Battery Stack Monitor Extends Life of Li-Ion Batteries in Hybrid Electric Vehicles

by Michael Kultgen and Jon Munson

Introduction

The cost of running a car on electricity is equivalent to paying $0.75/gallon for gasoline, and if that electricity comes from carbon neutral sources, car owners are saving both money and the environment (gasoline combustion produces 9kg of CO_2 per US gallon). Advancements in battery technology (see sidebar), especially with Lithium-based chemistries, hold the greatest promise for converting the worldwide fleet of cars to hybrid or fully electric.

Lithium battery packs offer the highest energy density of any current battery technology, but high performance is not guaranteed simply by design. In real world use, a battery management system (BMS) makes a significant difference in the performance and lifetime of Li-Ion batteries—arguably more so than the design of the battery itself. The LTC6802 multicell battery stack monitor is central to any BMS for the

Figure 1. 96-cell battery pack
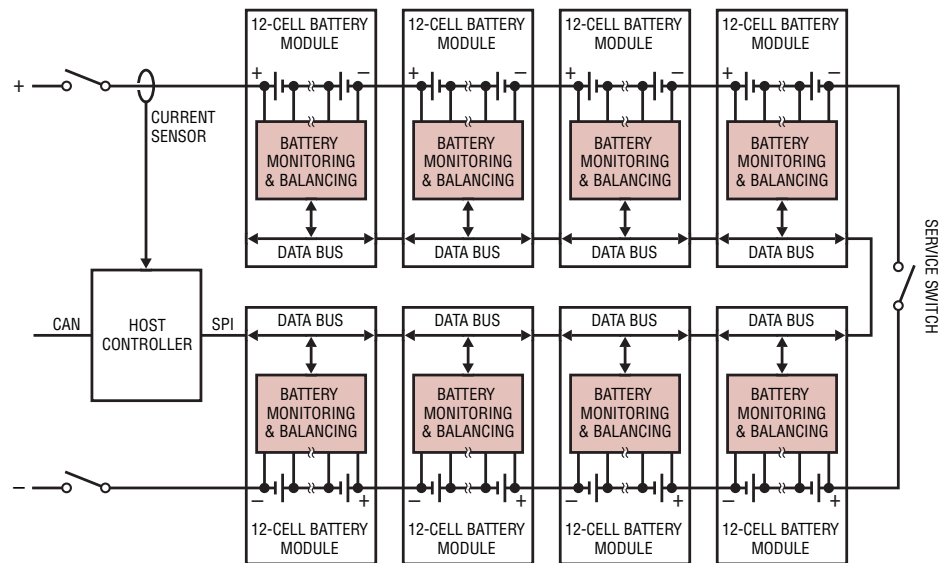
LINEAR TECHNOLOGY

Linear in the News…

EDN Highlights Linear for Innovation Awards

EDN magazine in February chose several Linear Technology products as finalists for their annual Innovation Awards, to be announced later this month. And the nominees are:

Best Contributed Article—"High Voltage, Low-Noise DC/DC Converters" by Jim Williams

You can find the article in its entirety on the *EDN* website at www.edn.com/jimwilliams.

### Battery ICs Category—LTC6802 Battery Stack Monitor

The LTC6802 is a highly integrated multicell battery monitoring IC capable of precisely measuring the voltages of up to 12 series-connected battery cells. Using a novel stacking technique, multiple LTC6802s can be placed in series without optocouplers or isolators. See the cover article of this issue for an overview of this part.

Power ICs Category—LTC3642 50mA Synchronous Step-Down Converter

The LTC3642 uses a unique high voltage synchronous rectification design, capable of continuous input voltages of 45V and offers transient protection up to 60V. Its internal synchronous rectification and its programmable peak current mode control feature enable it to deliver up to 93% efficiency, maximizing battery run time.

Power ICs: Modules—LTM4606 Ultralow EMI, 6A DC/DC µModule Regulator

The LTM4606 DC/DC µModule™ regulator significantly reduces switching regulator noise by attenuating conducted and radiated energy at the source. The µModule device is a complete DC/DC system-in-a-package, including the inductor, controller IC, MOSFETs, input and output capacitors and the compensation circuitry, housed in an enclosed surface-mount plastic package resembling an IC.
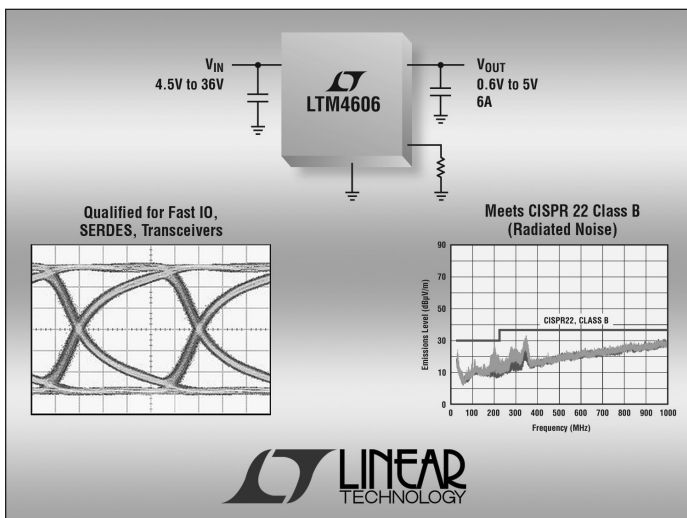
The LTM4606 reduces switching regulator noise at the source.

