MBR0540TL1 | 75Ω | 220pF | 5.9k |

**Figure 123. 5V to –9V/250mA Isolated LAN Supply**

A: $V_{SW}$ = 20V/DIV

B: $I_{SW}$ = 1A/DIV

C: $V_{SW}$ = 20V/DIV

D: $I_{SW}$ = 0.2A/DIV

1μs/DIV
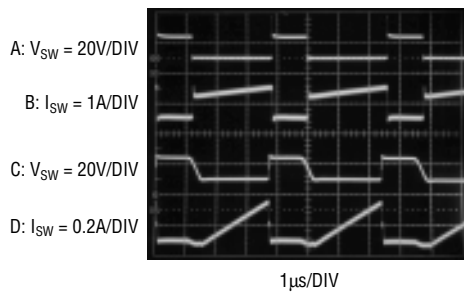
**Figure 125. Switch Voltage and Current for Figure 123's Circuit with Outputs of –9V/250mA and –9V/30mA**

even with one output fully loaded and the other unloaded (±1.5% with input voltages of 10V–15V). The isolation voltage is ultimately limited only by bobbin selection and transformer construction.

Figure 127 implements a 12V to 5V/1A step-down regulator with off-the-shelf magnetics. The circuit uses an external, cascoded 100V MOSFET to extend the LT1425's 35V maximum switch voltage limit. D1 and Q1 ensure the LT1425 does not start until almost 9V, guaranteeing adequate gate voltage for the MOSFET. The MUR120 prevents the source from rising above the gate at turn-off.

The circuit in Figure 128 achieves even higher input voltages, this time in the form of a –48V to 5V/2A isolated telecom supply. The input voltage is too high to directly run Q1 or the LT1425, so a bootstrap winding is used to provide feedback and power for the IC after start-up. The voltage to the $V_{IN}$ pin is controlled by D1, D2, Q2, Q3 and associated components, which form the necessary start-up circuitry with hysteresis. Nothing happens until C1 charges through R1 to 15V. At that point, Q2 turns on Q3, pulling the shutdown pin high. Q3, in turn, latches Q2 on, setting the turn-off voltage to approximately 11V. Switching begins and, before C1 has a chance to discharge to 11V, the bootstrap winding begins to supply power. If the output is shorted, R2 prevents C1 from being charged by the transformer's leakage energy, causing the supply to continually attempt to restart. This limits input and output current during a short circuit. Feedback voltage is fed directly through a resistor divider to the $R_{REF}$ pin. The sampling error amplifier still works, but the load compensation circuitry is bypassed. This results in a ±5% load regulation over line and load. A dedicated feedback winding referencing the feedback voltage to the $V_{IN}$ pin could be used to include the load compensation function and improve regulation.
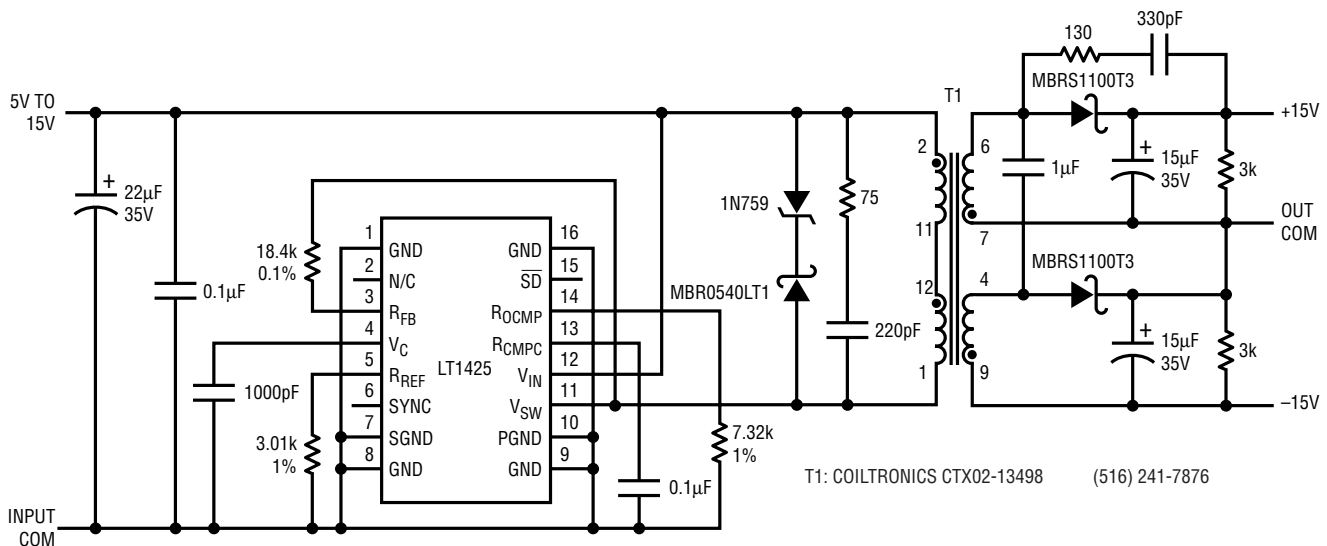


**Figure 126. Fully Isolated ±15V, ±60mA Supply**
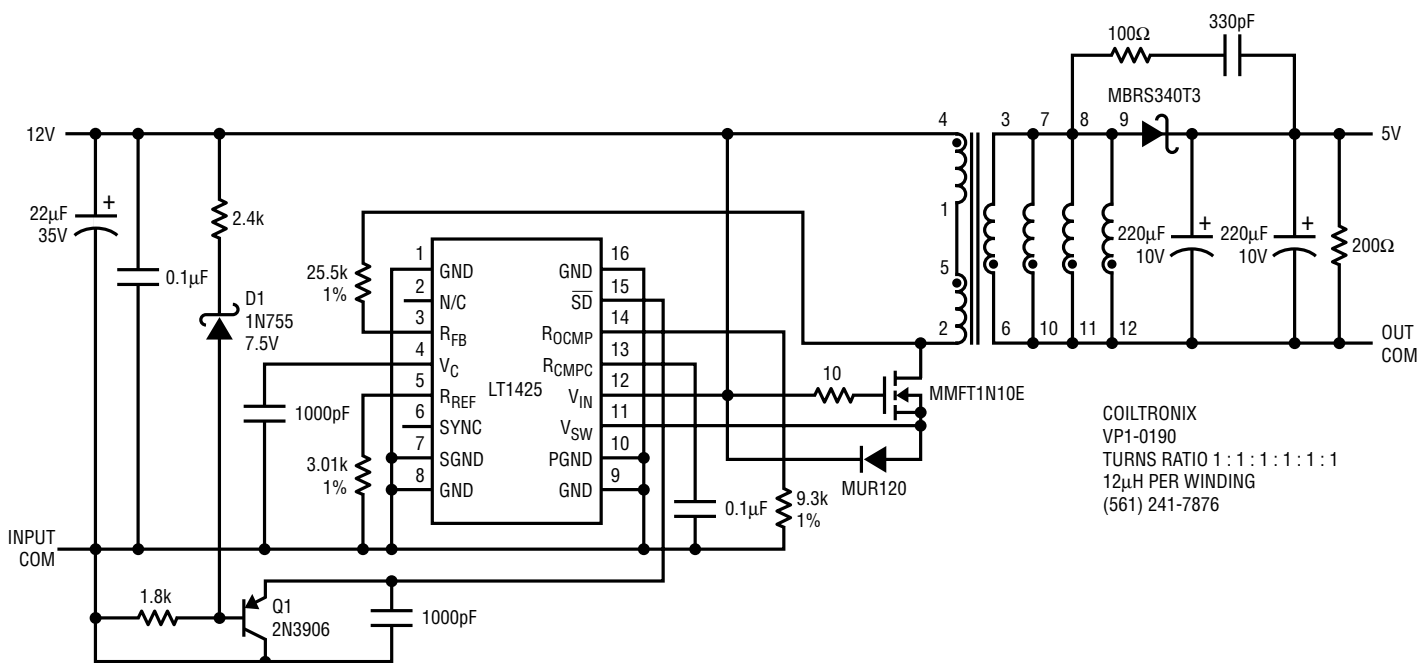
**Figure 127. 5V/1A Step-Down, Isolated Supply**
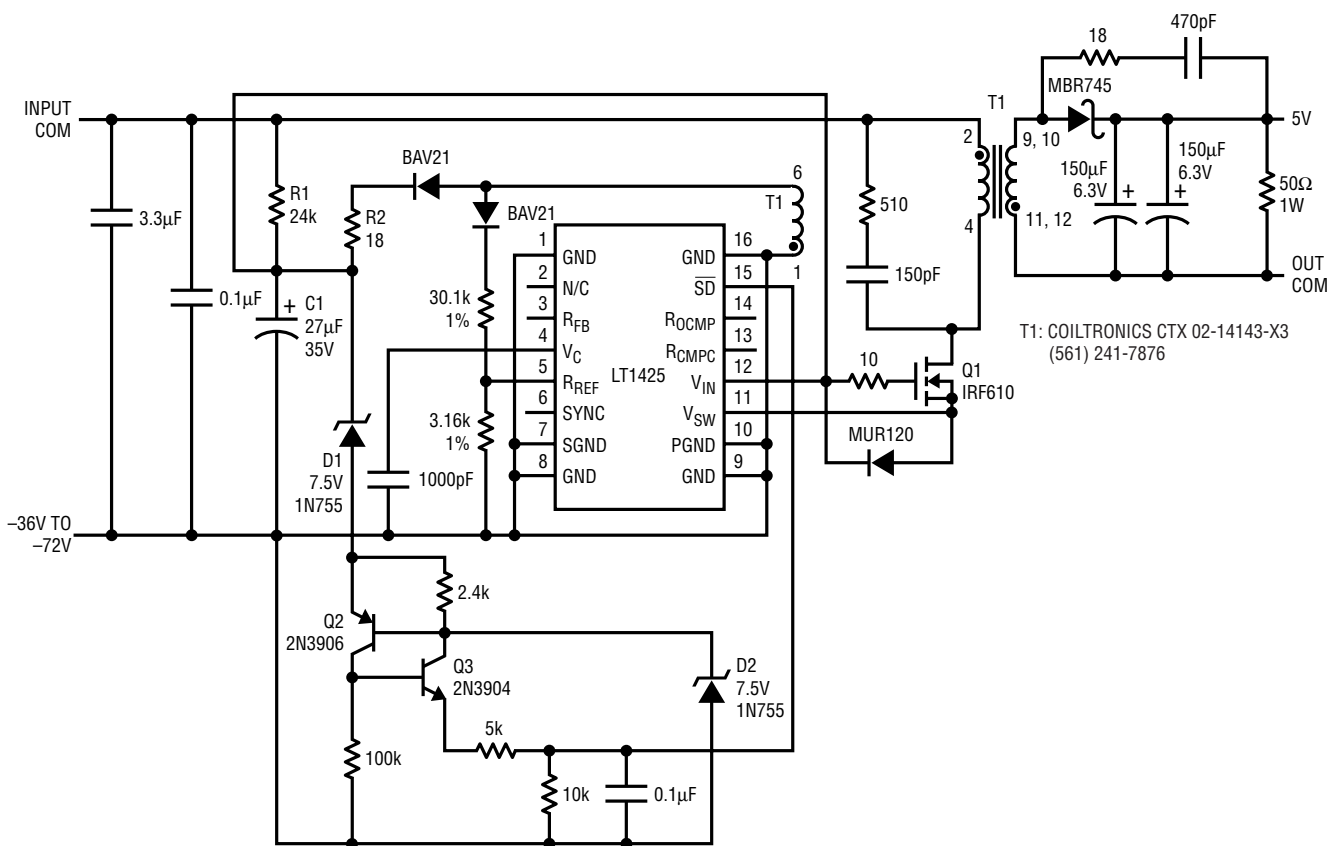
**Figure 128. 5V/2A Telecommunications Supply**

## HIGH ISOLATION CONVERTER USES OFF-THE-SHELF MAGNETICS
*by Mitchell Lee*

Isolated flyback converters usually evoke thoughts (or bitter memories) of custom transformers, slipped delivery schedules and agency approval problems. Off-the-shelf flyback transformers are available from several vendors, but these carry isolation ratings of only 300V–500V, and, rarely, of up to 1kV. Flyback transformers with isolation ratings of 3750V$_{RMS}$ are impossible to find, and if an application requires this level of isolation, an expensive, custom design is likely the only solution.

Gate-drive transformers, designed to couple switching regulator controllers to MOSFET gates, are readily available from stock with high isolation ratings and low cost. These are wound on ungapped cores and have very high inductance (500μH to 2mH), and will quickly saturate in a normal flyback converter circuit.

The transformer used in Figure 129's circuit handles significant current without saturating. The converter operates from a 12V battery-backed input supply and outputs 24V at 200mA. The key feature is that the second coil is not a coil at all, but rather an off-the-shelf gate drive
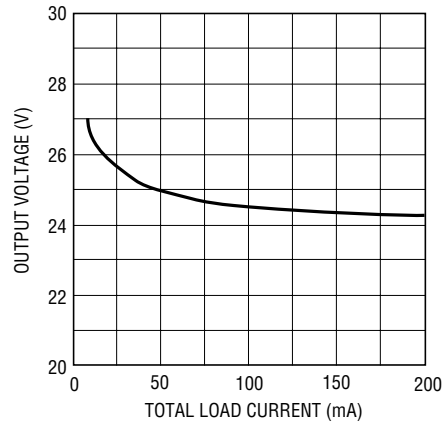


**Figure 130. Output Regulation for Figure 129's Circuit**

transformer. This component offers 3750V$_{RMS}$ isolation and full VDE approval at a lower cost than a comparable custom design.

Feedback is derived from the primary winding, through D3. R1 acts to filter the leakage-inductance spike at switch turn-off, and C4 smooths the recovered feedback voltage. Note that the transformer is wound 1:1; C4 peak detects a voltage roughly equal to the output. Sizing R1 and C4 is a trade-off between minimum load and load regulation. As shown, a minimum load of 3600Ω is recommended. Output regulation is shown in Figure 130. Line regulation from 10V to 20V input at full load is 0.13%/V.
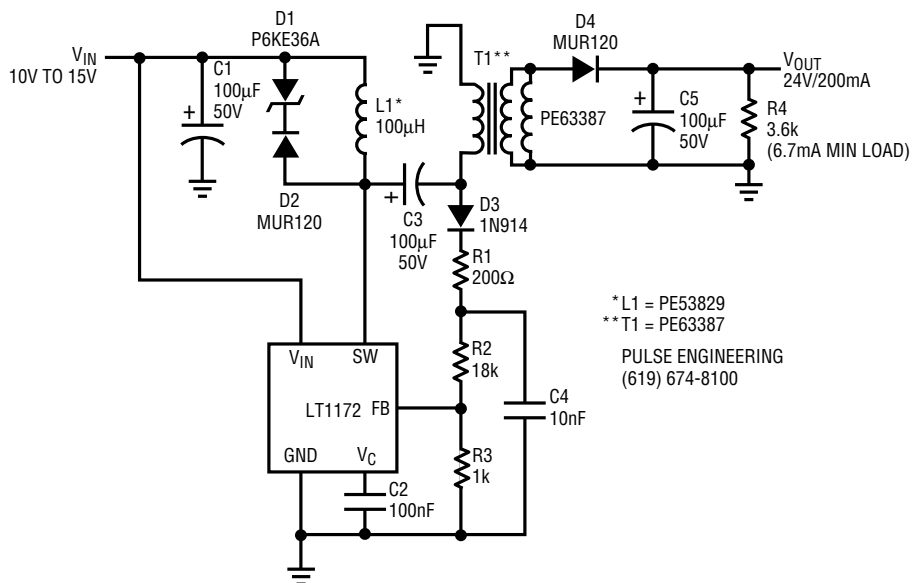


**Figure 129. 24V/200mA Bulk Supply with 3750V$_{RMS}$ Isolation**

## WIDE-INPUT-RANGE, LOW VOLTAGE FLYBACK REGULATOR

*by Kurk Mathews*

Many new switching regulators are designed with a specific application or topology in mind. If your requirements happen to fall within these parameters, all is well. Unfortunately, when faced with unusual requirements, the designer is often forced to choose bare-bones, universal regulators. The LTC1624 overcomes these issues by providing a full featured regulator that can operate in the step-down (buck), step-up (boost), buck-boost or flyback mode.

This constant-frequency current mode controller includes a high-side differential current sense amplifier and a floating high current N-Channel MOSFET driver. In the buck mode, an external bootstrap capacitor between the BOOST and SW pins works in conjunction with the internal 5.6V regulator and diode to provide a regulated supply for a high-side driver. In the boost, buck-boost or flyback mode, the SW pin is grounded, providing drive for a low-side MOSFET.

An example of a wide-input-range flyback is shown in Figure 131. The circuit provides ±50V at 75mA from a 4.75 to 24V source. The sum of line-, load- and cross-regulation is better than ±5%. The TG pin voltage is controlled by the internal 5.6V regulator, allowing the input voltage to be above Q1's 16V maximum gate-to-source voltage rating. 200kHz fixed frequency operation minimizes the size of T1. The R-C snubber formed by C1 and R1 in combination with T1's low leakage inductance keeps Q1's drain voltage well below its 100V rating. To improve cross-regulation, Q2, R2 and R3 were included to disable Burst Mode™ operation (a feature that improves efficiency at light load conditions by skipping switching cycles). The LTC1624's 95% maximum duty cycle accommodates the 5-to-1 input voltage range. Finally, by reconfiguring T1's secondaries, a variety of output configurations, such as 24V out (four windings in parallel), single 50V/150mA or a single 100V output, are possible with this same basic circuit.
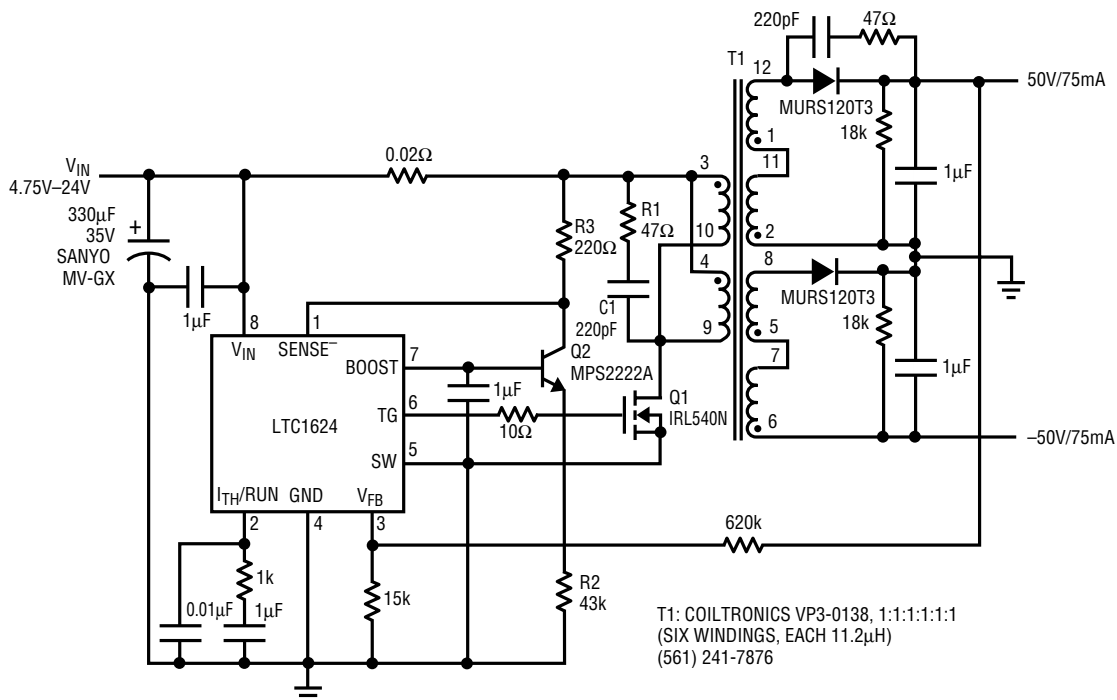


**Figure 131. Wide-Input-Range Flyback Regulator Provides ±50V at 75mA**

Regultors—Switching (Low Noise)

## THE LT1533 HERALDS A NEW CLASS OF
## LOW NOISE SWITCHING REGULATORS
*by Jeff Witt*

### Introducing the LT1533 Low Noise Switcher

The LT1533 is a switching regulator that provides a solution to EMI problems through two flexible approaches. First, the slew rates of both the current through the power switch and the voltage on it are easily programmed with external resistors. Limiting these slew rates will remove the highest harmonics from the switching waveforms. Second, the LT1533, with two 1A power switches, is designed to operate in push-pull circuits. Such circuits, with their low input and output current ripple, are inherently quiet. The result is an integrated switching regulator that provides very quiet output power and very low emissions. Figure 132 illustrates what can be achieved. The top trace shows the output of a push-pull boost regulator generating 120mA at 12V from an input of 5V. This trace was measured using a 10MΩ oscilloscope probe with a six-inch ground lead, demonstrating that there is no significant inductively or capacitively coupled noise. Probing the output of the LT1533 circuit with a 50Ω low noise
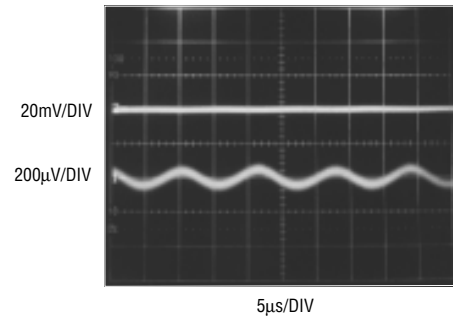


**Figure 132. Output Ripple of an LT1533 Switching Regulator Producing 120mA at 12V from a 5V Input**

amplifier reveals the real performance (second trace): peak-to-peak output ripple of the low noise switcher is only 150μV in a 10kHz to 100MHz bandwidth.

### A Closer Look at the LT1533

The LT1533 is a fixed frequency current mode PWM switching regulator. The output voltage is regulated by controlling the peak switch current on each cycle of the oscillator, resulting in good transient performance and rapid current limiting. The oscillator drives a toggle flip-flop, alternately enabling one of two 0.5Ω NPN power switches, QA and QB. The switch current is monitored by a sense resistor at the emitter of the switch. The output
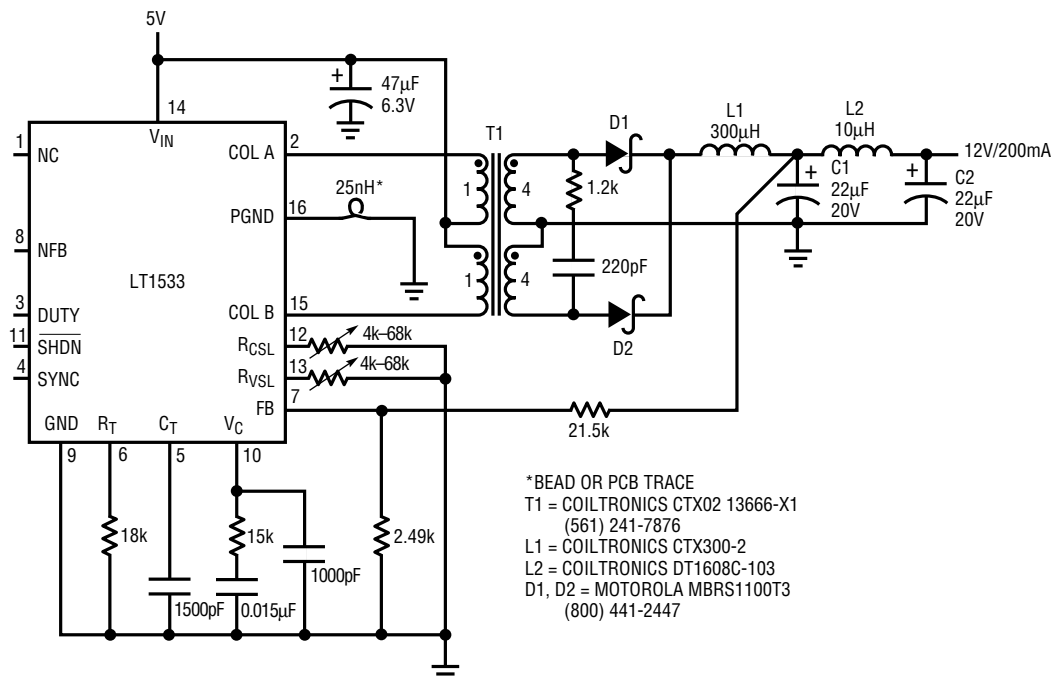


**Figure 133. 5V to 12V Push-Pull PWM Converter**
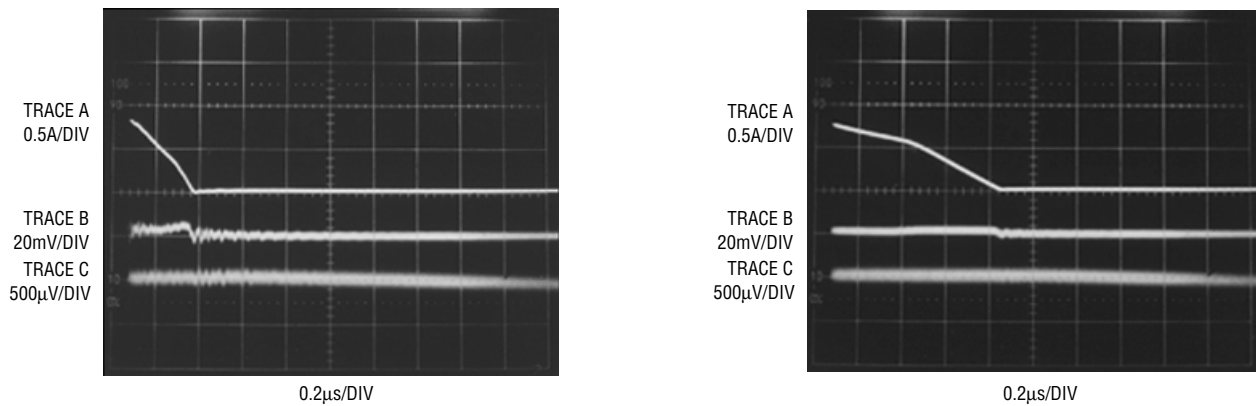
**Figure 134. Lowering the Slew Rates of the Power Switches (Trace A) Eliminates High Frequency Ripple at the Output (Traces B and C)**

voltage (either positive or negative) is compared with an accurate internal 1.25V reference voltage by an error amplifier whose current output, along with loop compensation components tied to the $V_C$ pin, determine the peak switch current required for regulation; a comparator turns off the switch when this current level is reached.

The slew-control circuitry monitors the collector voltages and emitter currents of the power switches and adjusts base drive to control both the voltage and current slew rates. The desired rates are programmed by tying the $R_{VSL}$ and $R_{CSL}$ pins to ground with resistors between 4k and 68k, corresponding to slew rates from ~80V/µs to 5V/µs and 7A/µs to 0.4 A/µs. This allows the circuit designer to

directly trade off quiet, low EMI operation with high efficiency: low slew rates result in slowly changing stray fields, which generate less interference, but increase the conduction losses in the switches.

The LT1533 oscillator presents additional opportunities for managing EMI. Its wide frequency range (20kHz to 250kHz) allows the designer to avoid sensitive frequencies. Operating frequency is set with a capacitor on the $C_T$ pin and a resistor of nominally 17k on the $R_T$ pin. The LT1533 can also be synchronized to an external clock, allowing accurate placement of both switching frequency and phase.
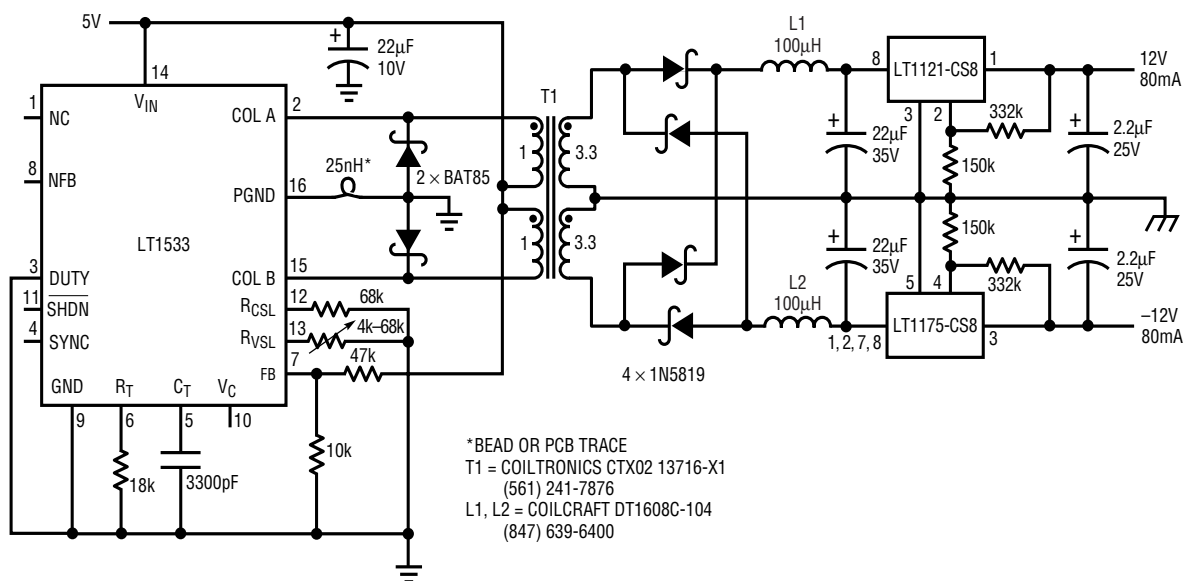


**Figure 135. 5V to ±12V DC/DC Transformer**
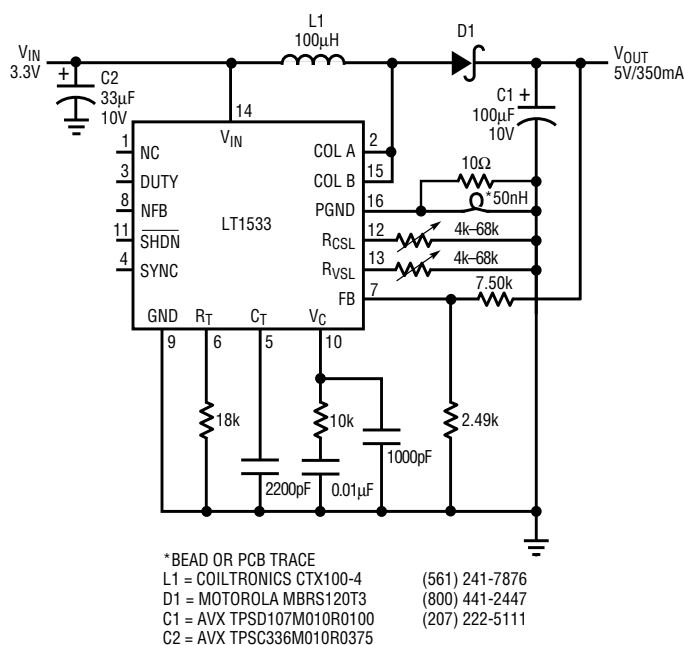
**Figure 136. 3.3V to 5V Boost Converter**

*BEAD OR PCB TRACE
L1 = COILTRONICS CTX100-4
D1 = MOTOROLA MBRS120T3
C1 = AVX TPSD107M010R0100
C2 = AVX TPSC336M010R0375

(561) 241-7876
(800) 441-2447
(207) 222-5111

## Push-Pull PWM Makes a Quiet Boost Converter

The push-pull converter in Figure 133 produces 200mA at 12V from an input of 5V. The oscillator is set to 80kHz (note that the circuit operates at half this frequency) and the LT1533 applies a pulse-width modulated 5V to the primary side of the transformer. The rectified secondary voltage is filtered by L1 to generate 12V on C1. In this circuit, L1 is the primary energy storage device, so the transformer can be made fairly small. Additional output filtering is provided by L2 and C2.

This topology is inherently quiet. Current through L1 into the primary output capacitor C1 is a continuous triangle wave with little high frequency content, resulting in low conducted output noise. With an appropriate transformer turns ratio, RMS input current is kept low, reducing the potential for conducted noise on the input.

It is advantageous to start with a good topology, but high frequency noise will still get around via stray capacitance and mutual inductance; the best way to deal with this is to eliminate fast edges. Figure 134 shows several waveforms from the circuit as it delivers 120mA of output current. The upper trace in each photo is the current in switch QA as it turns off. Trace B is the output voltage probed with a

10MΩ scope probe with a six-inch ground lead. The lower trace is the output measured with a low noise amplifier. In the left photo the switch slew rates are programmed to their highest values with 3.9k resistors on the $R_{CSL}$ and $R_{VSL}$ pins. The fast switch transients induce high frequency ripple on the output (the higher level of noise on the middle trace is due to the inductance of the scope probe's ground lead). By lowering the slew rates ($R_{CSL}$ = 24k and $R_{VSL}$ = 8.2k) this potentially troublesome output ripple is eliminated, as shown in the right photo. The efficiency penalty is minor; the slower slew rates reduce efficiency from 73% to 70%.

This combination of appropriate circuit topology and controlled slew rates produces the exceptionally clean output shown in Figure 132. This circuit is simply implemented with ordinary PCB construction, and can be placed in close proximity to sensitive circuits without the need for expensive electrostatic or magnetic shielding.

## DC Transformer with Civilized Edges

Grounding the Duty pin of the LT1533 disables the feedback loop and runs each switch at 50% duty cycle, allowing the LT1533's use in DC transformer circuits. Such circuits are useful for generating bipolar or isolated supplies; Figure 135 shows an example. The LT1533 switches 5V across a 3.3:1 transformer and a diode bridge rectifies the secondary side voltages to produce nominally 16V bipolar outputs that are regulated to ±12V. Short-circuit current limit at the output is provided by the LT1533's switch current limit; the 1A switch limit is transformed to 0.3A on the secondary.

A common problem with isolated-output switchers is that fast edges couple through stray capacitance between the primary and secondary windings of the transformer to create common mode noise on the outputs. Also, linear regulators are incapable of rejecting high frequency noise at their inputs. Both problems are greatly reduced by limiting the switch slew rates. Shielding between the windings can be eliminated, reducing transformer size and cost. LC filters on the isolated side are unnecessary with the linear regulators rejecting ripple at the operating frequency and the controlled slew rates eliminating high frequency ripple.
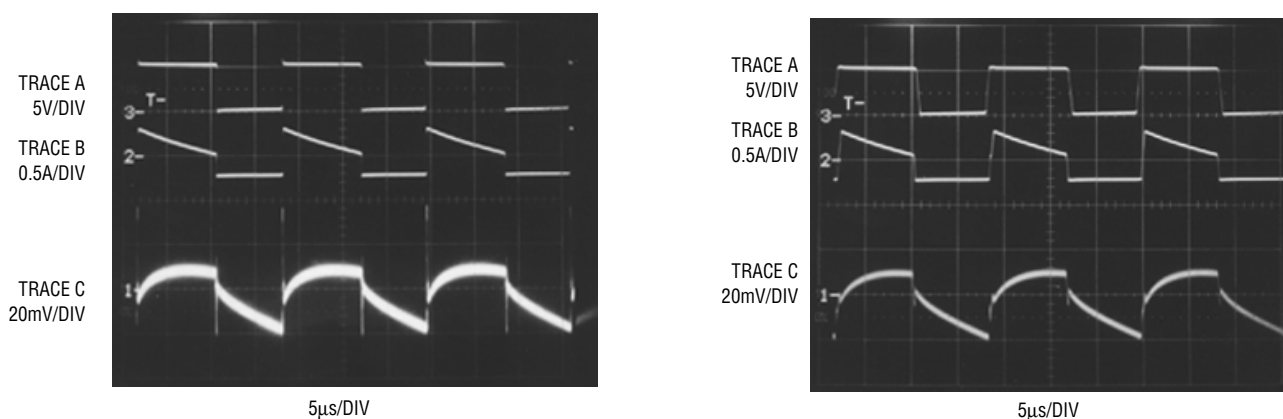
**Figure 137. Limiting Switch Slew Rates (Traces A and B) Lowers the High Frequency Content of the Boost Regulator's Output Ripple (Trace C)**

### 3.3V to 5V Boost Converter

Simple switching topologies can also benefit from the LT1533's low noise features. In a boost regulator, for example, the current into the output capacitor is a square wave, which contains the high frequency harmonics generated by a fast power switch. Even when the rectifying diode is off, fast voltage waveforms at the switch couple through the Schottky diode's capacitance. Fast switching can also excite high frequency resonant circuits formed by the diode's capacitance and parasitic inductance due to board traces. All of these effects can be reduced by controlling the slew rate of the switch. Figure 136 shows the LT1533 in a simple boost circuit generating 5.0V from a 3.3V input, a typical requirement when interfacing 3.3V logic systems to 5V high performance ADCs. The collectors of the two power switches are tied together and alternately energize the boost inductor. Figure 137 shows several waveforms at two different slew rate settings with the circuit delivering 200mA of output current. Trace A is the switch voltage, trace B is the current through the output capacitor and trace C is the AC-coupled output voltage in a 100MHz bandwidth. In the left photo, the slew rates are set to their maximum values ($R_{CSL} = R_{VSL} = 3.9k\Omega$). The rapidly switched current combined with the finite series inductance of the output capacitor result in large voltage spikes on the output. The right photo shows

the same waveforms with the slew rates lowered ($R_{CSL} = R_{VSL} = 22k$), eliminating the troublesome transients. The penalty is a drop in efficiency from 85% to 80%.

### Conclusion: a Switcher for Sensitive Systems

With two 1A power switches, the ability to control positive or negative outputs, and a wide input operating range (2.7 to 30V), the LT1533 is a highly flexible switching regulator. Thermal shutdown, in addition to switch-current limit, provides circuit protection. The LT1533 is packaged in the narrow 16-lead SO, and is available in commercial and industrial grades.

The LT1533 allows the circuit designer to add a switching regulator to sensitive analog systems without fear of introducing uncontrollable noise and interference. The programmable operating frequency and switch slew rates allow final tuning to occur in the circuit, when the system is running and interference problems may first become apparent. In addition to providing a way to deal with unforeseen problems, this flexibility means that sacrifices in efficiency will be limited to those needed for proper system performance. The LT1533 is the switching regulator of choice for high performance analog systems.

**LT1533 ULTRALOW NOISE SWITCHING REGULATOR FOR HIGH VOLTAGE OR HIGH CURRENT APPLICATIONS**
*by Jim Williams*

The LT1533 switching regulator[1,2] achieves 100μV output noise by using closed-loop control around its output switches to tightly control switching transition time. Slowing down switch transitions eliminates high frequency harmonics, greatly reducing conducted and radiated noise.

The part's 30V, 1A output transistors limit available power. It is possible to exceed these limits while maintaining low noise performance by using suitably designed output stages.

**High Voltage Input Regulator**

The LT1533's IC process limits collector breakdown to 30V. A complicating factor is that the transformer causes the collectors to swing to twice the supply voltage. Thus, 15V represents the maximum allowable input supply. Many applications require higher voltage inputs; the circuit in Figure 138 uses a cascoded[3] output stage to achieve such high voltage capability. This 24V to 5V ($V_{IN}$ = 20V–50V) converter is reminiscent of previous LT1533 circuits, except for the presence of Q1 and Q2.[4] These devices, interposed between the IC and the transformer, constitute a cascoded high voltage stage. They provide voltage gain while isolating the IC from their large drain voltage swings.
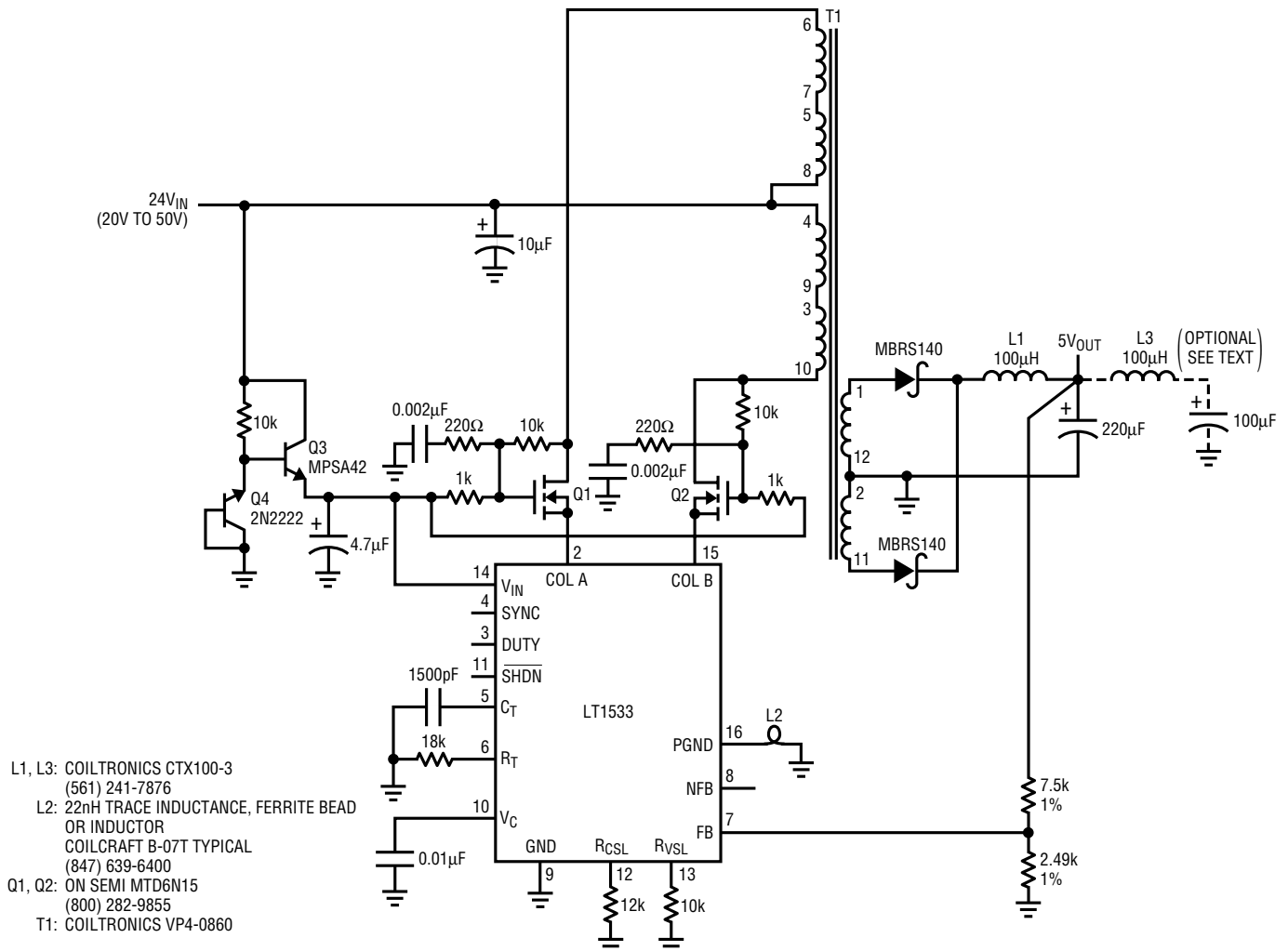


**Figure 138. A Low Noise 24V to 5V Converter ($V_{IN}$ = 20V–50V): Cascoded MOSFETs Withstand 100V Transformer Swings, Permitting the LT1533 to Control 5V/2A**

A = 20V/DIV

B = 5V/DIV
(AC COUPLED)

C = 100V/DIV

10µs/DIV
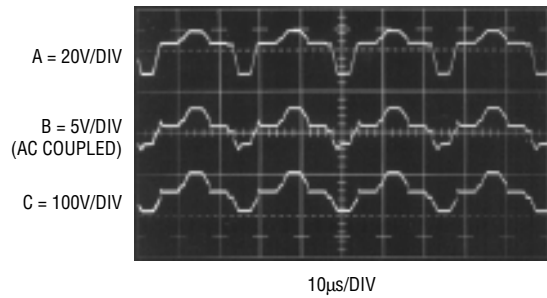
**Figure 139. MOSFET-Based Cascode Permits the Regulator to Control 100V Transformer Swings while Maintaining a Low Noise 5V output. Trace A is Q1's Source, Trace B is Q1's Gate and Trace C is the Drain. Waveform Fidelity through Cascode Permits Proper Slew-Control Operation**



A = 5mV/DIV

B = 100µV/
DIV

2µs/DIV

**Figure 141. Waveforms for Figure 139 at 10W Output: Trace A Shows Fundamental Ripple with Higher Frequency Residue Just Discernable. The Optional LC Section Results in Trace B's 180µV$_{P-P}$ Wideband Noise Performance**

### Current Boosting

Figure 140 boosts the regulator's 1A output capability to over 5A. It does this with simple emitter followers (Q1–Q2). Theoretically, the followers preserve T1's voltage and current waveform information, permitting the LT1533's slew-control circuitry to function. In practice, the transistors must be relatively low beta types. At 3A collector current, their beta of 20 sources ≈150mA via the Q1–Q2 base paths, adequate for proper slew-loop operation.[5] The follower loss limits efficiency to about 68%. Higher input voltages minimize follower-induced loss, permitting efficiencies in the low 70% range.

Figure 141 shows noise performance. Ripple measures 4mV (Trace A) using a single LC section, with high frequency content just discernable. Adding the optional second LC section reduces ripple to below 100µV (trace B), and high frequency content is seen to be inside 180µV (note ×50 vertical scale-factor change).

Normally, high voltage cascodes are designed simply for supply isolation. Cascoding the LT1533 presents special considerations because the transformer's instantaneous voltage and current information must be accurately transmitted, albeit at lower amplitude, to the LT1533. If this is not done, the regulator's slew-control loops will not function, causing a dramatic output noise increase. The AC-compensated resistor dividers associated with the Q1–Q2 gate-drain biasing serve this purpose, preventing transformer swings coupled via gate-channel capacitance from corrupting the cascode's waveform-transfer fidelity. Q3 and associated components provide a stable DC termination for the dividers while protecting the LT1533 from the high voltage input.

Figure 139 shows that the resultant cascode response is faithful, even with 100V swings. Trace A is Q1's source; traces B and C are its gate and drain, respectively. Under these conditions, at 2A output, noise is inside 400µV peak.



L1: COILTRONICS CTX300-4
     (561) 241-8786
L2: 22nH TRACE INDUCTANCE, FERRITE BEAD OR
     INDUCTOR. COILCRAFT B-07T TYPICAL
     (847) 639-6400
L3: COILTRONICS CTX33-4
Q1, Q2: MOTOROLA D45C1
     (800) 441-2447
T1: COILTRONICS CTX-02-13949-X1
○: FERRONICS FERRITE BEAD 21-110J

**Figure 140. A 10W, Low Noise, 5V to 12V Converter: Q1–Q2 Provide 5A Output Capacity while Preserving the LT1533's Voltage/Current Slew Control. Efficiency is 68%. Higher Input Voltages Minimize Follower Loss, Boosting Efficiency Above 71%**

Notes:

1  Witt, Jeff. The LT1533 Heralds a New Class of Low Noise Switching Regulators. *Linear Technology* VII:3 (August 1997).

2  Williams, Jim. *LTC Application Note 70: A Monolithic Switching Regulator with 100μV Output Noise.* October 1997.

3  The term "cascode," derived from "cascade to cathode," is applied to a configuration that places active devices in series. The benefit may be higher breakdown voltage, decreased input capacitance, bandwidth

improvement or the like. Cascoding has been employed in op amps, power supplies, oscilloscopes and other areas to obtain performance enhancement.

4  This circuit derives from a design by Jeff Witt of Linear Technology Corp.

5  Operating the slew loops from follower base current was suggested by Bob Dobkin of Linear Technology Corp.

## Regulators—Switching (Multioutput)



V$_{IN}$ 5.2–28V; SWITCHING FREQUENCY = 200kHz
5V-3A / 3.3V-3.5A / 12V-120mA
M1–M4 = Si4412DY
T1 = DALE LPE-6562-A092; 15μH; 1:2.2

L1 = SUMIDA CDRH125-100MC 10μH
INPUT AND OUTPUT CAPACITORS ARE AVX-TPS SERIES
HAVING A MAXIMUM ESR SPECIFICATION

**Figure 142. LTC1538-AUX Provides 3.3V/3.5A, 5V/3A, 12V/120mA and 5V/20mA Standby Power**

## LTC1538-AUX: A NEW ADDITION TO LTC'S ADAPTIVE POWER CONTROLLER FAMILY
*by Steve Hobrecht*

### Notebook Computer Power Solution

The circuit shown in Figure 142 is a power solution for a portable notebook computer. The switching controllers provide 5V at 3A, 3.3V at 3.5A and a regulated 12V/120mA output using the auxiliary regulator. See the LTC1538-AUX/ LTC1539 data sheet for techniques illustrating how to generate other voltage and current combinations using the auxiliary regulator. The circuit provides a standby 5V output to power a keyboard controller. The keyboard controller has the ability to control the run/soft-start

(RUN/SS1 and RUN/SS2) pins of the LTC1538-AUX using simple logic gates. The turn-on sequence is determined by the ratio of $C_{SS1}$ to $C_{SS2}$. The secondary winding of transformer T1 develops a somewhat unregulated voltage due to the loading on $V_{OUT1}$. The SFB1 control pin will keep the minimum voltage of the secondary output at approximately 13V, but the peak voltage is affected by the loading and leakage inductance of the transformer. The auxiliary regulator will keep the 12V supply well within its normal ±5% specified tolerance. Short-circuit protection can be added to this circuit if required, but it is assumed here that the protection will only be required at the user PCMCIA interface and will therefore be taken care of as part of the interface and not duplicated here.

## HIGH EFFICIENCY, LOW POWER, 3-OUTPUT DC/DC CONVERTER
*by John Seago*

The recent proliferation of battery powered products has created a lot of interest in low power, high efficiency DC/ DC converter designs. These products are small, lightweight and portable, so space for bulky batteries is limited. Often, operating time between charges is a major selling feature, making the efficient use of battery power very important. Since many products cannot function with a single regulated voltage, multiple-output DC/DC converters are required.

Although developed for somewhat higher power levels, the single output LTC1435 can be used in applications requiring a very efficient, very small, low power, multiple-output DC/DC converter (see Figure 143). This is accomplished through the use of an overwound buck inductor. With additional windings, the inductor can provide additional outputs, requiring only a diode and filter capacitor for each output. As with the less efficient flyback topology, the additional outputs are not as well regulated as the primary output, but the regulation is suitable for most applications.
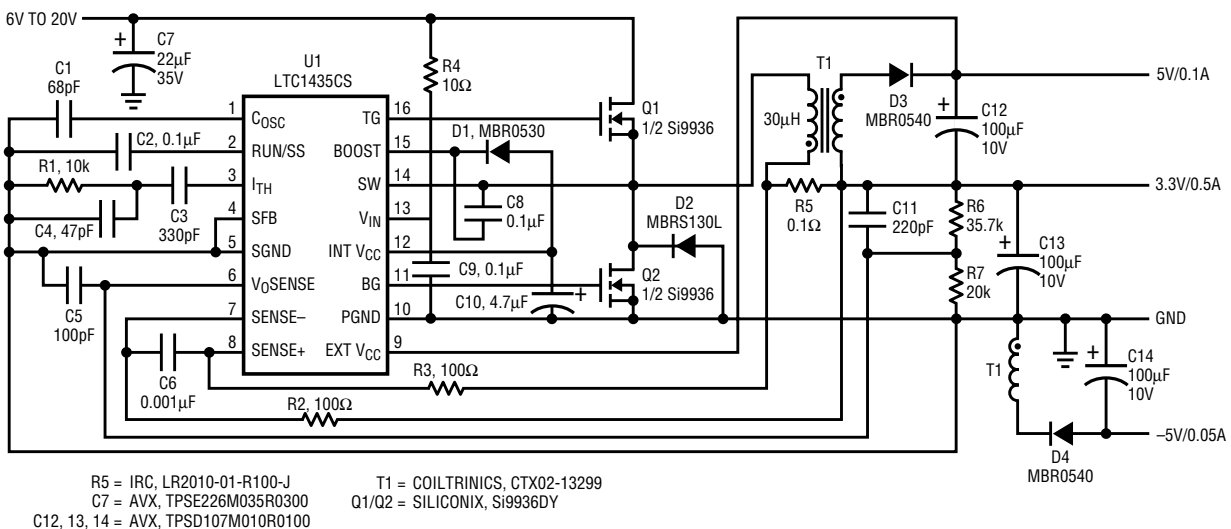


Figure 143. High Efficiency, 3-Output DC/DC Converter

The circuit of Figure 143 provides 3.3V at 0.5A, 5V at 0.1A and −5V at 0.05A, and has greater than 93% efficiency for test loads between 1.25W and 2.4W with a 6V input. Load and line regulation of the positive outputs are quite good. Each output voltage was measured with all output currents varied independently between 20% and 100% of their full load range, while the input voltage was varied from 6V to 20V. Table 3 shows the worst-case output voltages measured.

**Table 3. Worst-Case Output Voltages**

| Output | Minimum | Maximum |
|--------|---------|---------|
| 3.3V | 3.307V | 3.315V |
| 5V | 5.03V | 5.24V |
| −5V | −4.98V | −5.51V |

The buck regulator with an overwound inductor is a good solution for those applications that do not have large load current or line voltage variations. The smaller the load and line variations, the smaller the voltage variations on the overwound outputs. As a general rule, output voltage regulation is suitable for most applications if the switch duty cycle is kept between 15% and 50% and minimum load current is kept above 20% of maximum. Since load variation and line variation have an additive effect on output voltage, applications with relatively constant load current requirements can have a larger input voltage range and vice versa. For zero output current requirements, a small preload resistor can be used.

## DUAL-OUTPUT VOLTAGE REGULATOR
*by Peter Guan*

The LTC1266-3.3 and LTC1263, as shown in the schematic of Figure 144, are perfect complements for one another. The combination of these two parts provides two regulated outputs of 3.3V/5A and 12V/60mA from an input range of 4.75V to 5.5V. These two outputs are perfect for notebook and palmtop computers with microprocessors that burn several amps of current from a regulated 3.3V supply, flash memories that consume milliamps of current from a regulated 12V supply and interface and logic components that still run off the 5V supply. In fact, this quick and easy combination may well be the aspirin for many of the headaches caused by the rigorous power supply demands in today's electronics.

The LTC1263, using only four external components (two 0.47μF charge capacitors, one 10μF bypass capacitor and a 10μF output capacitor), generates the regulated 12V/60mA output from a 5V input using a charge pump tripler. During every period of the 300kHz oscillator, the two charge capacitors are first charged to $V_{CC}$ and then stacked in series, with the bottom plate of the bottom capacitor shorted to $V_{CC}$ and the top plate of the top capacitor connected to the output capacitor. As a result, the output capacitor is slowly charged up from 5V to 12V. The 12V output is regulated by a gated oscillator scheme that turns the charge pump on when $V_{OUT}$ is below 12V and turns it off when it exceeds 12V.

The LTC1266-3.3 then uses the 5V input along with the 12V output from the LTC1263 and various external components, including bypass capacitors, sense resistors and Schottky diodes, to switch two external N-channel MOSFETs and a 5μH inductor to charge and regulate the 3.3V/5A output. The charging scheme for this part, however, is very different from that of the LTC1263. The LTC1266-3.3 first charges the output capacitor by turning on the top N-channel MOSFET, allowing current to flow from the 5V input supply and through the inductor. By monitoring the amount of current flow in the inductor with a sense resistor, the 3.3V output is regulated by turning on and off the top and bottom N-channel MOSFETs to charge and discharge the output capacitor.

If we replaced the top external N-channel MOSFET with a P-channel, the LTC1266-3.3 could generate the same 3.3V/5A output without the help of the LTC1263. But, since N-channel MOSFETs have lower gate capacitance and lower $R_{DS(ON)}$, their higher efficiency at high currents more than compensates for the extra complexity in bringing in another higher input voltage, especially if that second input voltage is readily available.

**Figure 144. 5V to 3.3V/5A and 12V/60mA Supply**

Since both of these devices are very stingy on quiescent current, their combination is also very gentle to the main power supply, especially if that power supply is a battery. In standby mode, the LTC1263 and the LTC1266-3.3 have a total quiescent current of about 500µA. To conserve even more current, both of these parts can be put into shutdown mode by floating their shutdown pins or pulling them high. The total shutdown current is less than 40µA. When loaded, the LTC1263 has a 76% efficiency, whereas the

LTC1266-3.3 can squeeze out more than 90%. Together, with a 60mA load at the 12V output and a 5A load at the 3.3V output, the overall efficiency is 87%.

The LTC1266-3.3 is available in the 16-pin SO package and the LTC1263 is available in the 8-pin SO package. Together, these two parts provide an easy and efficient solution for multiple power supply demands.

## SWITCHER GENERATES TWO BIAS VOLTAGES WITHOUT TRANSFORMER
*by Jeff Witt*

LCD displays and CCD imaging circuits in today's portable products require several bias voltages of 10V to 20V at a few mA. When symmetric bipolar bias supplies are needed, the negative supply can be generated with a discrete charge pump operating from the power switch of the boost regulator that generates the positive supply. However, an asymmetric bipolar supply is typically required: for example 20V and –10V for LCD displays or 15V and –7.5V for CCDs. One possible solution is to add a linear regulator to the negative output; this adds cost and greatly reduces the efficiency of the switcher. Another possibility is a 2-output flyback circuit, but the added cost and bulk of a transformer make this solution unappealing. The circuit in Figure 145 avoids these penalties, producing 20V at 5mA and –10V at 5mA from 3.3V with 73% efficiency. The circuit uses standard surface mount parts.

The LT1316, a micropower Burst Mode switching regulator with an integrated 0.6A power switch, operates in an ordinary boost circuit to generate the 20V ($V_{OUT1}$) set by resistor divider R1 and R2. An internal comparator at the

FB pin regulates the output by gating the LT1316's oscillator. A charge pump (C2 and associated diodes) coupled to the LT1316's switch pin generates the negative output voltage. This negative output ($V_{OUT2}$) is monitored by the LT1316's low-battery detector through the resistor divider R3 and R4, using the positive 20V output as a reference. When the negative output falls below 10V, the low-battery detector output (LBO pin and lowest trace in Figure 146) turns Q1 on, enabling the charge pump and charging output capacitor C4. Note that the switch pin jumps between ground and ~10V during this period. Once the negative output has been charged enough to overcome the low-battery detector's hysteresis, Q1 turns off and the switch pin is free to fly to 20V, charging the positive output.

This circuit can also operate directly from two alkaline or NiCd cells. Slightly higher peak currents are necessary; change R5, which determines the peak switch current of the LT1316, to 6.8kW and change L1 to 15mH.
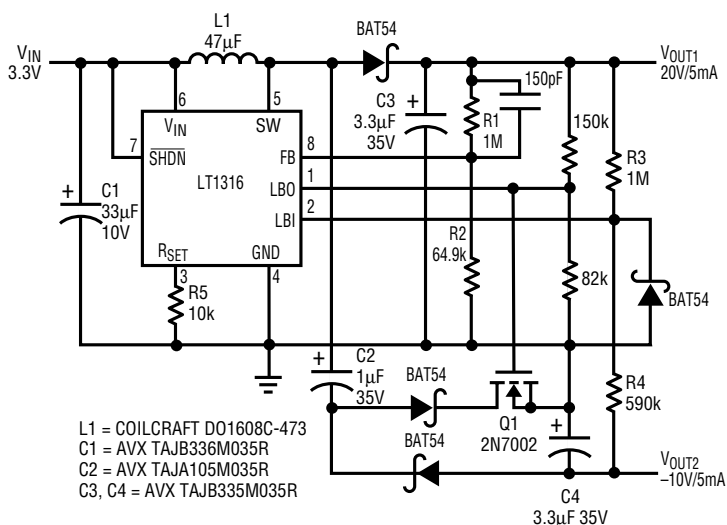


**Figure 145. By Gating the Charge Pump, this Circuit Generates a Regulated Negative Output with a Magnitude Different from that of the Positive Output**



0.1ms/DIV
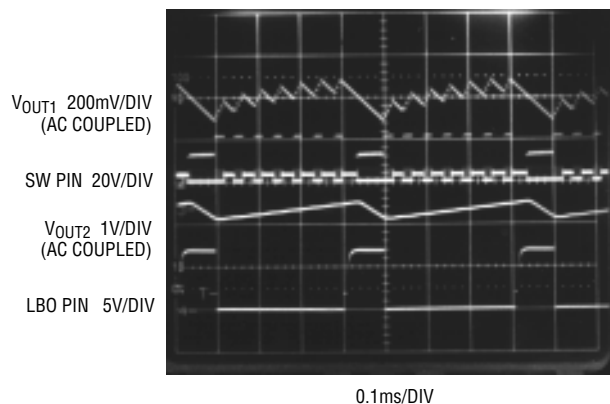
**Figure 146. Voltage Waveforms of Figure 145's circuit**

## NEW IC FEATURES REDUCE EMI
## FROM SWITCHING REGULATOR CIRCUITS
*by John Seago*

One disadvantage of using a switching regulator is that it generates electronic noise, known as EMI (electromagnetic interference). This noise can be conducted or radiated, and it can affect other circuits in your product or interfere with the operation of nearby products. The LTC1436-PLL, LTC1437, LTC1439 and LTC1539 have features that can be used to suppress this interference.

Frequently, EMI problems don't show up until the integration phase of product development. By using this EMI suppression capability, a resistor or capacitor value change may be all that is required to solve an interference problem. The LTC1436-PLL shown in the circuit of Figure 147 produces a switched 5V, 3A output and a 3.3V, 0.1A linear output. The circuit is configured to provide either switch-frequency synchronization or switch-frequency modula-

tion. Also, transistor Q3 ensures constant frequency at very low output current levels, thus eliminating audio frequencies and maintaining high efficiency using the internal Adaptive Power™ circuitry.

### Switch-Frequency Synchronization

Switching regulator noise results from switching high currents on and off. This creates high energy levels at the switching frequency and all of its harmonics. A common EMI-control technique is to synchronize the switching frequency to an external clock so that all harmonic frequencies can be controlled. The LTC1436-PLL uses a phase-locked loop for synchronization to avoid the loss of slope compensation common to other synchronizing techniques. In addition, the input to the VCO in the phase-locked loop is available at the PLL LPF (phase-locked loop lowpass filter) pin, so that a lowpass filter can be used to control how fast the loop acquires lock.
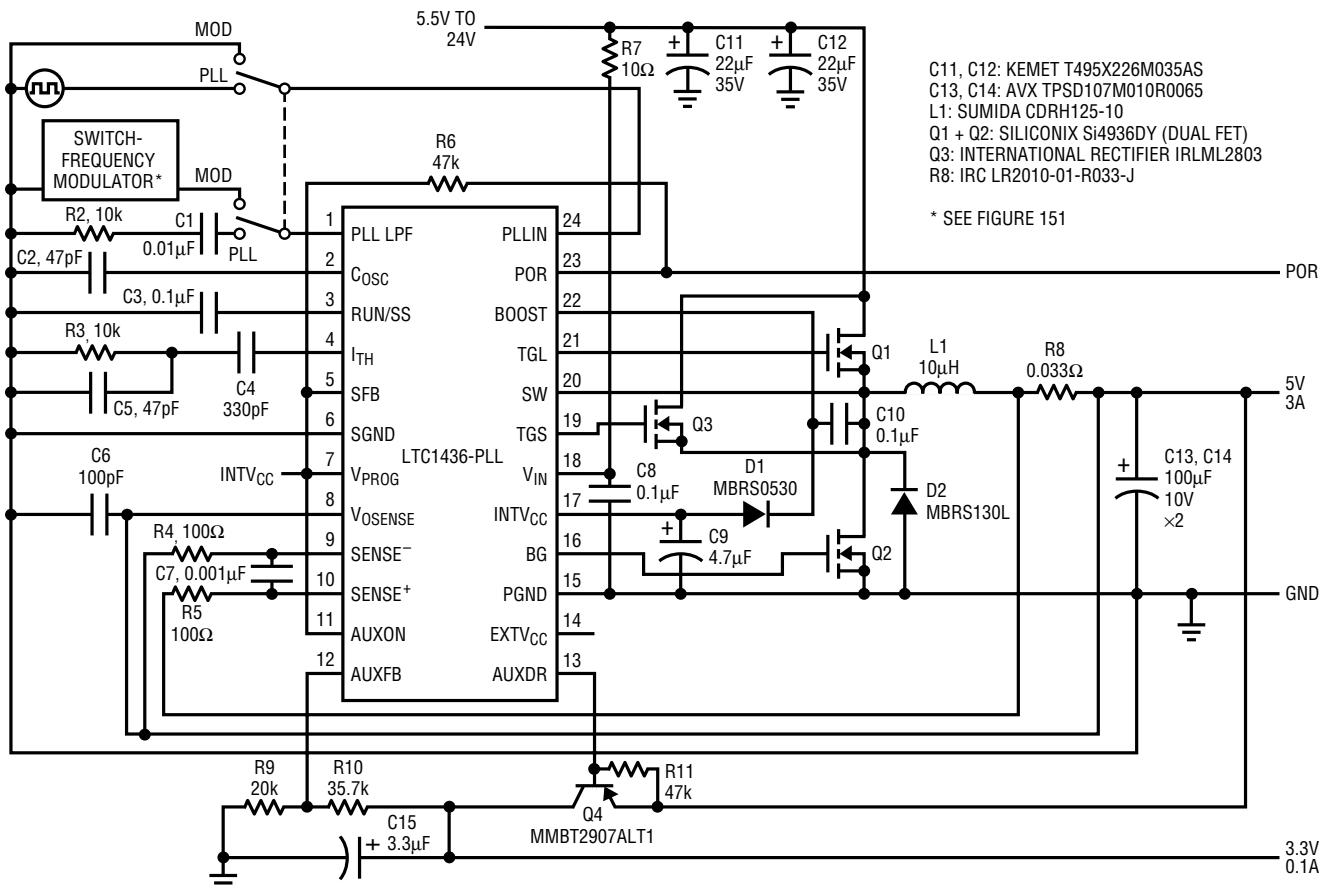


**Figure 147. 2-Output LTC1436-PLL Test Circuit**
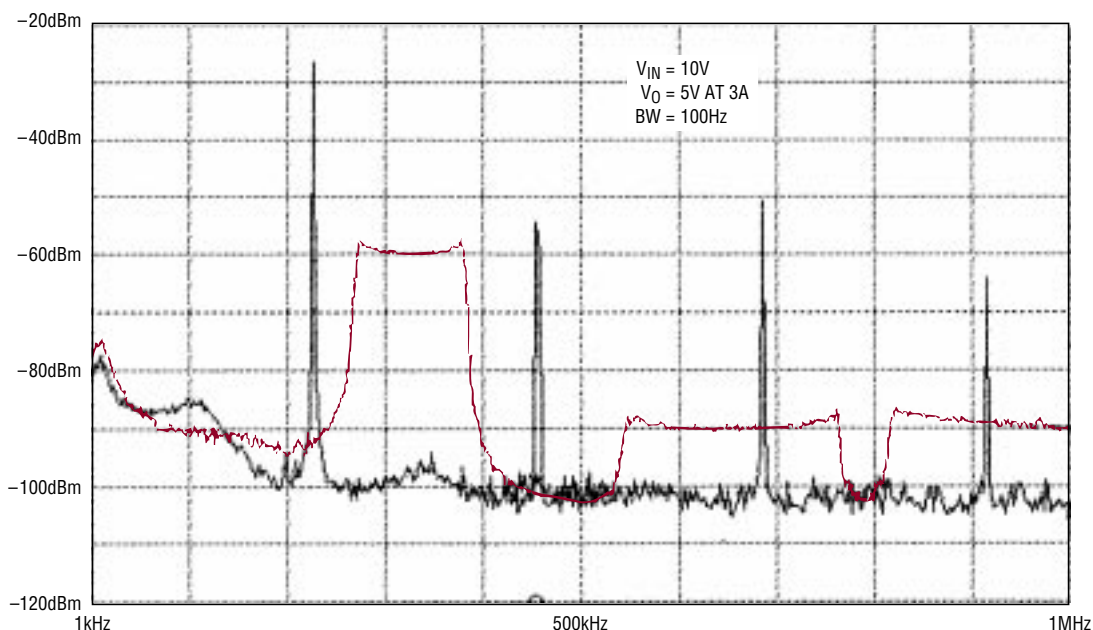
**Figure 148. Output Noise Before and After Switch-Frequency Modulation**
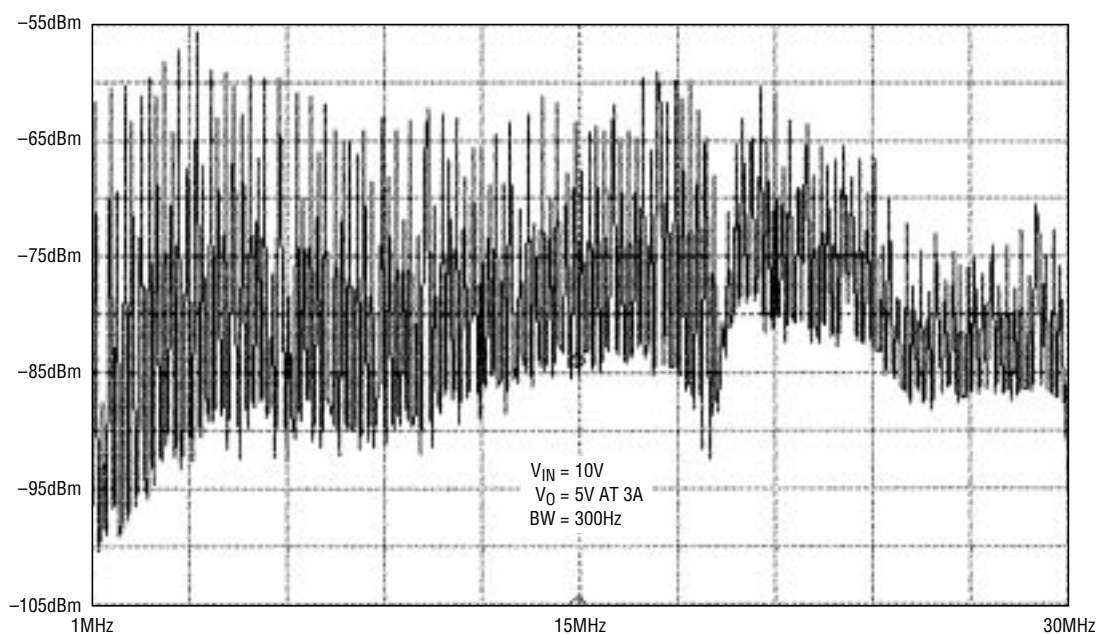


**Figure 149. Output High Frequency Noise Before Switch-Frequency Modulation**

## Switch-Frequency Modulation

Access to the VCO input also makes it possible to modulate the regulator's switching frequency. Through frequency modulation, the peak energy of the fundamental is spread over the frequency range of modulation, thus decreasing the peak energy level at any one frequency. This frequency spreading action increases with each harmonic, so that the second harmonic has twice the bandwidth and the third harmonic has three times the bandwidth until all the harmonics blend together, decreasing the signal strength at all frequencies. This can be seen in the spectrum analyzer plots shown in Figures 148–150.
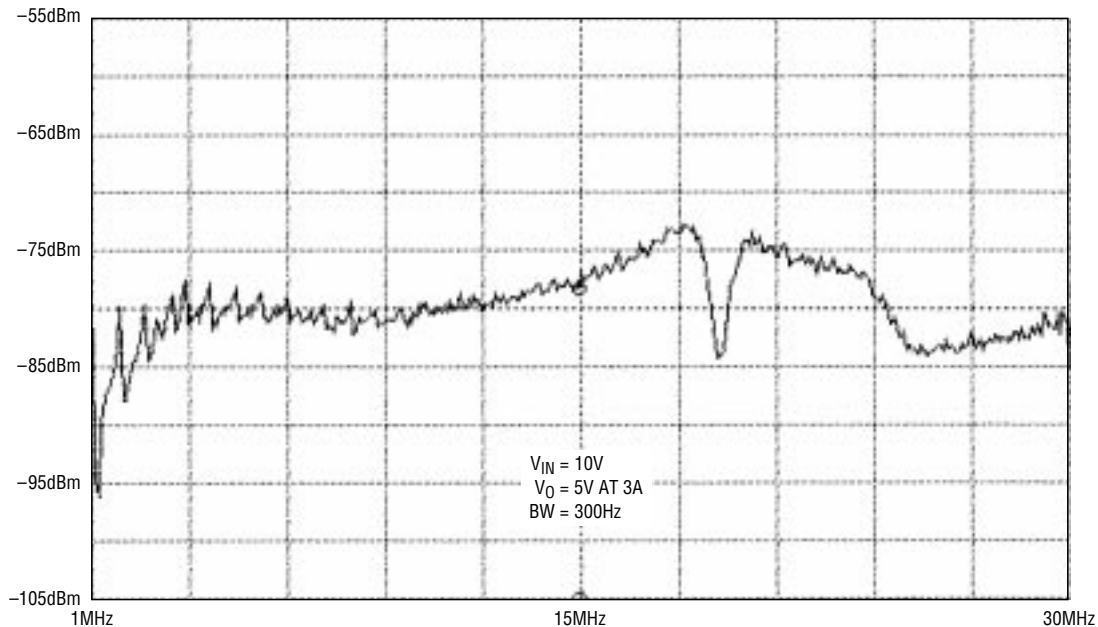
**Figure 150. Output High Frequency Noise After Switch-Frequency Modulation**

Figure 148 shows the full load output noise level from the circuit of Figure 147, before and after switch-frequency modulation. The black trace shows the normal output noise from 1kHz to 1MHz with the VCO at minimum frequency, whereas the colored trace shows output noise after modulation around the center frequency. The 228kHz unmodulated switch-frequency output noise decreased more than 30dB through modulation between 270kHz and 370kHz. Figures 149 and 150 show a 10dB to 15dB attenuation in full-load output voltage noise from 1MHz to 30MHz after modulation.

The VCO in the LTC1436-PLL has an input range from 0V to 2.4V. As shown in Figure 151, the switch frequency can be modulated at least ±30% around the center frequency $f_O$. The ideal modulating signal varies an equal amount above and below the center frequency voltage of 1.2V, with a constant slope. The reference circuit of Figure 152 develops a 100Hz sawtooth voltage from 0.9V to 1.5V that modulates the LTC1436-PLL in Figure 147 to generate the plots shown in Figures 148–150. Modulator circuit complexity is largely determined by functional requirements. For most applications, a precision modulating signal is not required, because high order harmonics blend together. Consequently, modulating frequency, slope and peak-to-peak voltage are not critical.
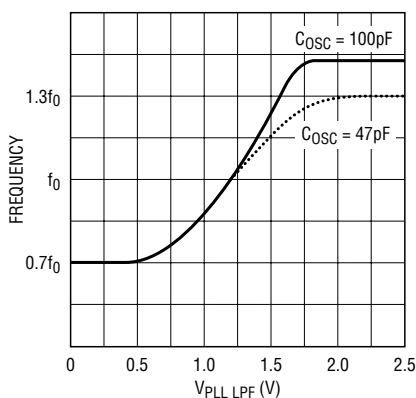


**Figure 151. Operating Frequency vs $V_{PLL\ LPF}$**
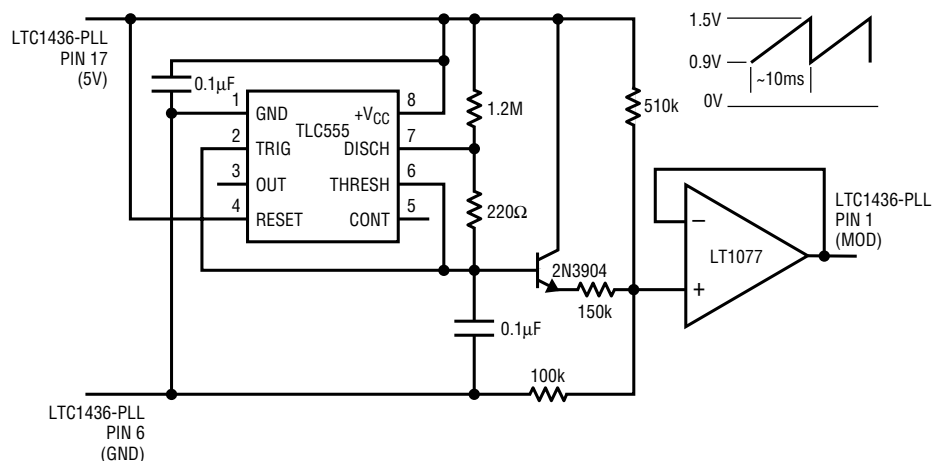


**Figure 152. Switch-Frequency Modulator**

Figure 153. Audio Frequencies in Output Noise during
Cycle-Skipping Operation



Figure 154. Output Noise with Adaptive Power Operation

## Audio Frequency Suppression

The Adaptive Power feature of the LTC1436-PLL significantly reduces audio frequency generation, while maintaining good efficiency under very light load conditions. Figure 153 shows the audio frequencies generated by the highly efficient cycle skipping mode of the LTC1436-PLL. Figu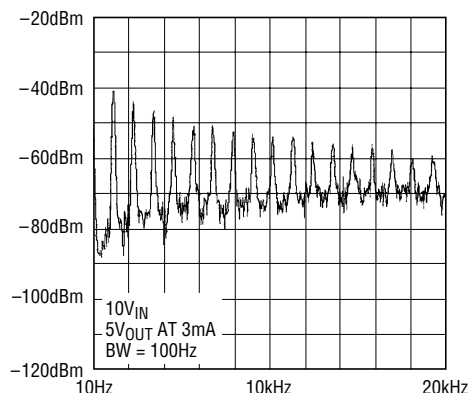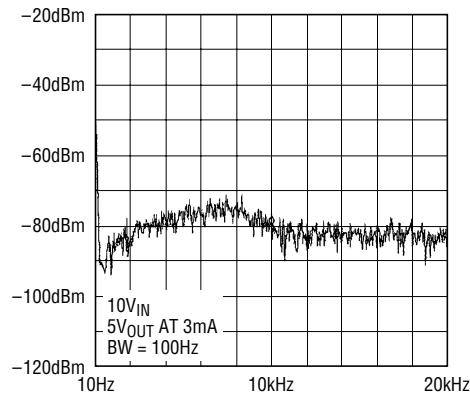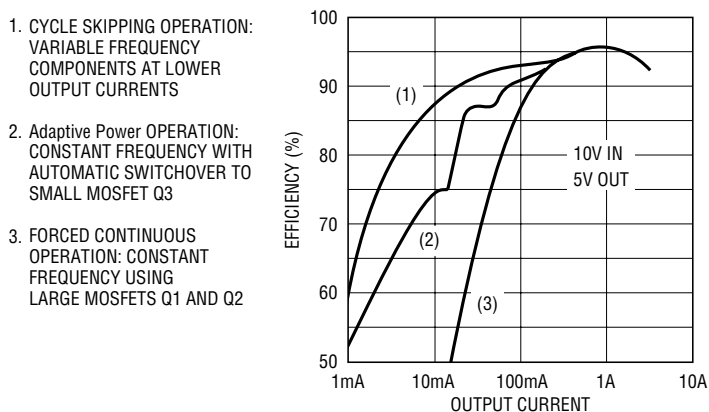re 154 shows the decrease in audio frequencies resulting from Adaptive Power operation. Figure 155 shows efficiency curves of both the cycle skipping and Adaptive Power modes along with the traditional, forced continuous mode of operation.

Cycle skipping is the most efficient mode during light-load operation, where the output capacitor supplies load current most of the time and is replenished by bursts of energy at a rate determined by the load. When load current is low enough, the burst rate falls into the audio-frequency range, which can cause problems. With the addition of Q3, an inexpensive SOT-23 size MOSFET, the Adaptive Power circuitry inside the LTC1436-PLL takes control during light load conditions, turning off high current MOSFETs Q1 and Q2. Q3 and D2 are then used in a conventional constant frequency buck mode, eliminating the power loss caused by charging and discharging the large input capacitance of both power MOSFETs.

The conventional way of avoiding audio-frequency interference is the forced current mode, where both high current MOSFETs continue to operate at full frequency and normal duty cycle under all load conditions. This causes the peak-to-peak inductor current to flow, even under no load conditions. The synchronous buck topology allows the top switch, Q1, to put current into the output capacitor, followed by the bottom switch, Q2, taking current out of the output capacitor while regulating the output voltage under no-load conditions. Although constant frequency is maintained, high current $I^2R$ losses and high gate charge losses continue under light load conditions. Forced-current operation is useful for fast transient response required for high di/dt loads like the Intel Pentium® processor.

Cycle skipping, Adaptive Power and forced current operation are all available on the LTC1436-PLL, so that the best operating mode can be selected for each application.

1. CYCLE SKIPPING OPERATION: VARIABLE FREQUENCY COMPONENTS AT LOWER OUTPUT CURRENTS

2. Adaptive Power OPERATION: CONSTANT FREQUENCY WITH AUTOMATIC SWITCHOVER TO SMALL MOSFET Q3

3. FORCED CONTINUOUS OPERATION: CONSTANT FREQUENCY USING LARGE MOSFETS Q1 AND Q2



Figure 155. Efficiency Curves for Light Load Currents

Regulators—Switching (Micropower)

## POWER MANAGEMENT AND HIGH EFFICIENCY SWITCHER MAXIMIZE NINE-VOLT BATTERY LIFE
*by LTC Applications Staff*

The LTC1174 (3.3V, 5V and adjustable versions) can convert a 9V battery source to system power with very high efficiency. Efficiency is over 90% at load currents from 20mA to 425mA and over 85% at a load current of 4mA. For a given load, maximum battery life can be obtained by minimizing shutdown current during system shutdown and maximizing converter efficiency during operation. A single control line to the LTC1174 can be used to select shutdown mode or operational mode, as required.

For this circuit (Figure 156), power-up is initiated by a low level signal on the NAND gate. This signal could come from any front-panel switch or from an external interrupt signal. The system power is turned off by means of a low level signal from a controller/logic device. In either case, the control signal to the LTC1174 must be latched. (A latched turn-off signal ensures a known state on the LTC1174 shutdown pin during the collapse of the 5V supply.)

The CD4012 and CD4013 are powered from the battery; the 2N2222 provides simple level shifting to the battery rail. R1 and C1 ensure that the circuit remains in power-down mode during battery replacement. The circuit shown here provides approximately 90% efficiency at 250mA load current, and consumes less than 1µA shutdown current. Turn-on and turn-off transitions are very clean.



**Figure 156. Schematic Diagram of High Efficiency DC/DC Converter**



**Figure 157. Load Profile**

## LT1307 MICROPOWER DC/DC CONVERTER ELIMINATES ELECTROLYTIC CAPACITORS
*by Steve Pietkiewicz*

The relentless push towards increasing miniaturization in portable electronic products has created the need for small, high speed, low voltage DC/DC converter ICs. The LT1307 combines a current-mode, fixed frequency PWM architecture with Burst Mode™ micropower operation to maintain high efficiency at light loads. It uses small, low cost ceramic capacitors for both input and output, minimizing board area. By employing fixed frequency 575kHz switching the LT1307 keeps spectral energy out of the 455kHz band. Dense, high speed bipolar process technology enables the LT1307 to fit in the MSOP package, and micropower circuitry results in just 60μA quiescent current at no load. Conversion efficiency exceeds 80%, and the device also includes a low battery detector.

### Single-Cell Boost Converter

A complete single-cell to 3.3V converter is shown in Figure 158. The circuit generates 3.3V at up to 75mA from a 1.0V input. The 10μF ceramic output capacitor can be obtained from several vendors. Efficiency, detailed in Figure 159, peaks at 80% and exceeds 70% over the 1:500 load range of 200μA to 100mA at a 1.25V input. Changing the value of R1 to 1.87MΩ moves the output to 5V. Efficiency of the 5V output converter is depicted in Figure 160. Figure 161's
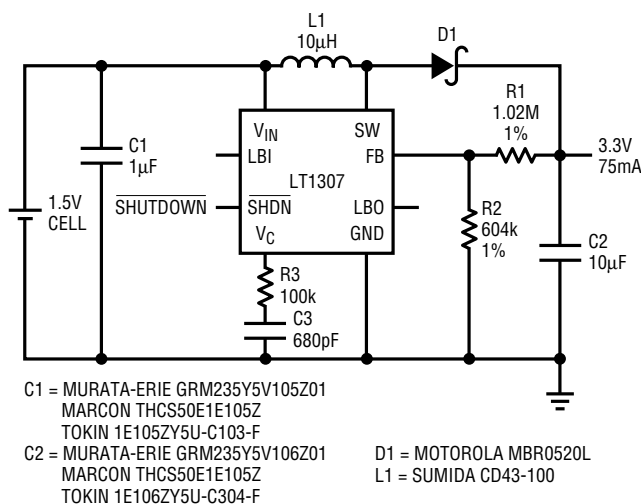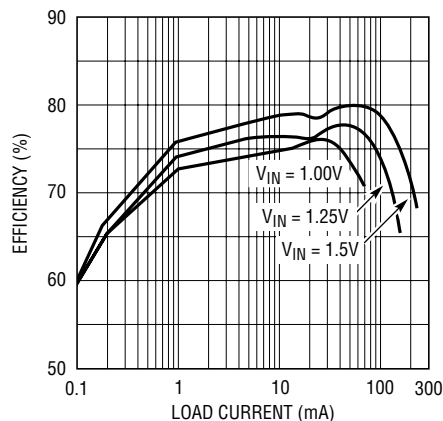


**Figure 159. 3.3V Efficiency**

oscillograph shows output voltage and inductor current as the load current is stepped from 5mA to 55mA, revealing substantial detail about the operation of the LT1307. With a 5mA load, $V_{OUT}$ (top trace) exhibits a ripple voltage of 60mV at 4kHz. The device is in Burst Mode at this output current level. Burst Mode operation enables the converter to maintain high efficiency at light loads by turning off all circuitry inside the LT1307 except the reference and error amplifier. When the LT1307 is not switching, quiescent current decreases to 60μA. When switching, inductor current (middle trace) is limited to approximately 100mA. Switching frequency inside the "bursts" is 575kHz. As the load is stepped to 55mA, the device shifts from Burst Mode to constant switching mode. Inductor current increases to about 300mA peak and the low frequency Burst Mode ripple goes away. R3 and C3 stabilize the loop.
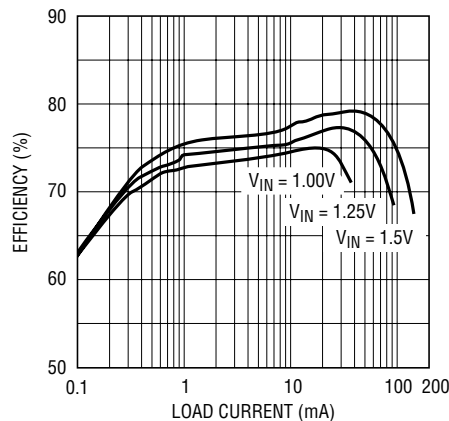


C1 = MURATA-ERIE GRM235Y5V105Z01
MARCON THCS50E1E105Z
TOKIN 1E105ZY5U-C103-F
C2 = MURATA-ERIE GRM235Y5V106Z01
MARCON THCS50E1E105Z
TOKIN 1E106ZY5U-C304-F

D1 = MOTOROLA MBR0520L
L1 = SUMIDA CD43-100

**Figure 158. Single Cell to 3.3V Boost Converter Delivers 75mA at 1.0V Input. Changing R1 to 1.87M Moves the Output to 5V**
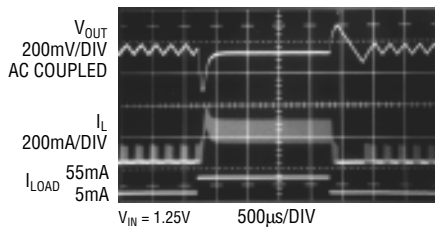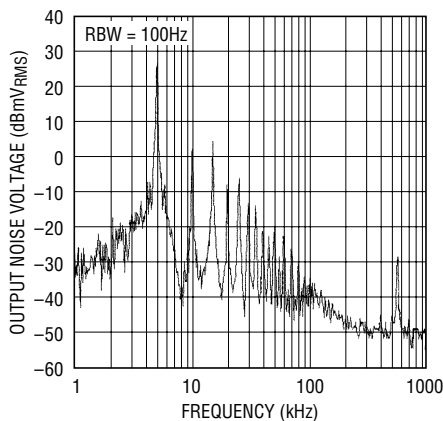


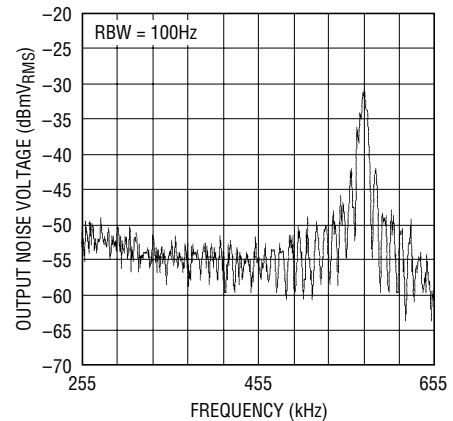**Figure 160. Efficiency at 5V Output**

**Figure 161. Transient Response with 5mA to 55mA Load Step**

## DC/DC Converter Noise Considerations

Switching regulator noise is a significant concern in many communications systems. The LT1307 is designed to keep noise energy out of the 455kHz band at all load levels while consuming only 60μW–100μW at no load. At light load levels, the device is in Burst Mode, causing low frequency ripple to appear at the output. Figure 162 details spectral noise directly at the output of Figure 158's circuit in a 1kHz to 1MHz bandwidth. The converter supplies a 5mA load from a 1.25V input. The Burst Mode fundamental at 5.1kHz and its harmonics are quite evident, as is the 575kHz switching frequency. Note, however, the absence of significant energy at 455kHz. Figure 163's plot reduces the frequency span from 255kHz to 655kHz with a 455kHz center. Burst Mode low frequency ripple creates sidebands around the 575kHz switching fundamental. These sidebands have low signal amplitude at 455kHz, measuring –55dBmV$_{RMS}$. As load current is further reduced, the Burst Mode frequency decreases. This spaces the sidebands around the switching frequency closer together, moving spectral energy further away from 455kHz. Figure 164 shows the noise spectrum of the converter with the
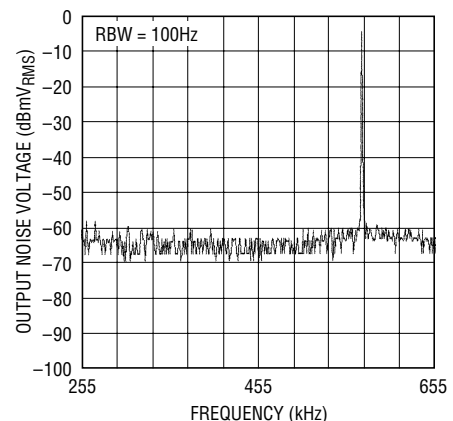


**Figure 163. Span Centered at 455kHz Shows –55dBmV$_{RMS}$ (1.8μV$_{RMS}$) at 455kHz. Burst Mode Creates Sidebands 5.1kHz Apart around the Switching Frequency Fundamental of 575kHz**

load increased to 20mA. The LT1307 shifts out of Burst Mode, eliminating low frequency ripple. Spectral energy is present only at the switching fundamental and its harmonics. Noise voltage measures –5dBmV$_{RMS}$ or 560μV$_{RMS}$ at the 575kHz switching frequency, and is below –60dBmV$_{RMS}$ for all other frequencies in the range. By combining Burst Mode with fixed frequency operation, the LT1307 keeps noise away from 455kHz, making the device ideal for RF applications where the absence of noise in the this band is critical.