Linear CEO Comments on Growth Markets

Last month in *EE Times*, Linear Technology CEO Lothar Maier discussed the challenging market conditions and the bright spots on the horizon: "In these times our customers will continue to invest in new products and new product development. Innovation will return growth to the semiconductor market—specifically to analog. Now is the time to get new products out, to be first to market and to have products that target emerging growth markets." He discussed several key markets:

❑ **Automotive.** "Automotive manufacturers are forecasting automotive electronic content to grow 2–3 times over the next few years, so we will continue to provide new products to the automotive area. In addition, every major automotive manufacturer in the world is now working on hybrid vehicles, which will add even more electronic content in cars. We have just introduced an innovative device, the LTC6802, a highly integrated battery stack monitor that significantly eases the design of battery monitoring systems for hybrid/electric vehicles."

❑ **Green Growth Markets.** "Products targeted toward energy conservation or energy harvesting will see growth opportunities and are insulated from the current market conditions. Energy costs and environmental concerns, as well as the need to extend battery life for mobile devices, have led to a focus on power optimization. Our energy-efficient products enable customers to convert power more efficiently, consume less power and extend battery life. Our LED drivers enable a new generation of low power lighting for a range of applications, from cars and medical instruments to laptops and office lighting. Our efficient analog solutions will help drive innovative cleantech markets such as solar and wind power systems."

❑ **Communications Infrastructure.** "Wireless systems continue to produce significant market opportunities for products in wireless and network infrastructure. Our high speed data converters and high frequency products are designed into the next generation of cellular basestations. And our Hot Swap™ and Power over Ethernet products are proliferating in networks."

❑ **Industrial.** "The broad industrial market continues to provide a solid core of business and is somewhat more insulated from market swings. Linear's analog products are used in a broad range of industrial systems, including factory automation, industrial process control, medical, instrumentation and security."

Lothar Maier concluded, "Finally, I believe that Linear's strategy of customer, market and geographic diversity will be a hedge against the current market conditions and will provide the conduit to future growth." *LT*

large battery stacks common in electric vehicles (EVs) and hybrid electric vehicles (HEVs). Its robust design and high accuracy helps guarantee the performance and lifetime of expensive battery packs.

For instance, to meet a 15-year, 5000 charge cycle goal, only a portion (say 40%) of the battery pack's cell-capacity can be used. Of course, using only 40% of the capacity essentially lowers the energy density of the pack. This is the problem: increasing battery lifetime is traded against the need to use as few kg of batteries as possible—the most expensive component in any EV. Only a well-designed BMS can maximize battery performance and lifetime in the face 200A peak charge and discharge currents.

Battery Management System Optimizes Li-Ion Run Time and Lifetime

In any battery stack, the more accurately you know state of charge (SOC) of each cell, the more cell capacity you can use while still maximizing cell life. In a laptop computer, gas gauging comes from monitoring cell voltage and counting coulombs in and out of the stack of four to eight cells. Voltage, current, time and temperature are combined in a robust algorithm to give an indication of the SOC. Unfortunately, it's nearly impossible to count coulombs in a car. The battery drives an electric motor, not a motherboard, so it must handle current spikes of 200A, followed by low level idling. Furthermore, you have from 96

Li-ion Batteries in Electric Vehicles and Hybrids

So why aren't all cars electric? One reason is energy density. Gasoline holds 80 times the energy per kg as Li-ion batteries (Table 1) and refuels in three minutes, essentially allowing indefinite driving. Even a big lithium pack only gives a passenger car about a 100-miles after an 8-hour charging cycle. To drive a passenger car further than 100 miles you still need a gasoline engine, but even so, batteries improve gas mileage in hybrid electric vehicles (HEVs). The peak efficiency of the Otto cycle engine is only 30% at high RPMs and the average efficiency is about 12%. Using batteries to supply torque during acceleration and recover joules during

| Table 1. Energy density comparison ||
Medium	Wh/kg
Diesel Fuel	12,700
Gasoline	12,200
Li-Ion Battery	150
NiMh Battery	100
Lead Acid Battery	25

regenerative braking means the gas engine runs less often and runs at a higher efficiency, effectively doubling the mpg.

In the 1970s the only available high power battery chemistry was lead acid, too heavy to reasonably power anything larger than a golf cart. Then came NiMh batteries, which improved energy density enough to enable the first commercially successful HEVs, like the Toyota Prius and Ford Escape. Li-ion batteries take energy density another step forward, by offering another 50% improvement. The safety of Li-ion was a concern, but new battery technologies like the A123 nanophosphate cell, the EnerDel Spinel-Titanate chemistry, the GS Yuasa EH6 design and others are as safe as NiMh, offer extremely high power (200A peak discharge rates), and last 10 to 15 years with proper charge management. By model year 2012, the majority of hybrid cars and trucks will use lithium battery technology.

Figure 1 shows a shows a block diagram of the battery pack with a BMS, and Figure 2 shows a typical HEV power train. The battery pack building block is a 2.5V to 3.9V, 4Ahr to 40Ahr Li-ion cell. 100 to 200 cells are connected in series to bring the battery pack voltage into the hundreds of volts. This DC power source drives a 30kW to 70kW electric motor. The pack voltage is high so that the average current is low for a given power level. Lower current reduces I^2R power losses, so cables can be smaller, thus reducing weight and cost. The pack should be able to deliver 200A under peak conditions and be quickly rechargeable. In other words, the battery needs to offer high energy density *and* high power density, specifications that can be met by Li-ion batteries. Systems for busses and tractor-trailers use up to four parallel packs of 640V each. LT
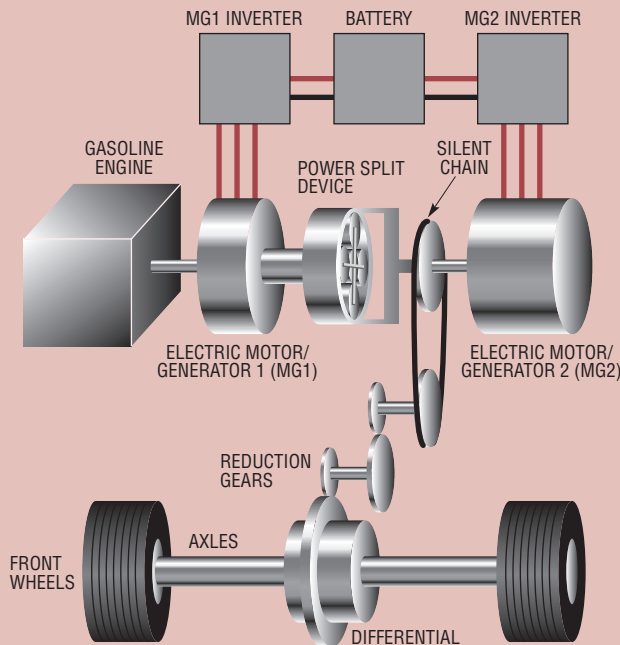