Output filtering can reduce output conducted noise. Figure 158's circuit, supplying a 50mA load at 3.3V from a 1.3V source, is shown with an output filter (R4 and C4) in Figure 165. The lowpass filter created by R4 and C4 places a pole at 34kHz, reducing high frequency spikes considerably.



**Figure 162. Spectral Noise Plot of 3.3V Converter Delivering 5mA Load; Burst Mode Fundamental at 5.1kHz is 23dBmV$_{RMS}$ or 14mV$_{RMS}$**



**Figure 164. With the Converter Delivering 20mA, Low Frequency Sidebands Disappear. Noise is Present Only at the 575kHz Switching Frequency**
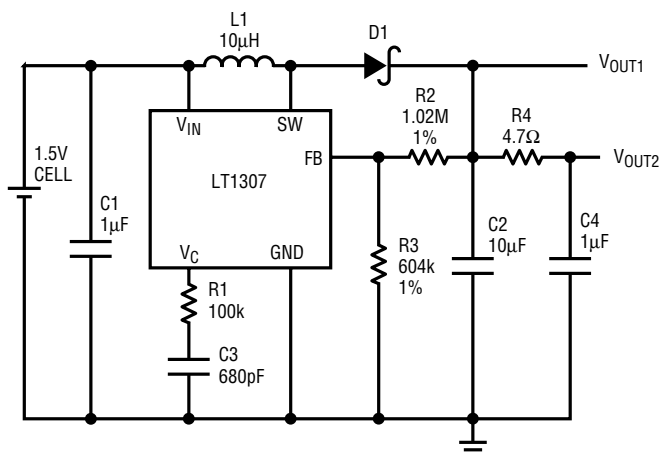
**Figure 165. Figure 158's Circuit with Output Filter R4/C4**

Viewed in a 50MHz bandwidth, the filter reduces switching spikes from about 10mV$_{P-P}$ to about 1mV$_{P-P}$, as detailed in Figure 166. Beware, though; the oscilloscope used in Figure 166's oscillograph (a Tektronix Type 547) is helping with the filtering by attenuating frequencies above 50MHz. Figure 167 shows the same circuit viewed on a 400MHz oscilloscope. The filter still attenuates but the magnitude of switching noise is far higher (140mV$_{P-P}$ unattenuated). A small amount of copper trace can be used in place of the resistor if the attendant voltage drop is unacceptable. A surprisingly small amount of trace is needed to create an effective filter; a PC trace of 1 oz. copper, 1 inch long by 10 mils wide, has an inductance of 29nH. Inductive reactance at 50MHz ($2\pi fL$) is 9.1$\Omega$. A combination of copper trace and 0.1$\mu$F ceramic capacitors will reduce high frequency spikes to acceptable levels in most systems.
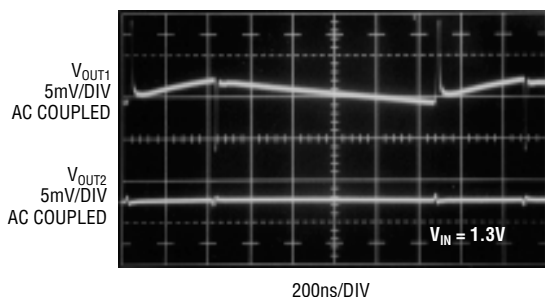
## LED Driver

LEDs require current source drive. Typically, a 5V supply with a series resistor to limit current is used to power the LED. Although simple, this approach has poor efficiency and requires a voltage source higher than the 2V–3V forward drop of most LEDs. Additionally, each LED requires its own ballast resistor. Figure 168's circuit uses the LT1307 configured as a current source to drive a series-connected pair of LEDs from a single-cell input. The IC's low battery detector monitors the voltage across sense resistor R1. LBO drives Q1; this provides correct phasing to the V$_C$ pin. Q1 and R2 drive the V$_C$ pin, overriding the internal error amplifier. With 200mV across R1, 25mA flows through the LED pair. C3 provides frequency compensation. For proper operation, the circuit must always supply enough power so as to not enter Burst Mode operation. This precludes driving most single LEDs (high brightness blue LEDs have a forward drop of 3.4V and can be driven singly). In shutdown mode, the circuit draws a only few microamperes. Start-up sequencing is detailed in Figure 169. The voltage at LBI stabilizes in about 200$\mu$s with minimal overshoot and ringing. The Lumex "Mega-Brite" red LEDs specified in Figure 168 provide enough light to act as a flashlight, providing young children with a high technology toy. Mounted on a small PC board with a push-button switch, the circuit entertained my two children for hours. They are both satisfied LT1307 customers.



200ns/DIV

**Figure 166. V$_{OUT1}$ is Output Voltage at 10$\mu$F Capacitor C2; V$_{OUT2}$ is After 4.7$\Omega$/1$\mu$F Output Filter. Circuit Supplies 50mA; Oscilloscope Bandwidth is 50MHz.**
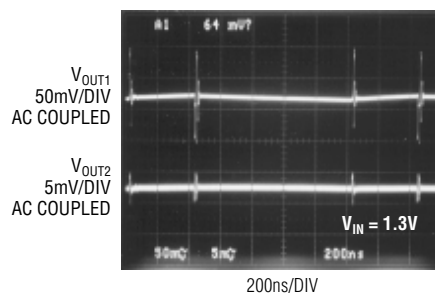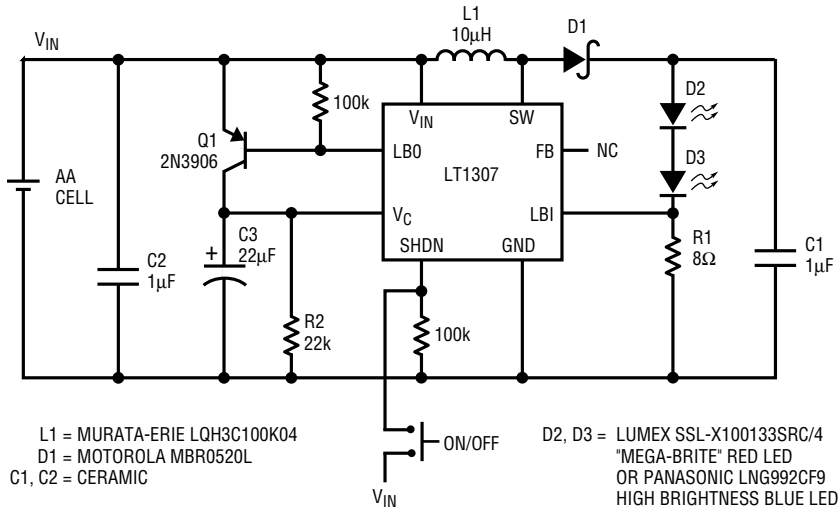


200ns/DIV

**Figure 167. A Faster Oscilloscope Shows More High Frequency Content at Both Outputs. Scope Bandwidth is 400MHz.**

L1 = MURATA-ERIE LQH3C100K04
D1 = MOTOROLA MBR0520L
C1, C2 = CERAMIC

D2, D3 = LUMEX SSL-X100133SRC/4
"MEGA-BRITE" RED LED
OR PANASONIC LNG992CF9
HIGH BRIGHTNESS BLUE LED

**Figure 168. Single-Cell LED Driver Supplies 25mA to LED String. Two Red LEDs Can Be Replaced by One Blue LED**
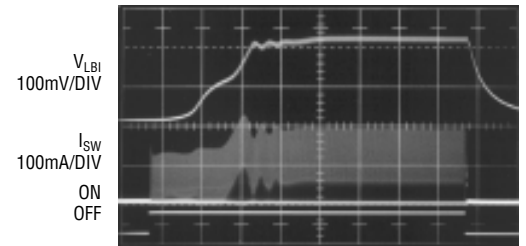


**Figure 169. Start-Up Response of LED Circuit. Many Switching Cycles Elapse before Current Flows in LEDs Because of C1 Charging**
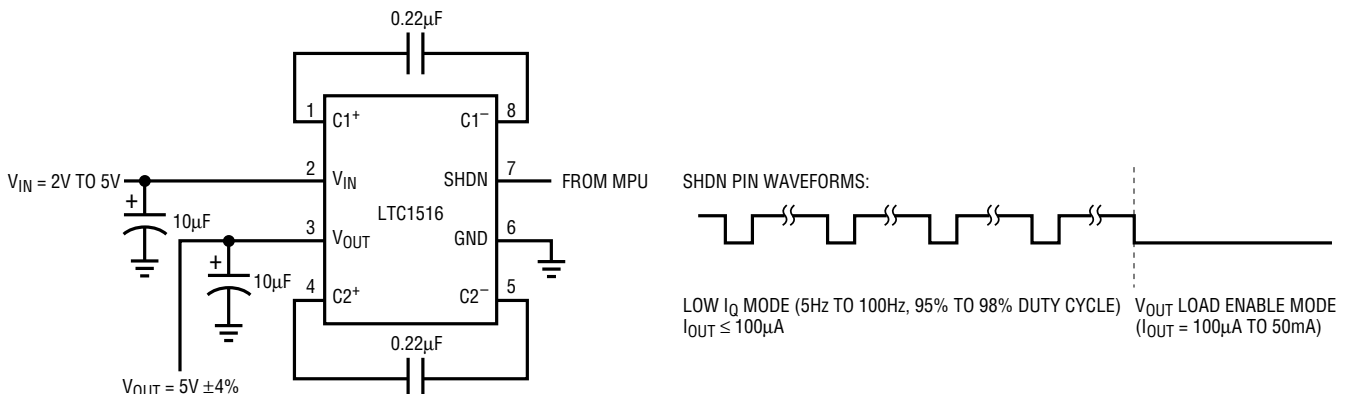
## AN ULTRALOW QUIESCENT CURRENT, 5V BOOST REGULATOR
*by Sam Nork*

Many battery-powered applications require an auxiliary 5V supply to power infrequently used circuitry, such as smart card readers, wireless i.d. tags, or the like. Keeping the 5V supply permanently active is desirable, since this eliminates timing delays and inrush currents due to supply start-up. The downside is that most 5V boost converters consume an unacceptable amount of quiescent current under no-load conditions. This problem is addressed by the SHDN features of the LTC1516 micropower, charge-pump DC/DC converter. Toggling the SHDN pin of the LTC1516 allows the 5V supply to remain in regulation with a typical no-load input current of less than 5µA. When the 5V output load is enabled, the part can supply up to 50mA of load current.

The LTC1516 produces a regulated 5V output from a 2V to 5V input In shutdown mode, the output load is disconnected from $V_{IN}$ and the quiescent current drops below 1µA.

When the output is in regulation, the internal sense resistor draws only 1.5µA (typical) from $V_{OUT}$. During no-load conditions, this internal load causes a droop rate of only 150mV per second on $V_{OUT}$ with $C_{OUT}$ = 10µF. Applying a 5Hz–100Hz, 95%–98% duty-cycle signal to the SHDN pin ensures that the circuit in Figure 170 comes out of shutdown frequently enough to maintain regulation during no-load (or low-load) conditions. Since the part is kept in shutdown mode for the majority of the time, the no-load quiescent current (see Figure 171) is approximately equal to $(V_{OUT} \times (1.5\mu A + I_{LOAD}))/(V_{IN} \times efficiency)$.



**Figure 170. Ultralow Quiescent Current (<5µA) Regulated Supply**

The LTC1516 must be taken out of shutdown mode for a minimum of 200µs to allow the internal sense circuitry to start up and keep the output in regulation. As the $V_{OUT}$ load current increases, the frequency with which the part is taken out of shutdown must also be increased to prevent $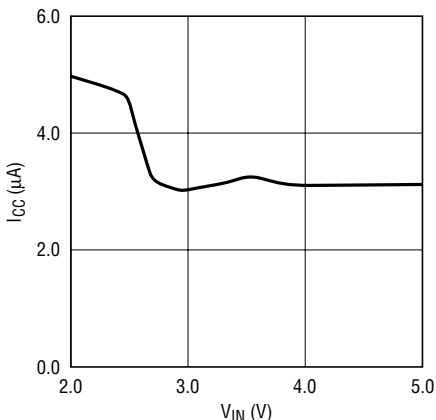V_{OUT}$ from drooping below 4.8V during the OFF phase (see Figure 172). A 100Hz, 98% duty cycle signal on the SHDN pin ensures proper regulation with load currents as high as 100µA. When load current greater than 100µA is needed, the SHDN pin must be forced low, as in normal operation. The typical no-load supply current for this circuit with $V_{IN}$ = 3V is only 3.2µA.



**Figure 171. No-Load $I_{CC}$ vs Input Voltage for Figure 170's Circuit**
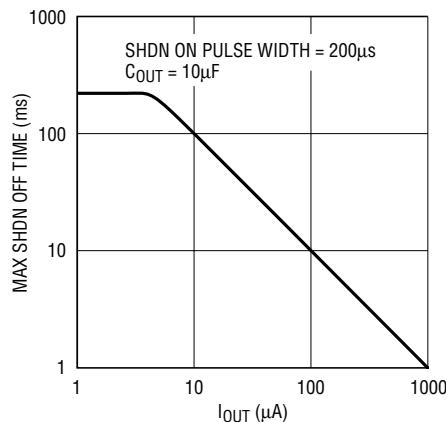


**Figure 172. Maximum SHDN OFF-Time vs Output Load Current for Ultalow $I_Q$ Operation**

## CAPACITIVE CHARGE PUMP POWERS 12V VPP FROM 5V SOURCE
*by Mitchell Lee*

The LTC1263, a regulating charge pump tripler, converts a 5V input to a regulated 12V, 60mA output. No inductors are required; charge pumps operate with capacitors only. Figure 173 shows the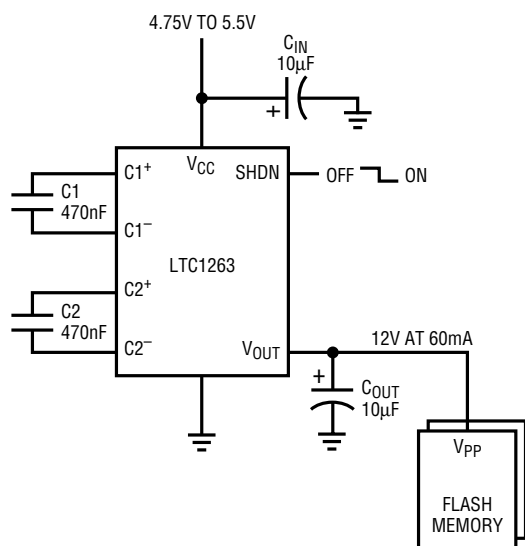 LTC1263 configured to provide VPP for two flash memory chips. The "flying" capacitors in the charge pump, C1 and C2, are sized well within the surface mount ceramic range. $C_{IN}$ and $C_{OUT}$, as shown, are surface mount tantalum capacitors, such as Sprague 595D series. In the 10µF capacitance range, tantalum capacitors cost less than ceramic units. The chip operates by charging C1 and C2 in parallel across 5V and ground and then discharging them in series across 5V and the output. In theory, the output could reach 15V, but an internal regulation loop maintains the output at a constant 12V.

SHUTDOWN reduces the quiescent current of the LTC1263 to less than 1µA under logic control. In shutdown mode, the output is held at 5V by an internal 500Ω, $V_{CC}$-to-$V_{OUT}$ switch. Output-voltage fall time is guaranteed to be less than 15ms for the component values shown. Output rise time coming out of shutdown is guaranteed to be less than 800µs.

Designing a circuit to generate a split supply from a single 5V source is usually an unpleasant chore; one to be avoided at all costs. If load current requirements are modest, the LTC1263 can generate both 12V and –7V for op amps and biasing needs. Figure 174 shows how. The



**Figure 173. Programming Two Flash Chips with the LTC1263 Charge Pump: In Shutdown Mode, the Output is Held at 5V**
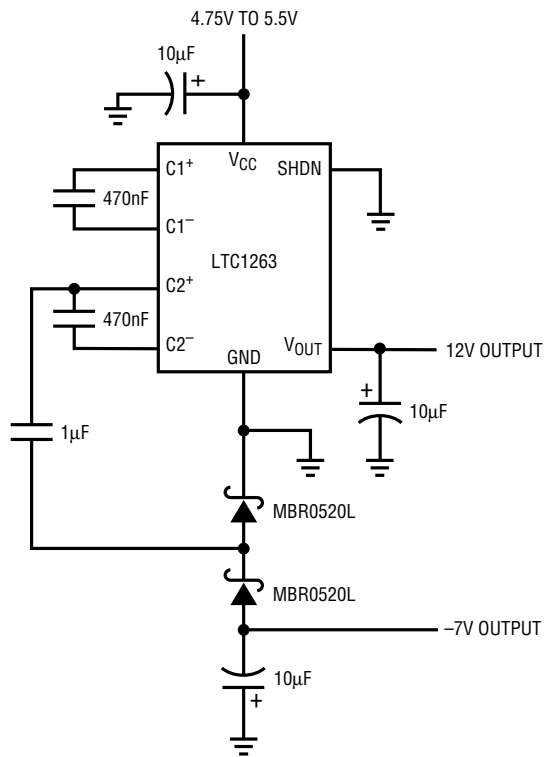
Figure 174. Split-Supply Generator:  Cross Regulation is Improved by Driving the Inverting Charge Pump from C2+.



Figure 175. Cross Regulation with a Constant 12V Load



Figure 176. Output Regulation with a Common Load

LTC1263 is connected in the usual way to produce a regulated, 12V output, but a 2 diode, 2-capacitor charge pump is added to the C2+ pin. This pin switches between $V_{CC}$ and $V_{OUT}$, swinging approximately 7$V_{P-P}$. The result is an outboard charge pump inverter with a –7V output.
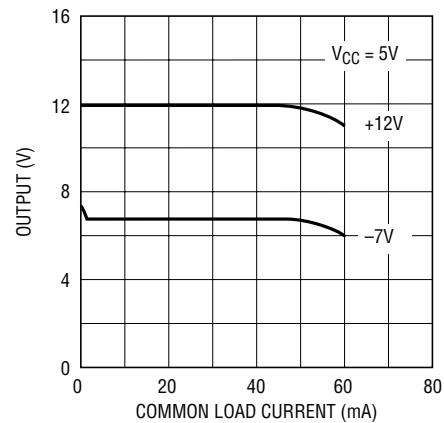
Schemes like this one often suffer from poor cross regulation. Although the inverting output is not directly regulated, the –7V load does affect the 12V output, thereby improving cross regulation (see Figure 175). The regulation with a common load (such as op amps) is shown in Figure 176.

## LTC1474 AND LTC1475 HIGH EFFICIENCY SWITCHING REGULATORS DRAW ONLY 10μA SUPPLY CURRENT
*by Greg Dittmer*

### Introduction

Maximizing battery life, one of the key design requirements for all battery-powered products, is now easier with Linear Technology's new family of ultralow quiescent current, high efficiency step-down regulator ICs, the LTC1474 and LTC1475. The LTC1474/LTC1475 are step-

down regulators with on-chip P-channel MOSFET power switches. These regulators draw only 10μA supply current at no load while maintaining the output voltage. Wide supply voltage range (3V–18V) and 100% duty cycle capability for low dropout allow maximum energy to be extracted from the battery, making the LTC1474/LTC1475 ideal for moderate current (up to 300mA) battery-powered applications.

Other features include Burst Mode™ operation to maintain high efficiency over almost four decades of load current, an on-chip low-battery comparator and a shutdown mode to further reduce supply current to 6μA. The LTC1475 provides on/off control with push-button switches for use in handheld products.

The LTC1474/LTC1475 are available in adjustable and fixed 3.3V/5V output voltage versions, in 8-pin MSOP and SO packages.

### 3.3V/200mA Step-Down Regulator

A typical application circuit using the LTC1474 is shown in Figure 177. This circuit supplies a 200mA load at 3.3V with an input supply range of 4V–18V (3.3V at no load). The 0.1Ω sense resistor reduces the peak current to about 285mA, which is the minimum level necessary to meet the 200mA load current requirement with a 100μH inductor. The peak current can be reduced further if a higher value inductor is used. Since the output capacitor dominates the output voltage ripple, an AVX TPS series low ESR (150mΩ) output capacitor is used to provide a good compromise between size and low ESR. With this capacitor the output ripple is less than 50mV.

### Efficiency Considerations

The efficiency curves for the 3.3V/200mA regulator at various supply voltages are shown in Figure 178. Note the flatness of the curves over the upper three decades of load current and that the efficiency remains high down to extremely light loads. Efficiency at light loads depends on low quiescent current. The curves are flat because all significant sources of loss except for the 10μA standby current—$I^2R$ losses in the switch, catch diode losses, gate charge losses to turn on the switch and burst cycle DC supply current losses—are identical during each burst cycle. The only variable is the rate at which the burst cycles occur. Since burst frequency is proportional to load, the loss as a percentage of load remains relatively constant. The efficiency drops off as the load decreases below about 1mA because the non-load-dependent 10μA standby current loss then constitutes a more significant percentage of the output power. This loss is proportional to $V_{IN}$ and thus its effect is more pronounced at higher $V_{IN}$.

### LTC1475 Push-Button On/Off Operation

The LTC1475 provides the option of push-button control of run and shutdown modes for handheld products. In contrast to the LTC1474's run/shutdown mode, which is controlled by a voltage level at the RUN pin (ground = shutdown, open/high = run), the LTC1475 run/shutdown mode is controlled by an internal S/R flip-flop that is set (run mode) by momentarily shorting the $\overline{ON}$ pin to ground and reset (shutdown mode) by a momentary ground at the LBI pin (see Figure 179). This provides simple on/off control with two push-button switches. The simplest implementation of this function is shown in Figure 180, with normally open push-button switches connected to the ON and LBI pins. Note that because the switch on LBI is normally open, it doesn't affect the normal operation of this input to the low-battery comparator. With a resistor
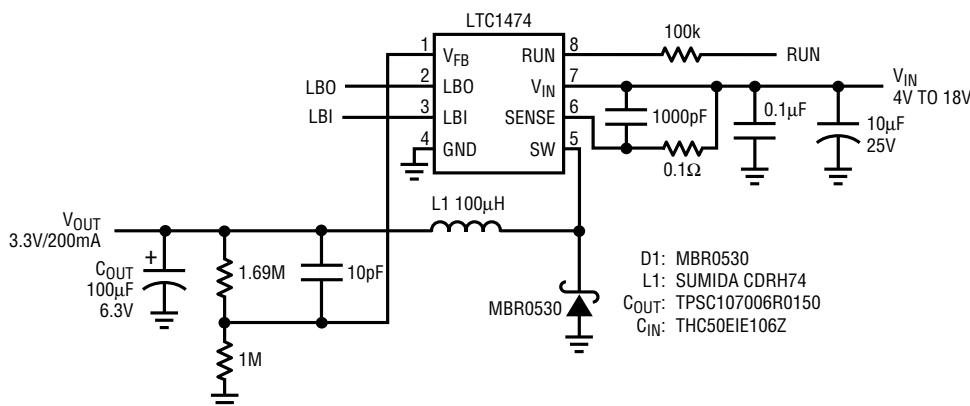
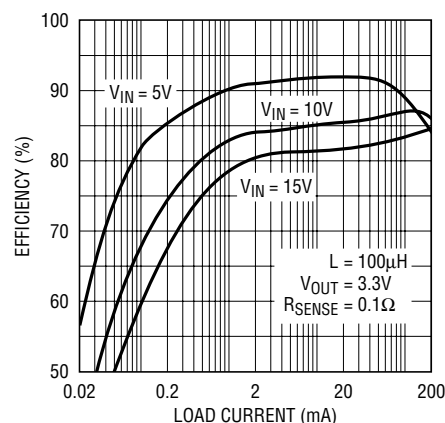

**Figure 177. LTC1474 3.3V/200mA Step-Down Regulator**
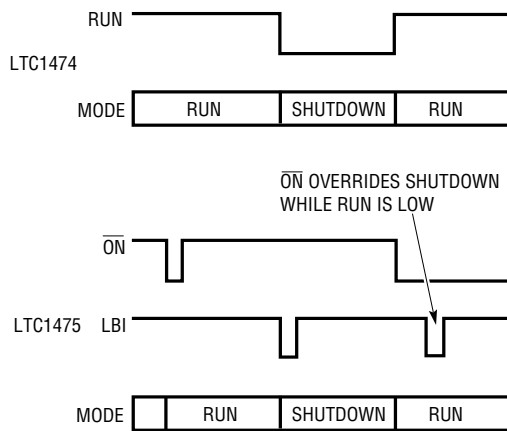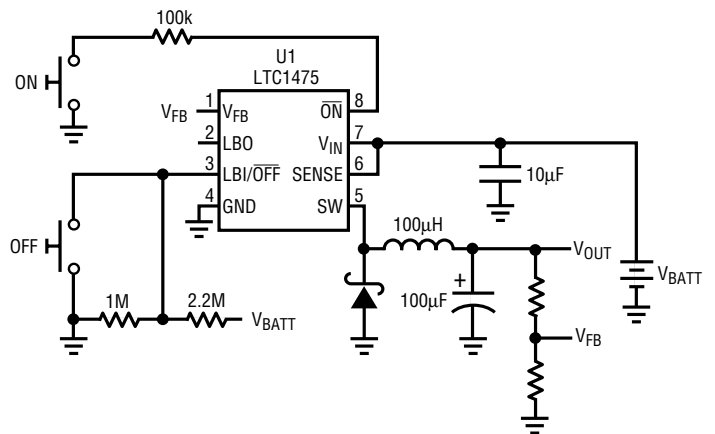


**Figure 178. Efficiency vs Load Current for Figure 177's Circuit**

**Figure 179. Comparison of RUN/SHUTDOWN Operation for the LTC1474 and LTC1475**



**Figure 180. LTC1475 Step-Down Regulator with Push-Button On/Off Control**

divider network connected to the LBI to monitor the input supply voltage level, the voltage at this pin will normally be above the low-battery trip threshold of 1.23V. When this pin is pulled below 0.7V by depressing the switch, the internal flip-flop is reset to invoke shutdown.

Figure 181 shows an example of push-button on/off control of a LTC1475 microcontroller application with a single push button. The push button is connected to the microcontroller as a discrete input so that the microcontroller can monitor the state of the push button. The LTC1475 LBI pin is connected to one of the microcontroller's open-drain discrete outputs so that it

can force the LTC1475 off when it detects a depressed push button. Because the LTC1475 supplies power to the microcontroller, once the microcontroller is off, it can no longer turn the LTC1475 back on. However, since the push button is also connected directly to the $\overline{ON}$ pin, the LTC1475 can be turned back on directly from the push button without the microcontroller. The LTC1475 then powers up the microcontroller. The discrete inputs of most microcontrollers have a reverse biased diode, D2, between the input and supply; thus a blocking diode with less than 1μA leakage is necessary to prevent the powered down microcontroller from pulling down on the $\overline{ON}$ pin.
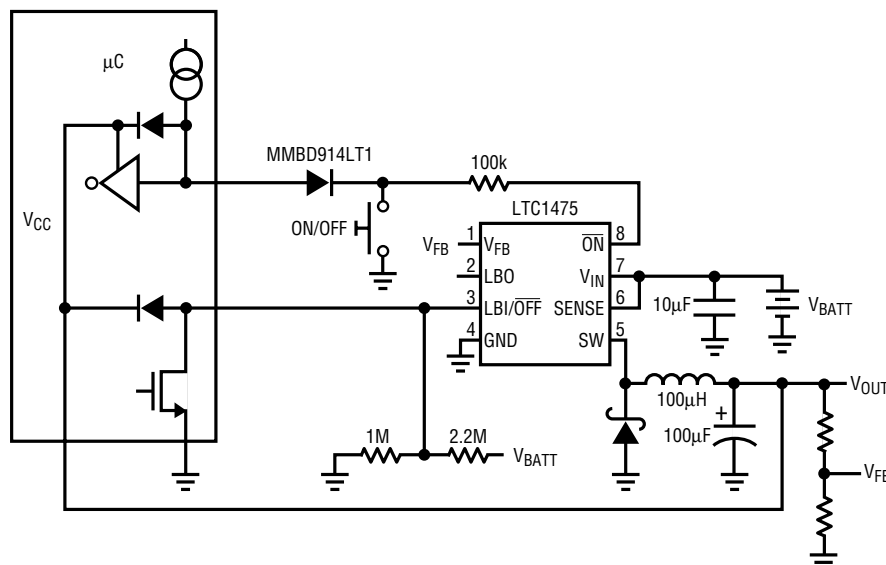


**Figure 181. A Single Push-Button Controls On/Off for the LTC1475 Regulator and Microcontroller**

## FREE DIGITAL PANEL METERS FROM THE OPPRESSIVE YOKE OF BATTERIES
*by Mitchell Lee*

Digital panel meters (DPMs) have dropped in price to well under $10 for 3-1/2 digit models, even in single-piece quantities. These make excellent displays for many instruments, but suffer from one major flaw: they require a floating power supply, usually in the form of a 9V battery. This renders inexpensive meters useless for most applications because no one wants multiple 9V batteries in their product.

The circuit shown in Figure 182 powers up to five meters from a single 1.8V to 6V source. The source need not be floating, yet all five outputs are fully floating, isolated and independent in every respect. The circuit consists of an LT1303 micropower, high efficiency DC/DC converter driving a 5-output flyback converter. An off-the-shelf surface mount coil, Coiltronics' VERSA-PAC™ VP1-0190, is used as the transformer. This device is hipot tested to 500V$_{RMS}$—more than adequate for most applications.

Feedback is extracted from the primary by Q1, which samples the flyback pedestal during the switch off time. Typical DPMs draw approximately 1mA supply current. The primary is also loaded with 1mA for optimum regulation and ripple. Primary snubbing components, a necessity in most flyback circuits, are obviated by the primary feedback rectifier and smoothing capacitor. Although this circuit has been set up for 9V output (9.3V, to be exact), some DPMs need 5V or 7V. Use a 4.3kΩ or 6.2kΩ resistor in place of R1 for these voltages. The output voltage is set by

R1 = (V$_{OUT}$ – 0.7)/1mA.

Do not attempt to regulate the output beyond 10V or you will exceed the maximum switch rating of the LT1303. The LT1111 is better suited for higher voltage applications.

Output ripple measures 200mV$_{P–P}$ and can be proportionately reduced by increasing the output capacitance. If more ripple is acceptable, the output capacitors can be reduced in value. A shutdown feature is available on the LT1303, useful where a "sleep" function is included to save power.

With each output loaded at 1mA, the input current is 16.5mA on a 5V supply. This figure rises to about 45mA on a 1.8V (2-cell) input. If the system is battery operated and if the battery voltage does not exceed 7V, operate the circuit directly from the battery for best efficiency. In line-operated equipment, use a regulated 5VDC or 3.3VDC supply.
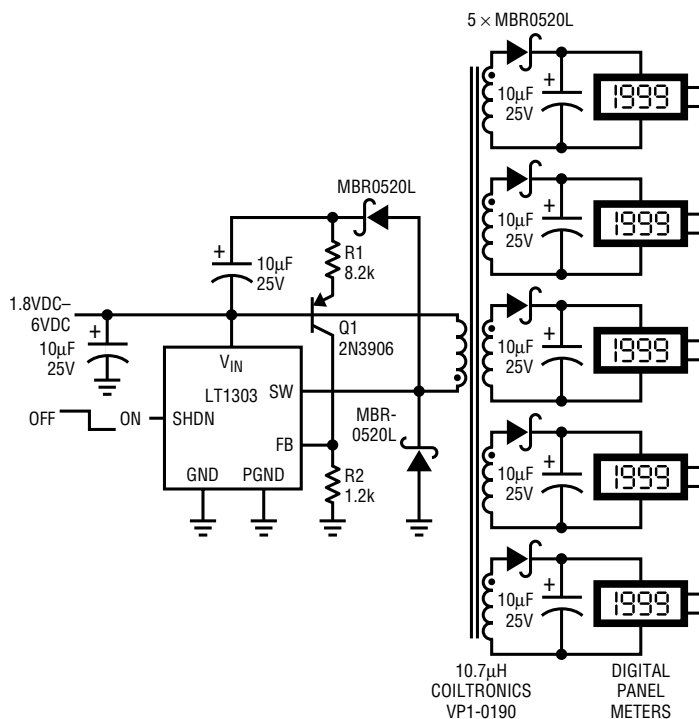


**Figure 182. LT1303 Flyback Regulator Provides Fully Floating and Isolated 9V Supplies to Five Independent Digital Panel Meters. Substitute 4.3k for R1 if 5V Meters are Used.**

# THE LTC1514/LTC1515 PROVIDE LOW POWER STEP-UP/STEP-DOWN DC/DC CONVERSION WITHOUT INDUCTORS
*by Sam Nork*

## Introduction

Many applications must generate a regulated supply from an input source that may be above or below the desired regulated output voltage. Such applications place unique constraints on the DC/DC converter and, as a general rule, add complexity (and cost) to the power supply. A typical example is generating 5V from a 4-cell NiCd battery. When the batteries are fully charged, the input voltage is around 6V; when the batteries are near end of life, the input voltage may be as low as 3.6V. Maintaining a regulated 5V output for the life of the batteries typically requires an inductor-based DC/DC converter (for example, a SEPIC converter) or a complex hybrid step-up/step-down solution. The LTC1514/LTC1515 family of switched capacitor DC/DC converters handles this task with only three external capacitors (Figure 183).

A unique architecture allows the parts to accommodate a wide input voltage range (2.0V to 10V) and adjust the operating mode as needed to maintain regulation. Hence, the parts can be used with a wide variety of battery and/or adapter voltages. Low power consumption ($I_Q = 60\mu A$ typ) and low parts count make the parts well suited for space-conscious low power applications, such as cellular phones, PDAs and portable instruments. The parts come in adjustable and fixed output-voltage versions and include additional features such as power-on reset capability (LTC1515 family) and an uncommitted comparator that is kept alive in shutdown (LTC1514 family).
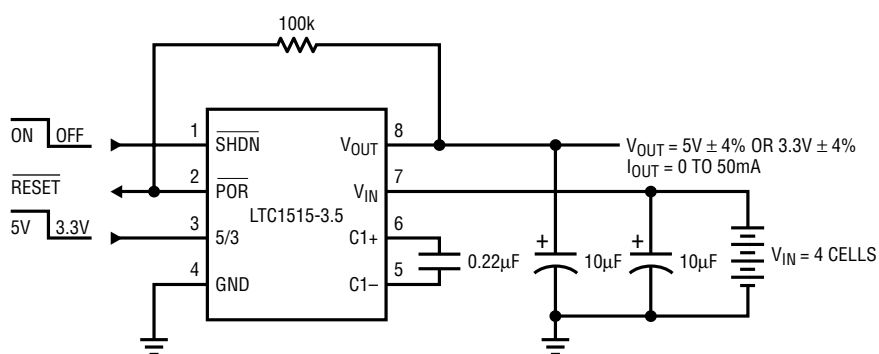
## Dual Output Supply from a 2.7V to 10V Input

The circuit shown in Figure 185 uses the low-battery comparator as a feedback comparator to produce an auxiliary 3.3V regulated output from the $V_{OUT}$ of the LTC1514-5. A feedback voltage divider formed by R2 and R3 connected to the comparator input (LBI) establishes the output voltage. The output of the comparator (LBO) enables the current source formed by Q1, Q2, R1 and R4. When the LBO pin is low, Q1 is turned on, allowing current to charge output capacitor C4. Local feedback formed by R4, Q1 and Q2 creates a constant current source from the 5V output to C4. Peak charging current is set by R4 and the $V_{BE}$ of Q2, which also provides current limiting in the case of an output short to ground. R5 pulls the gate of Q1 high when the auxiliary output is in regulation. C5 is used to reduce output ripple. The combined output current from the 5V and 3.3V supplies is limited to 50mA. Since the regulator implements a hysteretic feedback loop in place of the traditional linear feedback loop, no compensation is needed for loop stability. Furthermore, the high gain of the comparator provides excellent load regulation and transient response.

## Conclusion

With low operating current, low external parts count and robust protection features, the LTC1514 and LTC1515 are well-suited to low power step-up/step-down DC/DC conversion. The shutdown, POR and low-battery detect features provide additional value and functionality. The simplicity and versatility of these parts make them ideal for low power DC/DC conversion applications.
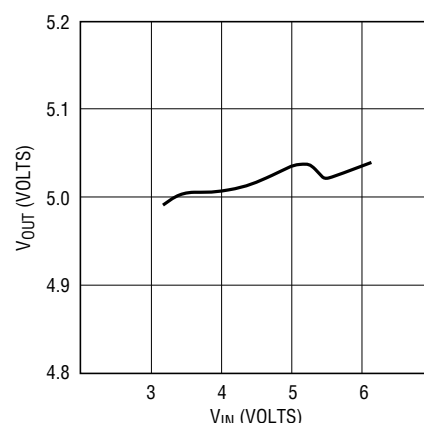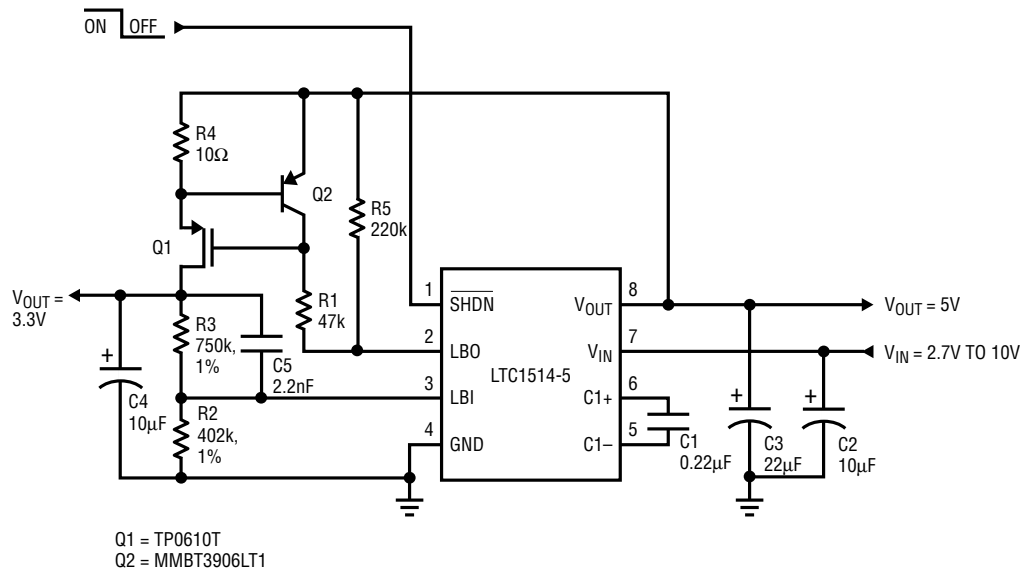
**Figure 183. Programmable 5V/3V Power-Supply with Power-On Reset**

**Figure 184. $V_{OUT}$ vs $V_{IN}$ for Figure 183's Circuit**

**Figure 185. Using the Low-Battery Comparator as a Feedback Comparator to Produce an Auxiliary 3.3V Regulated Output from the V$_{OUT}$ of the LTC1514-5**

## LTC1626 LOW VOLTAGE MONOLITHIC STEP-DOWN CONVERTER OPERATES FROM A SINGLE Li-Ion CELL
*by Tim Skovmand*

### Introduction

The LTC1626 is a monolithic, low voltage, step-down current mode DC/DC converter with an input supply voltage range of 2.5V to 6V, making it ideal for single-cell Li-Ion or 3- to 4-cell NiCd/NiMH applications. A built-in 0.32Ω P-channel switch (V$_{IN}$ = 4.5V) allows up to 0.6A of output current. The maximum peak inductor current is externally programmable to minimize component size in lower current applications.

The LTC1626 incorporates automatic power saving Burst Mode operation to reduce gate-charge losses when the load current drops below the level required for continuous operation. With no load, the converter draws only 160μA; in shutdown it draws a mere 1μA—making it ideal for current-sensitive applications.
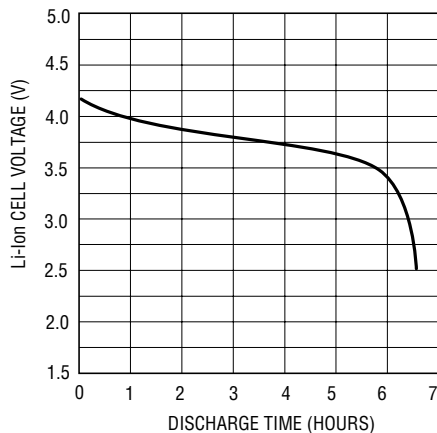


**Figure 186. Typical Single-Cell Li-Ion Discharge Curve**
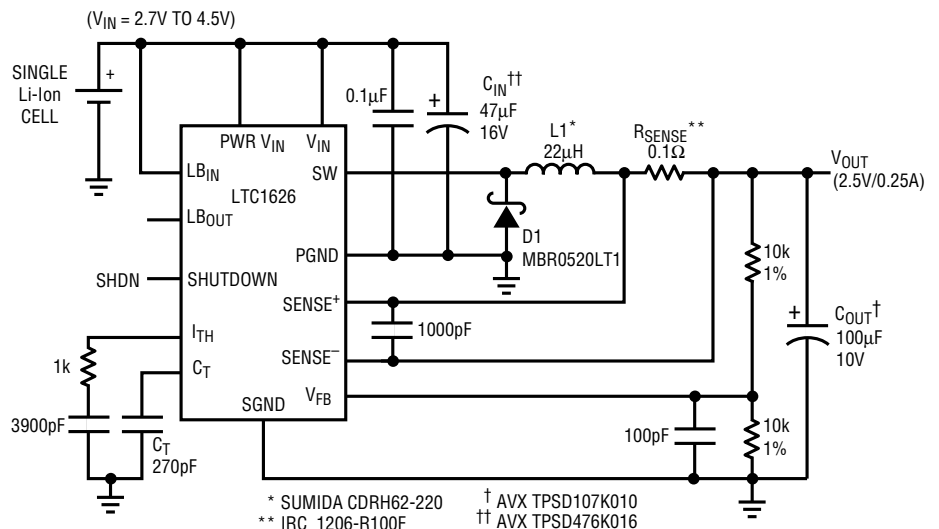


**Figure 187. Single-Cell Li-Ion Battery to 2.5V Converter**

**Figure 188. High Efficiency 2.5V Step-Down Converter**



**Figure 189. Efficiency vs Output Load Current**

## Single-Cell Li-Ion Operation

As shown in Figure 186, a fully charged single-cell Li-Ion battery begins the discharge cycle at either 4.1V or 4.2V (depending upon the manufacturer's charge voltage specifications). During the bulk of the discharge time, t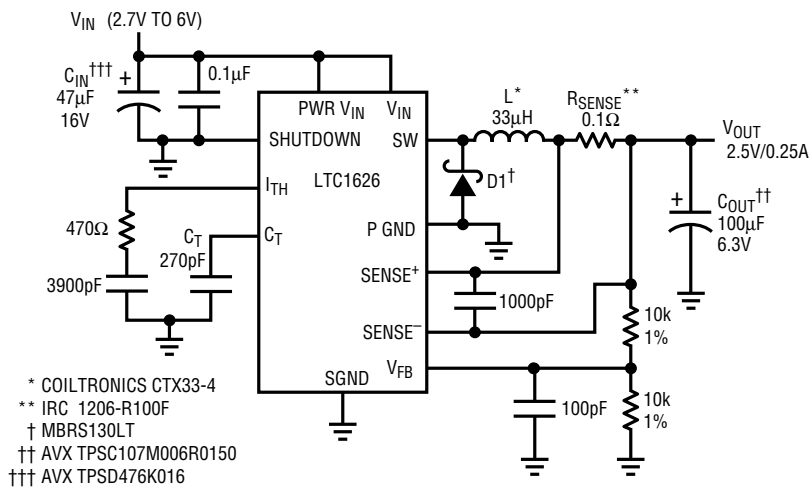he cell produces between 3.5V and 4.0V. Finally, toward the end of discharge, the cell voltage drops fairly quickly below 3V. It is recommended that the discharge be terminated somewhere between 2.2V and 2.8V (again, depending upon the manufacturer's specifications).

The LTC1626 is specifically designed to accommodate a single-cell Li-Ion discharge curve. For example, using the circuit shown in Figure 187, it is possible to produce a stable 2.5V/0.25A regulated output voltage with as little as a 2.7V from the battery—thus obtaining the maximum possible run time.

## High Efficiency Operation

Using the circuit shown in Figure 188, efficiencies of greater than 90% are maintained from 20mA to 250mA of load current with a 3.5V input supply voltage, as shown in Figure 189.
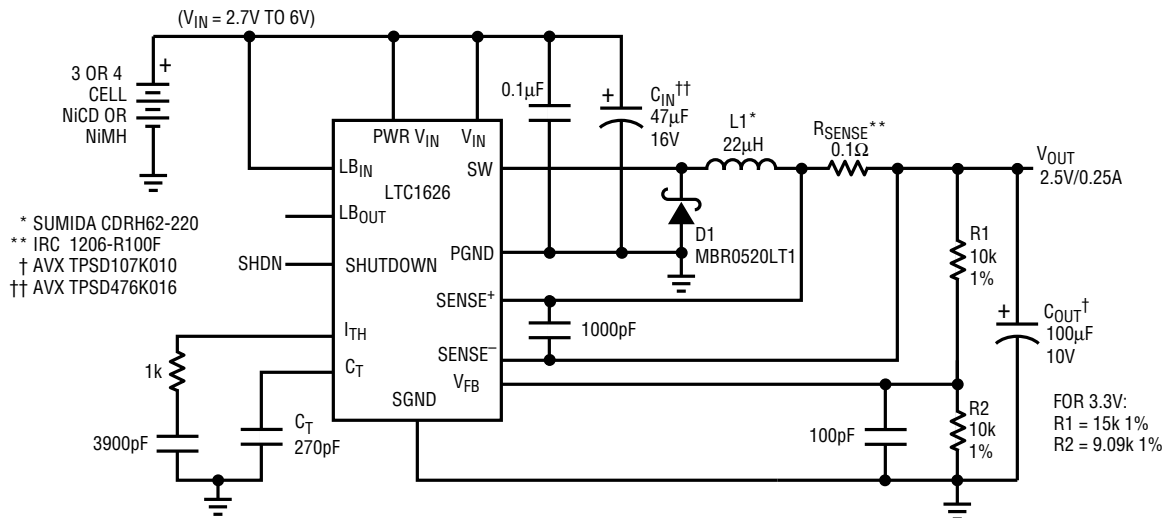


**Figure 190. 3- or 4-Cell NiCd/NiMH to 2.5V Converter**

## Typical Applications

### *3- or 4-Cell NiCd/NiMH DC/DC Converter*

Figure 190 is a schematic diagram that shows the LTC1626 being powered from a 3- or 4-cell NiCd or NiMH battery pack. (This circuit is also suitable for operation from three or four alkaline cells.) All the components shown in this schematic are surface mount and have been selected to minimize the board space and height. The output voltage is set at 2.5V, but is easily programmed to 3.3V for 4-cell applications. Simply modify the two output ladder resistors, R1 and R2, from 10k each to 15k and 9.09k, respectively, as shown in Figure 190.

### *Single Li-Ion 3.3V Buck/Boost Converter*

The circuit shown in Figure 191 produces 3.3V from an input voltage ranging from 2.5V to 4.5V. The two windings of a common inductor core are used to implement this circuit. Note that the current sense resistor is connected to ground. The table in Figure 191 shows the output current capability as a function of battery voltage.

## Conclusion

The LTC1626 is specifically designed to operate from a single-cell Li-Ion battery. With its low dropout, high efficiency and micropower operating modes, it is ideal for battery operated products and efficiency-sensitive devices such as cellular phones and handheld industrial and medical instruments.

| $V_{IN}$ (V) | $I_{OUT}$ (mA) |
|---|---|
| 2.5 | 200 |
| 3.0 | 350 |
| 3.5 | 500* |
| 4.0 | 500* |
| 4.2 | 500* |

\* DESIGN LIMIT

| MANUFACTURER | PART NO. |
|---|---|
| COILTRONICS | CTX33-4 |
| DALE | LPT4545-330LA |

\* SANYO OS-CON CAPACITOR
\*\* IRC 1206-R100F
† AVX TPSD107M010R0100
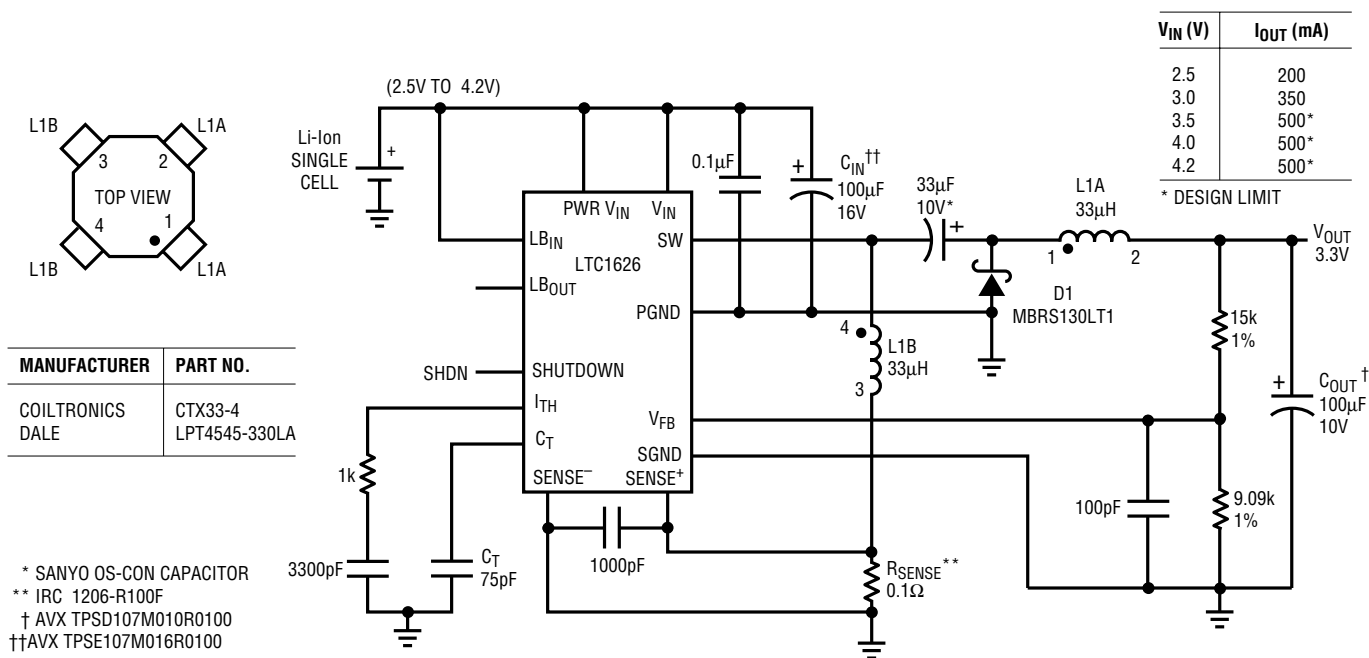††AVX TPSE107M016R0100

**Figure 191. Single-Cell Li-Ion to 3.3V Buck/Boost Converter**

# 12V WALL CUBE TO 5V/400mA DC/DC CONVERTER IS 85% EFFICIENT

*by Steve Pietkiewicz*

The ubiquitous 12V wall cube, power source of countless electronic products, generates an unregulated DC voltage between 8V and 18V, depending on line voltage and load. If you use a linear regulator to drop the voltage to 5V, a 400mA load means the linear regulator must dissipate 5W under worst-case conditions. To deal with this heat, you must provide adequate heat sinking, increasing your product's size and weight. Additionally, the heat is some-times objectionable to customers. These factors can ne-gate the cost advantage of a linear regulator. Figure 192's circuit, a negative buck converter, delivers 5V at loads up to 400mA from a 7V–25V input with peak efficiency of 85%, eliminating the need for a heat sink. Since the LT1307B (U1) is intended for use with a low input voltage, Q1 and Q2 are used to make a simple preregulator, providing 1.9V for U1's $V_{IN}$ pin. The IC switches at 600kHz, allowing a low cost 22μH inductor and 10μF ceramic output capacitor to be used. Q3 is needed to level shift the output voltage because U1's feedback pin is referenced to the negative input. Output ripple measures 10mV$_{P-P}$ at a load of 400mA. The circuit's efficiency is detailed in Figure 193, and response to a load step from 150mA to 300mA is shown in Figure 194. Input bypass capacitor C1 sees worst-case RMS ripple current equal to one-half the output current and should have an ESR of less than 0.5Ω. Take care during construction to keep R1–R3 and Q3 close to U1's FB pin and away from the SW pin to prevent unwanted coupling. Use a ground plane and keep traces for the power components short and direct.
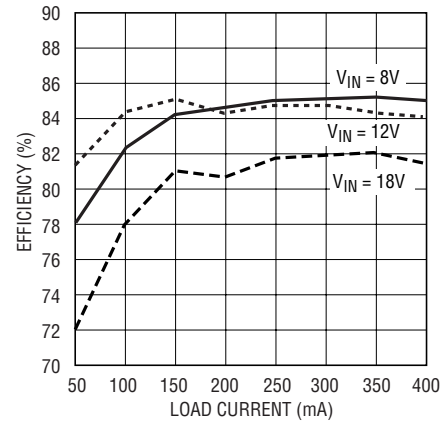


**Figure 193. Efficiency Peaks at 85%; It Is Above 80% Over an Input Range of 8V–18V**

Although it might seem unsettling that the negative side of the wall cube is not grounded, remember that the 9V wall cube floats. The circuit merely regulates the negative side, rather than the more conventional positive side.



50μs/DIV
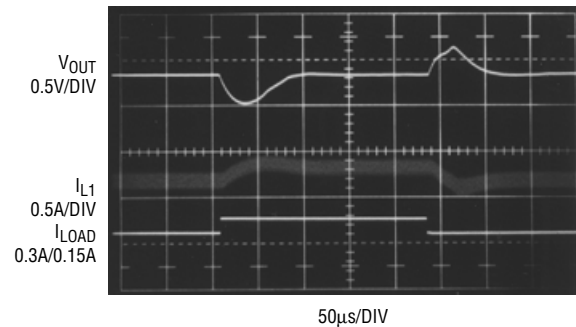
**Figure 194. Load-Step Response; the Load Changes from 150mA to 300mA**
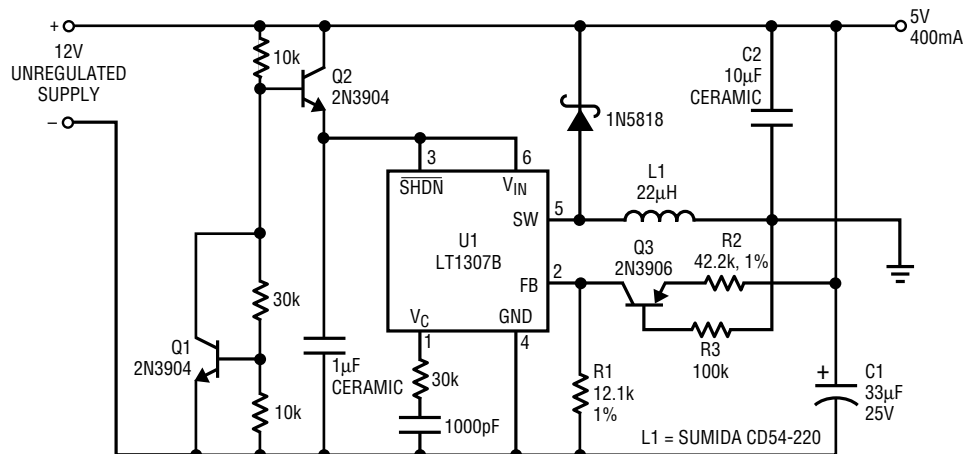


**Figure 192. This Negative Buck Converter Delivers 5V at 400mA from a 7V–25V Input**

## MICROPOWER 600kHz FIXED-FREQUENCY DC/DC CONVERTERS STEP UP FROM A 1-CELL OR 2-CELL BATTERY
*by Steve Pietkiewicz*

Linear Technology introduces two new micropower DC/DC converters designed to provide power from a single-cell or higher input voltage. The LT1308 features an onboard switch capable of handling 2A with a voltage drop of 300mV and operates from an input voltage as low as 1V. The LT1317, intended for lower power requirements, operates from an input voltage as low as 1.5V. Its internal switch handles 600mA with a drop of 360mV. Both devices feature Burst Mode operation at light load; efficiencies are above 70% for load currents of 1mA. Both devices switch at 600kHz; this high frequency keeps associated power components small and flat; additionally, troublesome interference problems in the sensitive 455kHz IF band are
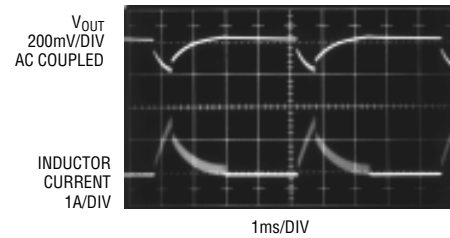


**Figure 197. Transient response of DC/DC converter: $V_{IN}$ = 3V, 0A–1A load step**

avoided. The LT1308 is intended for generating power on the order of 2W–5W. This is sufficient for RF power amplifiers in GSM or DECT terminals or for digital-camera power supplies. The LT1317, with its smaller switch, can generate 100mW to 2W of power. The LT1317 is available in LTC's smallest 8-lead package, the MSOP. This package is approximately one-half the size of a standard 8-lead SO package. The LT1308 is available in the 8-lead SO package.
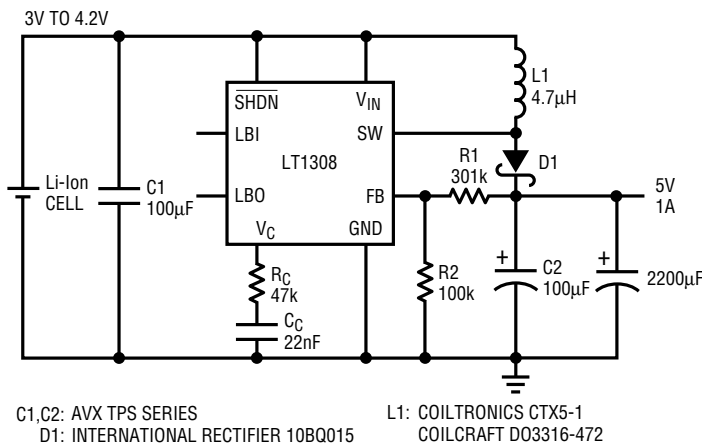


**Figure 195. Single Li-Ion Cell to 5V/1A DC/DC Converter**
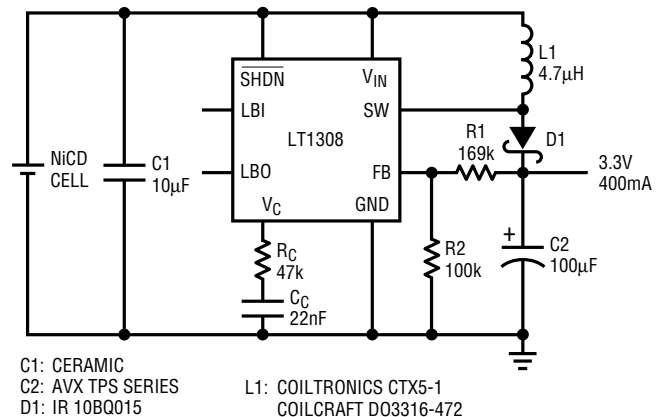


**Figure 198. Single NiCd Cell to 3.3V 400mA DC/DC Converter**
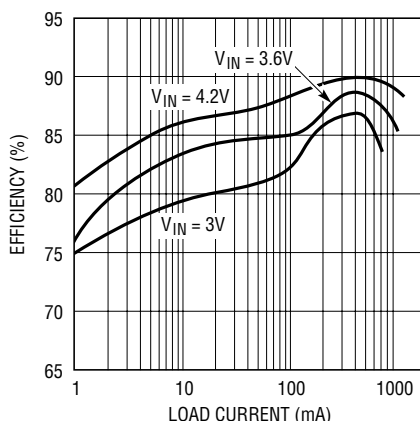


**Figure 196. Efficiency of Figure 195's Circuit**



**Figure 199. Efficiency of Figure 198's Circuit Reaches 81%**

**Figure 200. DECT Load Transient Response: with a Single NiCd Cell, the LT1308 Provides 3.3V with a 400mA Pulsed Load. The Pulse Width = 416µs**



**Figure 201. DECT Load Transient Response: Faster Sweep Speed (100µs/DIV) Details V$_{OUT}$ and Inductor Current of a Single DECT Transmit Pulse**

### Single Li-Ion Cell to 5V/1A DC/DC Converter for GSM

GSM terminals have emerged as a worldwide standard. A common requirement for these products is an efficient, compact, step-up converter to develop 5V from a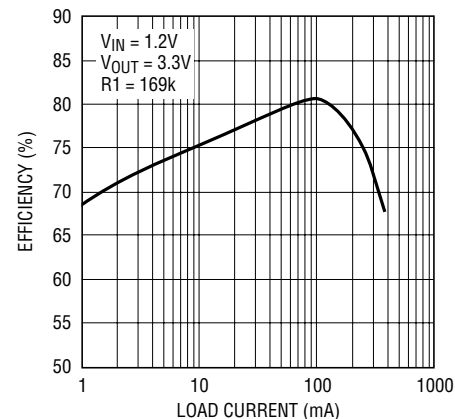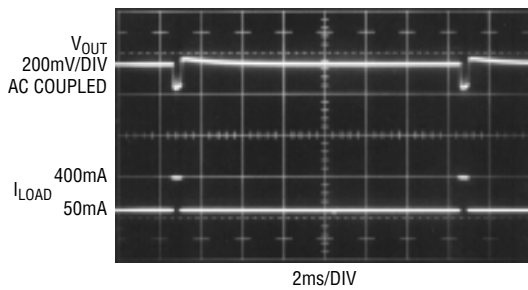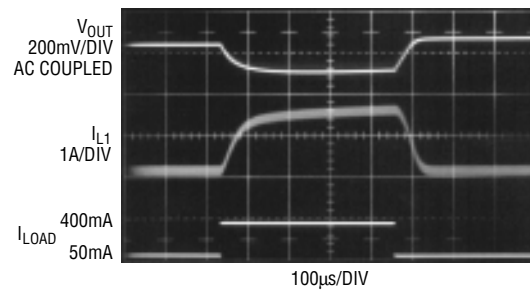 single Li-Ion cell to power the RF amplifier. The LT1308 performs this function with a minimum of external components. The circuit is detailed in Figure 195. Many designs use a large aluminum electrolytic capacitor (1000µF to 3300µF) at the DC/DC converter output to hold up the output voltage during the transmit time slice, since the amplifier can require more than 1A. The output capacitor, along with the LT1308 compensation network, serves to smooth out the input current demanded from the Li-Ion cell. Efficiency, which reaches 90%, is shown in Figure 196. Transient response of a 0A to 1A load step with typical GSM profiling (1:8 duty cycle, 577µs pulse duration) is depicted in Figure 197. Voltage droop (top trace) is 200mV. Inductor current

(bottom trace) increases to 1.7A peak; the input capacitor supplies some of this current, with the remainder drawn from the Li-Ion cell.

### Single NiCd Cell to 3.3V/400mA Supply for DECT

Only minor changes are required in Figure 195's circuit to construct a single-cell NiCd to 3.3V converter. The large output capacitor is no longer required as the output current can be handled directly by the LT1308. Figure 198 shows the DECT DC/DC converter circuit. Efficiency, reaching 81% from a 1.2V input, is pictured in Figure 199. Transient response of a typical DECT load of 50mA to 400mA is detailed in Figure 200. Output voltage droop (top trace) is under 200mV. Figure 201 zooms in on a single pulse to show the output voltage and inductor current responses more clearly.
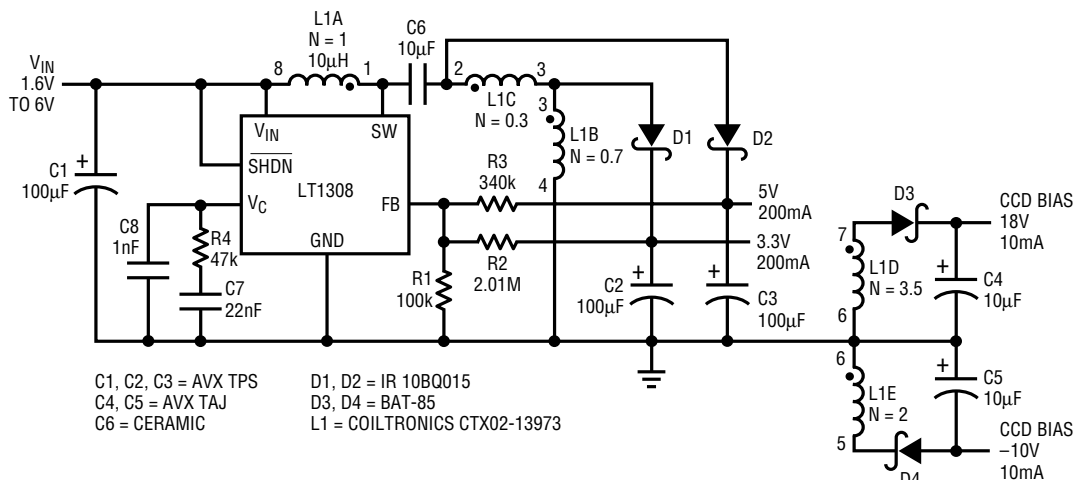


**Figure 202. This Digital Camera Power Supply Delivers 5V/200mA, 3.3V/200mA, 18V/10mA and −10V/10mA from 2 AA Cells**

## 2-Cell Digital Camera Supply
## Produces 3.3V, 5V, 18V and –10V

Power supplies for digital cameras must be small and efficient while generating several voltages. The DSP and logic need 3.3V, the ADC and LCD display need 5V and biasing for the CCD element requires 18V and –10V. The power supplies must also be free of low frequency noise, so that postfiltering can be done easily. The obvious approach, to use a separate DC/DC converter IC for each output voltage, is not cost-effective. A single LT1308, along with an inexpensive transformer, generates 3.3V/200mA, 5V/200mA, 18V/10mA and –10V/10mA from a pair of AA or AAA cells. Figure 202 shows the circuit. A coupled-flyback scheme is used, actually an extension of the SEPIC (single ended primary inductance converter) topology. The addition of capacitor C6 clamps the SW pin, eliminating a snubber network. Both the 3.3V and 5V outputs are fed back to the LT1308 FB pin, a technique known as split feedback. This compromise results in better overall line and load regulation. The 5V output has more influence than the 3.3V output, as can be seen from the relative values of R2 and R3. Transformer T1 is available from Coiltronics, Inc. (561-241-7876). Efficiency vs input voltage for several load currents on both 3.3V and 5V outputs is pictured in Figure 203. The CCD bias voltages are loaded with 10mA in all cases.

### LT1317 2-Cell to 5V DC/DC Converter

Figure 204 shows a simple 2-cell to 5V DC/DC converter using the LT1317. This device generates a clean, low ripple output from an input voltage as low as 1.5V. Designed for 2-cell applications, it offers better perfor-



C1: CERAMIC
D1: MOTOROLA MBRO520L
L1: 22μH SUMIDA CD43-220

**Figure 204. 2-Cell to 5V Boost Converter Using the LT1317**

mance than its 1-cell predecessor, the LT1307. More gain in the error amplifier results in lower Burst Mode ripple, and an internal preregulator eliminates oscillator variation with input voltage. For comparison, Figure 205 details transient responses of both the LT1307 and the LT1317 generating 5V from a 3V input. The load step is 5mA to 200mA. Output capacitance in both cases is 33μF. The LT1307 has low frequency ripple of 100mV, whereas the LT1317 Burst Mode ripple of 20mV is the same as the 600kHz ripple resulting from the output capacitor's ESR with a 200mA load.

### Single Li-Ion Cell to ±4V DC/DC Converter

By again employing the SEPIC topology, a ±4V supply can be designed with one IC. Figure 206's circuit generates 4V at 70mA and –4V at 10mA from an input voltage ranging from 2.5V to over 5V. Maximum component height is 2mm. This converter uses two separate inductors (L1 and L2), so it is an uncoupled SEPIC converter. This reduces the overall cost, but requires that all output current pass



**Figure 203. Camera Power Supply Efficiency Reaches 78%**
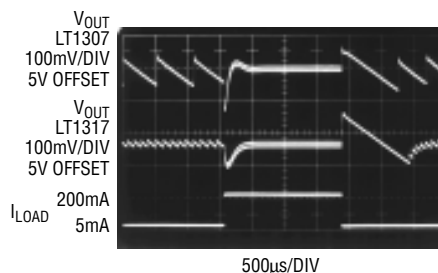


**Figure 205. The LT1317 Has Reduced
Burst Mode Ripple Compared to the LT1307**

through C1. Since C1 is ceramic, its ESR is low and there is no appreciable efficiency loss. C5 is charged to –$V_{OUT}$ when the switch is off, then its bottom plate is grounded when the switch turns on. The negative output is fairly well regulated, since the diode drops tend to cancel. The circuit is switching continuously at rated load, where efficiency is 75%. Output ripple is under 40mV and can be reduced further with conventional postfiltering techniques.

## Conclusion

The LT1308 and LT1317 provide low noise compact solutions for contemporary portable-product power supplies.



L1, L2 = MURATA LQH3C220
C1 = MURATA GRM235Y5V106Z01
D1 = MBR0520
D2 = BAT54S (DUAL DIODE)
C2 = AVX TAJB33M6010
C3 = AVX TAJA156MO1O
C4, C5 = CERAMIC
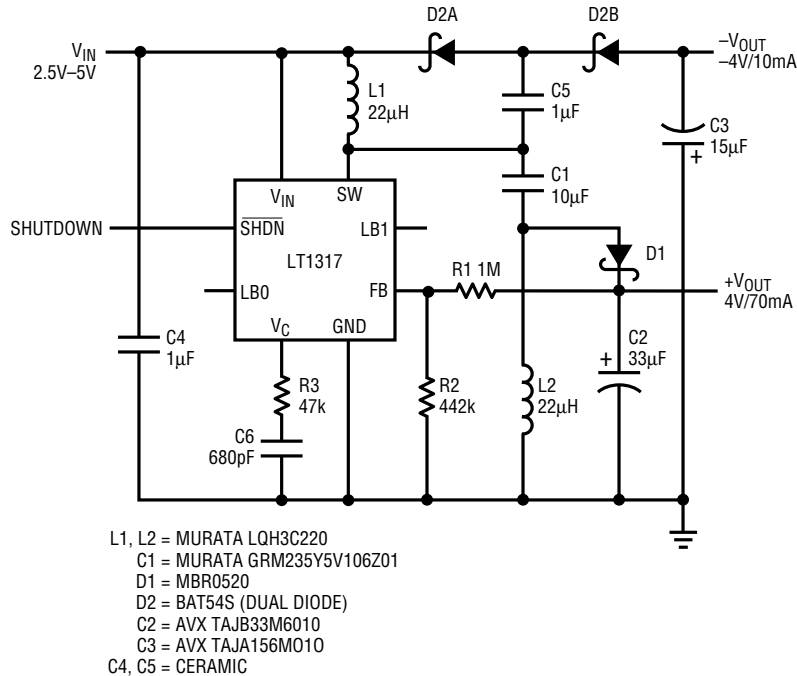
**Figure 206. This SIngle Li-Ion Cell to ±4V DC/DC Converter Has a Maximum Height of 2mm**

## LT1610 MICROPOWER STEP-UP DC/DC CONVERTER RUNS AT 1.7MHZ
*by Steve Pietkiewicz*

### Introduction

The LT1610, a micropower DC/DC converter IC, addresses the issue of footprint in several ways. First, the switching frequency is 1.7MHz, allowing the use of small, inexpensive, minimal-height inductors and capacitors. Second, the frequency-compensation components have been integrated, eliminating the requirement for an external RC network in most applications. Finally, the device comes in LTC's 8-lead MSOP package, one-half the size of the 8-lead SO package.

The LT1610's input voltage ranges from 1V to 8V, and the 30V, 300mA switch allows several different configurations, such as boost, SEPIC and flyback, to be successfully



C1, C2: AVX TAJA226M010R
D1: MOTOROLA MBR0520
L1: MURATA LQH3C4R7M24
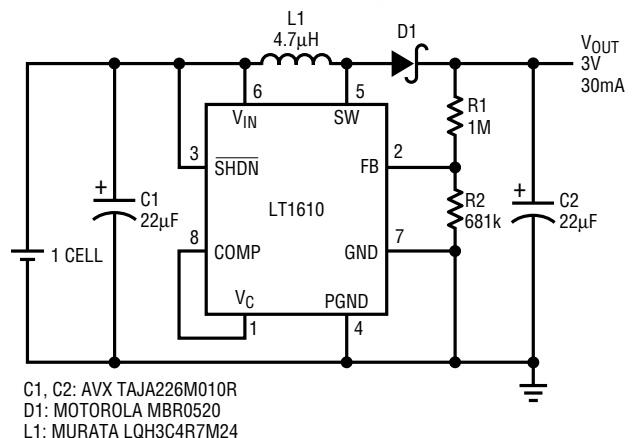
**Figure 207. This Single Cell to 3V Converter Delivers 30mA**
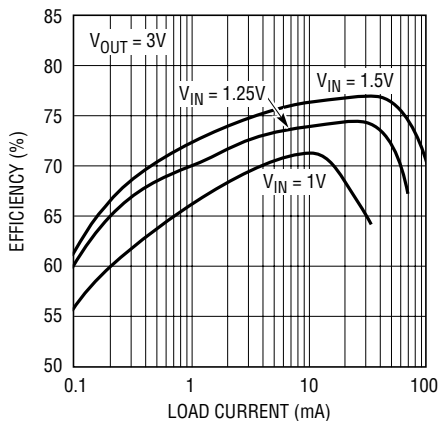
**Figure 208. Single-Cell Converter Efficiency Reaches 77%**

implemented. Output voltage can be up to 28V in boost mode. Operating quiescent current is 50μA unloaded; grounding the shutdown pin reduces the current to 0.5μA. The device can generate 3V at 30mA from a single (1V) cell, or 5V at 100mA from two cells (2V). Configured as a Li-Ion cell to 3.3V SEPIC converter, the LT1610 can deliver 100mA. In boost mode, efficiency ranges from 60% at a 100μA load to 83% at full load.

## Single-Cell to 3V DC/DC Converter

A 1V to 3V boost converter is shown in Figure 207. The specified components take up very little board space. The 4.7μH Murata inductor specified measures 2.5mm by 3.2mm and is only 2mm high. The 22μF AVX "A" case tantalum capacitors measure 1.6mm by 3.2mm and are 1.6mm tall. Circuit efficiency, which reaches 77%, is detailed in Figure 208. Transient response to a 1mA to 31mA load step is pictured in Figure 209. The device features Burst Mode operation at light loads. This can be seen at a load of 1mA. When the load is increased to 31mA, the device shifts to constant-frequency switching and peak switch current is controlled to achieve output regulation.
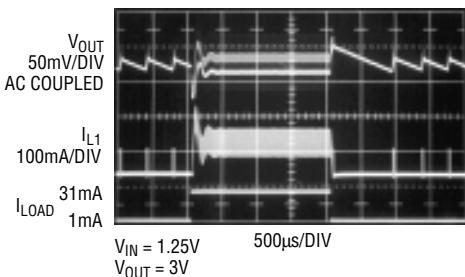


**Figure 209. Transient Load Response of Single-Cell Converter, Load Stepped from 1mA to 31mA**



C1, C2: AVX TAJA156M010R
D1: MOTOROLA MBR0520
L1: SUMIDA CD43-4R7
    MURATA LQH3C4R7M24
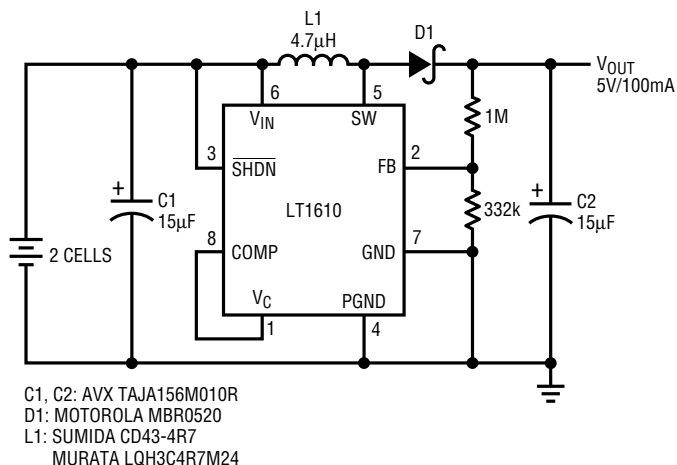
**Figure 210. 2 Cell to 5V Converter Delivers 100mA at 2V Input**

## 2-Cell to 5V DC/DC Converter

By simply changing the feedback resistor values, the LT1610 can generate 5V. Figure 210's circuit generates 5V at a load of up to 100mA from a 2-cell input. Figure 211's graph shows efficiency the of the circuit, which reaches 83%. This circuit is also suitable for 3.3V to 5V conversion, supplying over 200mA.

## Li-Ion to 3.3V SEPIC Converter

Figure 212 employs the SEPIC (single ended primary inductance converter) topology to provide a regulated 3.3V output from an input that can range above or below the output voltage. Although the circuit requires two inductors and a ceramic coupling capacitor, the total footprint of this solution is still attractive compared with alternative methods of generating 3.3V, such as a boost converter followed by a linear regulator. The circuit can
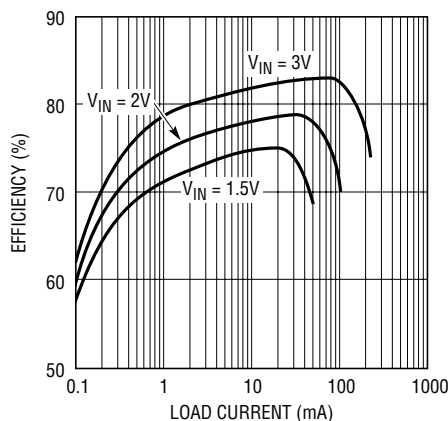


**Figure 211. 2-Cell Converter Efficiency Reaches 83%**

supply up to 100mA. Efficiency, while lower than that of a standard boost converter, reaches approximately 73%. Unlike a boost converter, this topology provides input-to-output isolation. The output is completely disconnected

from the battery in shutdown mode, preventing inadvertent battery discharge through the load. The LT1610's sub-µA shutdown current reduces standby losses, increasing battery life.



C1, C2: AVX TAJA226M010R
C3: AVX 1206YG475
D1: MOTOROLA MBR0520
L1, L2: MURATA LQH3C4R7M24
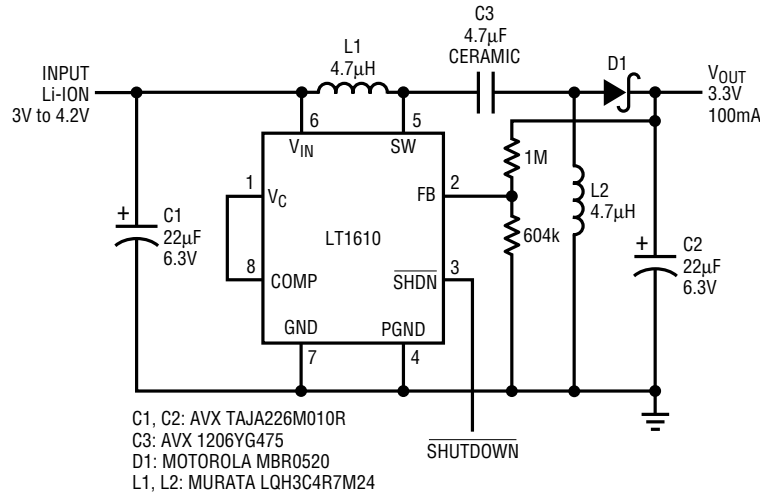
**Figure 212. Li-Ion to 3.3V SEPIC Converter Delivers 100mA**

## LOW NOISE 33V VARACTOR BIAS SUPPLY
*by Jeff Witt*

Wideband tuning circuits, such as those used in cable television systems, require a power supply for driving a varactor. This bias supply is usually at a voltage higher than the system supply voltage, allowing a large tuning range. The supply must have very little noise; voltage

ripple, for example, can appear as sidebands on a local oscillator. This circuit takes advantage of the fixed operating frequency of the LT1317B boost regulator to generate a low noise 33V bias voltage.



D1 TO D3: MOTOROLA MMBD914LT1
C1: AVX TAJ156M010
C2: SANYO 35CV33GX
L1: MURATA LQH3C220
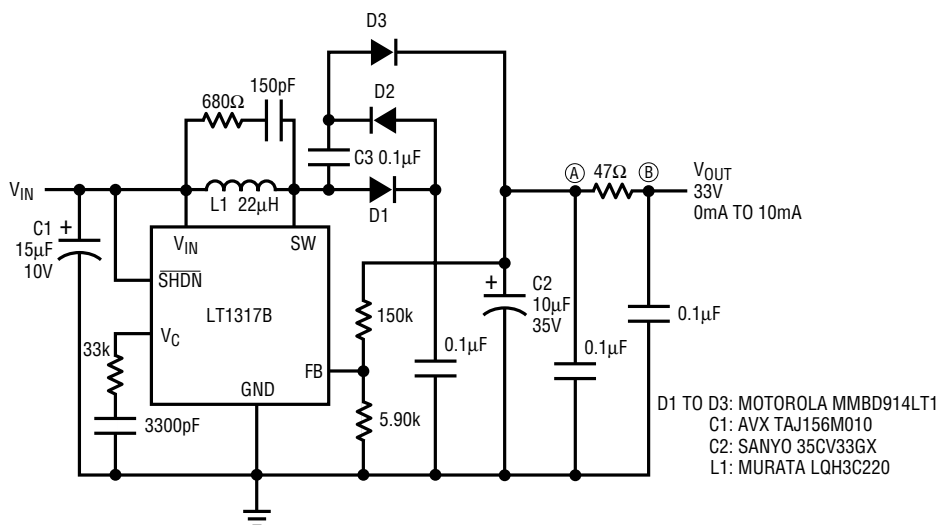
**Figure 213. This Circuit Generates a Low Noise Bais Supply for Varactor-Based Tuning Circuits**

The circuit (Figure 213) is a simple boost regulator with its output voltage doubled by diodes D2 and D3 and capacitor C3. With this doubler, the circuit can generate an output voltage greater than the voltage rating of the LT1317B's internal power switch. This supply can deliver 10mA at 33V from a 3V to 6V input, allowing operation from either 3.3V or 5V logic rails. The high operating frequency (600kHz) results in low, easily filtered output ripple, as shown in Figure 214. The high frequency also allows the use of small, low cost external components.



NODE A
OUTPUT RIPPLE
20mV/DIV

NODE B
OUTPUT RIPPLE
20mV/DIV

1µs/DIV
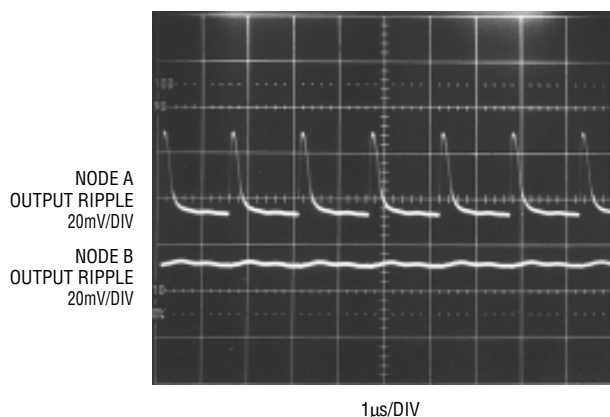
**Figure 214. The Output Ripple of Figure 213's Supply as it Delivers 5mA at 33V from a 5V Input; Traces A and B Show Ripple Before and After the RC Output Filter, Respectively**

## THE LTC1516 CONVERTS TWO CELLS TO 5V WITH HIGH EFFICIENCY AT EXTREMELY LIGHT LOADS
*by Sam Nork*

Many battery-powered applications require very small amounts of load current from the regulated supply over long periods of time, followed by moderate load currents for short periods of time. In these types of applications (for example, remote data-acquisition systems, hand-held remote controls, and the like), the discharge rate of the battery is dominated by the overall current demands under low load conditions. In such low load systems, a primary source of battery drain is the DC/DC converter that converts the battery voltage to a regulated supply.

The circuit shown in Figure 215 converts an input voltage from two cells to 5V using a switched-capacitor charge-pump technique. An integral comparator on the LT1516 senses the output voltage and enables the charge pump as
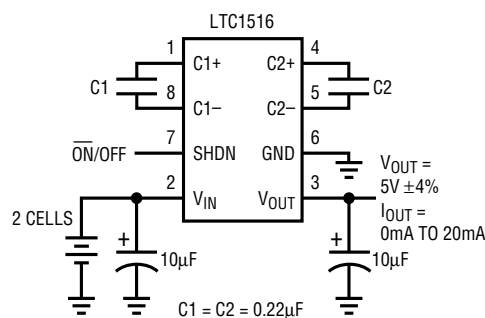


**Figure 215. 2 Cell to 5V Converter**

the output begins to droop. The charge pump's 2-phase clock controls the internal switching of flying caps C1 and C2. (See Figure 216.) On phase one of the clock, the flying caps are connected between $V_{IN}$ and GND. On phase two, the negative plate of C1 is connected to $V_{IN}$, the negative plate of C2 is connected to the positive plate of C1, and the positive plate of C2 is connected to the output. During this phase of the clock, the potential on the top plate of C2 is approximately 3 • $V_{IN}$ and the charge is dumped from C2 onto the output cap to raise the output voltage. The repeated charging and discharging of C1 and C2 continues at a nominal frequency of 600kHz until the output voltage has risen above the internal comparator's trip point.