Figure 2. Toyota Prius "split power" hybrid drive train

to 200 cells in series, in groups of 10 or 12. The cells age at different rates, were manufactured from multiple lots, and vary in temperature. Their capacities diverge constantly. Different cells with the same coulomb count can have wildly different charge levels.

That's why the BMS focuses on cell voltage. If you can accurately measure the voltage of every cell, you can know the cell's SOC with reasonable accuracy (Figure 3). The trick is to improve the accuracy of the voltage measurement by taking into account temperature effects on battery ESR and capacity. By constantly measuring each cell's voltage, you keep a running estimation of each cell's charge level. If some cells are overcharged and some under, they can be balanced by bleeding off charge (passive balancing) or redistributing charge (active balancing).

Accurate Monitoring is Key to Raising Battery Performance while Lowering Costs

The LTC6802 (Figure 4) is a precision data acquisition IC optimized for measuring the voltage of every cell in a large string series-connected batteries. In the BMS, the LTC6802 does the heavy lifting analog function, passing digital voltage and temperature measurements to the host processor for SOC computation. The LTC6802's high accuracy, excellent noise rejection, high voltage tolerance, and extensive self-diagnostics make it robust and easy-to-use. The high level of integration means a substantial cost savings for customers when compared to discrete component data acquisition designs.

Increasing measurement accuracy reduces battery cost, as illustrated by the following example. Figure 5 shows the typical performance of the LTC6802, where 0.1% total error from −20°C to 60°C translates to 4mV precision for a 3.7V cell. Suppose that to achieve a 15-year battery lifetime, you are limited to 40% of a cell's capacity per charge cycle, and assume the cell voltage vs charge level of the battery is very flat, e.g., 1.25mV/%SOC. A measurement error of 4mV means the
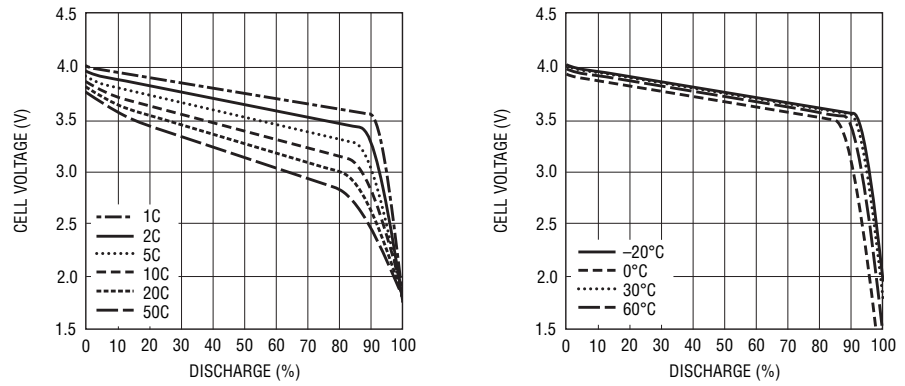
Figure 3. State of charge vs current and temperature for a typical Li-ion cell
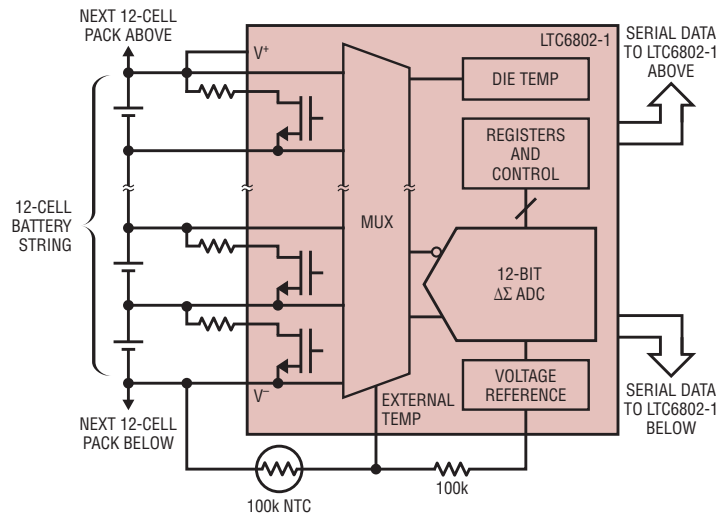
Figure 4. Simplified block diagram of the LTC6802

estimation of SOC is accurate to 3%. The BMS must charge cells to no more than 37% (40% – 3%) of their capacity to guarantee the 15-year lifetime.