When the battery cells are fully charged (approximately 1.5V per cell, for a nominal 3V $V_{IN}$), the circuit operates as a voltage doubler to maintain regulation. In doubler mode, only C2 is charged to $V_{IN}$ and discharged onto $V_{OUT}$ when the charge pump is enabled. As the batteries discharge and/or the load increases, the circuit will change from doubler mode to tripler mode. Under light load conditions, the part will remain in doubler mode until $V_{IN}$ has dropped below 2.55V. Under heavier loads, the part will go into tripler mode at a higher $V_{IN}$ to maintain regulation. By switching operating modes as the $V_{IN}$ and the load conditions change, the LTC1516 optimizes overall efficiency for the life of the batteries. As shown in Figure 217, Figure 215's circuit achieves better than 70% efficiency with load currents from 50μA to 20mA for almost the entire life of the batteries.



Figure 216. LT1516 Charge Pump
in Trippler Mode, Discharge Cycle



Figure 217. Efficiency vs $V_{OUT}$ for Figure 215's Circuit

Regulators—Linear

## LOW DROPOUT REGULATOR DRIVER HANDLES FAST LOAD TRANSIENTS AND OPERATES ON A SINGLE 3V–10V INPUT
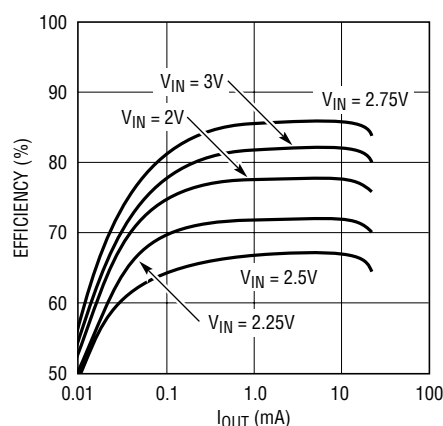*by Lenny Hsiu*

### Introduction

The LT1573 is designed to provide a low cost solution to applications requiring high current, low dropout and fast transient response. When combined with an external PNP power transistor, this device provides up to 5A of load current with dropout voltages as low as 0.35V. The LT1573's

circuitry is designed for extremely fast transient response. This greatly reduces the bulk storage capacitance required when the regulator is used in applications with fast, high current load transients.

Base-drive current to the external PNP is limited for instantaneous protection and a time-delayed latch protects the regulator from continuous short circuits. The latch time-out period can be varied by an external capacitor. Guaranteed minimum available base-drive current to the external PNP is 250mA. The LT1573 is equipped with an active-high shutdown and a thermal shutdown function. The shutdown function can be used to reset the
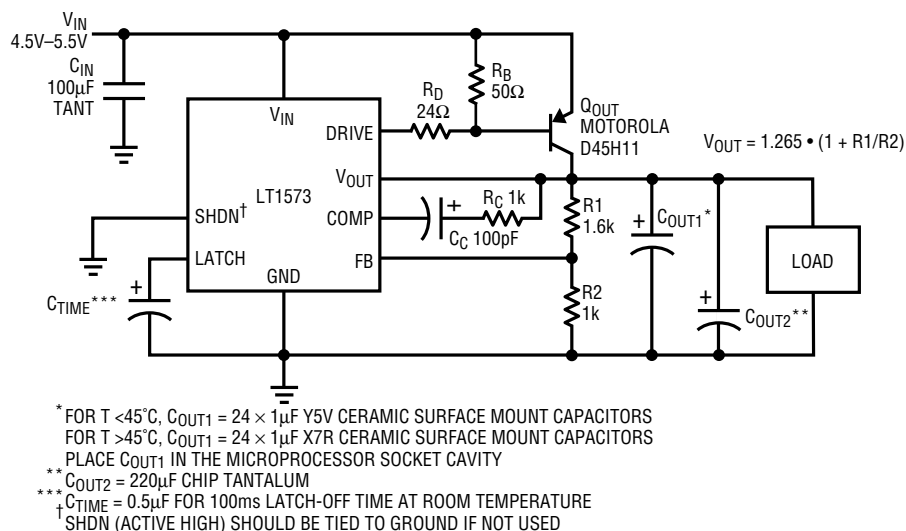
**Figure 218. 3.3V, 5A Microprocessor Supply**

overcurrent latch. The thermal shutdown function can be used to protect the PNP power transistor if it is thermally coupled to the LT1573.

## Basic Regulator Circuit

The adjustable-output LT1573 circuit shown in Figure 218 senses the regulator output voltage from its feedback pin via the output voltage divider and drives the base of the external PNP transistor to maintain the regulator output at the specified value. For fixed-output versions of the LT1573,

the regulator output voltage is sensed from the feedback pin via an internal voltage divider. In this case, the FB pin is left unconnected. The resistor $R_D$ is required for the overcurrent latch-off function. $R_D$ is also used to limit the drive current available to the external PNP transistor and to limit the power dissipation in the LT1573. Limiting the drive current to the external PNP transistor will limit the output current of the regulator, thereby minimizing the stress on the regulator circuit under overload conditions.

See the LT1573 Data Sheet for additional design details.

## THE LT1575/LT1577 UltraFast LINEAR REGULATOR CONTROLLERS ELIMINATE BULK TANTALUM/ ELECTROLYTIC OUTPUT CAPACITORS
*by Anthony Bonte*

### Introduction

The LT1575/LT1577 family of single/dual controller ICs are new, easy-to-use devices that drive discrete N-channel MOSFETs as source followers to produce extremely low dropout, UltraFast™ transient response regulators. These circuits achieve superior regulator bandwidth and transient load performance, and completely eliminate expensive tantalum or bulk electrolytic capacitors in the most demanding microprocessor applications. For example, a 200MHz Pentium® processor can operate with only the twenty-four 1µF ceramic capacitors that Intel already

requires for the microprocessor. Users realize significant savings because all additional bulk capacitance is removed. The additional savings of insertion cost, inventory cost and board space are readily apparent.

Precision-trimmed adjustable and fixed-output voltage versions accommodate any required microprocessor power supply voltage. Dropout voltage can be user defined via selection of the N-channel MOSFET $R_{DS(ON)}$. The only output capacitors required are the high frequency ceramic decoupling capacitors. The regulator responds to transient load changes in a few hundred nanoseconds—a great improvement over regulators that respond in many microseconds. The ceramic capacitor network generally consists of ten to twenty-four 1µF capacitors, depending on individual microprocessor requirements. The LT1575/ LT1577 family also incorporates current limiting at no
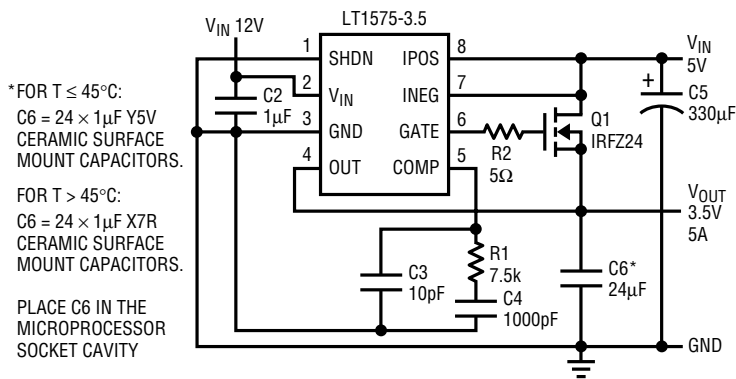
Figure 219. UltraFast Transient Response 5V to 3.3V, Low Dropout Regulator



Figure 220. Transient Response for 0.2A–5A Output Load Step

additional system cost, provides on/off control and can provide overvoltage protection or thermal shutdown with the addition of a few simple external components. The LT1575 is available in 8-pin SO or PDIP and the LT1577 is available in 16-pin narrow-body SO.

**UltraFast 5V to 3.3V Low Dropout Regulator**

Figure 219 shows the basic regulator control circuit. The input voltage is a standard 5V "silver box" and the output voltage is set to 3.5V, the Pentium P54 VRE microprocessor supply voltage. The typical maximum output current is about 5A in most Pentium microprocessor applications. The output capacitor network consists of only twenty-four inexpensive 1µF ceramic, surface mount capacitors. Proper layout of this decoupling network is critical to proper operation of this circuit. Consult Linear Technology Application Note 69: *LT1575 UltraFast Linear Controller Makes Fast Transient Response Power Supplies*, for details on board layout.

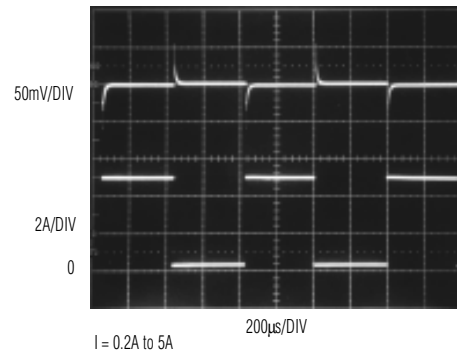The photo in Figure 220 shows the transient response performance for an output load current step of 0.2A to 5A. The main loop compensation in Figure 219's regulator circuit is provided by R1 and C4 at the COMP pin. Capacitor C3 introduces a high frequency pole and provides adequate gain margin beyond the unity-gain crossover frequency of 1MHz. This compensation network limits overshoot/undershoot to 50mV under worst-case load transient conditions. With a 1% specified worst-case output voltage tolerance, the 100mV output voltage error

budget for a P54 VRE microprocessor is easily met with production margin to spare. All bulk tantalum/electrolytic capacitors are completely eliminated.

The discrete N-channel MOSFET chosen is a low cost International Rectifier IRFZ24 or equivalent. The input capacitance is approximately 1000pF with $V_{DS} = 1V$. The specified on-resistance is $0.1\Omega$ at room temperature and about $0.15\Omega$ at 125°C. At 7A output current, the dropout voltage is only 1.05V. This eases the restriction on local input decoupling capacitor requirements because significant droop in the typical 5V input supply voltage is permitted before dropout voltage operation is reached. (Note that 5V supply tolerance restrictions are typically limited by a ±5% tolerance so that 5V logic systems will operate correctly.) However, a simple LC input filter can eliminate the need for large input bulk capacitance at the regulator 5V supply for additional system cost savings.

Figure 221 shows a more complete system configuration that incorporates current limiting and current limit time-out with latch-off. Current limit is incorporated for no additional system cost by manufacturing the current limit resistor from a Kelvin-sensed section of PC board trace. In this example, current limit is set to 7A. A capacitor from the SHDN pin to ground sets a fault condition time-out period that latches off the drive to the external MOSFET if the time-out period is exceeded. The regulator is reset by pulling the SHDN pin low. The output voltage in this application is set to 3.3V. The ±5% tolerance permitted in 3.3V systems translates to a ±165mV output-voltage

tolerance. This permits a 50% reduction in the number of ceramic capacitors required from twenty-four to twelve. Loop compensation is adjusted accordingly.

Figure 222 shows an application circuit using the LT1577, a dual regulator. All functions for each regulator are identical to those of the LT1575. One section is configured for a 3.3V output and the other section is configured for a 2.8V output. This circuit provides all the power requirements for a split-plane system: 3.3V for the logic supply and 2.8V for the processor-core supply. Note that both SHDN pins are tied to a common time-out capacitor. If either or both regulators encounter a fault condition, both regulator sections are latched off after the time-out period is exceeded.



**Figure 221. 5V to 3.3V Regulator**



**Figure 222. LT1577 Dual Regulator for Split-Plane Systems**

## LT1579 BATTERY-BACKUP REGULATOR PROVIDES UNINTERRUPTIBLE POWER
*by Todd Owen*

### Introduction

Designed for a multitude of applications, the LT1579 is a dual input, single output, low dropout regulator that provides an uninterruptible output voltage from two independent input voltage sources on a priority basis. All power supplied to the load is drawn from the primary input ($V_{IN1}$) until the device senses that the primary source is failing. At this point, the LT1579 smoothly switches from the primary input to the secondary input ($V_{IN2}$) to maintain output regulation. The LT1579 is capable of providing 300mA from either input at a dropout voltage of 0.4V. Total quiescent current is 50μA: 45μA from the primary input source, 2μA from the secondary input source, and an additional 3μA from the higher voltage of the two.

### Circuit Examples

The basic application of the LT1579 is shown in Figure 223. It uses two independent voltage sources for the inputs. These voltage sources may be batteries, wall adapters or any other DC source. The low-battery comparators are configured to give a low output if either input voltage drops below 5.5V. The trip points can be adjusted by changing the values of the divider resistors (R1 and R2 for LB1, R3 and R4 for LB2). All logic outputs (LBO1,
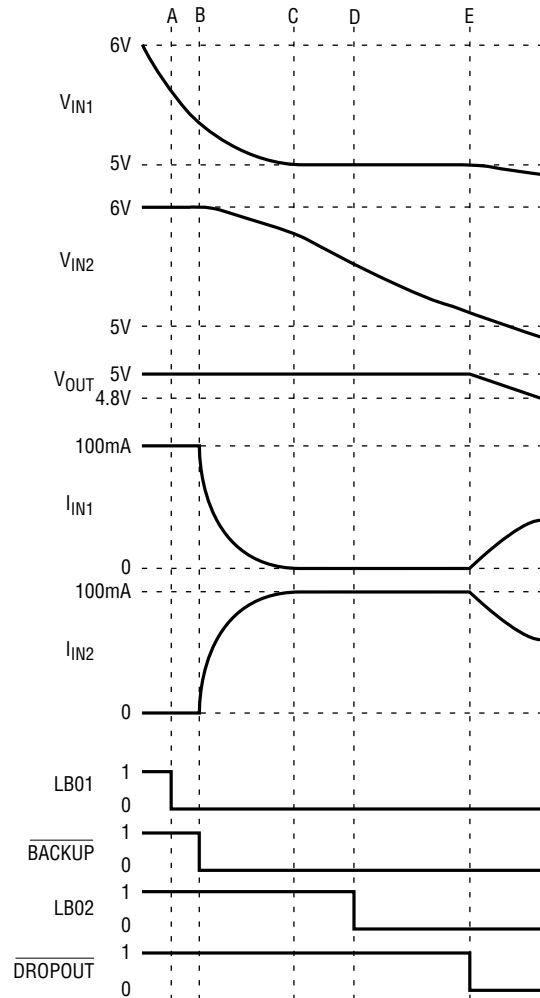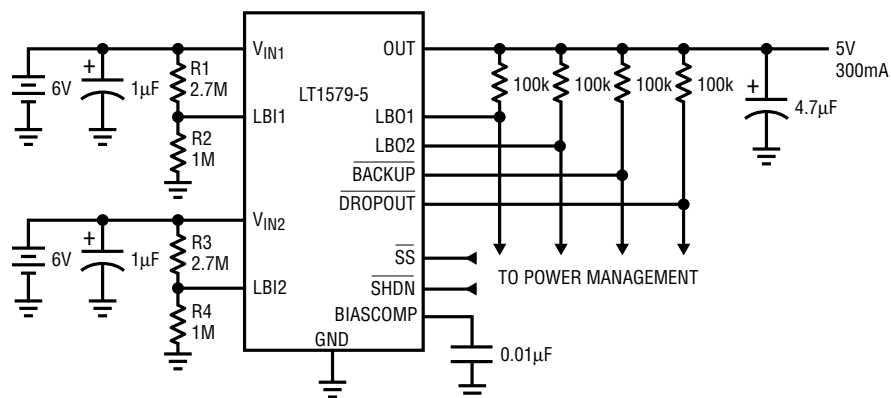


Figure 224. Basic Application Timing Diagram



Figure 223. LT1579 Basic Application

LBO2, $\overline{\text{BACKUP}}$ and $\overline{\text{DROPOUT}}$) are open-collector outputs that require an external pull-up resistor. They are capable of sinking 20µA at a maximum output voltage of 0.32V, which is useful for driving both CMOS and TTL logic families. For driving LED's, all logic outputs can sink 5mA at a maximum output voltage of 1.2V.

Figure 224 is the timing diagram for the basic circuit. No time scale is shown for the timing diagram because actual discharge rates are a function of the load current and the type of batteries used. The timing diagram is meant as a tool to help in understanding the LT1579's basic operation.

Five milestones are noted on the timing diagram. Time A is where the primary input voltage drops enough to trip the low-battery detector, LB1. The trip threshold for LB1 is set at 5.5V, slightly above the dropout voltage of the primary input. At time B, the $\overline{\text{BACKUP}}$ flag goes low, signaling the beginning of the transition from the primary source to the secondary source. Between times B and C, the input current makes a smooth transition from $V_{IN1}$ to $V_{IN2}$. By

time C, the primary battery has exhausted most of its useful charge. The primary input will still deliver a small amount of current to the load, diminishing as the primary input voltage drops. By time D, the secondary battery has dropped to a low enough voltage to trip the second low-battery detector, LB2. The trip threshold for LB2 is also set at 5.5V, slightly above where the secondary input reaches dropout. At time E, both inputs are low enough to cause the LT1579 to enter dropout, with the $\overline{\text{DROPOUT}}$ flag signaling the impending loss of output regulation.

Some interesting things can be noted on the timing diagram. The amount of current available from a given input is determined by the input/output voltage differential. As the primary voltage drops, the amount of current drawn from the input also drops, slowing discharge of the battery. Dropout-detection circuitry will maintain the maximum current draw from the input for the given input/output voltage differential, based on the impedance of the pass transistor. In the case shown, this causes the current drawn from the primary to approach zero, although it
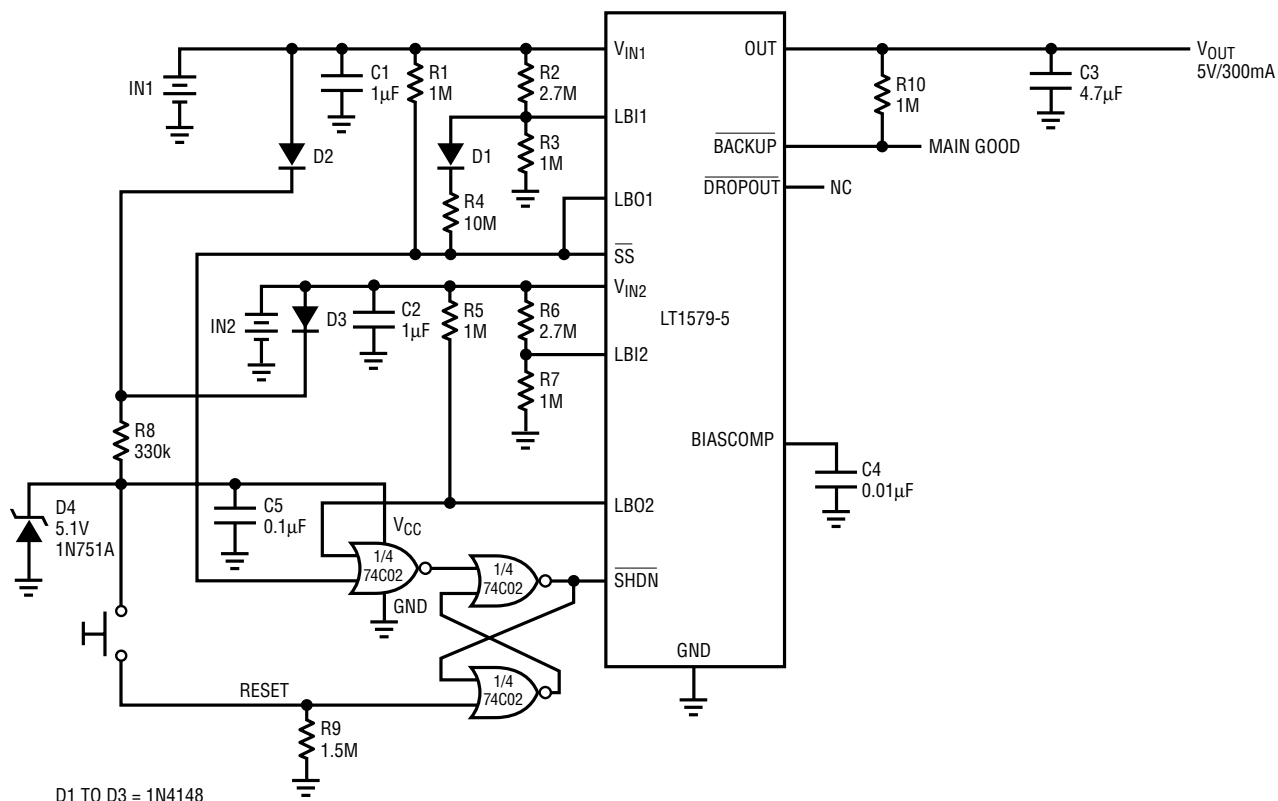


**Figure 225. Added SR Latch Shuts the LT1579 Off when Both Low-Battery Detectors are Tripped**

never reaches that point. Note that the primary begins to supply significant current again when the secondary input drops low enough to cause a loss in output regulation. This occurs because the input/output voltage differential of the primary input increases as the output voltage drops. The LT1579 will automatically maximize the power drawn from the inputs to maintain the highest possible output voltage.

A final circuit example is shown in Figure 225. This circuit has a few notable changes from the basic application. First, the Secondary Select pin is connected directly to LBO1. When the primary input voltage drops below the threshold level for LB1, the comparator output will pull the Secondary Select pin low. This forces the device to switch completely over to the secondary input, limiting the discharge voltage of the cells. Second, the logic gates used form an SR latch. When both batteries are below the threshold level for their respective comparators, the latch will be set, forcing the part into shutdown. The latch is reset by pulling up on the RESET node, allowing the part to come out of shutdown.

The series resistance of a battery can cause its terminal voltage to rise as its current decreases. This effect can reset the low-battery detector and cause the LT1579 to oscillate between the primary and secondary inputs. To combat this, the low-battery comparators have up to 18mV of built-in hysteresis at the input to the comparator (LBI1, LBI2). The hysteresis is determined by the amount of load current on the comparator output. At no load, the comparator hysteresis is zero, increasing to a maximum of 18mV for load currents above 20µA. For the pull-up resistor shown, load current on the output of the comparator is 5µA, so hysteresis will be 5mV. With the values shown for resistor divider R2/R3, this translates to 19mV

of hysteresis at the primary input of the LT1579. Additional hysteresis can be added by connecting D1 and R4. The values shown will give an additional 200mV of hysteresis.

When LBO1 and LBO2 are high impedance and either input is greater than 6.5V, the logic-flag voltages can be above the maximum voltage rating. Internal clamps on the logic flags limit the output voltage to approximately 6.5V and the pull-up resistor values shown will limit the current into the logic flags to less than the maximum current rating.

**Conclusion**

The LT1579 can provide a continuous regulated output voltage to critical circuits from any of a number of different input sources. It will provide up to 300mA of output current at a dropout voltage of 0.4V. Should the primary input fail, the device switches seamlessly to the secondary input, maintaining output regulation. A single error amplifier controls both output stages so regulation remains tight regardless of which input is providing power. The LT1579 can handle instantaneous removal of either one of its inputs without losing regulation. System power management is aided by two status flags, which provide information about which input is providing power and signal the loss of output regulation. Two independent low-battery comparators can be used to monitor input voltages. Also, an external pin can be used to force the switch to the secondary input. Total quiescent current of the LT1579 is 50µA, dropping to a mere 7µA in its low power shutdown state. Internal circuitry guards against a number of fault conditions, including current limit, thermal limit and reverse voltages, protecting sensitive circuitry and inputs. Whether the application is simple or complex, the LT1579 is truly a "smart" regulator.

Battery Chargers

## THE LT1511 3A BATTERY CHARGER CHARGES ALL BATTERY TYPES, INCLUDING LITHIUM-ION
*by Chiawei Liao*

The LT1511 current mode PWM battery charger is the simplest, most efficient solution for fast charging modern rechargeable batteries, including lithium-ion (Li-Ion), nickel-metal-hydride (NiMH) and nickel-cadmium (NiCd) that require constant-current and/or constant-voltage charging. The internal switch is capable of delivering 3A DC current (4A peak current). Full charging current can be programmed by resistors or by a DAC to within 5%, and the trickle charge current can be programmed to 10% accuracy. With 0.5% reference voltage accuracy, the LT1511 meets the critical constant-voltage charging requirement for lithium cells.

The LT1511 is equipped with a voltage-control loop to control charging voltage and a current-control loop to control charging current. A third control loop is provided to regulate the current drawn from the AC adapter. This allows simultaneous equipment operation and battery charging without overloading the adapter. Charging current

is reduced to keep the adapter current within specified levels.

The LT1511 can charge batteries ranging from 1V to 20V. Ground sensing of current is not required and the battery's negative terminal can be tied directly to ground.

### LT1511 Applications

#### Lithium-Ion Charging

The 3A lithium battery charger (Figure 226) charges lithium-ion batteries at a constant 3A until the battery voltage reaches a limit set by R3 and R4. The charger will then automatically go into a constant-voltage mode, with the current decreasing to zero over time as the battery reaches full charge. This is the normal regimen for lithium-ion charging, with the charger holding the battery at "float" voltage indefinitely. In this case no external sensing of full charge is needed.

Current though the R3/R4 divider is set at 15µA to minimize battery drain when the charger is off. The input current to the OVP pin is 3nA and this error can be neglected.
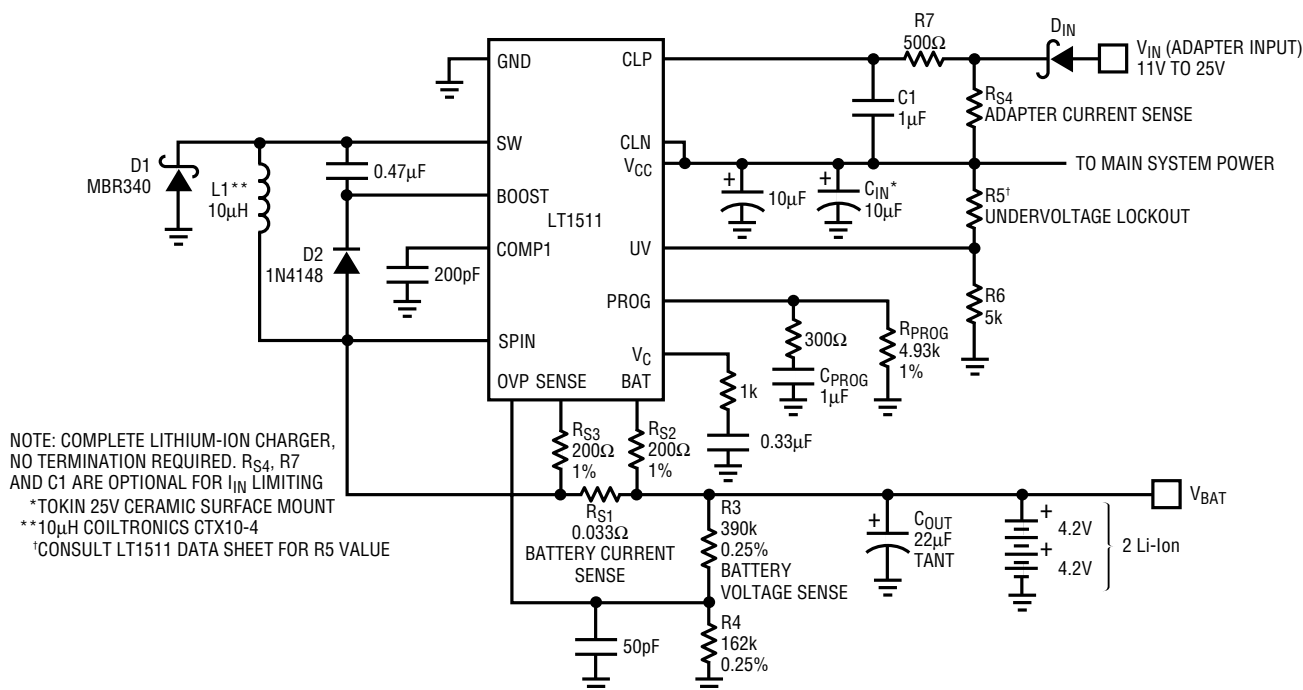


**Figure 226. 3 Amp Lithium-Ion Battery Charger**

With divider current set at 15µA, R4 = 2.465/15mA = 162k and

$$R1 = \frac{V_{OUT} - 1.245}{\frac{1.245}{R2} + (3 \times 10^{-7})}$$

where $V_{OUT}$ = battery float voltage

Lithium-ion batteries typically require float-voltage accuracy of 1% to 2%. The accuracy of the LT1511 OVP voltage is ±0.5% at 25°C and ±1% over full temperature. This leads to the possibility that very accurate (0.1%) resistors might be needed for R3 and R4. Actually, the temperature of the LT1511 will rarely exceed 50°C in float mode because charging currents have tapered off to a low level, so 0.25% accuracy resistors will normally provide the required level of overall accuracy.

## Nickel-Cadmium and Nickel-Metal-Hydride Charging

The circuit in the 3A lithium battery charger (Figure 226) can be modified as shown in Figure 227 to charge NiCd or NiMH batteries. Two-level charging is needed; 2A when Q1 is on and 200mA when Q1 is off. For 2A full current, the current sense resistor ($R_{S1}$) should be increased to 0.05Ω, so that enough signal (10mV) will be across $R_{S1}$ at 0.2A trickle charge to keep charging current accurate.

For a two-level charger, R1 and R2 are found from

$$R1 = \frac{(2.465)(4000)}{I_{LOW}} \qquad R2 = \frac{(2.465)(4000)}{I_{HI} - I_{LOW}}$$

All battery chargers with fast charge rates require some means to detect the full-charge state in the battery in order to terminate the high charging current. NiCd batteries are typically charged at high current until temperature rise or battery voltage decrease is detected as an indication of nearly full charge. The charging current is then reduced to a much lower value and maintained as a constant trickle charge. An intermediate "top off" current may be used for a fixed time period to reduce 100% charge time.

NiMH batteries are similar in chemistry to NiCd but have two differences related to charging. First, the inflection characteristic in battery voltage as full charge is approached is not nearly as pronounced. This makes it more difficult to use dV/dt as an indicator of full charge, and temperature change is more often used, with a temperature sensor in the battery pack. Second, constant trickle charge may not be recommended. Instead, a moderate level of current is used on a pulse basis (1% to 5% duty cycle) with the time-averaged value substituting for a constant low trickle.

If overvoltage protection is needed, R3 and R4 should be calculated according to the procedure described in lithium-ion charging section. The OVP pin should be grounded if not used. When a microprocessor DAC output is used to control charging current, it must be capable of sinking current at a compliance up to 2.5V if connected directly to the PROG pin.
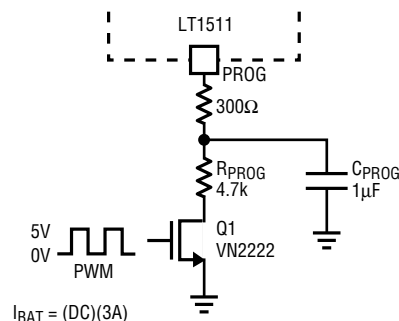


**Figure 227. 2-Step Charging**

## LT1512/LT1513 BATTERY CHARGERS OPERATE WITH INPUT VOLTAGES ABOVE OR BELOW THE BATTERY VOLTAGE
*by Bob Essaff*

### Introduction

The LT1512 and LT1513 form a unique family of constant-current, constant-voltage battery chargers that can charge batteries from input voltages above or below the battery voltage. This feature can help simplify system design and add product flexibility by allowing battery charging from multiple sources, such as a wall adapter, a 12V automotive system or a 5V power supply, all with the same circuit. The constant-current, constant-voltage architecture makes the LT1512 and LT1513 well suited for charging NiCd, NiMH, lead-acid or lithium-ion batteries.

Both devices are current mode switching regulators that operate at a fixed frequency of 500kHz. Product features include a ±1% reference-voltage tolerance, 2.7V minimum input voltage, easy external synchronization and 12µA supply current in shutdown mode. The LT1512 and LT1513 also include low loss on-chip power switches rated for 1.5 amps and 3 amps respectively. High frequency switching allows the use of small surface mount inductors and capacitors, and the battery can be directly grounded.

### Applications

The LT1512 and LT1513 are specifically optimized to use the SEPIC converter topology, which is shown in Figure 228's typical application. The SEPIC (single-ended primary inductance converter) topology has several advantages for battery-charging applications. It will operate with input voltages above or below the battery voltage, has no path for battery discharge when turned off, and eliminates the snubber losses of flyback designs. It also has a current sense point that is ground referred and need not be connected directly to the battery. The two inductors shown are actually two identical windings on one inductor core, although two separate inductors can be used.

The topology is essentially identical to a 1:1 transformer-flyback circuit except for the addition of capacitor C2, which forces identical AC voltages across both windings. This capacitor performs three tasks: it eliminates the power loss and voltage spikes usually caused by a flyback-converter's leakage inductance; it forces the input current and the current in resistor R3 to be a triangle wave riding on top of a DC component instead of forming a large amplitude square wave; and it eliminates the voltage spikes across the output diode when the switch turns on.

When the battery is below its float voltage, set by R1 and R2, the charger is in the constant-current mode. The suggested value for R2 is 12.4k. R1 is calculated from:
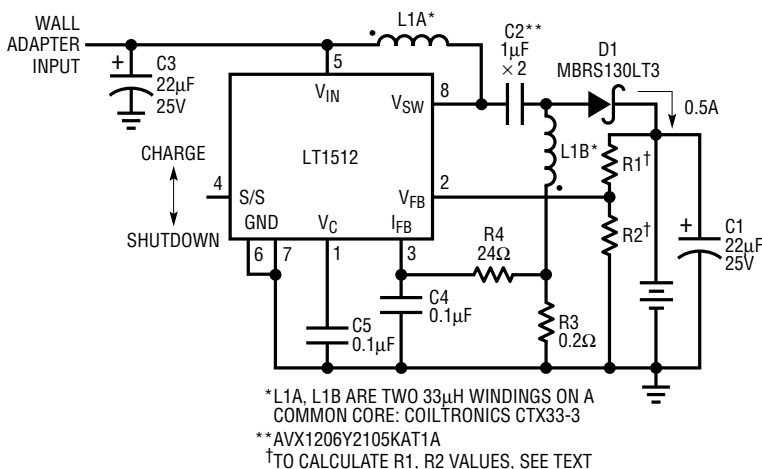


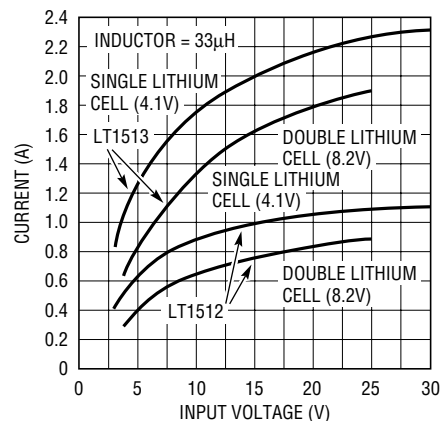Figure 228. Battery Charger with 0.5A Output Current



Figure 229. Maximum Charging Current

$$R1 = \frac{V_{OUT} - 1.245}{\frac{1.245}{R2} + (3 \times 10^{-7})}$$

where $V_{OUT}$ = battery float voltage

Charging current in the battery, which also flows through R3, develops a voltage on the $I_{FB}$ pin. The $I_{FB}$ pin's 100mV sense voltage sets the programmed charging current to $I_{CHG}$ = 100mV/R3. The RC filter formed by R4 and C4 smoothes the signal presented to the $I_{FB}$ pin.

Charging current remains constant until the battery reaches its float voltage, at which point the LT1512/LT1513 changes to the constant-voltage mode. In this mode, the charging current will taper off as required to keep the battery at its float voltage. The circuit's maximum input voltage is partly determined by the battery voltage. When the switch is off, the voltage on the $V_{SW}$ pin is equal to the input voltage, which is stored across C2, plus the battery voltage. Both the LT1512 and LT1513 have a maximum input voltage rating of 30V and a maximum rated switch voltage of 35V, thereby limiting input voltage to 30V or 35V minus the battery voltage, whichever is less.

Figure 229 shows the maximum available charging current for a single-cell or double-cell lithium battery pack. Note that the actual programmed charging current will be independent of the input voltage if it does not exceed the values shown.
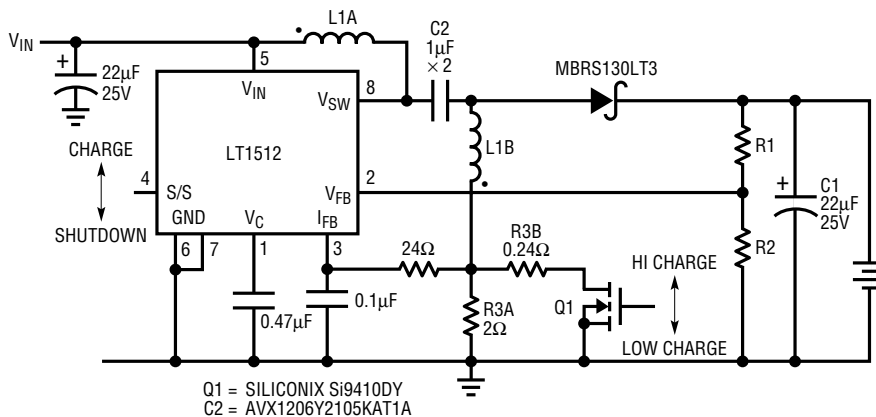
## Programming the Charge Current

As mentioned earlier, charging current is set by R3, where $I_{CHG}$ = 100mV/R3. The charge current is programmed by changing the effective value of R3, as shown in Figure 230. In the low charge mode, Q1 is off, setting charge current to $I_{CHG\ LOW}$ = 100mV/R3A, or 100mV/2$\Omega$ = 50mA. In the high-charge mode, Q1 is on, and charge current is $I_{CHG\ HI}$ = 100mV/R3A + 100mV/(R3B + Q1's $R_{DS(ON)}$), or 100mV/2$\Omega$ + 100mV/(0.24$\Omega$ + 0.04$\Omega$)) = 50mA + 357mA = 407mA. Note that Q1's $R_{DS(ON)}$ is a factor in the high-charge mode, requiring the use of a low $R_{DS(ON)}$ FET.

## Off-State Leakage

Charging can be terminated by placing the LT1512/LT1513 into shutdown mode. If the battery remains connected to the charger when in the off state, two leakage paths that load the battery must be considered.

The first is the 100$\mu$A resistor-divider feedback current that flows through R1 and R2. This current can be eliminated with the addition of a FET, Q1, between R1 and the R2/$V_{FB}$ junction, as shown in Figure 231. In this example, pulling the charge/shutdown input above 3.75V will activate charging and turn on Q1, whereas driving the charge/shutdown input below 0.6V will shut down the LT1512/LT1513 and turn off Q1.
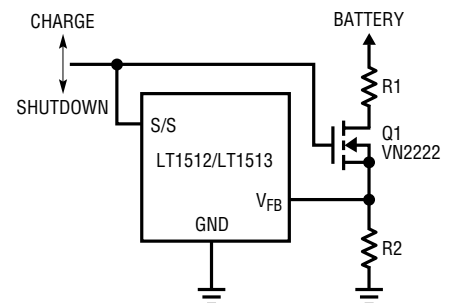


**Figure 230. 50mA/400mA Programmable Battery Charger**



**Figure 231. Shutdown-Controlled Disconnect**

The second leakage path to consider is in the output diode, D1 (Figure 228). When the charger is in the off state, the output diode sees a reverse voltage equal to the battery voltage. Though the Schottky diode reverse leakage may typically be only 10µA, its guaranteed specifications are much worse, up to 1mA. One solution is to change the output diode to an ultra-fast silicon diode, such as an MUR-110. The higher forward voltage of the silicon diode will decrease the circuit's efficiency, but these diodes have reverse leakage specifications below 5µA.

## Li-Ion BATTERY CHARGER DOES NOT REQUIRE PRECISION RESISTORS
*by LTC Applications Staff*

In constant-voltage mode charging, a Li-Ion cell requires 4.1V ±50mV. This 1.2% tolerance is tight. In a regulation loop where a voltage divider is compared against a reference, the accuracy is achieved by selecting a 0.7% reference and a voltage divider with 0.25% tolerance resistors. Unfortunately, 0.25% precision resistors cost three times as much as 1% resistors and have very long lead times.

One solution for moderate volume production involves adding two 1% resistors and two jumpers to the charger circuit, as shown in Figure 232. The jumpers are removed as necessary to bring the constant voltage to the required accuracy of 1.2%.

The charger selected for this example is the LT1510 and the number of Li-Ion cells in the battery is three. Select a value for R4 (20k) and calculate the values for resistors R1, R2 and R3 using the equations in Figure 232. K is the relative change required for a circuit with all its tolerances in one direction. For example, in the case of a 0.5% reference and two 1% resistors, the total tolerance is 2.5%. In order to bring it back to 1.2%, the percentage change required is 2.5% − 1.2% = 1.3% and K = 0.013.

The jumpers J1 or J2 need to be opened based on the following:

If $V_{OUT}$ is K/2 below nominal, remove J1.

If $V_{OUT}$ is K/2 above nominal, remove J2.



$$R1 = R4 \times \frac{V_{OUT} - V_{REF}}{V_{REF}}$$

$$R2 = (R1+R2) \times K$$

$$R3 = \frac{R4 \times K}{1 - (1 - K)\frac{V_{REF}}{V_{OUT}}}$$

TO CALCULATE K, SEE TEXT
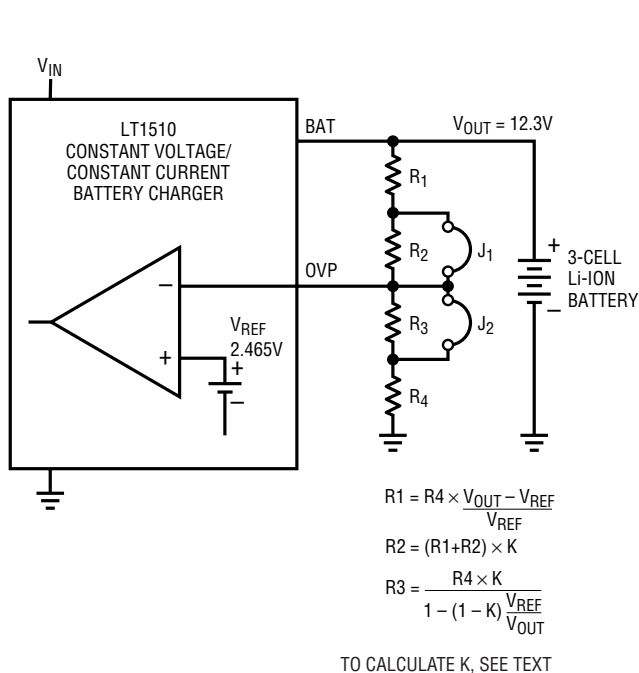
**Figure 232. R2, R3, J1 and J2 Eliminate the Need for Precision Resistors**



*TOKIN OR MARCON CERAMIC SURFACE MOUNT
** COILTRONICS CTX33-2

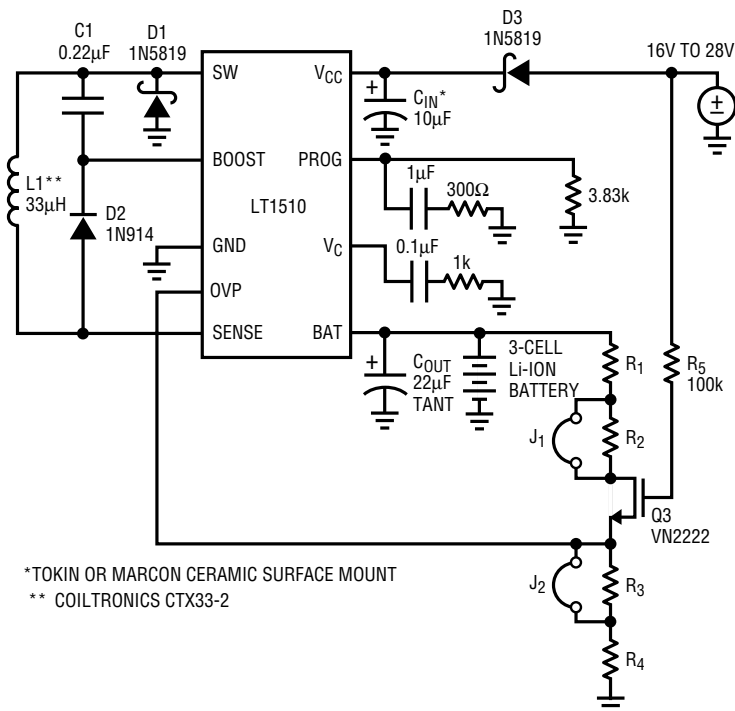**Figure 233. 3-Cell Li-Ion Charger without Precision Resistors**

The following values were calculated: R1 = 20k, R2 = 324Ω, R3 = 80.6Ω and R4 = 4.99k.