Now consider a monitor IC with 10mV error over similar conditions. In this case, the BMS can only use 32% (40% – 10mV • 1%/1.25mV) of the cells' capacity and still guarantee a 15-year life. This seemingly negligible increase in measurement error results in a significant 14% reduction in the usable capacity. That is, a vehicle requires least 14% more batteries, or
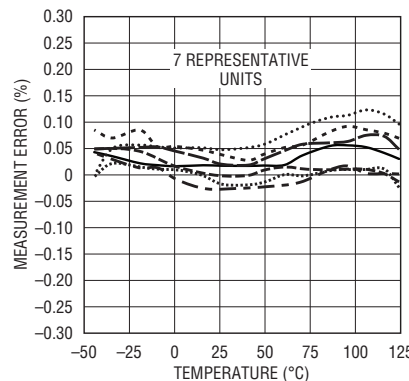
Figure 5. Typical measurement accuracy vs temperature of seven samples
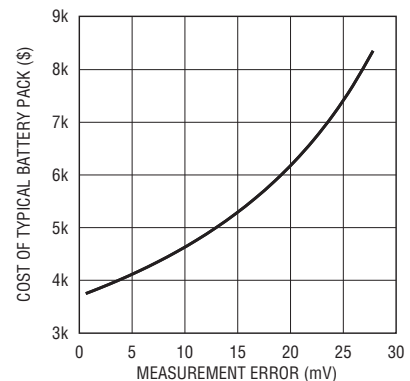
Figure 6. High BMS accuracy is important to keeping battery costs in check, as shown in this cost vs measurement error model.

Figure 7. Improving accuracy with calibration

Figure 8. Inverter noise example

the common mode voltage. In a pack like Figure 2, each LTC6802 (one per module) sees up to 12V common mode transients and 1V differential transients per cell. The transients are at the PWM frequency of 10kHz to 20kHz. The LTC6802 has excellent common mode rejection (Figure 9) to eliminate this error term. The SINC2 filter inherent in the delta-sigma ADC attenuates the differential noise by 40dB (Figure 10). External filtering or measurement averaging can be used to further reduce the differential noise.

Diagnostic Features of the LTC6802 Improve Robustness

Automotive systems require that "no bad cell reading be misinterpreted as a good cell reading." Two of the more common faults that can cause false readings are open circuits and IC failures. If there is an open circuit in the wiring harness and if there is a fi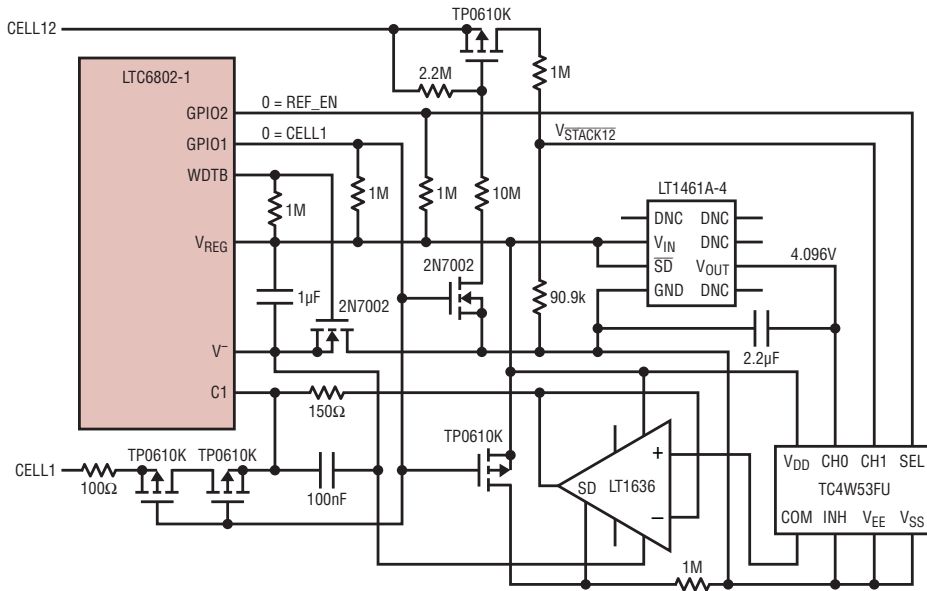lter capacitor on the ADC input (Figure 11), the capacitor will tend to hold the input voltage at a point midway between the adjacent cells. Some type of open wire detection or cell resistance measuring function is necessary. The LTC6802 includes 100µA current sources to load the cell inputs. The current source will cause large changes in cell readings if there is an open circuit in the harness.

at least 14% more weight, cost and electronics to travel an equivalent distance as a vehicle with the more accurate BMS. Batteries are expensive. It takes about $4000 worth of batteries to drive 50 miles, so the increased measurement error means $560 in additional cells. This is why BMS designers scrutinize every 0.01% of measurement error. Figure 6 shows a simple battery cost model as a function of BMS accuracy.

Adding a low drift reference, an initial factory calibration, and a periodic self-calibration routine can improve the measurement accuracy of the LTC6802 to 0.03%. For example, in Figure 7 the LT1461A-4 is periodically applied to channel C1. The temperature stable LT1461 measurement is used to correct temperature drift in the LTC6802. The initial error of the LTC6802 and LT1461A is corrected by measuring and storing a calibration reference after board assembly.

Inverter noise can seriously interfere with cell voltage measurements. When a 100-cell stack is loaded by an electric motor it can have a 370V open circuit voltage and up to 100V switching transients (Figure 8). Spreading the transient equally over the 100 cells means the top cell has 370V of common mode voltage, 100V of common mode transients, 1V of differential transients and an average DC value

of 3.7V, which we need to measure to 4mV. Breaking the battery stack into 12-cell modules further reduces

The LTC6802's 0.1% total measurement error from –20°C to 60°C translates to 4mV precision for a 3.7V cell. Batteries are expensive. It takes about $4000 worth of batteries to drive 50 miles, so just increasing measurement error to 10mV means $560 in additional cells. This is why BMS designers scrutinize every 0.01% of measurement error.
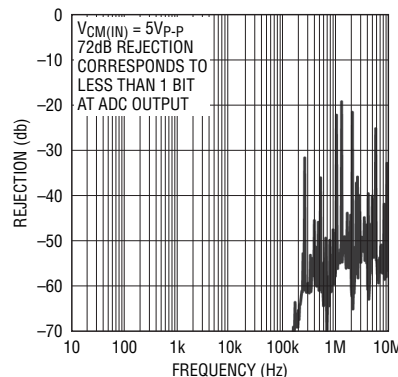

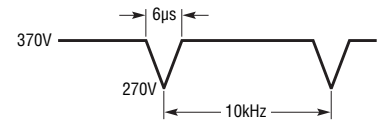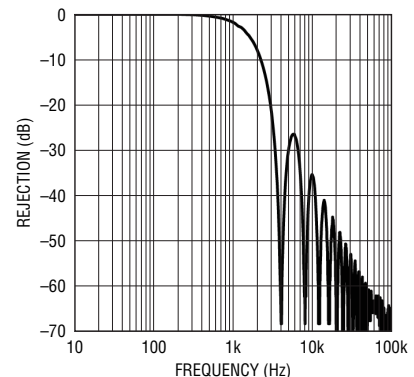**Figure 9. Cell measurement common mode rejection**

Figure 10. Cell measurement filtering

The host controller must be able to run diagnostics on all the modules during normal operation to detect IC failures. If these periodic self-tests fail, then the control algorithm is suspect and the battery pack must be taken off line. The LTC6802 includes a built-in self-test in combination with external support circuits to allow the BMS to completely verify the data acquisition system. See the LTC6802 data sheets for more details.

The LTC6802 Isolates Communications from Swings in Ground Potential

Breaking a ~100 cell pack into modules makes it easier to integrate the analog circuits. Unfortunately, we are left with the task of getting the data from measurement IC to the host controller when the difference in ground potential exceeds 300V. The LTC6802 can solve this problem in a number of ways, depending on the specific needs of the application.

The LTC6802 comes in two flavors, depending on the desired data communication scheme. The LTC6802-1 offers a built-in stackable serial peripheral interface (SPI) solution designed for easy daisy chaining of the interface. The addressable LTC6802-2 is designed for bus-oriented (parallel) SPI communication, but it can also be used in a parallel-addressable, daisy chained interface for a robust and relatively inexpensive solution. All three schemes are described below.

Figure 11. Current sources help detect open circuits.

SPI Bus Communication with the Addressable LTC6802-2 and Digital Isolators

The most straightforward approach is to use a bus communications scheme, with a digital isolator between each module and the host controller. Figure 12 shows a 96-cell pack using eight multicell modules monitored by the LTC6802. The physical layer is a 4-wire SPI bus. An addressing scheme allows the control module to talk to the battery modules separately or in unison. The data buses on the modules are isolated from one another. This is a robust scheme, but it has one major drawback: digital isolators are expensive and require an isolated power supply so that the battery cells don't have to provide the power to the cell side of the isolator.

Daisy Chaining the SPI Interface with the LTC6802-1

The LTC6802-1 provides fixed 1mA signaling between stacked devices to enable easy implementation a daisy chained SPI interface with inexpensive support circuitry. The digital isolators are eliminated as shown in Figure 13. The interface exploits the fact that the positive supply of module "N" is the same voltage as the ground of module "N+1." A 1mA current is used to transmit data between adjacent modules. Like the analog circuits, the modular approach means the data bus has to deal with a fraction of the total pack voltage.
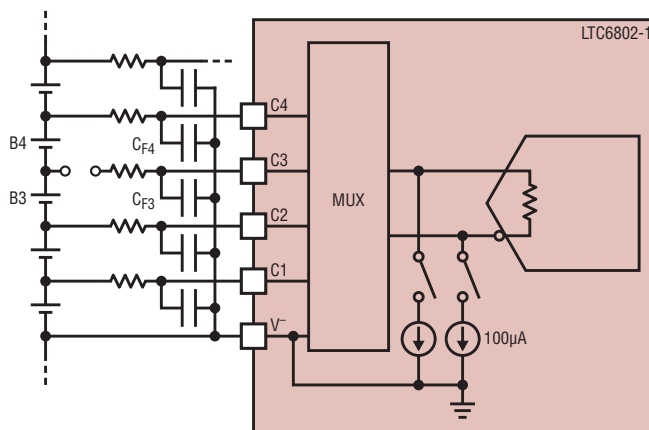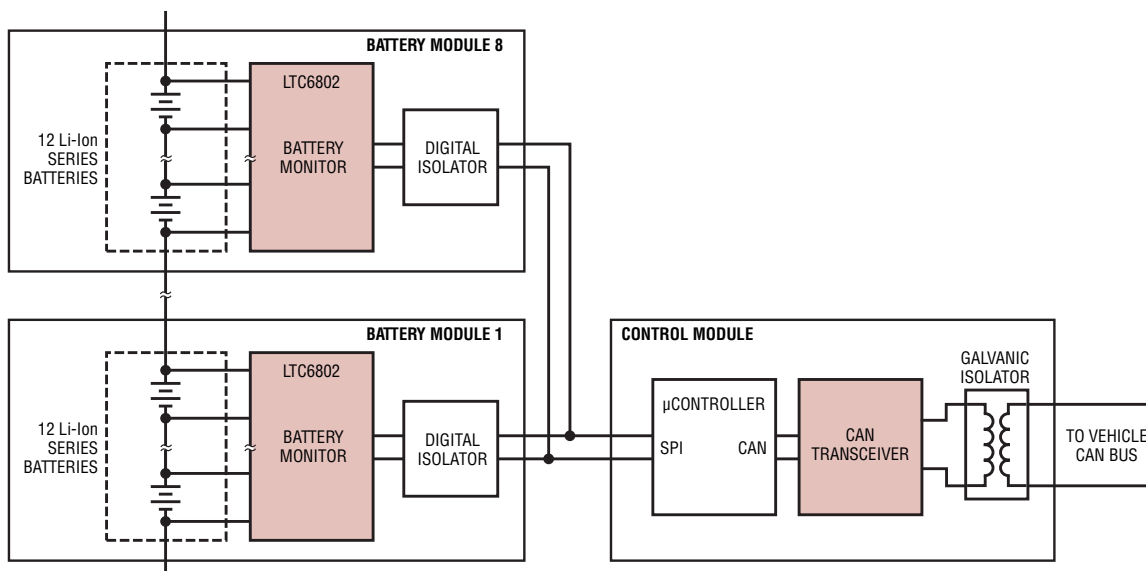