The voltage below which J1 should be opened is 12.34V – 1.3%/2 = 12.22V.

The voltage above which J2 should be opened is 12.34V + 1.3%/2 = 12.42V.

The complete schematic can be seen in Figure 234. Q3 is off when the charger is not powered, preventing current drain from the battery through the voltage divider. R5, a 100k resistor, isolates the OVP pin from any high frequency noise on $V_{IN}$. The charger in Figure 233 is programmed for 1.3A constant current.

## LT1510 CHARGER WITH –∆V TERMINATION
*by LTC Applications Staff*

Any portable equipment that requires fast charge needs proper charge termination. Commonly, a LT1510 constant-voltage, constant-current type charger controlled by a microcontroller is used. Sometimes, however, a microcontroller is not available or is not suitable for fast-charge termination.

When fast charging NiCd batteries with constant current, the internal battery temperature rises toward the end of the charge. Since the temperature coefficient of NiCd is negative, the temperature rise causes the battery voltage to drop. The drop can be detected and used for termination (called –∆V termination). The circuit in Figure 234 is a solution for a 3-cell (Panasonic P140-SCR) NiCd battery charger with –∆V termination.

U1 in Figure 234 is programmed by resistor R2 for a conservative charge current of 0.8A, which is 0.57**C**. Typical fast-charge current is 1**C**. (The boldfaced **C** represents a normalization concept used in the battery industry. A **C** rate of 1 is equal to the capacity of the cell in ampere-hours, divided by 1 hour. Since the capacity of the P140-SCR is 1.4 ampere-hours, **C** is 1.4 amperes.)

To determine the voltage droop rate, the battery was connected to an LT1510 charger circuit programmed for a 0.8A constant-current. The data was plotted as voltage versus time and the results are shown in Figure 235. The voltage slope is calculated to be –0.6mV/s. After the battery voltage dropped 300mV from the peak of 4.93V (100mV per cell), the charger was disabled.

At the heart of the circuit in Figure 234 is U3, a sample-and-hold IC (LF398). For every clock pulse at pin 8, the output of U3 (pin 5) updates to the input level on pin 3. When the battery voltage drops, the input to U3 also drops. If the update step at the output of U3 is sufficiently negative, U2B latches in the high state and Q1 turns on. Q1 terminates the charge by pulling down the LT1510's $V_C$ pin, and thereby disabling it.

U2A and the associated passive components smooth, amplify and level shift the battery voltage. The timer (U4) updates the hold capacitor (C8) every fifteen seconds. The timer signal stays high for 7ms, sufficient time for the hold capacitor to be charged to the input level. U2B and the associated parts form a latch that requires a momentary negative voltage at pin 6 to change state. R15 supplies the negative feedback and Q2, R16, R17 and C10 reset the latch on turn-on.

U3's output voltage droops at a rate proportional to the hold capacitor's internal leakage and the leakage current at pin 6 (10pA typical). This droop is very low and does not affect the operation of the circuit.

The minimum negative battery voltage slope required to trigger termination (–dV/dT) is 0.3mV/s. It can be calculated from:

$-dV/dT = V_{TRIG}/(T_{CLK} \times G_{U2A})$ where:
$V_{TRIG}$ is the trigger voltage of U2B,
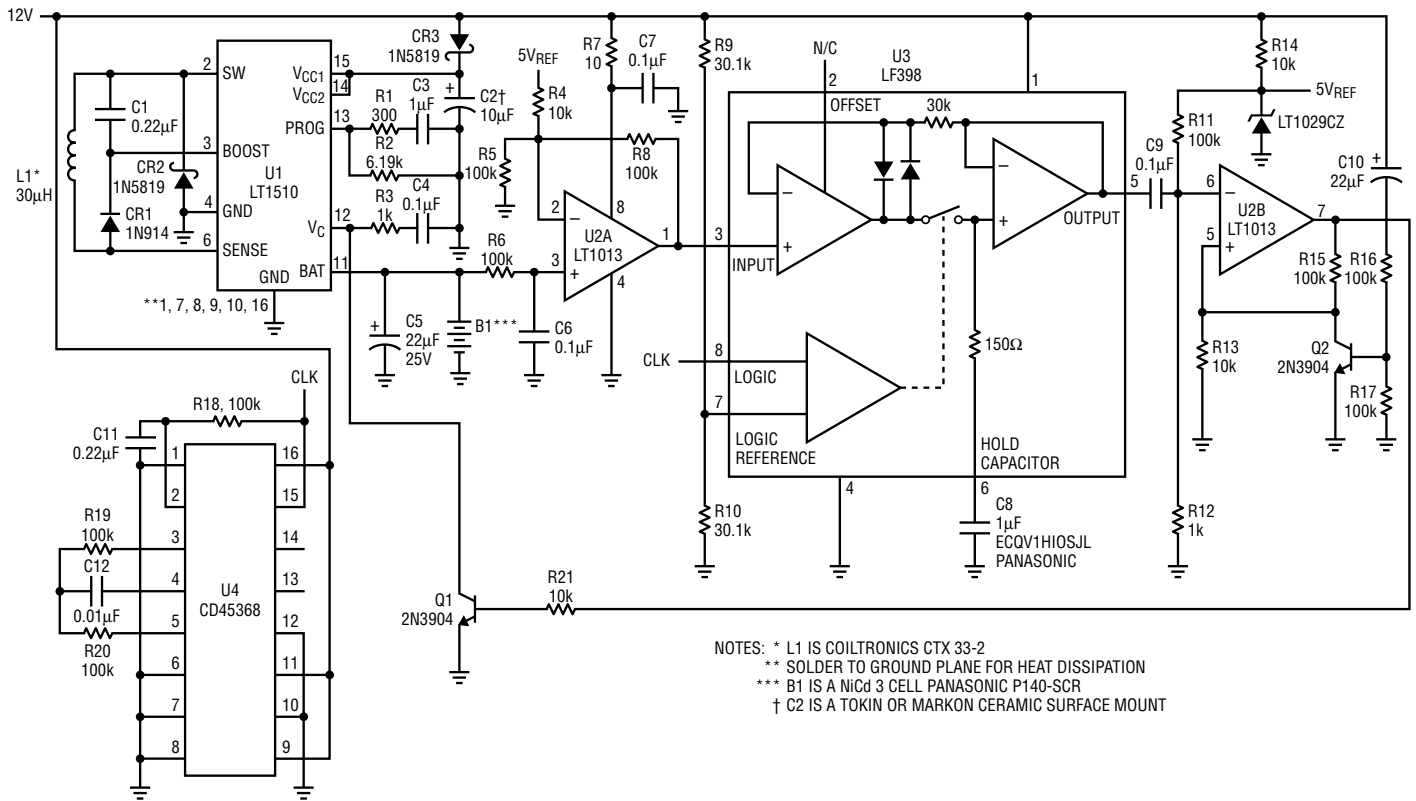$V_{TRIG} = V_{REF} \times R12/(R11 + R12) = 5 \times 1/101 = 49.5mV$
$V_{REF} = 5V$
$T_{CLK}$ is the clock period, 15 seconds,
$G_{U2A}$ is the gain of the first stage, = R8/(R4 || R5) = 11

**Figure 234. Schematic Diagram: 3-Cell NiCd Charger with −ΔV Termination**

The circuit in Figure 234 was built and connected to a system that discharges the battery to 3V after termination, at constant current of 0.8A. Once the battery drops to 3V, the system reenables charging, and thus the complete system repeats charge/discharge cycles indefinitely. The duration of 70 charge discharge cycles was recorded. The following is condensed data from the test:

1. Average Charge Time:                 2:00:55 Hours

2. Standard Deviation of Charge Time:     5:37 Minutes

3. Average Discharge Time:             1:59:14 Hours

4. Standard Deviation of Discharge Time:    48 Seconds.

The ratio of standard deviation of charge time to average charge time proves that the charger has good repeatability. However, the ratio of standard deviation of discharge time to average discharge time shows that the charge level at

the time of termination is very consistent because the discharge time at constant current is a better measure of charge level than charge time. A secondary termination method, such as time, battery temperature, or the like, is also recommended.
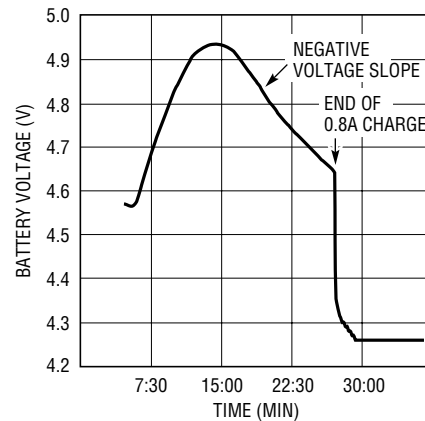


**Figure 235. Voltage-Droop Rate, 3-Cell NiCd Battery**

## CONSTANT-VOLTAGE LOAD BOX FOR BATTERY SIMULATION
*by Jon Dutra*

Linear Technology has developed many new switcher-based battery charger ICs. Testing accuracy, regulation and efficiency in the lab with a battery load is inconvenient because the terminal voltage of a battery constantly changes as it is being charged. If much testing is to be done, a large supply of dead batteries will be needed, since one set of cells can quickly become overcharged. This article describes an active load circuit that can be used to simulate a battery in any state of charge. The battery simulator provides a constant-voltage load for a battery-charging circuit, independent of applied charging current. The simulator's impedance is less than 500mΩ at all reasonable input frequencies. Best of all, the simulator can never be overcharged, allowing long-term testing and debugging of a charger system without the possibility of battery damage.

### Circuit Operation

The simulator (Figure 236) uses an LT1211 high speed, single-supply op amp to drive the base of a high gain PNP transistor-stage active load. Power for the LT1211—a portion of the charging current—is supplied through a diode so the op amp and reference can survive brief periods of zero charging current. The op amp is configured for a DC gain of four, so the voltage on its noninverting input is one fourth of the voltage that the load box is set to. With S1 open, the load-voltage adjust range will be from 10V to 20V, and with S1 closed it will be approximately 3.5V–10V. Low voltage operation could be improved by replacing the top LT1004-2.5 with an LT1004-1.2 and reducing R1, the reference bias resistor, to 1k. The 510Ω and 1.1k resistors are required for high frequency stability; they suppress a 1MHz oscillation. The 1N5400 diode and 4-amp fuse protect the circuit from reverse voltages.

### Results

The battery simulator circuit has been tested "swallowing" currents from 30mA to 3A with the output voltage essentially unchanged. When simulating a battery, the voltage adjust can be increased until the charger thinks the battery is fully charged and reduces the current into the simulator. Conversely, as the voltage is adjusted down, the battery charger may think the battery is becoming discharged and increase the current into the simulator.

Figure 237 shows the circuit's capacity for current absorption at two voltages, 5V and 15V, from 50mA to 3 amps.
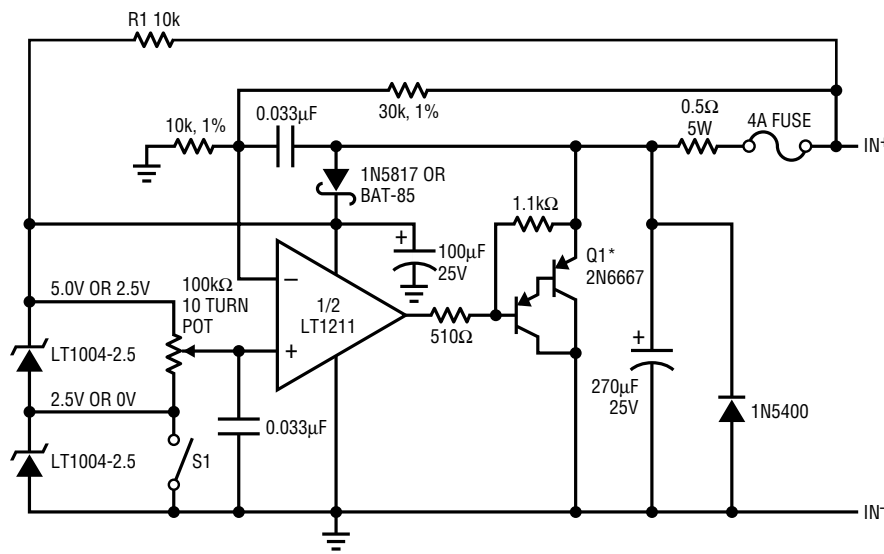


S1 CLOSED ≥ 0 TO 10V RANGE
S1 OPEN ≥ 10V TO 20V RANGE
ALL RESISTORS 5% UNLESS NOTED

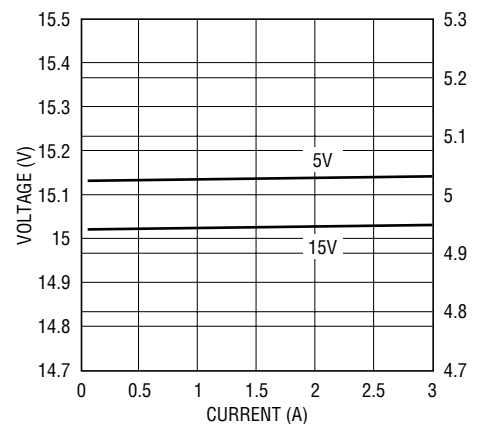* Q1 DISSIPATES MOST OF THE POWER, MOUNT ON AN ADEQUATE HEAT SINK

**Figure 236. Schematic Diagram of the Battery Simulator**



**Figure 237. Current Absorption Capacity of the Battery Simulator at 5V and 15V**

## HIGH EFFICIENCY, LOW DROPOUT LITHIUM-ION BATTERY CHARGER CHARGES UP TO FIVE CELLS AT 4 AMPS OR MORE
*by Fran Hoffart*

### Introduction

Rechargeable lithium batteries feature higher energy density per volume, higher energy density per weight and higher voltage per cell than any of the competing battery chemistries. For these reasons, manufacturers of portable equipment are adopting the lithium-ion rechargeable battery as the battery of choice for high performance portable equipment. Lighter weight and increased operating time between charges are important features that customers want and need from portable products.

Increased demands from laptop computers have forced manufacturers to use multiple cells in a combination of series and parallel configurations. Paralleling cells increases the amount of current that can be drawn from the battery and/or increases the operating time between charges, but it also increases the current requirements of the charger.
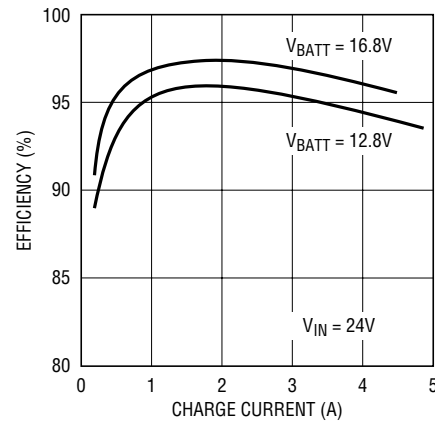


Figure 239. Charger Efficiency for 3- and 4-Cell Applications

### Higher Charge Currents

Paralleling cells, regardless of cell chemistry, requires relatively high charge currents to bring the battery up to full charge in a short period of time. When charging needs exceed the 3A maximum rating of the LT1511 or LT1513, the circuit shown in Figure 238 can provide much higher current solutions, and very high efficiency. This circuit uses the LTC1435 and LT1620 in a charger that delivers 4A or more with exceptional efficiency and low dropout voltage (Figures 238 and 239).
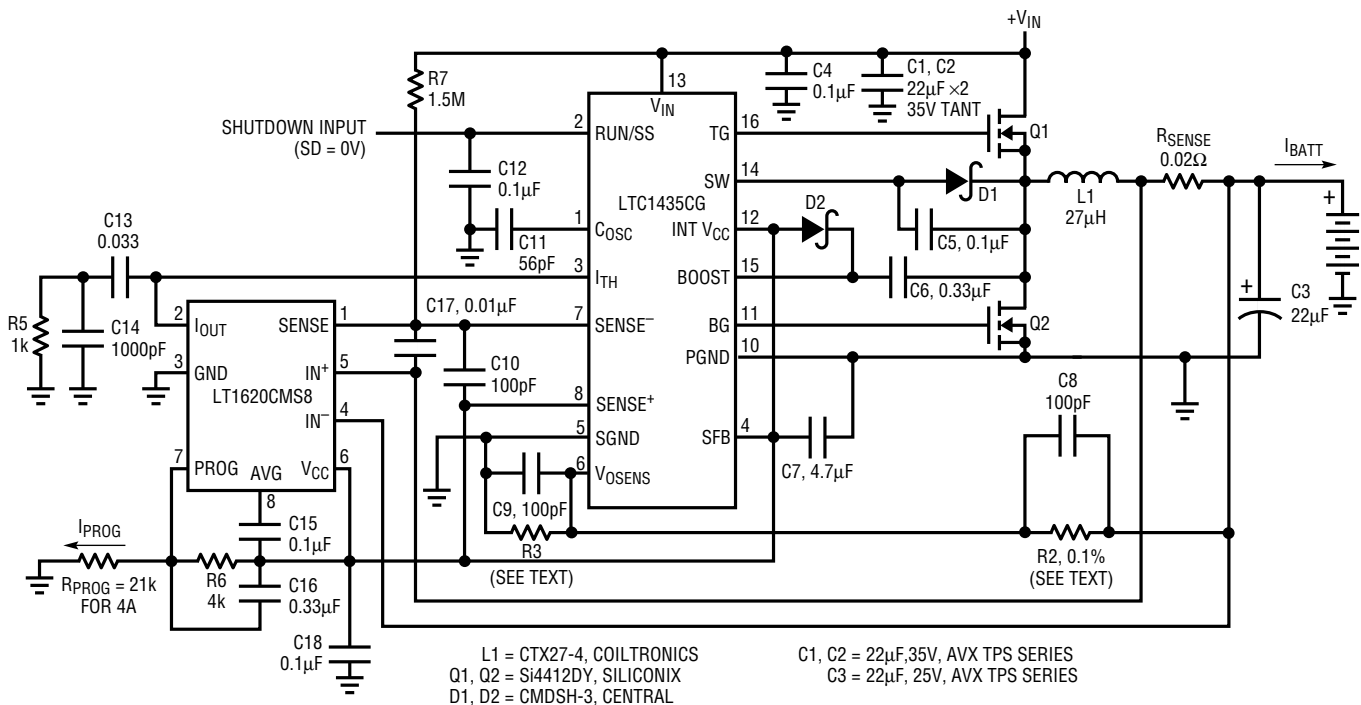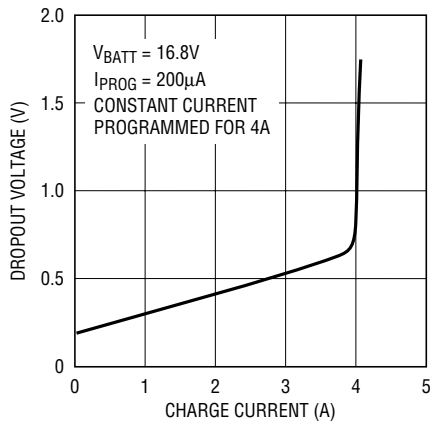


Figure 238. Complete Schematic of the High Efficiency, 4A, Constant-Voltage/Constant-Current Charger Using All Surface Mount Components, with a Circuit Board Area of 1.5in$^2$

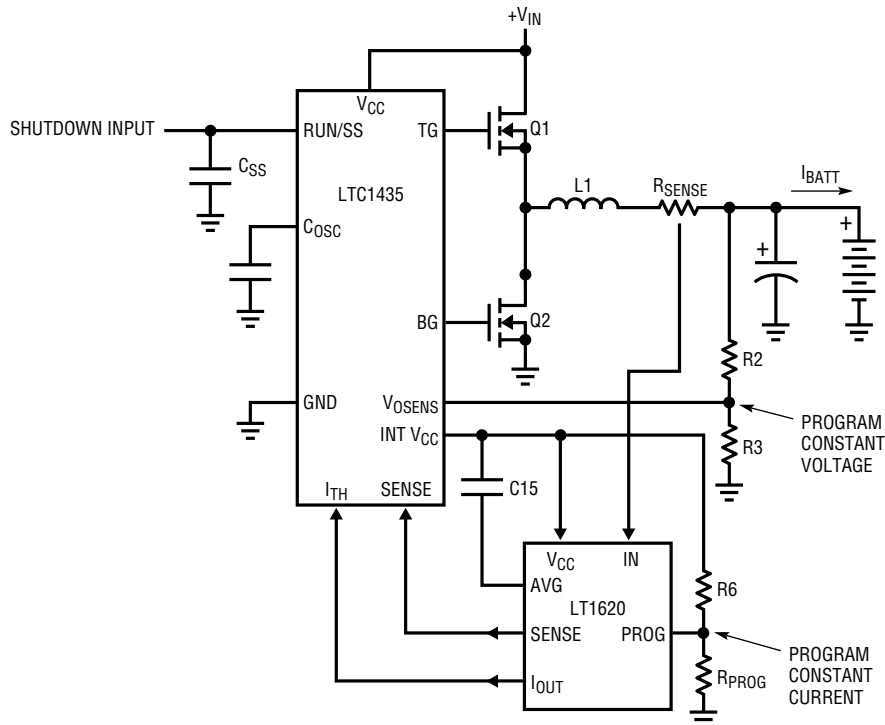**Figure 240. Charger Dropout Voltage vs Charge Current**

9V), programmable soft start, logic-controlled micro-power shutdown and a secondary feedback control pin. Because external MOSFET switches are used, the maximum output load current is determined by the current capabilities of the selected FETs.

## The LTC1435 as a Battery Charger

The low dropout voltage, high current capability and high efficiency of the LTC1435 switching regulator would seem to make it an appropriate choice for high current battery chargers, but it has several limitations. The absolute maximum output voltage of 10 volts allows only two series-connected lithium cells to be charged and the output current is not readily programmable.

## The LT1435 Switching Regulator Controller

The LTC1435 is a step-down current mode switching regulator controller designed to drive two external N-channel power MOSFETs. Operating from input voltages between 3.5V and 36V, this device includes a programmable switching frequency, synchronous rectification, Burst Mode™ operation and a 99% maximum duty cycle for low dropout voltage. Additional features include a 1% tolerance output voltage (adjustable between 1.2V and
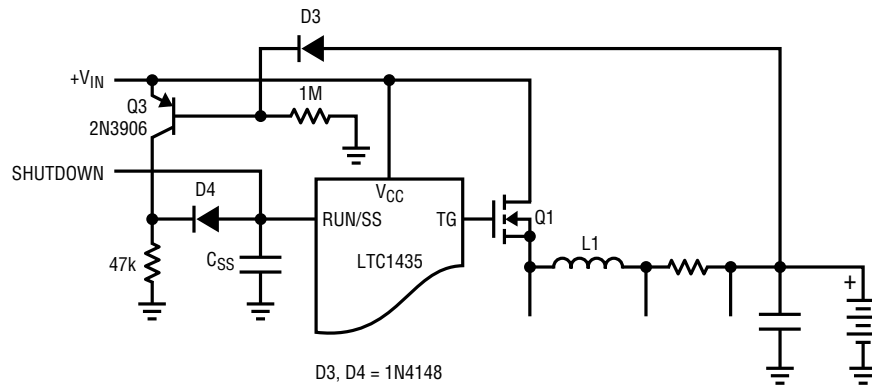
## Introducing the LT1620

The LT1620 is an IC designed to be used with a current mode PWM controller (such as the LTC1435 and similar products) to increase the output voltage range and optimize the circuit for battery charging applications. Used together, these two products overcome the voltage and



**Figure 241. Simplified Diagram of the Constant-Voltage/Constant Current Charger**

**Figure 242. Simplified Digram of Constant-Current Control Loop**

current programming limitations previously mentioned, to produce a high current, high performance constant-voltage/constant-current battery charger for lithium-ion and other battery types.

## How They Work Together

To understand how the two parts work together, a brief review of the LTC1435 operation is necessary. See Figure 241. During each cycle of operation, the series MOSFET switch Q1 is turned on by the LTC1435 oscillator (Q2 is off). This causes a current to begin ramping up in inductor L1. When the current in L1 reaches a peak level determined by the voltage at the $I_{TH}$ pin, Q1 is turned off and the synchronous MOSFET Q2 is turned on, causing the current in L1 to ramp down to the level at which it started. Thus, a sawtooth of inductor ripple current is generated, with a peak level set by the voltage on the $I_{TH}$ pin. This inductor current is sensed via an external, low value sense resistor in series with the inductor and is used to drive the LTC1435 internal current sense amplifier as the current mode feedback signal. This current sense amplifier has a maximum common mode voltage limit of 10V, which limits the maximum output voltage to 10V.

Enter the LT1620. The LT1620 also contains a current sense amplifier, which has a common mode range that

extends up to 28V. This amplifier is used to level shift the differential sense voltage, which is riding on the battery voltage, and reference it to the internal 5V $V_{CC}$ voltage generated by the LTC1435. This level-shifted signal is used to drive the LTC1435 current sense pins, thus providing current mode feedback for the constant-voltage feedback loop. This signal is also used to control the constant output current feedback loop, as explained below.

## Constant Charge Current

The LT1620 also provides a simple method of accurately programming the constant-current output. Sinking an adjustable current from the PROG pin to ground controls the charge current from zero current to maximum current. This program current can be derived from a variety of sources, such as a single resistor to ground or the output of a DAC.

The constant-current feedback loop operates as follows. With a discharged battery connected to the charger, and assuming that the battery voltage is less than the float voltage programmed by R2 and R3, the error amplifier in the LTC1435 begins pulling up on the $I_{TH}$ pin. This increases the peak inductor current in an effort to force the battery voltage to be equal to the programmed voltage. By

**Figure 243. Circuitry that Shuts Down the Charger when Input Power is Removed, Minimizing Reverse Battery Current Drain**

limiting the voltage on the $I_{TH}$ pin, the peak inductor current and the average output current can be controlled. The $I_{TH}$ pin has an internal 2.4V clamp that sets the peak inductor to its maximum level. This 2.4V clamp provides some degree of current regulation, but the average battery current will vary considerably as a result of dependence on inductor ripple current and LTC1435 parameter variations. By adding the LT1620 to the circuit, the constant charging current control performance is considerably improved. As shown in Figure 242, the signal from the current sense amplifier in the LT1620 is amplified by 10, averaged by $C_{AVG}$ (C15 in Figure 238) and compared to the voltage drop across R6. This voltage is developed by a current, $I_{PROG}$, flowing through R6. When the voltage at the LT1620 AVG pin approaches the voltage on the PROG pin, $I_{OUT}$ begins to pull the $I_{TH}$ pin of the LTC1435 down, limiting the peak inductor current and completing the constant-current feedback loop.

**Complete Charger Circuit**

The circuit shown in Figure 238 can charge up to five series-connected lithium-ion cells at currents up to 4A. Using low $R_{DS(ON)}$ MOSFET switches for the switch and synchronous rectifier results in efficiency exceeding 95% and allows all surface mount components to be used, resulting in a design that occupies less than 1.5 in$^2$ of board space. This circuit operates at a switching frequency of 200kHz and is capable of up to 99% duty cycle; it can operate over a very wide input voltage range, from a minimum input of only 600mV greater than the battery

charging voltage to a maximum of 28V (limited by the MOSFETs).

Constant-voltage charging with better than 1.2% accuracy and constant-current charging with 7.5% accuracy provides almost ideal lithium-ion battery charging conditions.

In battery charger designs, an important issue is reverse battery drain current caused by the charger when the input power is removed or the charger is shut down, or both. If the battery will remain connected to the charger for extended periods of time, it is important to minimize this reverse drain current to prevent discharging the battery. The charger can be shut down by using the RUN/SS pin on the LTC1435. This stops the charging current and results in a reverse battery drain current in the tens of microamps.

The LTC1435 and LT1620 have been configured so that the battery can remain connected to the charger when the input power is removed, but because of the inherent body diode in the Q1 MOSFET, current can flow from the battery, through the Q1 body diode, to the LTC1435's $V_{IN}$ pin, keeping it powered up. In this situation, because the charger is effectively powered by the battery, the reverse battery drain can be several mA, which could discharge the battery over an extended period. Figure 243 contains circuitry that automatically shuts down the LTC1435 when the input power is removed and puts it into a low quiescent current condition. Because the LT1620 is powered from the LTC1435 INT $V_{CC}$ pin, it is also turned off.

When input power is applied, the charger can still be shut down with an external signal to the RUN/SS pin. Shutdown occurs by pulling this pin low; releasing it allows the capacitor to charge up via the internal 3µA current source, producing a soft start.

By substituting higher current MOSFETs and changing some component values, much higher charging currents can be obtained.

**Selecting Battery Voltage Programming Resistors**

The charging voltage of lithium-ion cells is either 4.1 or 4.2 volts per cell, depending on the battery chemistry. Contact the battery manufacturer for the recommended charge voltage. To program battery charging voltage (float voltage) use the following equation (for best accuracy and stability, use 0.1% resistors).

$$V_{BATT} = V_{REF} \left(1 + \frac{R2}{R3}\right)$$

$V_{REF} = 1.19V$; USE APPROXIMATELY 100kΩ FOR R3

$$R2 = R3 \left(\frac{V_{BATT}}{V_{REF}} - 1\right)$$

**Selecting R_SENSE**

$R_{SENSE}$ is an external, low value resistor that is placed in the inductor current path to develop a signal representative of the inductor or charge current ($I_{BATT}$). This signal is used as feedback to control the switching regulator constant-voltage and constant-current loops. To minimize overall dropout voltage and power dissipation in the sense resistor, a sense voltage of 80mV was chosen to represent maximum charging current. Use the following equation to select current sense resistor $R_{SENSE}$. The maximum battery charge current (MAX $I_{BATT}$) must be known.

$$R_{SENSE} = \frac{0.08V}{MAX\ I_{BATT}}$$

**Selecting I_PROG**

$I_{PROG}$ is a current from the PROG pin to ground that is used to program the maximum charging current. $I_{PROG}$ can be derived from a resistor to ground, from the output of a DAC or by other methods. This program current is generated using resistors and the 5V $V_{CC}$ available from the LTC1435.

Refer to the simplified diagram of the constant-current control loop shown in Figure 242. The DC voltage across $C_{AVG}$ is proportional to the average charge current. This voltage drives one input of a transconductance ($g_m$) amplifier. A program voltage (relative to the 5V $V_{CC}$ line) proportional to the desired, or programmed charge current is applied to the other input of the transconductance amplifier. This voltage should be selected to be ten times the average voltage dropped across $R_{SENSE}$ when the charger is in a constant-current mode.

If the voltage across $C_{AVG}$ increases to a level equal to the voltage at the PROG pin, the transconductance amplifier begins pulling down on the $I_{TH}$ pin of the LTC1435, thereby limiting the peak inductor current, and thus the average charge current.

The program voltage needed on the program pin can easily be generated by two resistors, as shown in Figure 242. A current ($I_{PROG}$) is generated by these resistors and the 5V $V_{CC}$ voltage. This $I_{PROG}$ develops a voltage across R6, which is used to set the maximum constant charge current level. The circuit is designed for an approximate PROG voltage of 800mV (don't exceed the maximum spec of 1.25V), referenced to the LT1620 $V_{CC}$ pin. Because of the gain-of-10 amplifier, this corresponds to a typical voltage across $R_{SENSE}$ of 80mV (with a maximum of 125mV).

The recommended range of resistor values for R6 is approximately 2kΩ to 10kΩ. With 0.8V across R6, this will result in program currents ($I_{PROG}$) between 400µA and 80µA.

*LT* LINEAR
TECHNOLOGY

The LT1620 was designed to reduce the charging current to zero under all conditions when the $I_{PROG}$ is set to zero. To ensure that the charging current will always go to zero, an offset was designed into the transconductance amplifier. In the equations for R6 and $R_{PROGRAM}$, this offset is represented by using 840mV rather than 800mV.

Example:

GIVEN: MAXIMUM $I_{BATT}$ = 4A
$I_{PROG}$ = 200µA (FOR MAXIMUM $I_{BATT}$)

$$R_{SENSE} = \frac{0.08V}{MAX\ I_{BATT}} = \frac{0.08V}{4A} = 0.02\Omega$$

$$R6 = \frac{0.84V}{I_{PROG}} = \frac{0.84V}{200\mu A} = 4.2k\Omega$$

$$R_{PROG} = \frac{5V - 0.84V}{I_{PROG}} = \frac{5V - 0.84V}{200\mu A} = 20.8k\Omega$$

Once $R_{PROG}$ and R6 are known, the following equations can be used to determine $R_{PROG}$ and $I_{PROG}$ for lower $I_{BATT}$ currents:

$$R_{PROG} = \frac{R6\ [5 - 10(I_{BATT})(R_{PROG})]}{0.04 + 10(I_{BATT})(R_{PROG})}$$

$$I_{PROG} = \frac{10(I_{BATT})(R_{PROG}) + 0.04}{R6}$$

**PC Board Layout**

As with any high frequency switching regulator, layout is important. Switching current paths and heat producing thermal paths should be identified and the printed circuit board designed using good layout practices.

Even with efficiency numbers in the mid 90s, under some charging conditions power losses can be as high as 4 watts. These losses are primarily in the two MOSFETs, the inductor and the current sensing resistor. Since these are surface mount components, the major thermal paths are through the pc board copper to the surrounding air. Maximizing copper area around the heat producing components, increasing board area and using double-sided board with feedthrough vias all contribute to heat dissipation. Remember, the pc board is the heat sink.

One exception to the maximum copper area rule is the switch node consisting of Q1's source, Q2's drain and the left side of L1. This node switches between ground and $V_{IN}$ at a 200kHz rate. To minimize radiation from this node, it should be short and direct. Other copper traces related to input and output capacitors and MOSFET connections should also be as short as practical. See the LTC1435 data sheet for information on good layout practices and additional applications information.

**BATTERY CHARGER IC CAN ALSO SERVE AS MAIN STEP-DOWN CONVERTER**
*by LTC Applications Staff*

Using a power adapter with the highest feasible output voltage is attractive to portable system designers for a couple of reasons. Lower current is required to maintain the same system power, which translates into a smaller cable and input connector. If the adapter output voltage is considerably higher than the battery voltage, the adapter output voltage does not need to be regulated or well filtered, resulting in lower adapter cost.

A portable system with a high output-voltage adapter, however, requires that the system's DC-to-DC converter functions over a very wide range of input voltage: from fully discharged battery voltage to the highest adapter output voltage.

This problem can be resolved by using the LT1510 as both the battery charger and the main step-down converter, as shown in Figure 244. An important feature of the circuit in Figure 244 is the glitch-free transfer from AC operation to battery operation and back.

**Figure 244. LT1510 Battery Charger/Main Step-Down Converter Provides Glitch-Free Transfer between AC and Battery Operation**

The LT1510 battery charger IC is capable of charging currents up to 1.5A and output (battery) voltages up to 20V. High efficiency and small inductor size are achieved by a saturating switch running at 200kHz. The LT1510 is capable of charging lithium-ion and sealed-lead-acid batteries in the constant-voltage/constant-current configuration, and nickel-cadmium and nickel-metal-hydride batteries in the constant-current configuration. The LT1510 contains an internal switch and current sense resistor. All the designer needs to do in order to program the current and voltage is select the current-programming resistor and the voltage-divider resistors.

In the circuit shown in Figure 244, the system's DC-to-DC converter is connected to the SENSE pin. This way, the internal sense resistor is bypassed for the system load but is active in regulating the charging current. The sum of the charging current and system current should not exceed the maximum output current allowed (limited by thermal considerations or peak switch current). Since the DC-to-DC converter circuit has a large input capacitor, it cannot

be connected directly to the SENSE pin. This is because the internal sense resistor between SENSE and BAT pins will see a large capacitance across it, which will cause instability. A 2.2mH inductor, such as the DT1608C-222 by Coilcraft (L2), is used to isolate the input capacitance of the DC-to-DC converter. CR5 limits the transient current through the LT1510's internal sense resistor when the



**Figure 245. System Efficiency vs Input Voltage**

system is operating from the battery and turned on. Q2 (Si9433) is required if the series resistance of 0.2Ω between the BAT pin and SENSE pin is too high. The Si9433's on resistance is 0.075Ω. The charge pump comprising C8, C9, CR6, CR7 and R2 biases the gate of Q2. Q1 and R1 turn Q2 off on AC operation ($V_{IN}$ active). R7 programs the trickle-charge current (maximum value is about 100k) and the equivalent value of R7 and R8 programs the charge current. The Charge input must be pulled low at the end of the charge.

The charger in Figure 244 is connected to a 2-cell NiCd battery, BAT1. The system switching regulator is LT1300 (U2) based and powers a 5V/250mA load. The efficiency, η, of the complete system is defined as:

$$\eta = \frac{\text{LT1300 Output Power + Battery Charger Power}}{\text{LT1510 Input Power}}$$

The efficiency plot is shown in Figure 245. For the purpose of measurement, the battery voltage was 3.2V, the charging current was 0.4A and the trickle charge was 40mA.

## LT1635 1A SHUNT CHARGER
*by Mitchell Lee*

Most battery chargers comprise nothing more than a series-pass regulator with current limit. In solar-powered systems, you can't count on sufficient headroom to keep a series regulator alive, so a shunt method is preferred. A simple shunt battery charger is shown in Figure 246. It consists of an op amp driving a shunt transistor and ballast resistor, and is built around an LT1635. This device contains both an op amp and a reference, making it perfectly suited for regulator and charger applications.

Operation is straightforward: the battery voltage is sensed by a feedback divider composed of two 1M resistors. The internal 200mV reference is amplified to 7.05V and compared against the feedback. $R_{T1}$ introduces a TC that accurately tracks the battery's correct charging voltage over a wide temperature range. Because $R_{T1}$ is designed to compensate for changes in battery temperature, it should be located close to the battery and as far as possible from the shunt elements. When the battery charges to 14.1V, the op amp output voltage begins to rise, turning on the Darlington shunt and resisting further increases in voltage. Full panel power is divided equally between the transistor and 7.5Ω resistor when the battery is completely charged. Don't forget to provide adequate heat sinking and air flow for up to 15W dissipation.
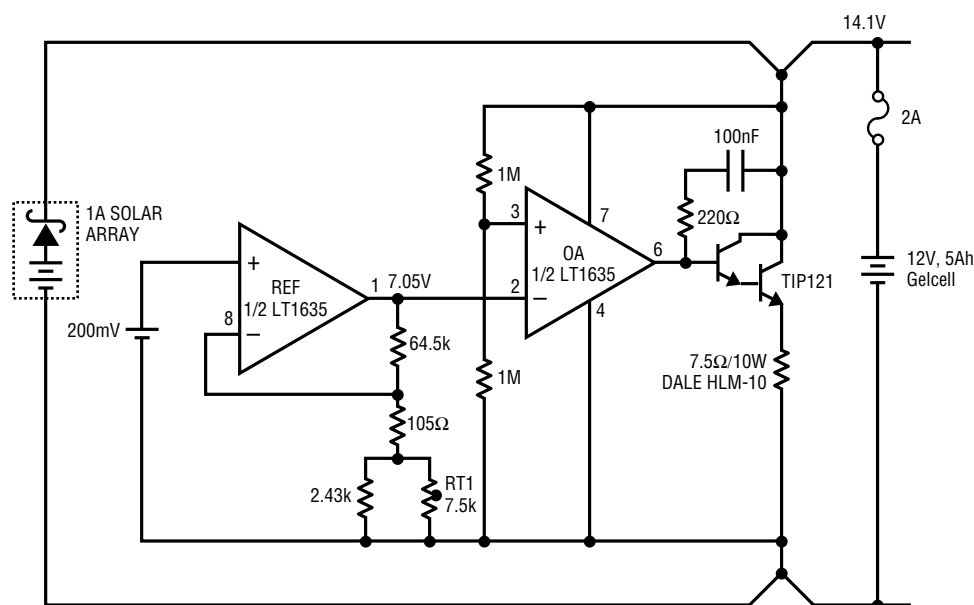


RT1 = THERM-O-DISC 1K752J

**Figure 246. 1A Shunt Battery Charger ($I_{DARK}$ = 230μA; $V_{FLOAT}$ = 14.1V)**

The charger is designed to handle 1A continuous, which is compatible with a "20W" panel. There is no need to disconnect or diode isolate the charger during periods of darkness, because the standby current is only 230µA— less than 10% of the self discharge of even a small battery.

If a different or adjustable output is desired, the feedback ratio can be easily modified at the 1M divider. 14.1V is a compromise between an aggressive charge voltage and a conservative float voltage. Given the cyclic nature of

insolation, allowing periodic charging at 14.1V is not detrimental to Gelcell™ batteries. The circuit in Figure 246 will work with larger or smaller batteries than that shown. As a rule of thumb, the panel should be sized from 1W per 10Ah battery capacity (a float charge under good conditions with a good battery) to 5W per 1Ah battery capacity (1 day recharge of a completely discharged battery under favorable conditions of insolation).

Gelcell is a trademark of Johnson Controls, Inc.

## 800mA Li-Ion BATTERY CHARGER OCCUPIES LESS VOLUME THAN TWO STACKED QUARTERS
*by Fran Hoffart*

Each new generation of cell phones, PDAs, portable instruments and other handheld devices is invariably more powerful, smaller and, most likely, thinner than the last. The circuit shown in Figure 247 is designed to charge one or two Lithium-Ion cells at currents up to 800mA, with all components equal to or less than 2.2mm (0.086 inches) tall. Using 0.031 inch PC board material, the total circuit thickness for this charger is 3.4mm (0.136in) or the thickness of two quarters. The complete 800mA constant-current/constant-voltage charger, including the PC board, occupies less volume than two quarters. This compact, low profile construction is ideal for cell phones or other applications where circuit height is restricted.

## LT1510-5CGN High Efficiency 500kHz Switch Mode Battery Charger IC

The charger consists of an LT1510 constant-voltage/constant-current PWM IC, which includes an onboard 1.5A switch. The LT1510 is available in either 200kHz or 500kHz versions; the higher frequency version allows lower value, smaller-sized inductors to be used. An internal 0.5% reference allows precision battery-voltage programming and a current programming pin allows a single resistor, PWM signal or a programming current from a DAC to control the charging current. Also included are undervoltage lockout and a low quiescent current sleep mode that is activated when input power is removed.



$$I_{BAT} = 2000\left(\frac{2.465V}{R1}\right)$$

$$V_{BAT} = 2.465V\left(1 + \frac{R5 + R6}{R4}\right)$$
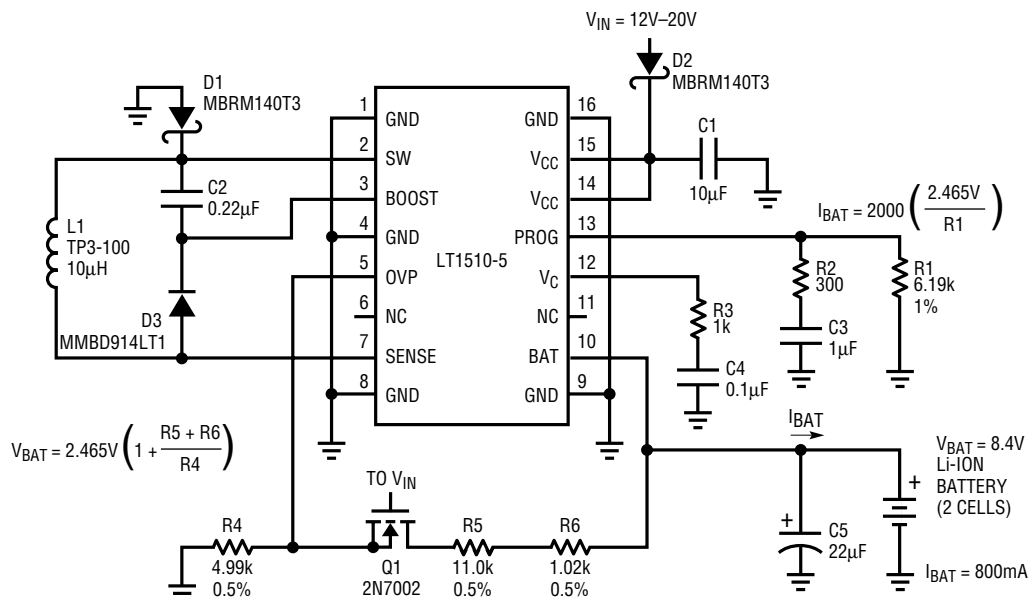
**Figure 247. Compact, Low Profile, Constant-Current/Constant-Voltage Charger for Li-Ion Batteries**

### Fused-Lead Package Offers Lower Thermal Resistance

The LT1510-5 is available in a specially constructed 16-lead plastic SSOP package that has the die-attach paddle connected (fused) directly to the four corner leads and fits in the same area as an SO-8 package. This low profile fused-lead package provides a lower thermal resistance by conducting much of the heat generated by the die through the copper leads to the PC board copper.