Figure 12. Using digital isolators to communicate to the LTC6802
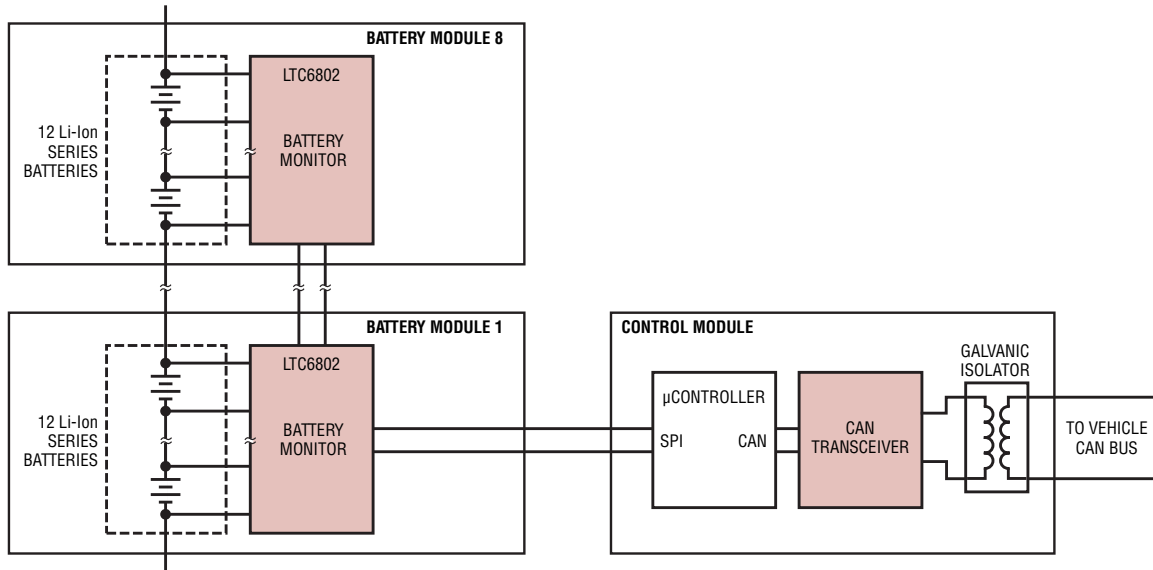
Figure 13. Using the daisy chained SPI to eliminate digital isolators

The disadvantage of any pure daisy chain is that a fault in one module results in a loss of communications with all the modules above it in the stack. Also, since there is no galvanic isolation between modules, the interface needs to handle large voltages that occur during fault conditions. For example if the "service switch" in Figure 1 is open and there is a load on the pack then the data bus connection between modules 4 and 5 will see a reverse voltage equal to the total pack voltage (–300V to –400V). The LTC6802 interface relies on external discrete diodes to block the reverse voltage during fault conditions.

The Best of Both Worlds: Daisy Chained, Addressable Interface with the LTC6802-2

With inexpensive external circuitry, the LTC6802-2 can also be used in a stacked SPI configuration like the LTC6802-1, but with more flexibility in the operating parameters.

The SPI port of the LTC6802-2 is a 4-wire connection: chip select in (CSBI), clock in (SCKI), data in (SDI), and data out (SDO). The inputs are conventional CMOS levels and the output is an open-drain NMOS. The SDO pin must have an external pull-up current or added resistance suitable for the intended data rate. The IC also provides a versatile always-on 5V output (V_{REG}), which can produce up to

4mA to energize low power auxiliary circuitry.

Figure 14 shows a complete stacked LTC6802-2 SPI interface for a 36-cell application. The stack can be increased in size by replicating the

circuit of the middle IC. In Figure 14, the V_{REG} and V^- pins of each stacked IC are used to bias common-base connected transistors to form a signal translation current for each SPI data line. Each LTC6802 can monitor up