To take advantage of the improved thermal properties of this fused-lead package, it is important to provide as much PC board copper around the package leads as practical. Back-side copper and internal copper layers interconnected by feed-through vias all contribute to the overall effectiveness of the PC board as a heat sink.

### Charger Operation

A typical charge profile for a discharged Li-Ion battery is an initial constant-current charge at 800mA until the battery voltage rises to the programmed voltage. It then changes to a constant-voltage charge, with the charging current gradually decreasing to near 0mA as the battery approaches full charge. If complete charge termination is required, pulling the $V_C$ pin low or sinking zero current from the program pin stops the charge current. These signals could be supplied by an external timer or microprocessor.

When the input power is removed, the LT1510-5 goes into a low quiescent current (3µA) sleep mode, with this current coming from the battery. This low battery drain current allows the battery to remain connected to the charger for an extended period of time without appreciably discharging the battery. Additional battery-drain current can result from reverse leakage current in the Schottky catch diode D1. Many Schottky diodes have relatively high leakage currents, so care must be exercised in their selection.

Refer to the LT1510 data sheet for complete product specifications and to design notes DN111 and DN124 and application note AN68 for additional application information.

**Table 4. Low Profile Components Used in Figure 247's Circuit**

| Reference Designator | Quantity | Part Number | Description | Vendor | Phone |
|---|---|---|---|---|---|
| C1 | 1 | THCR50E1E106ZT | 10µF, 25V, 20% Y5U Ceramic | Marcon | (847) 696-2000 |
| C2 | 1 | 12063C224MAT1A | 0.22µF, 25V, 20% X7R Ceramic | AVX | (207) 282-5111 |
| C3 | 1 | 0805ZC105MAT | 1µF, 10V, 20% X7R Ceramic | AVX | (207) 282-5111 |
| C4 | 1 | 08055G104MAT1A | 0.1µF, 50V, 20% X7R Ceramic | AVX | (207) 282-5111 |
| C5 | 1 | EEFCD1B220R | 22µF, 12.5V, 20% Polymer Aluminum Electrolytic | Panasonic | (408) 945-5660 |
| D1, D2 | 2 | MBRM140T3 | 1A, 40V Schottky | Motorola | (800) 441-2447 |
| D3 | 1 | MMBD914LT1 | 0.2A 100V Silicon | Motorola | (800) 441-2447 |
| L1 | 1 | TP3-100 | 10µH Thin-Pac | Coiltronics | (561) 241-7876 |
| Q1 | 1 | 2N7002 | SOT-23 N-Channel MOSFET | Zetex | (516) 543-7100 |
| R1 | 1 | | 6.19k. 1% Chip Resistor | IRC | (512) 992-7900 |
| R2 | 1 | | 300Ω, 5% Chip Resistor | IRC | (512) 992-7900 |
| R3 | 1 | | 1k, 5% Chip Resistor | IRC | (512) 992-7900 |
| R4 | 1 | | 4.99k, 0.5% Chip Resistor | IRC | (512) 992-7900 |
| R5 | 1 | | 11.0k, 0.5% Chip Resistor | IRC | (512) 992-7900 |
| R6 | 1 | | 1.02k, 0.5% Chip Resistor | IRC | (512) 992-7900 |
| U1 | 1 | LT1510-5CGN | Battery Charger IC | LTC | (408) 432-1900 |

## SINGLE-CELL Li-Ion BATTERY SUPERVISOR
*by Albert Lee*

Recently introduced precision products from Linear Technology allow designers to implement high precision applications at supermicropower levels. Among these devices are the LT1496 quad precision input/rail-to-rail output op amp and the LT1634 precision shunt voltage reference, which operate at only 1.5µA and 10µA, respectively. Even at such low power levels, precision performance is not compromised. The LT1496 features 475µV maximum input offset voltage and 1nA maximum input bias current. The LT1634 achieves 0.05% initial accuracy and 25ppm/°C maximum temperature drift.

Figure 248 shows a single-cell Li-Ion battery supervisory circuit. The building blocks of this circuit are the LT1496 precision op amp and LT1634 voltage reference. The useful region of operation of a single-cell Li-Ion battery is between 4.2V and 3V. The cell voltage drops fairly quickly below 3V. System operation below this voltage can be erratic. Although Li-Ion battery use is becoming widespread, it is costly to damage the battery. The supervisory circuit protects the battery from overcharging and/or overdraining and prevents the battery voltage from falling out of its operating region. The LT1496 operates down to 2.2V, ensuring that circuit operation is maintained when the battery voltage falls below 3V.

The Li-Ion battery is monitored via a voltage divider off the battery voltage (node A). The divided voltage is fed into the positive inputs of comparators A2 and A3 and compared to the threshold voltages of 1.75V and 1.25V, respectively. These voltages are selected so that the minimum battery charge voltage is 3V and the maximum is 4.2V. The LT1634 1.25V reference is buffered by op amp A1. The constant 1.25V across R2 creates a 1µA constant current, so that the output of A1 is amplified to 1.75V. This output drives $R_S$ to provide constant bias current for the LT1634.

Depending on the battery voltage, the circuit is in one of the three states, as shown in Table 5.
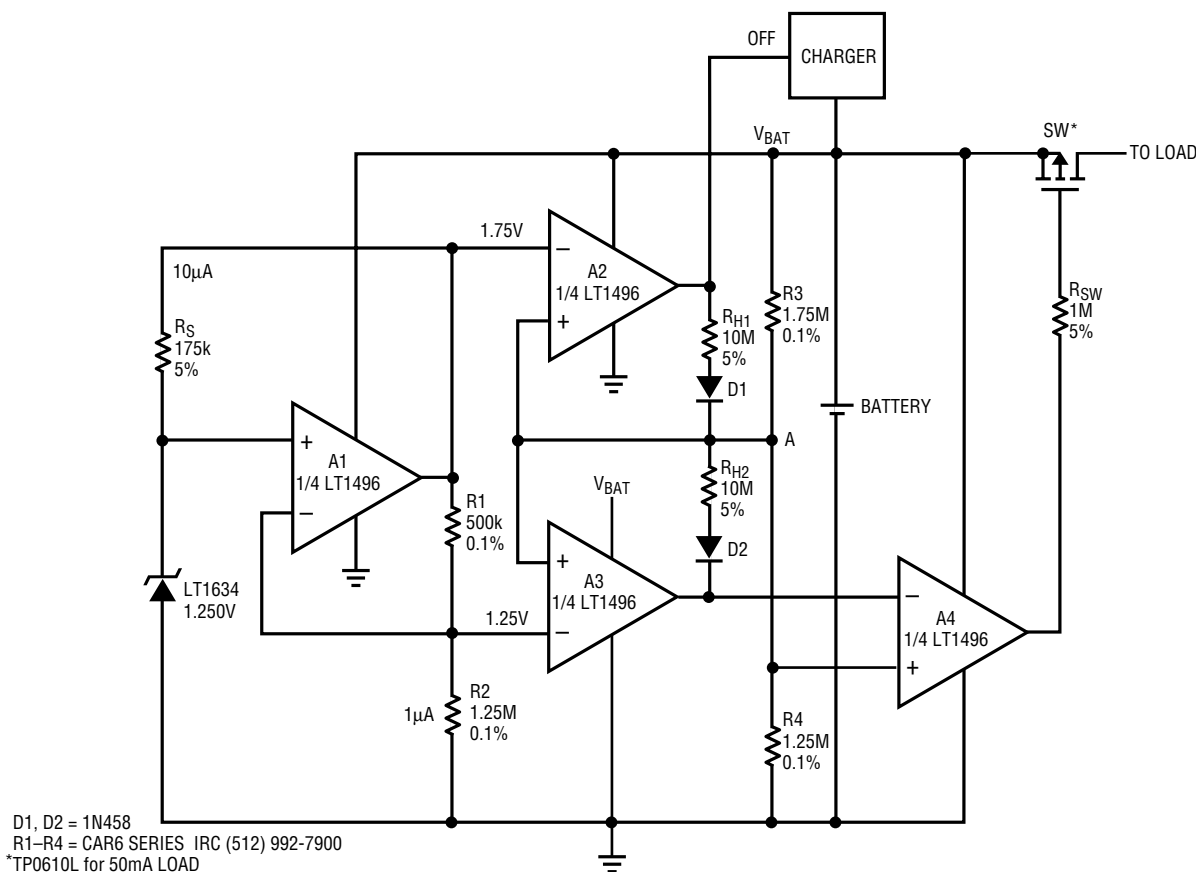


**Figure 248. Single-Cell Li-Ion Battery Supervisor Circuit**

The voltage at node A is compared to the two threshold voltages to determine the state of the circuit. For instance, when node A reaches or exceeds 1.75V (battery voltage reaches 4.2V), the outputs of A2 and A3 will swing to the positive rail, terminating the charger and connecting the load to the battery. When node A falls between 1.25V and 1.75V (battery voltage between 3V and 4.2V), the output of A2 swings low, turning the charger on, while the output of A3 stays high, leaving the load connected. When node A falls below 1.25V (battery voltage less than 3.0V), the output of A2 stays low, keeping the charger on. The output of A3 will also swing low, which, in turn, will cause the output of A4 to go high, turning off the FET SW that disconnects the load from the battery.

If node A were to bounce around at either threshold voltage, the circuit would bounce between states. To avoid this problem, hysteresis is added via the resistor and diode networks connected between the outputs of A2 and A3 and their positive inputs. Figure 249 shows the behavior of $V_{BAT}$ vs node A entering the trip points with hysteresis. When $V_{BAT}$ rises to 4.2V (node A increases to 1.75V), op amp A2's output will switch from low to high, causing current to flow through $R_{H1}$. The additional current will raise node A by an amount $\Delta VA_{HYS1}$, which will clearly put the circuit in state 3. The circuit will not exit state 3 until $V_{BAT}$ falls to $\Delta V_{HYS1}$ (310mV for the circuit shown) below 4.2V, which will cause node A to fall back to the upper trip point of 1.75V (point 1 of Figure 2). Similarly, when $V_{BAT}$ drops below 3V (node A falls below 1.25V), op amp A3's output will switch low, causing current to conduct through $R_{H2}$. This will drag node A an amount $\Delta VA_{HYS2}$ below 1.25V, which will put the circuit in state 1. The circuit will not exit state 1 until the battery voltage is charged to an amount $\Delta V_{HYS2}$ (149mV for circuit shown) above 3V (point 2). This will bring node A back up to the lower trip point, 1.25V, bringing the circuit out of state 1. The

amount of hysteresis desired can be calculated using the following formulas:

High Trip Point:

$$V_{BAT} = \frac{\left( \frac{R3}{R_{H1}} \bullet (V_{OHMIN} + V_{BE} + 1.75V) + (1.75V \bullet \frac{R3}{R4}) + 1.75 \right)}{1 + \frac{R3}{R_{H1}}}$$

$$\Delta V_{HYS1} = 4.2V - V_{BAT}$$

Low Trip Point:
$$I_{RH2} = (1.25V - V_{OLMAX} - V_{BE})/R_{H2}$$
$$\Delta V_{HYS2} = I_{RH2} \bullet R3$$
where:
$V_{OHMIN}$ = output voltage swing high (LT1496)
$V_{OLMAX}$ = output voltage swing low (LT1496)
$V_{BE}$ = diode voltage of 1N458

Using an automobile analogy, if the LT1496 op amp is the transmission of the circuit (switching from one state to the next), the LT1634 voltage reference is the engine. It not only generates the threshold voltages, but also the amount of error that the circuit will have. How much accuracy and error you get depends on the car you drive. Maximum input offset voltage and input bias current for the LT1496 are 475µV and 1nA, respectively. The LT1634 is a 0.05% initial accuracy, 25ppm/°C tempco, 10µA precision shunt reference. Its 1.250V output voltage will appear at the input of A3 with an accuracy of 0.088% (initial accuracy + input offset voltage). R1 and R2 being 0.1% resistors, the worst-case ratio error will be 0.2%. The worst-case volt-

**Table 5. Circuit States**

| State | $V_{BAT}$ | Node A | Output A2 | Output A3 | Output A4 | Status |
|-------|-----------|--------|-----------|-----------|-----------|--------|
| 1 | < 3V | < 1.25V | Low | Low | High | Load Off, Charge State |
| 2 | 3V< V < 4.2V | 1.25V < V < 1.75V | Low | High | Low | Load On, Charge State |
| 3 | > 4.2V | > 1.75V | High | High | Low | Load On, Charge Terminated |
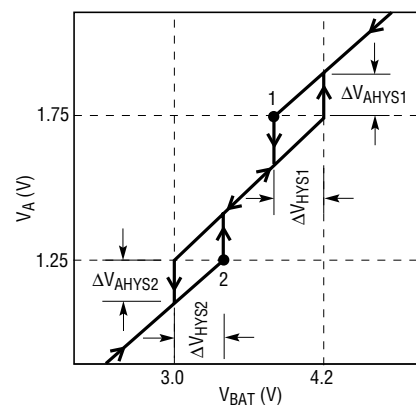


**Figure 249. $V_{BAT}$ vs $V_A$ with Hysteresis**

age error across R1 will then be 0.2% or 1mV. This error compared to the 1.75V threshold voltage is 0.057%. Similarly, error at 1.75V due to worst-case 2nA input bias current is 0.057%. Total worst-case error at 1.75V will be 0.202%.

$V_{BAT}$ error contributed by the voltage divider branch will consist of three terms: resistor matching, op amp input bias current and input offset voltage. The amount of error is different at the two trip points when $V_{BAT}$ is 3V or 4.2V. Similar calculations as above result in 0.328% when $V_{BAT}$ = 3V and 0.268% when $V_{BAT}$ = 4.2V. Therefore, total

battery voltage error at either trip points is better than 0.47%. Since only the ratios of R1 to R2 and R3 to R4 are critical, precision matched resistors with ten times better performance can be used to reduce the overall error by 33%.

This supervisory circuit demonstrates unparalleled performance achievable only with Linear Technology's supermicropower precision devices. The supervisory circuit consumes only 20mA. Battery voltage monitoring and control accuracy is better than 0.5%.

Power Management

## LTC1479 PowerPath CONTROLLER SIMPLIFIES PORTABLE POWER MANAGEMENT DESIGN
*by Tim Skovmand*

### Introduction

The LTC1479 PowerPath™ controller drives low loss N-channel MOSFET switches to direct power in the main power path of a dual rechargeable battery system, the type found in most notebook computers and other portable equipment.

Figure 250 is a conceptual block diagram that illustrates the main features of an LTC1479 dual-battery power management system, starting with the three main power sources and ending at the input of the DC/DC switching regulator.

Switches SWA/B, SWC/D and SWE/F direct power from either the AC adapter (DCIN) or one of the two battery packs (BAT1 and BAT2) to the input of the DC/DC switching regulator. Switches SWG and SWH connect the desired battery pack to the battery charger. These five switches are intelligently controlled by the LTC1479, which interfaces directly with the power management microprocessor.



**Figure 250. Dual PowerPath Controller Conceptual Block Diagram**

**Figure 251. Dual Li-Ion Battery Power-Management System (Simplified Schematic)**

## Typical Application Circuit

A typical dual Li-Ion battery power management system is illustrated in Figure 251. If "good" power is available at the DCIN input (from the AC adapter), both MOSFETs in switch pair SWA/B are on—providing a low loss path for current flow to the input of the LTC1538-AUX DC/DC converter. Switch pairs SWC/D and SWE/F are turned off to block current from flowing back into the two battery packs from the DC input.

## Battery Charging

The LTC1479 works equally well with both Li-Ion and NiMH batteries and chargers. In this application, an LT1510 constant-voltage, constant-current (CC/CV) battery charger circuit is used to alternately charge two Li-Ion battery packs.

The power management microprocessor decides which battery is in need of recharging by either querying a smart battery pack directly or by more indirect means. After the

determination is made, switch pair SWG or SWH is turned on by the LTC1479 to pass charger current to one of the batteries. Simultaneously, the selected battery voltage is returned to the voltage feedback input of the LT1510 CV/CC battery charger via a built-in switch in the LTC1479.

After the first battery is charged, it is disconnected from the charger circuit. The second battery is then connected through the other switch pair and the second battery charged. (The LTC1479 works equally well with the LT1511 3A CC/CV Battery Charger and LTC1435/LT1620 4A CC/CV Battery Charger.)

## Running on Batteries

When the AC adapter is removed, the LTC1479 instantly informs the power management microprocessor that the DC input is no longer "good" and the desired battery pack is connected to the input of the LTC1538-AUX high efficiency switching regulator through either switch pair SWC/D or SWE/F.

## Back-Up Power and System Recovery

Backup power is provided by a standby switching regulator, which is typically powered from a small rechargeable battery and ensures that the DC/DC input voltage does not drop below a predetermined level (for example, 6V).

## The "3-Diode Mode"

When the system is powered by the backup regulator, the LTC1479 enters a unique operating state called the "3-diode mode," as illustrated in Figure 252. Under normal operating conditions, both halves of each switch pair are turned on and off simultaneously. For example, when the input power source is switched from a good DC input (AC adapter) to a good battery pack, BAT1, both gates of switch pair SWA/B are turned off and both gates of switch pair SWC/D are turned on. The back-to-back body diodes in switch pair SWA/B block current flow in or out of the DC input connector.

In the 3-diode mode, only the first half of each power path switch pair, that is, SWA, SWC and SWE, is turned on; and the second half , that is, SWB, SWD and SWF, is turned off. These three switch pairs now act as 3-diodes connected to

the three main input power sources. The power path diode with the highest input voltage passes current through to the input of the DC/DC converter to ensure that the system cannot lock up regardless of how power is initially applied.

After "good" power is reconnected to one of the three main inputs, the LTC1479 drives the appropriate switch pair on fully as the other two are turned off, restoring normal operation.

## Interfacing to the Power Management Microprocessor

The LTC1479 takes logic level commands directly from the microprocessor and makes changes at high current and high voltage levels in the power path. Further, it provides information directly to the microprocessor on the status of the AC adapter, the batteries and the charging system.

The LTC1479 logic inputs and outputs are TTL level compatible and therefore interface directly with standard power management microprocessor. Because of the direct interface via five logic inputs and two logic outputs, there is virtually no latency (time delay) between the microprocessor and the LTC1479. In this way, time-
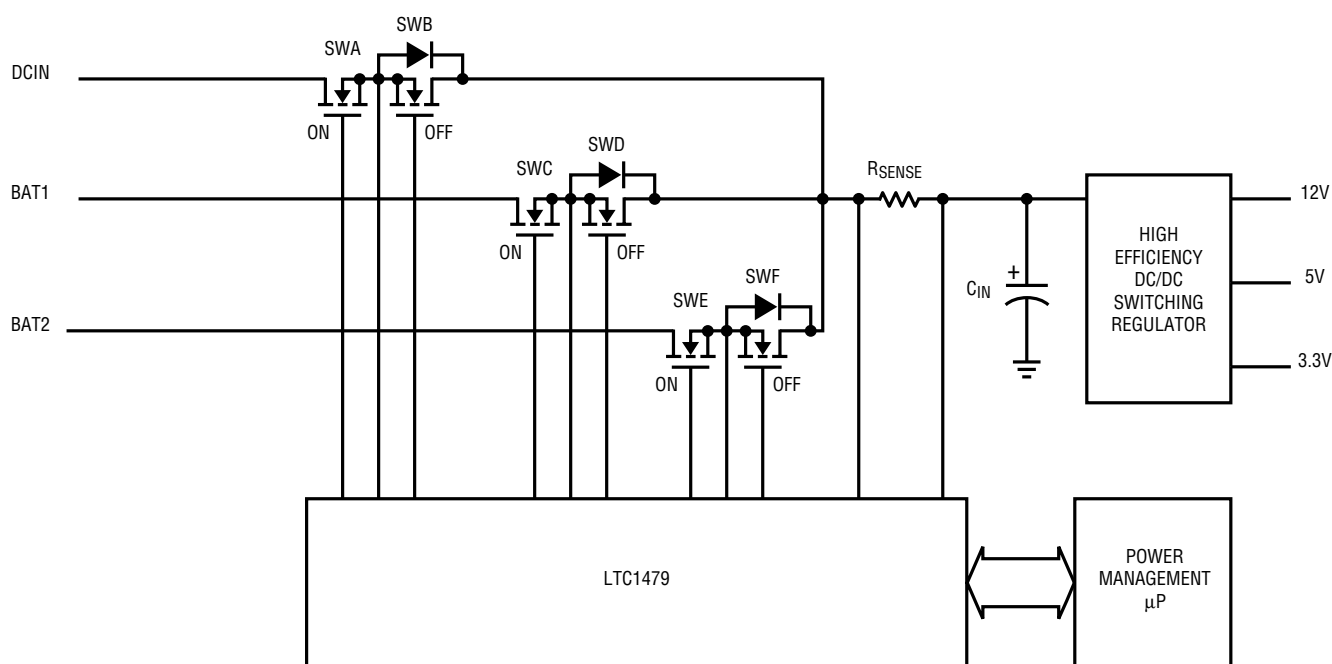


**Figure 252. LTC1479 PowerPath Controller in "3-Diode Mode"**

critical decisions can be made by the microprocessor without the inherent delays associated with bus protocols and the like. These delays are acceptable in certain portions of the power management system, but it is vital that the power path switching control be made through a direct connection to the power management microprocessor. The remainder of the power management system can be easily interfaced to the microprocessor through either parallel or serial interfaces.

## The Power Management Microprocessor

The power management microprocessor provides intelligence for the overall power system, and is easily programmed to accommodate the custom requirements of each system and to allow performance updates without resorting to costly hardware changes. Many inexpensive microprocessors are available that can easily fulfill these requirements.

## THE LTC1473 DUAL PowerPath SWITCH DRIVER SIMPLIFIES PORTABLE POWER MANAGEMENT DESIGN
*by Jaime Tseng*

## Introduction

The LTC1473 is the latest addition to Linear Technology's new family of power management controllers, which simplify the design of circuitry for switching between two batteries or a battery and an AC adapter. The LTC1473 dual PowerPath™ switch driver drives low loss N-channel MOSFET switches that direct power in the main power path of a single or dual rechargeable battery system, the type found in most notebook computers and other portable equipment.

## Overview

The power management system in Figure 253 shows the LTC1473 driving two sets of back-to-back N-channel MOSFET switches connecting the two batteries to the system DC/DC regulator. Each of the switches is controlled by a TTL/CMOS compatible input that interfaces directly with a power management system microprocessor. An internal boost regulator provides the voltage to fully enhance the logic-level N-channel MOSFET switches.
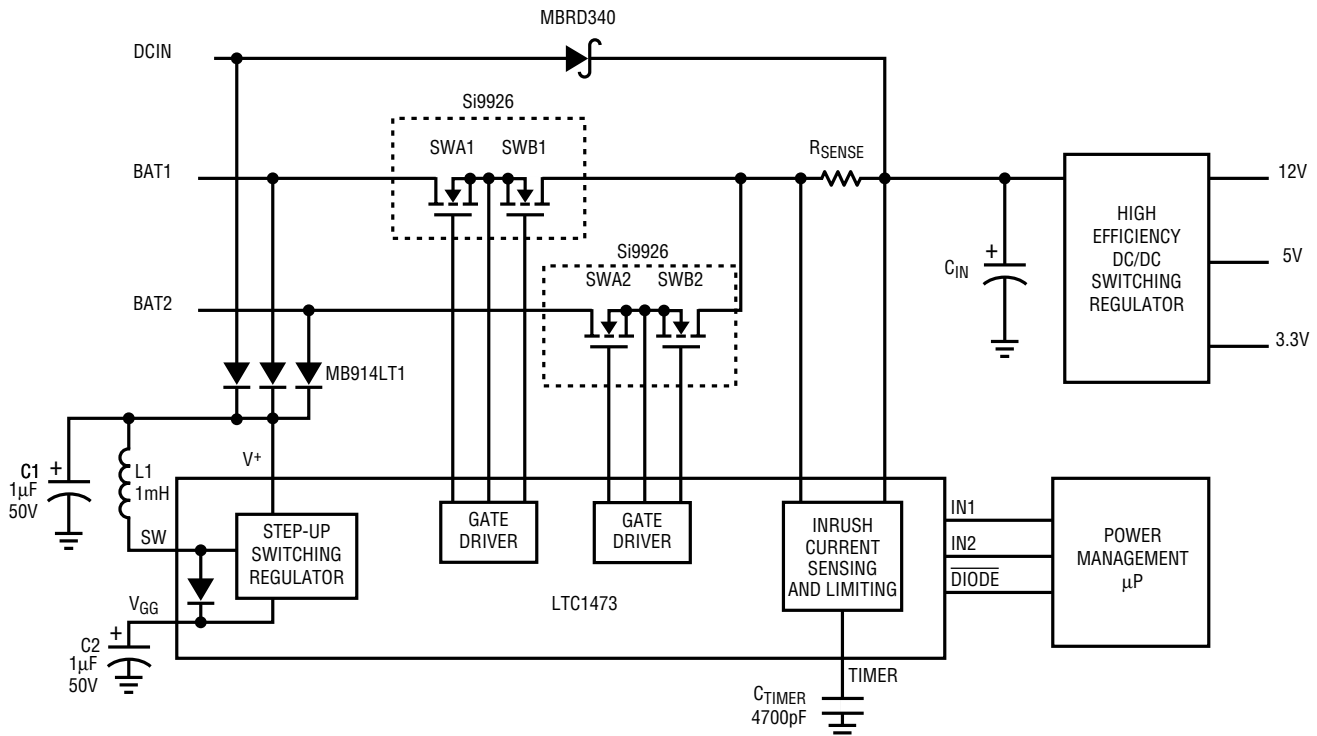


**Figure 253. Dual-Battery PowerPath Switch Driver: V$_{GG}$ Regulator, Inrush Limiting and Switch-Gate Drivers**

The LTC1473 uses a current sense loop to limit current rushing in and out of the batteries and the system supply capacitor during switch-over transitions or during a fault condition. A user programmable timer monitors the time during which the MOSFET switches are in current limit and latches them off if the programmed time is exceeded. A unique "2-diode logic mode" ensures system start-up, regardless of which input receives power first.

**Typical Application**

A typical dual-battery system is shown in Figure 254. The LTC1473 accepts commands from a power management microprocessor to select the appropriate battery. The microprocessor monitors the presence of batteries and the AC adapter through a supply monitor block, or, in the case of some battery packs, through a thermistor sensor. This block comprises a resistor divider and a comparator for each supply. If the AC adapter is present, the two switches are turned off by the microprocessor and the power is delivered to the input of the system DC/DC switching regulator via a Schottky diode.



**Figure 254. Dual-Battery Power-Management System**

## SHORT-CIRCUIT-PROOF ISOLATED
## HIGH-SIDE SWITCH
*by Mitchell Lee*

Figure 255 shows a MOSFET switch, driven by the LTC1177–5 2.5kV$_{RMS}$ isolator. This device allows a logic signal to control a power MOSFET and provides complete galvanic isolation. The device includes an internal current limiting circuit, but at higher voltages limiting the current is just not enough for effective protection of the MOSFET. Foldback (shown on the LTC1177 data sheet) helps, but the part has trouble starting certain types of loads when foldback current limiting is used. The circuit shown here latches off in an overcurrent condition and is restarted by cycling the logic input.

Q1 and Q2 form an SCR with a holding current of less than 100nA. If the load current exceeds approximately 1A, the SCR fires, shorting the MOSFET gate to source. The LTC1177 output current (about 7μA) is more than adequate to hold the SCR on indefinitely. The circuit resets when the logic input briefly cycles off.

Inductive loads present a special problem. If the load creeps up on the overcurrent threshold and fires the SCR, the load's inductance will carry the MOSFET source far below ground, which could destroy the MOSFET. Diode D1 clamps the gate at ground, turning the MOSFET back on, and safely dissipates the stored magnetic energy in the MOSFET.

As shown the output rise time is about 2ms, allowing the circuit to successfully charge capacitors of up to 100μF. Increase C1 proportionally to handle higher value load capacitors.



**Figure 255. Short-Circuit Protected, Isolated High-Side Switch**

## TINY MSOP DUAL SWITCH DRIVER IS SMBus CONTROLLED
*by Peter Guan*

### Introduction

The LTC1623 SMBus switch controller offers an inexpensive, space-saving alternative for controlling peripherals in today's complex portable computer systems. Pin-to-pin connections between the system controller and each peripheral device not only result in complicated wiring, but also limit the number and type of peripheral devices connected to the system controller. Using the SMBus architecture, the LTC1623 eliminates these problems by requiring only two bus wires and allowing easy upgrades and additions of new peripherals.

### Applications

The main application of the LTC1623 is to control two external high-side N-channel switches (Figure 256). As seen in the figure, a 0.1µF capacitor and a 1k resistor are placed on each gate-drive output to respectively slow down the turn-on time of the external switch and to eliminate any oscillations caused by the parasitic capacitance of the external switch and the parasitic inductance of the connecting wires.

Tracking the growing popularity of portable communication systems, the LTC1623 makes a very handy single-slot 3.3V/5V PC Card switch matrix. As shown in Figure 257, this circuit enables a system controller to switch either a 3.3V or a 5V supply to any of its SMBus-addressed peripherals. Besides N-channel switches, the LTC1623 can also be used to control a P-channel switch, as shown in Figure 258. As a result, the load connected to the P-channel switch will be turned on upon power-up of the LTC1623, whereas the other load must wait for a valid address and command to be powered.
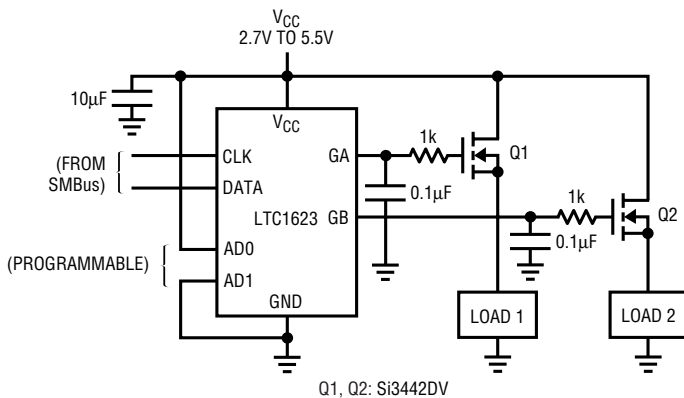


**Figure 256. LTC1623 Controlling Two High-Side Switches**
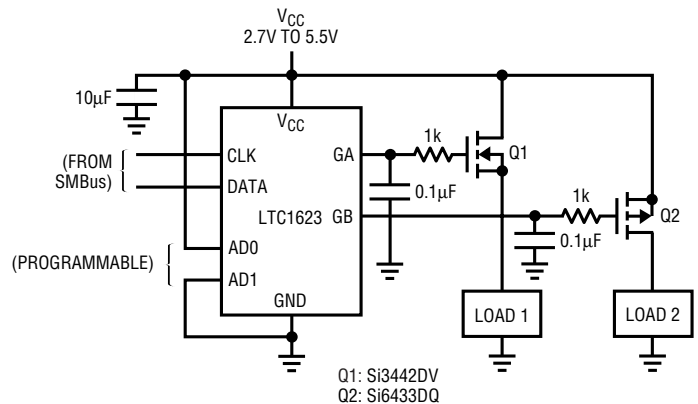


**Figure 258. LTC1623 Controlling a P-Channel Switch (Q2)**
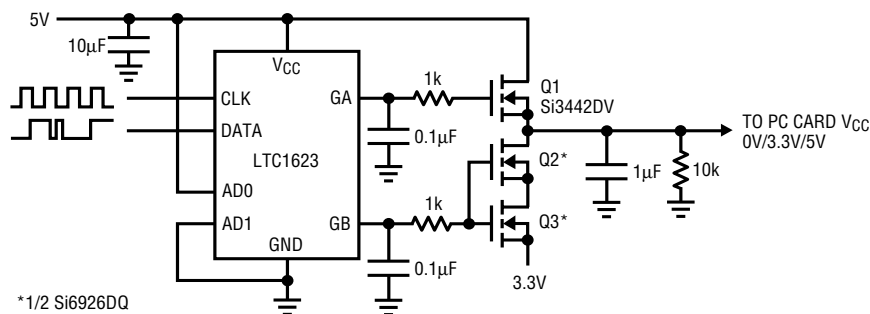


**Figure 257. PC Card 3.3V/5V Switch Matrix**

## LTC1710: TWO 0.4Ω SWITCHES WITH SMBus CONTROL FIT INTO TINY MSOP-8 PACKAGE

*by Peter Guan*

### Introduction

The LTC1710 SMBus dual switch (Figure 259) is a complete solution for supplying power to portable-equipment peripherals without the need for external switches. Two internal high-side N-channel switches, each capable of delivering 300mA at an $R_{DS(ON)}$ of 0.4Ω, are available in the tiny MSOP-8 package. With a low standby current of 14μA, the LTC1710 operates on an input voltage of 2.7V to 5.5V while maintaining the SMBus-specified 0.6V $V_{IL}$ and 1.4V $V_{IH}$ input thresholds.

Figure 260 shows a circuit using SMBus peripherals requiring different input voltages can be simultaneously switched by the LTC1710.



**Figure 259. Typical Application: The LTC1710 Switches Two SMBus Peripherals**
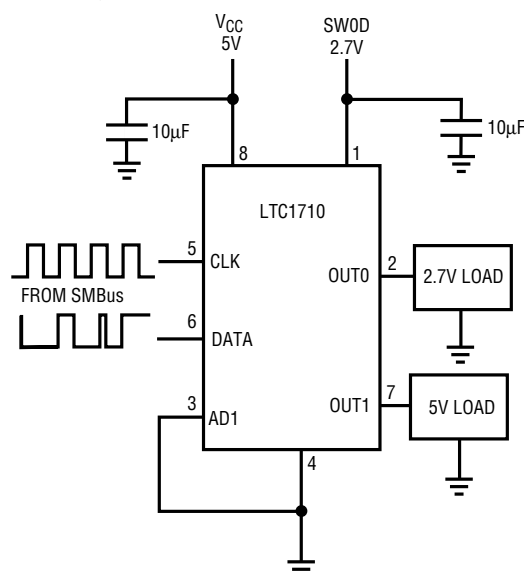


**Figure 260. LTC1710 Switches Two SMBus Peripherals with Different Input Voltages**

Miscellaneous

## VID VOLTAGE PROGRAMMER FOR INTEL MOBILE PROCESSORS
*by Peter Guan*

Microprocessor manufacturers' relentless push for higher speed and lower power dissipation, especially in areas of mobile laptop computer processors, is forcing supply voltages to these processors to a level previously thought impossible or impractical. In fact, the supply voltage has become so critical that different microprocessors demand different yet precise supply voltage levels in order to function optimally.

To accommodate this new generation of microprocessors, LTC introduces the LTC1706-19 VID (voltage identification) voltage programmer. This device is a precision, digitally programmable resistive divider designed for use with an entire family of LTC's DC/DC converters with onboard 1.19V references. These converters include the LTC1433, LTC1434, LTC1435, LTC1435A, LTC1436, LTC1438, LTC1439, LTC1538-AUX, LTC1539 and LTC1624. (Consult the factory for future compatible DC/DC converter products.) The LTC1706-19 is fully compliant with the Intel mobile VID specifications and comes in a tiny SO-8 package. Four digital pins are provided to program output voltages from 1.3V to 2.0V in 50mV steps with an accuracy of ±0.25%.
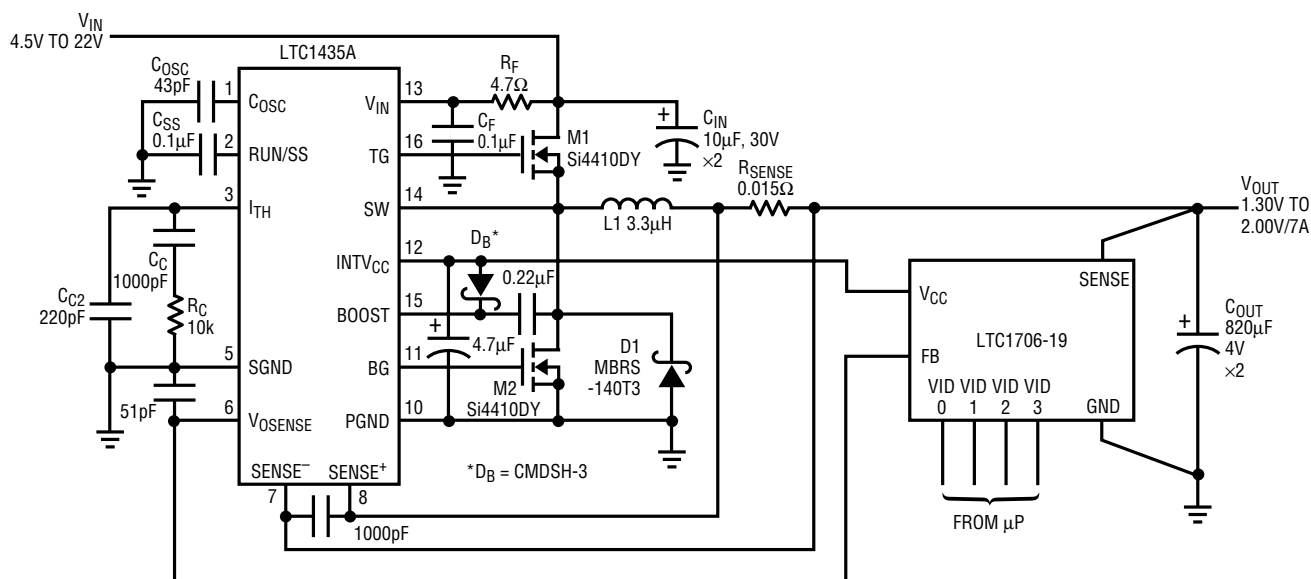
**Figure 261. Intel Mobil Pentium II Processor VID Power Converter**

## Applications

Figure 261 shows a VID-programmed DC/DC converter for an Intel mobile processor that uses the LTC1435A and LTC1706-19 to deliver 7A of output current with a programmable $V_{OUT}$ of 1.3V to 2.0V from a $V_{IN}$ of 4.5V to 22V. Simply connecting the LTC1706-19's FB and SENSE pins to the LTC1435A's $V_{OSENSE}$ and SENSE⁻ pins, respectively, closes the loop between the output voltage sense and the feedback inputs of the LTC1435A regulator with the appropriate resistive divider network, which is controlled by the LTC1706-19's four VID input pins.

Table 6 shows the VID inputs and their corresponding output voltages. VID3 is the most significant bit (MSB) and VID0 is the least significant bit (LSB). When all four inputs are low, the LTC1706-19 sets the regulator output voltage to 2.00V. Each increasing binary count is equivalent to decreasing the output voltage by 50mV. Therefore, to obtain a 1.30V output, the three MSBs are left floating while only VID0 is grounded. In cases where all four VID inputs are tied high or left floating, such as when no processor is present in the system, a regulated 1.25V output is generated at $V_{SENSE}$.

**Table 6. VID Inputs and Corresponding Output Voltages**

| Code | VID3 | VID2 | VID1 | VID0 | Output |
|------|------|------|------|------|--------|
| 0000 | GND | GND | GND | GND | 2.00V |
| 0001 | GND | GND | GND | Float | 1.95V |
| 0010 | GND | GND | Float | GND | 1.90V |
| 0011 | GND | GND | Float | Float | 1.85V |
| 0100 | GND | Float | GND | GND | 1.80V |
| 0101 | GND | Float | GND | Float | 1.75V |
| 0110 | GND | Float | Float | GND | 1.70V |
| 0111 | GND | Float | Float | Float | 1.65V |
| 1000 | Float | GND | GND | GND | 1.60V |
| 1001 | Float | GND | GND | Float | 1.55V |
| 1010 | Float | GND | Float | GND | 1.50V |
| 1011 | Float | GND | Float | Float | 1.45V |
| 1100 | Float | Float | GND | GND | 1.40V |
| 1101 | Float | Float | GND | Float | 1.35V |
| 1110 | Float | Float | Float | GND | 1.30V |

Figure 262 shows a combination of the LTC1624 and the LTC1706-19 configured as a high efficiency step-down switching regulator with a programmable output of 1.3V to 2.0V from an input of 4.8V to 20V. Using only one N-channel power MOSFET, the two SO-8 packaged LTC parts offer an extremely versatile, efficient, compact regulated power supply.

Figure 263 shows the LTC1436A-PLL and the LTC1706-19, a combination that yields a high efficiency low noise synchronous step-down switching regulator with programmable 1.3V to 2V outputs and external frequency synchronization capability.
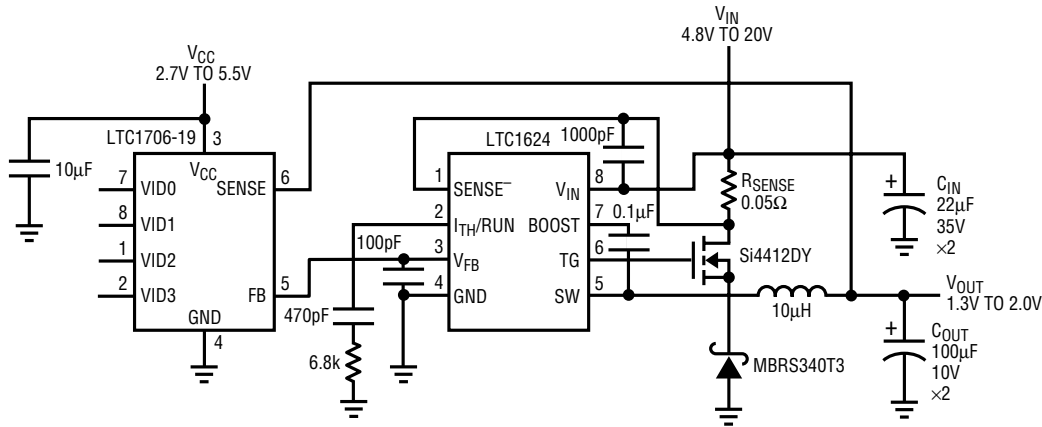
**Figure 262. High Efficiency SO-8, N-Channel Switching Regulator with Programmable Output**
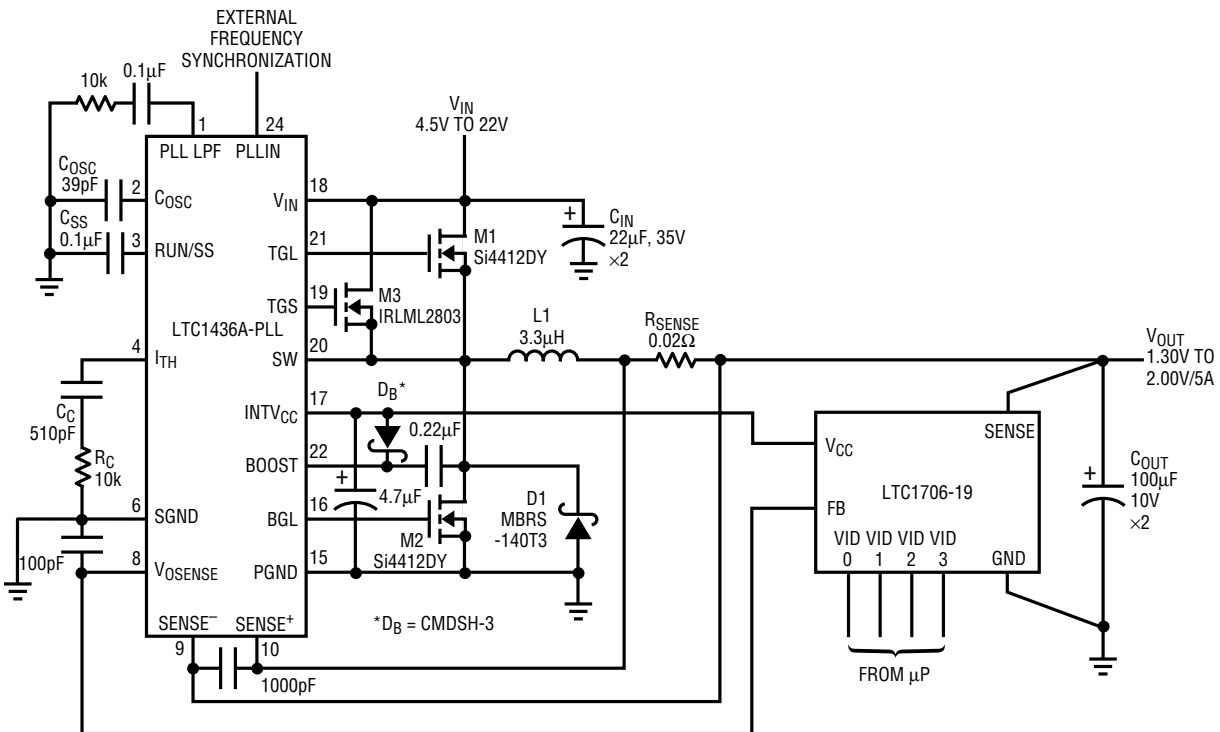
**Figure 263. High Efficiency, Low Noise, Synchronous Step-Down Switching Regulator with Programmable Output and External Synchronization**

Besides the LTC family of 1.19V-referenced DC/DC converters, the LTC1706-19 can also be used to program the output voltages of regulators with different onboard references. Figure 264 shows the LTC1706-19 programming the output of the LT1575, an UltraFast™ transient response,

low dropout regulator that is ideal for today's power-hungry desktop microprocessors. However, since the LT1575 has a 1.21V reference instead of a 1.19V reference, the output will range from 1.27V to 2.03V in steps of 50.8mV.
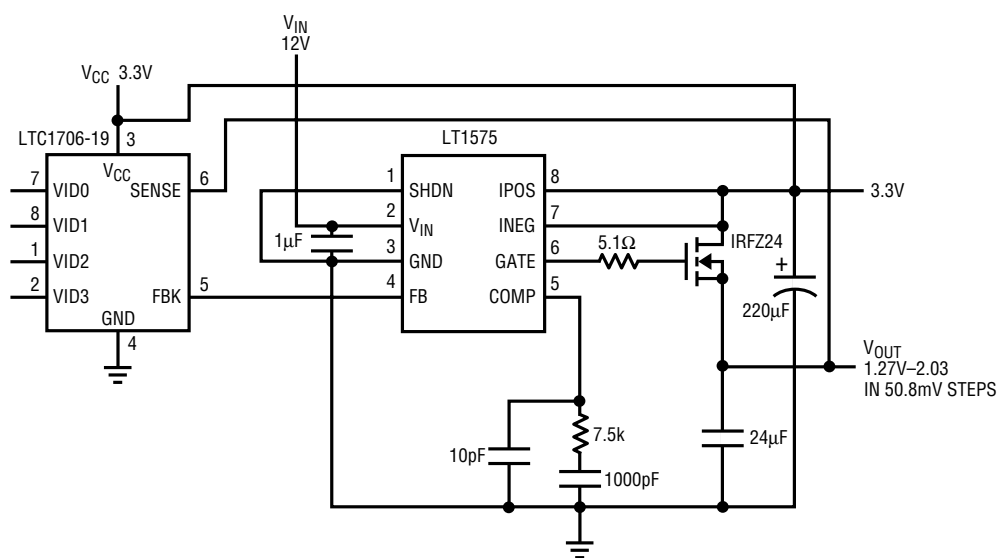


**Figure 264. UltraFast Transient Response, Low Dropout Regulator with Adjustable Output Voltage**

## BATTERY CHARGER IC
## DOUBLES AS CURRENT SENSOR
*by Craig Varga*

It's always fun to find applications for an IC that its designer never intended. The circuit shown in Figure 264 is such a design. In many cases, a circuit is required to provide a ground-referenced output voltage that is proportional to a measured current. Frequently, the current must be measured with a shunt in the positive rail that may be well above ground and, worse yet, may vary considerably with time. The LT1620 was originally intended as a controller for a synchronous buck regulator in battery-charger applications. The normal operating mode for this IC is to mirror a current signal down to a 5V reference supply. By adding a single small-signal MOSFET and a few resistors, it is possible to again mirror this signal to provide a ground referenced output.

Circuit operation is as follows: The LT1620 operates by producing a voltage between the $V_{CC}$ pin and the AVG pin that is 10× the voltage across sense resistor R5. C2 filters this voltage. An internal op amp has its noninverting input at the AVG pin (pin 8), its inverting input at the PROG pin (pin 7) and its output at the $I_{OUT}$ pin (pin 2). With the circuit connected as shown in Figure 265, this amplifier will force enough current through R4 to make the voltage drop on R4 equal to the voltage across C2. This current is mirrored through R3 and is filtered by C3, producing a clean, ground-referenced, DC output voltage. Resistor R2 cancels a small built-in offset in the LT1620's amplifiers. The output voltage obeys the following relationship: $V_O = I_L$ (R5 • R3 • 10)/R4. Changing the value of R3 selects different scale factors.

The circuit yields excellent linearity over a wide range of loads and input voltages. The curve shown in Figure 266 was measured with the sense resistor referenced to a 5V input source. The curve looks the same even at inputs over 25V, so only one curve is presented. Maximum input voltage is 36V. There is a small offset at no load, but in a typical microprocessor-based data acquisition system, only a simple 2-point calibration is needed to obtain absolute accuracy.
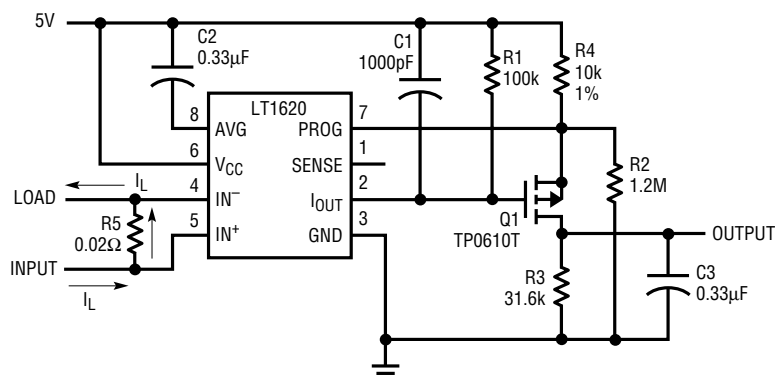


**Figure 265. Current Sensor Schematic**
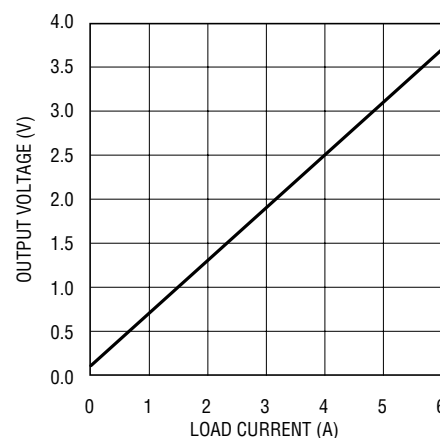


**Figure 266. Transfer Function**

## 100V, 2A, CONSTANT-VOLTAGE/ CONSTANT-CURRENT BENCH SUPPLY

*by Mitchell Lee and Jesus Rosales*

Most engineering labs are well stocked with low voltage, moderate current power supplies, but higher voltage supplies capable of several amperes of output current are hard to find. We solved this problem in our lab by building the supply shown in Figure 267.

The circuit is based on U1, an LT1270 high efficiency switching regulator configured in a SEPIC topology, which allows the output to be adjusted higher or lower than the input voltage. Operation is similar to that of a flyback converter, but the primary and secondary windings are coupled together by capacitor C1. This allows the primary and secondary windings to share current, reducing copper loss; it also eliminates the snubbing circuitry and losses found in flyback converters.

The converter is designed to operate from an input of 40V to 60V, supplied by a line transformer, diode bridge and filter capacitor (not shown). Output voltage is linearly adjustable from zero to 100V via potentiometer R20.

The current is limited by two independent loops. The first current limit loop is user controlled over a range of zero to 8A by setting potentiometer R21. This setting does not interact with changes in output voltage. A second current limit loop limits the maximum available current as a function of voltage (components R1–R5 and U2), minimizing component stress. Under any given operating condition, the lower of the two loops takes control. Maximum available output current is highest at low output voltage settings (about 8A), and decreases to 2A at 100V output.
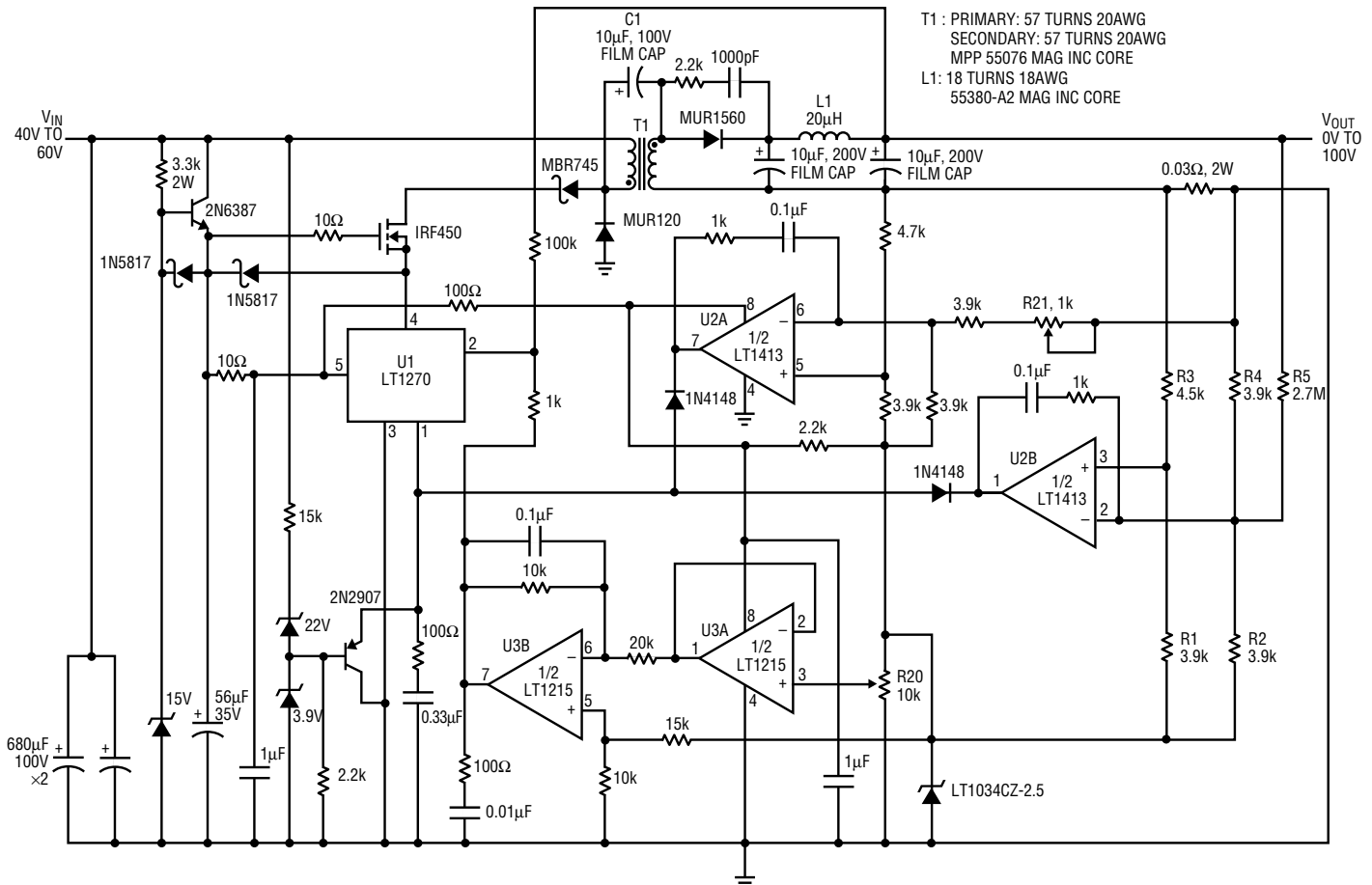


**Figure 267. Constant-Voltage/Constant-Current Bench Supply**

# A COMPLETE BATTERY BACKUP SOLUTION USING A RECHARGEABLE NiCd CELL

*by L.Y. Lin and S.H. Lim*

Battery-powered systems, including notebook computers, personal digital assistants (PDAs) and portable instruments, require backup systems to keep the memory alive while the main battery is being replaced. The most common solution is to use an expensive, nonrechargeable lithium battery. This solution requires low-battery detection, necessitates battery access and invites inadvertent battery removal. The LTC1558 battery backup controller eliminates these problems by permitting the use of a single, low cost 1.2V rechargeable Nickel-Cadmium (NiCd) cell. The LTC1558 has a built-in fast-/trickle-mode charger that charges the NiCd cell when main power is present.
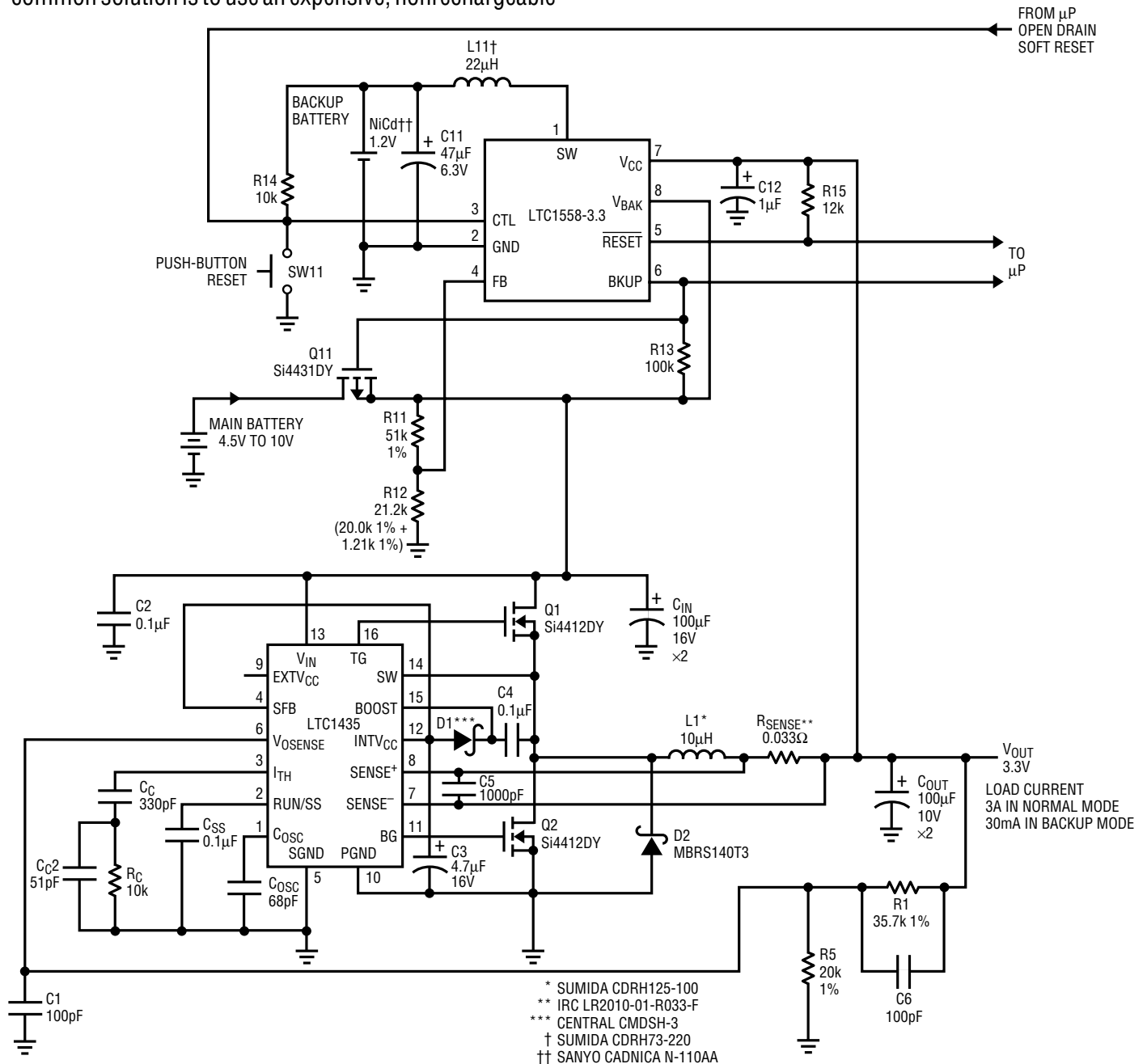


**Figure 268. LTC1558 Backup System with an LTC1435 as the Main System Regulator**
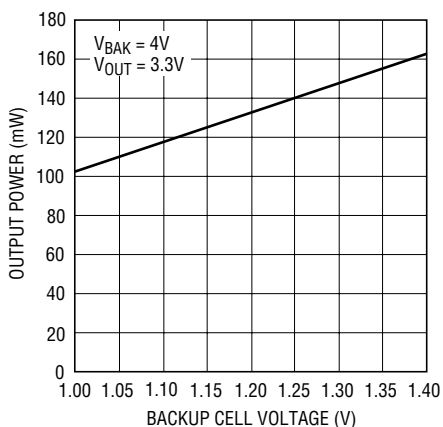
**Figure 269. 3.3V Output Power vs Backup Cell Voltage**
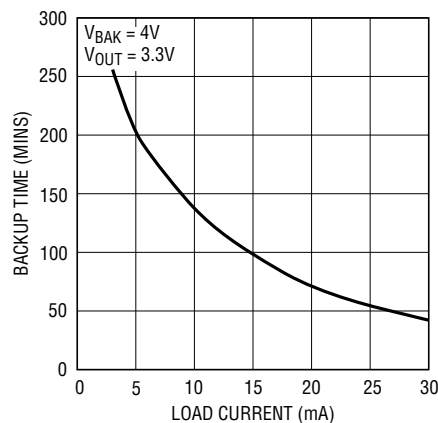


**Figure 270. Backup Time vs 3.3V Output Load Current**

Figure 268 shows a typical application circuit with an LTC1558-3.3 providing backup power to an LTC1435 synchronous step-down switching regulator. The backup circuit components consist of the NiCd cell, R11–R14, C11–C12, L11 and Q11. SW11 and R15 provide a soft or hard reset function.

## Normal Mode (Operation from the Main Battery)

During normal operation, the LTC1435 is powered from the main battery, which can range from 4.5V to 10V (for example, a 2-series or 2-series × 2-parallel Li-Ion battery pack, or the like) and generates the 3.3V system output. The LTC1558 operates in standby mode. In standby mode, the LTC1558 BKUP (backup) pin is pulled low and P-channel MOSFET Q11 is on. The NiCd cell is fast charged by a 15mA current source connected between the LTC1558's $V_{CC}$ and SW pins. Once the NiCd cell is fully charged (according to the LTC1558's gas-gauge counter), the LTC1558 trickle charges the NiCd cell. R14 sets the trickle-charge current according to the formula $I_{(TRICKLE)}$ = 10 • ($V_{NiCd}$ – 0.5)/R14. The trickle-charge current is set to overcome the NiCd cell's self-discharge current, thereby maintaining the cell's full charge.

## Backup Mode (Operation from the Backup Battery)

The main battery voltage is scaled down through resistor divider R11–R12 and monitored by the LTC1558 via the FB pin. If the voltage on the FB pin drops 7.5% below the internal 1.272V reference voltage (due to discharging or exchanging the main battery), the system enters backup

mode. In backup mode, the LTC1558's internal switches and L11 form a synchronous boost converter that generates a regulated 4V at $V_{BAK}$. The LTC1435 operates from this supply voltage to generate the 3.3V output voltage. The BKUP pin is pulled high by R13 and Q11 turns off , leaving its body diode reverse biased. The BKUP pin also alerts the system microprocessor. C11, a 47µF capacitor, provides a low impedance bypass to handle the boost converter's transient load current; otherwise, the voltage drop across the NiCd cell's internal resistance would activate the LTC1558's undervoltage-lockout function. Table 7 shows several values of $V_{FB}$ vs the $V_{BAK}$ voltage. Figure 269 shows the maximum output power available at the 3.3V output vs the NiCd cell voltage. Over 100mW of output power is achieved for a NiCd cell voltage greater than 1V. Figure 270 shows the backup time vs the 3.3V load current using a Sanyo Cadnica N-110AA cell (standard series with a capacity of 110mAhrs). Over one hour of backup time is realized for less than 80mW of 3.3V output power.

**Table 7. $V_{FB}$ and $V_{BAK}$ Voltages**

| Relative % Below $V_{REF}$ | % of $V_{REF}$ | $V_{FB}$ | $V_{BAK}$ |
|---|---|---|---|
| –0% | 100% | 1.272V | 4.325V |
| –6% | 94% | 1.196V | 4.065V |
| –7.5% | 92.5% | 1.177V | 4.000V |

**Recovery from Backup Mode to Normal Mode**

When a new main battery pack is inserted into the system, Q11's body diode forward biases. Once the voltage at the FB pin increases to more than 6% below $V_{REF}$, the boost converter is disabled and the system returns to normal mode. The BKUP pin pulls low and turns Q11 back on. This allows the new battery pack to supply input power to the LTC1435. The LTC1558 now accurately replenishes the amount of charge removed from the NiCd cell through the internal charger and gas-gauge counter.

## WHAT EFFICIENCY CURVES DON'T TELL
*by San-Hwa Chee*

### Introduction

In switching regulators' data sheets, there are always efficiency curves that show how efficient the regulators are in transforming one voltage to another. Although these curves are useful in comparing one regulator to another, they don't allow a system designer to determine accurately how long batteries will last before they need to be replaced or recharged when they are used as the power source. This complication arises because the type of batteries used to power the system and the regulator load characteristic strongly affect the lifetime of the batteries.

In this article, battery lifetime curves are obtained for the LTC1174 and the LTC1433.
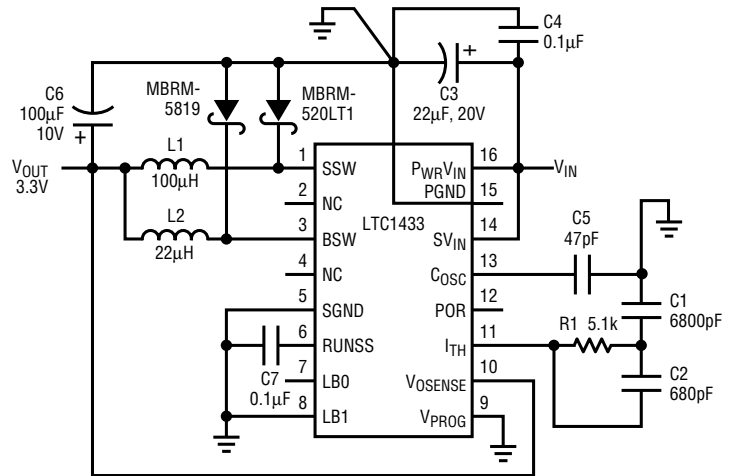


L1 = SUMIDA CD54-101
L2 = SUMIDA CD54-220

**Figure 271b. LTC1433 Dual-Inductor Configuration**

### A Short Introduction to the LTC1174 and LTC1433

The LTC1174 uses a constant off-time architecture to switch its internal P-channel power MOSFET. The input-to-output voltage ratio sets the on time and requires the inductor current to reach a preset limit. Even at low load current, the LTC1174 still requires the inductor current to reach the preset limit before it initiates the off-time cycle. Burst Mode operation of the LTC1174 enhances efficiency



L1 = SUMIDA CD54-220

**Figure 271a. LTC1433 Single-Inductor Configuration**
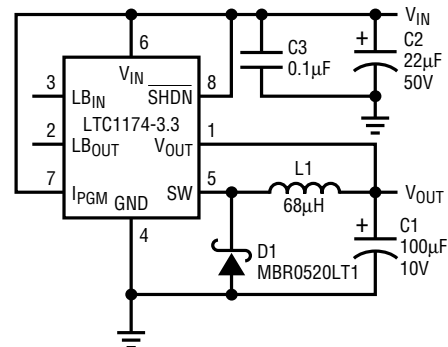


L1 = SUMIDA CDRH74-680

**Figure 271c. LTC1174 Test Circuit**
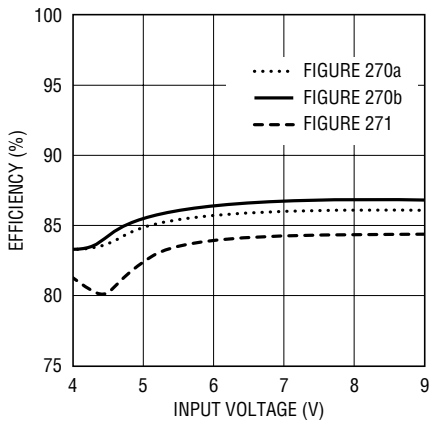
**Figure 272a. Efficiency Curves for
Figure 271's Circuits, I$_{LOAD}$ = 400mA**
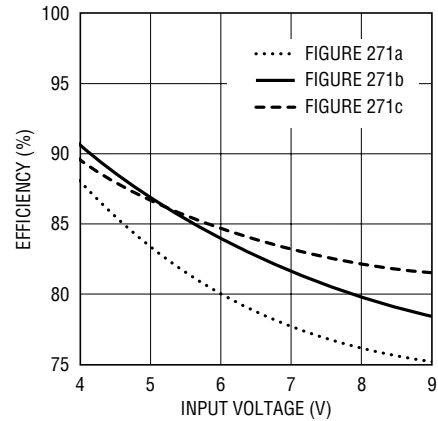


**Figure 272b. Efficiency Curves for
Figure 271's Circuits, I$_{LOAD}$ = 10mA**

throughout the load-current range by switching only the required number of cycles to bring the output into regulation and then stopping switching (going into sleep mode). When the output voltage has dropped slightly, the switching sequence resumes. By doing this, switching losses are

reduced and are minimized when the load current is low, because the sleep duration is long.

The LTC1433 is a constant-frequency, current mode, monolithic switching regulator in which the inductor peak
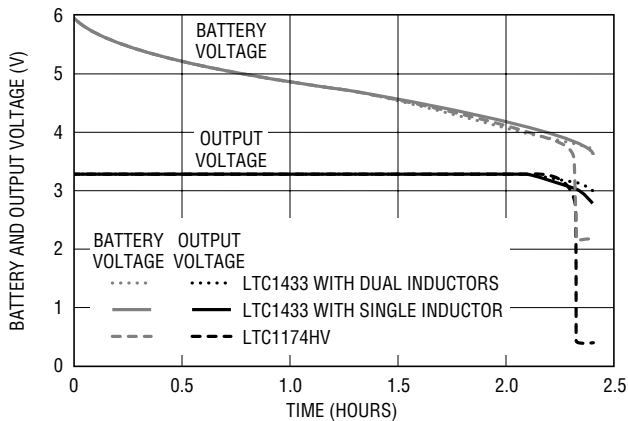


**Figure 273. Lifetime at I$_{LOAD}$ =
400mA—Four AA Alkaline Batteries**
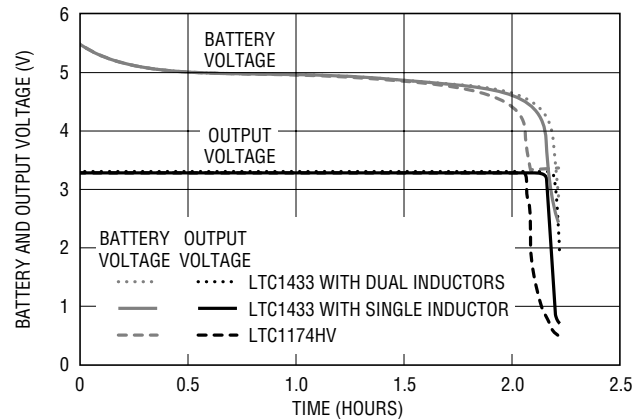


**Figure 274. Lifetime at I$_{LOAD}$ =
400mA—Four AA NiCd Batteries**
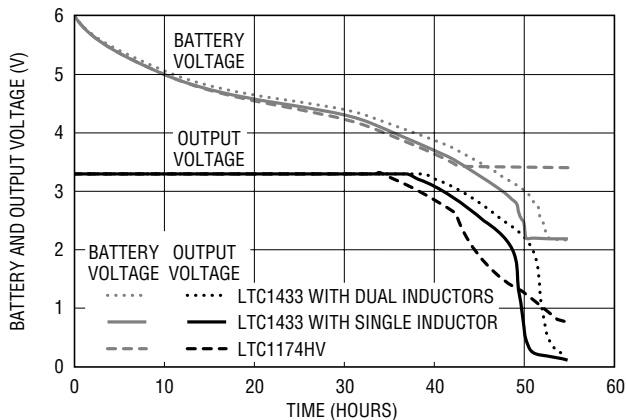


**Figure 275. Lifetime with Load Step from 10mA to 410mA,
10% Duty Cycle, T$_{PERIOD}$ = 20s—Four AA Alkaline Batteries**
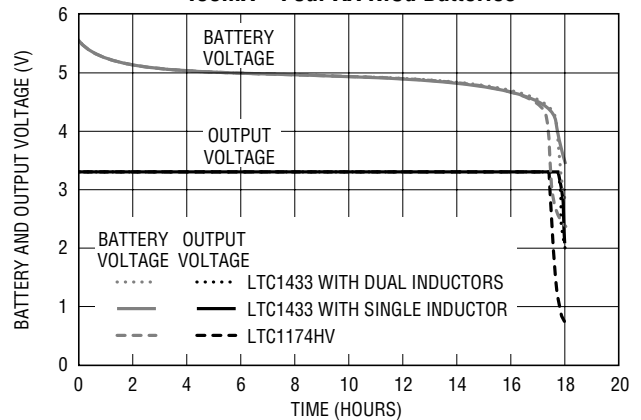


**Figure 276. Lifetime with Load Step from 10mA to 410mA,
10% Duty Cycle, T$_{PERIOD}$ = 20s—Four AA NiCd Batteries**
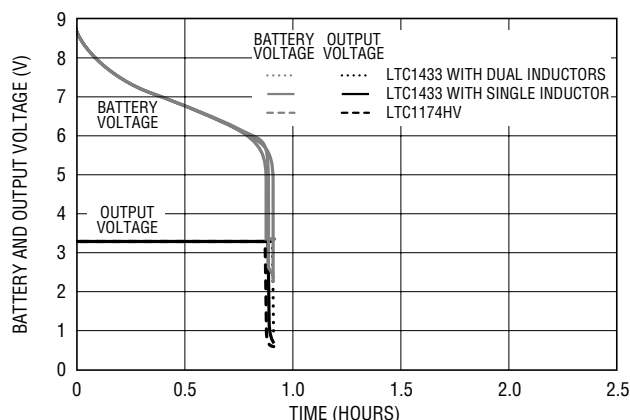
**Figure 277. Lifetime at I$_{LOAD}$ = 400mA—on a 9V Alkaline Battery**



**Figure 278. Lifetime with Load Step from 10mA to 410mA— one 9V Alkaline Battery (10% Duty Cycle, T$_{PERIOD}$ = 20s)**

current varies according to the load current. In place of Burst Mode operation, the LTC1433 has an Adaptive Power output stage to enhance its efficiency at low load current. Under low load conditions, the LTC1433 uses only a fraction of its power MOSFET, effectively reducing switching losses without introducing low frequency noise components.

For more information on both parts, consult the data sheets.

## The Setup

The circuits in Figures 271a, b and c were used to obtain the lifetime data. All outputs were set at 3.3V and the power was supplied by either four AA alkaline (Eveready No. EN91) or four AA NiCd (Eveready No. CH15) cells or a single 9V alkaline (Eveready No. EN22) battery. A current-sink load was set up to either draw a constant 400mA or provide a load-step characteristic. The load stepping operated at 0.05Hz, going from 10mA to 410mA with a duty cycle of 10%, providing an average load current of 50mA.

In Figure 271b, the LTC1433 was set up to optimize low load current efficiency by configuring the Adaptive Power output stage with separate inductors for low and high current operation.

Efficiency curves for each circuit are shown in Figure 272a and 272b. Figures 273 through 278 show the battery voltage and regulator output voltage versus time for various battery and load combinations.

## 4-Cell to 3.3V Configuration

Figures 273 and 274 were obtained with a load current of 400mA. For Figure 273, the input power to the regulator was provided by four AA alkaline batteries, whereas four AA NiCds were used in Figure 274. The alkaline batteries lasted longer than the NiCds, due to their higher energy capacity. From Figure 274, it is apparent when the NiCd gives up, from the cliff-like shape of the output voltage.

For Figures 275 and 276, a step load was applied to the regulators instead of a DC load. Figure 275 and 276 are the data obtained for alkaline and NiCd AA cells, respectively. With the average load one-eighth of the previous experiment, it would be expected that the lifetime of the alkaline batteries would be eight times longer or approximately 18 hours, but Figure 275 shows a significantly better result. The main reason for this improvement has to do with the internal resistance of the alkaline cell. At high constant DC load current, heat is dissipated by the internal resistance of the alkaline batteries. The internal resistance increases as the batteries voltage decreases, and hence causes more heat to be dissipated, thus lowering the lifetime.

For the NiCd battery, internal resistance is low and remains relatively constant over its life span. Therefore, the lifetime of the NiCd batteries for the load step case comes out to be approximately the expected eight times that of a constant DC load current.

The above result indicates that if the load is intermittent in nature, the user can operate the device much longer if the power is provided by alkaline batteries. Again, the NiCd exhibits a sudden "death" at the end of its life, whereas the alkaline shows a much gentler decay. The gentle sloping of the output voltage of Figure 275 towards the end of the battery life can be attributed to the on-resistance of the switch when the regulator is in dropout.

For the above load characteristic, where the load is light most of the time, making full use of the Adaptive Power mode of the LTC1433 by means of the dual inductor configuration helps to squeeze an additional 1.5 hours of life compared to the single inductor LTC1433 configuration.

Another important point to note is that although the efficiency for the LTC1174 is better than that of the single inductor configuration of the LTC1433 at 10mA load current, the LTC1433 lasted 2.9 hours longer than the LTC1174 in Figure 275. The reason for this is that the LTC1174 inductor's current always ramps up to the preset value of 600mA whether the load current is at 10mA or at 410mA. This high peak inductor current, combined with the high internal resistance of the alkaline AA cells, shortens the lifetime. Figure 276 shows that the use time is about the same for the LTC1174 and the LTC1433 because of the low, constant internal resistance of the NiCd batteries.

### 9V-to-3.3V

The lifetime graphs are shown in Figures 277 and 278. Comparing the data between the 9V and the AA alkaline cells, the lifetime of the AA cells is about 2.5 times longer. This is because the energy capacity of the 9V alkaline is much smaller than that of the AA cells. In addition, the internal resistance of the 9V alkaline is much higher than the AA cells, causing more energy to be dissipated as heat. For the load step case, the battery lasted 13.8 times longer than a constant 400mA load. The dual inductor configuration of the LTC1433 lasted about an hour longer than the single inductor one.

## APPENDIX A: COMPONENT VENDOR CONTACTS

The tables on this and the following pages list contact information for vendors of non-LTC parts used in the application circuits in this publication. In some cases, components from other vendors may also be suitable. For information on component selection, consult the text of the respective articles and the appropriate LTC data sheets.

| Capacitors | | | |
|---|---|---|---|
| Vendor | Product | Phone | URL |
| AVX | Chip Capacitors | (843) 946-0362 | www.avxcorp.com/products/capacitors |
| AVX | Tantalum Capacitors | (207) 282-5111 | |
| Electronic Concepts | 400V Film Capacitors | (908) 542-7880 | www.eci-capacitors.com |
| Kemet | Tantalum Capacitors | (408) 986-0424 | www.kemet.com |
| Marcon | High C/V Capacitors | (847) 696-2000 | www.chemi-con.com/main/company/marcon.html |
| Murata Electronics | Capacitors | (814) 237-1431 | www.iijnet.or.jp/murata/products/english |
| Nichicon | Electrolytic Capacitors | (847) 843-7500 | www.nichicon-us.com |
| Panasonic | Poly Capacitors | (714) 373-7334 | www.panasonic.com/industrial_oem/electronic_components/electronic_components_capacitors_home.htm |
| Sanyo | Oscon Capacitors | (619) 661-6835 | www.sanyovideo.com |
| Sprague | Capacitors | (207) 324-4140 | www.comsprague.com |
| Taiyo Yuden | Chip Capacitors | (408) 573-4150 | http://www.t-yuden.com |
| Tokin | Capacitors | (408) 432-8020 | www.tokin.com |
| United Chemicon | Electrolytic Capacitor | (847) 696-2000 | www.chemi-con.com/main |
| Vitramon | Ceramic Chip Capacitor | (203) 268-6261 | www.vishay.com |
| Wima | Paper/Film Capacitors | (914) 347-2474 | www.wimausa.com |

| Diodes | | | |
|---|---|---|---|
| Vendor | Product | Phone Number | URL |
| Agilent (formerly Hewlett Packard) | IR LEDs | (800) 235-0312 | www.semiconductor.agilent.com/ir |
| Central Semiconductor | Small Signal Discretes | (516) 435-1110 | www.centralsemi.com |
| Chicago Miniature Lamp | LEDs | (201) 489-8989 | www.sli-lighting.com/cml |
| Data Display Products | LEDs | (800) 421-6815 | www.ddp-leds.com |
| Fuji | Schottky Diodes | (201) 712-0555 | www.fujielectric/co/jp/eng/index-e.html |
| General Semiconductor | Diodes | (516) 847-3000 | www.gensemi.com |
| Motorola* | Discretes | (800) 441-2447 | www.mot-sps.com/products/index.html |
| ON Semiconductor* | Discretes | (408) 749-0510 | www.onsemi.com/home |
| Panasonic | LEDs | (201) 348-5217 | www.panasonic.com/industrial_oem/semiconductors/semiconductor_home.htm |
| Temic | IR Photo Diodes | (408) 970-5700 | www.temic.com |
| Vishay | Zener/Small Signal Diodes | (650) 665-9301 | www.vishay.com |
| Zetex | Small Signal Discretes | (516) 543-7100 | www.zetex.com |

*Discretes formerly manufactured by Motorola are now manufactured by ON Semiconductor. Part numbers have not been changed as of January 2000

### Inductors and Transformers

| Vendor | Product | Phone Number | URL |
|---|---|---|---|
| API Delevan | Inductors | (716) 652-3600 | www.delevan.com |
| BH Electronics | Inductors | (612) 894-9590 | www.bhelectronics.com |
| BI Technologies | Transformers | (714) 447-2656 | www.bitechnologies.com |
| Coilcraft | Inductors | (847) 639-6400 | www.coilcraft.com |
| Coiltronics | Inductors/ Transformers | (561) 241-7876 | www.coiltronics.com |
| Dale | Inductors/ Transformers | (605) 665-1627 | www.vishay.com/fp/fp.html#inductors |
| Gowanda | Inductors | (716) 532-2234 | www.gowanda.com |
| Midcom | Inductors/ Transformers | (800) 643-2661 | www.midcom-inc.com |
| Murata Electronics | Inductors, | (814) 237-1431 | www.murata.com |
| Panasonic | Inductors/ Transformers | (714) 373-7334 | www.panasonic.com/industrial_oem/electronic_components/ electronic_components_inductors_coils_and_transformers.htm |
| Philips | Inductors | (914) 246-2811 | www.acm.components.philips.com |
| Philips | Planar Inductors | (914) 247-2036 | www.acm.components.philips.com |
| Pulse | Inductors | (619) 674-8100 | www.pulseeng.com |
| Sumida | Inductors | (847) 956-0667 | www.japanlink.com/sumida |
| Tokin | Inductors | (408) 432-8020 | www.tokin.com |

### Logic

| Vendor | Product | Phone Number | URL |
|---|---|---|---|
| Fairchild | Logic | (207) 775-4502 | www.fairchildsemi.com |
| Intersil (formerly Harris) | Logic | (800) 442-7747 | www.intersil.com |
| *Motorola | Logic | (800) 441-2447 | www.mot-sps.com/products/index.html |
| *ON Semiconductor | Logic | (408) 749-0510 | www.onsemi.com/home |
| Toshiba | Logic Single Gate Logic | (949) 455-2000/ (714) 455-2000 | www.toshiba.com/taec |

*Logic Devices formerly manufactured by by Motorola are now manufactured by ON Semiconductor; there have been no changes in part numbers as of January 2000

### Resistors

| Vendor | Product | Phone Number | URL |
|---|---|---|---|
| Allen Bradley | Carbon Resistors | (800) 592-4888 | www.ab.com |
| AVX | Chip Resistors | (843) 946-0524 | www.avxcorp.com/products/resistors/chiprstr.htm |
| BI Technologies | Resistors/Resistor Networks | (714) 447-2345 | www.bitechnologies.com |
| Bourns | Potentiometers, SIPs | (801) 750-7253 | www.bourns.com |
| Dale | Sense Resistors | (605) 665-9301 | www.vishayfoil.com or www.vishay.com |
| IRC | Sense Resistors | (361) 992-7900 | www.irctt.com |
| RG Allen | Metal Oxide Resistors | (818) 765-8300 | www.rgaco.com |
| TAD | Chip Resistors | (800) 508-1521 | www.tadcom.com |
| Taiyo Yuden | Chip Resistors | (408) 573-4150 | www.t-yuden.com |
| Thin Film Technology | Thin Film Chip Resistors | (507) 625-8445 | www.thin-film.com |
| Tocos | SMD Potentiometers | (847) 884-6664 | www.tocos.com |

**Transistors**

| Vendor | Product | Phone Number | URL |
|---|---|---|---|
| Central Semiconductor | Small Signal Discretes | (516) 435-1110 | www.centralsemi.com |
| Fairchild | MOSFETs | (408) 822-2126 | www.fairchildsemi.com |
| IR | MOSFETs | (310) 322-3331 | www.irf.com |
| Motorola* | Discretes | (800) 441-2447 | www.mot-sps.com/products/index.html |
| ON Semiconductor* | Discretes | (408) 749-0510 | www.onsemi.com/home |
| Philips | Discretes | (401) 767-4427 | www-us.semiconductors.philips.com |
| Siliconix | MOSFETs | (800) 554-5565 | www.siliconix.com |
| Zetex | Small Signal Discretes | (516) 543-7100 | www.zetex.com |

*Discretes formerly manufactured by Motorola are now manufactured by ON Semiconductor; There are no changes in part numbers as of January 2000.

**Miscellaneous**

| Vendor | Product | Phone Number | URL |
|---|---|---|---|
| Aavid | Heat Sinks | (714) 556-2665 | www.aavid.com |
| Epson | Crystals | (310) 787-6300 | www.eea.epson.com |
| Infineon (formerly Siemens Semiconductor) | Optoelectronics | (108) 257-7910 | www.infineon.com/us/opto/content.htm |
| Magnetics, Inc. | Toroid Cores, etc. | (800) 245-3984 | www.mag-inc.com |
| MF Electronics | Crystal Oscillators | (914) 576-6570 | www.mfelec.com |
| Murata Electronics | RF Devices | (770) 433-5789 | www.murata.com |
| QT Optoelectronics | RF Switches | (408) 720-1440 | www.qtopto.com |
| Raychem | Fuses | (800) 227-4856 | www.raychem.com |
| RF Micro Devices | RF Semiconductors | (336) 664-1233 | www.rfmd.com |
| RTI/Ketema | Surge Suppressors | (714) 630-0081 | www.rtie.rti-corp.com |
| Schurter | Fuses and Holders | (707) 778-6311 | www.schurterinc.com |
| Thermalloy | Heat Sinks | (972) 243-4321 | www.thermalloy.com |
| Toko | RF Products | (847) 699-3430 | www.tokoam.com |

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