Figure 14. Inexpensive SPI daisy chain for parallel-addressed LTC6802-2

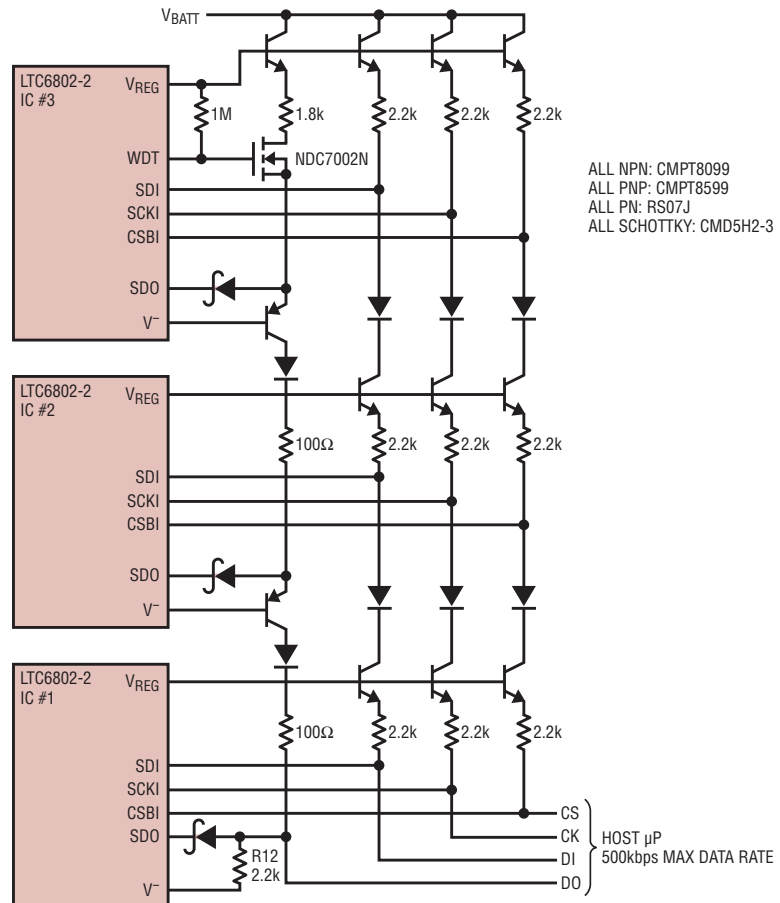to 12 cell-potentials, which could sum to 60V in certain instances, so the transistors selected for the SPI translation need to have a V_{CBO} over 60V, but they should be the highest available f_T to prevent undue slowing of the logic signals. A suitable NPN candidate is the CMPT8099, while the CMPT8599 is its PNP complement, both from Central Semiconductor. These are fast 80V devices (f_T > 150MHz).

Sending Signals Upwards
At the bottom-of-stack IC, the logic signal is furnished by the host connection, be it a microprocessor or an SPI isolation device. By simply pulling down the emitter leg of an NPN having a V_{REG} base potential through a known resistance, a specific current is formed for a logic low input signal. In the case of the component values shown, the current is about 2mA for a logic low, and conversely, the transistor is essentially turned off with a logic high (~0mA for 5V logic).

Since the collector current is nearly identical to the emitter current, the same current pulls on the next higher cascode circuit. Since that next circuit is the same as the first, the voltage on the upper emitter resistor reproduces that of the bottom circuit logic level for the upper IC. This continues up the daisy chain, eventually terminating at the top potential of the battery stack. Since each IC is provided the same input waveforms, this structure forms a parallel bus from a logical perspective, even though each IC is operating at a different potential in the stack.

The NPN transistors at the top IC source the logic current directly from the battery stack. Only small base currents flow from any V_{REG} output. The 600V collector diodes provide reverse-voltage protection in the event a battery group interconnection is lost, perhaps during service (these are not required for functionality and could be omitted in some situations).

Bringing Data Down the Stack
The SDO cascode chain is similar in concept, except the current starts at the top of the stack and flows downward. At the top IC, a PNP transistor with its base connected to the local V^- pin has current injected into its emitter by a pullup resistor. Here again, the collector current is essentially identical to the emitter current, and so current flows downward through each successive PNP and terminates into a resistor at the bottom of stack. In this case, the presence of the current in the termination resistor, about 2mA for the component values shown, forms a logic high potential for the host interface.

A Schottky diode is connected from each SDO pin to the emitter of a local PNP thereby allowing any LTC6802 on the stack to divert the pullup current to the local V^- when outputting a logic low. This effectively turns off the emitter current to the local PNP transistor and all points lower in the stack, so the voltage on the bottom termination resistor then drops to a logic low level. Since each SDO pin can force a low level, this forms a wire-OR function that is equivalent to paralleled connections as far as the host interface is concerned. Note the bottom of stack SDO diode is connected slightly differently; it forms a direct wire-OR at the host interface. Since the LTC6802-2 is designed to use addressed readback commands, this line is properly multiplexed and no inter-IC contention occurs.

To eliminate the pull-up current during standby, a general purpose N-channel MOSFET is used to interrupt the top PNP emitter current when the watchdog timer bit goes low. The watchdog timeout will release when clock activity is present, so the SDO line will reactivate as needed. Here again, an NPN is used at the top of stack to ensure the pull-up current comes directly from the battery, rather than loading V_{REG}.

Collector diodes are added here as well to provide a high reverse voltage protection capability, plus some added series resistance is included to protect the lower transistor emitters from transient energy (once again, these protection parts don't add any other functionality to the data transmission and could be omitted in some circumstances).

External SPI Advantages
Since the LTC6802-2 uses a parallel addressable SPI protocol, the conventional method of connecting multiple devices in a stack is to provide isolation for each SPI connection, then parallel the signals on the host side. Isolators are relatively expensive and often need extra power circuitry, thus adding significantly to the total solution cost. The transistor circuitry shown here is quite inexpensive and offers the option to make certain design tradeoffs as well. With the propagation delays involved and desire to keep power fairly low, this circuit as shown still communicates at over 500kbps. Lower SPI currents could be chosen in applications that don't demand the high data rate by simply raising the resistance values accordingly.

The main feature of the transistorized SPI bus is the wide compliance range that is afforded by the unconstrained collector-base operating range of the transistors. In normal operation the V_{CB} ranges from just less than the cells connected to the LTC6802, to some five volts below that, depending on the logic level transmitted. This becomes important since voltage fluctuations on the battery, due to load dynamics or switching transients, affect the V_{CB} of the transistors even though the V^+ and ADC cell inputs may be filtered. Some vehicle manufacturers are requiring that a BMS tolerate 1V steps with 200ns rise/fall time per cell in the stack, so this is a 12V waveform edge as seen by the transistors in a typical application. With the low collector capacitance and 2mA logic level of the transistor chain, SPI transmissions remain error free with even this high level of noise.

Conclusion
EVs and HEVs are here to stay. Inherently safe lithium batteries, which combine energy density, power density, and cycle life, will continue to evolve to improve the performance of these vehicles. Battery management systems using the LTC6802 extract the most driving distance and lifetime from the battery pack while lowering system cost. *LT*

DC/DC Converter, Capacitor Charger Takes Inputs from 4.75V to 400V

by Robert Milliken and Peter Liu

Introduction

High voltage power supplies and capacitor chargers are readily found in a number of applications, including professional photoflashes, security control systems, pulsed radar systems, satellite communication systems, and explosive detonators. The LT3751 makes it possible for a designer to meet the demanding requirements of these applications, including high reliability, relatively low cost, safe operation, minimal board space and high performance.

The LT3751 is a general purpose flyback controller that can be used as either a voltage regulator or as a capacitor charger. The LT3751 operates in boundary-mode, between continuous conduction mode and discontinuous conduction mode. Boundary-mode operation allows for a relatively small transformer and an overall reduced PCB footprint. Boundary-mode also reduces large signal stability issues that could arise from using voltage-mode or PWM techniques. Regulation is achieved with a new dual, overlapping modulation technique using both
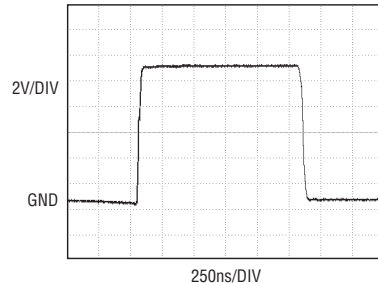

Figure 1. Gate driver waveform in a typical application

peak primary current modulation and duty-cycle modulation, drastically reducing audible transformer noise.

The LT3751 features many safety and reliability functions, including two sets of undervoltage lockouts (UVLO), two sets of overvoltage lockouts (OVLO), no-load operation, over-temperature lockout (OTLO), internal Zener clamps on all high voltage pins, and a selectable 5.6V or 10.5V internal gate driver voltage clamp (no external components needed). The LT3751 also adds a start-up/short-circuit protection circuit to protect against transformer or external FET damage. When used as a regulator, the LT3751's feedback loop is internally compensated to ensure stability. The LT3751 is available in two packages, either a 20-pin exposed pad QFN or a 20-lead exposed pad TSSOP.

New Gate Driver with Internal Clamp Requires No External Components

There are four main concerns when using a gate driver: output current drive capability, peak output voltage, power consumption and propagation delay. The LT3751 is equipped with a 1.5A push-pull main driver, enough to drive +80nC gates. An auxiliary 0.5A PMOS pull-up only driver is also integrated into the LT3751 and is used in parallel with the main driver for V_{CC} voltages of 8V and below. This PMOS driver allows for rail-to-rail operation. Above 8V, the PMOS driver must be deactivated by tying its drain to V_{CC}.

Most discrete FETs have a V_{GS} limit of 20V. Driving the FET higher than 20V could cause a short in the internal gate oxide, causing permanent
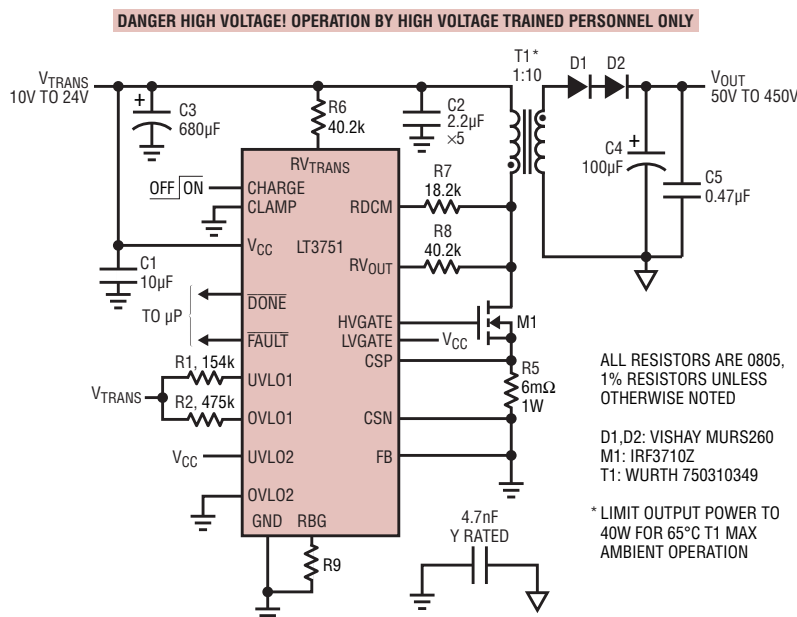

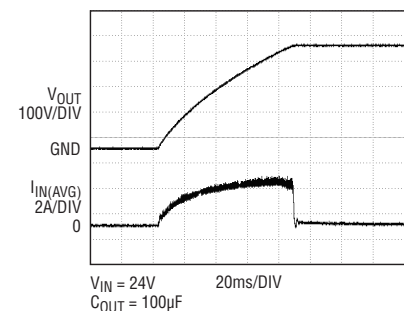**Figure 2. Isolated high voltage capacitor charger from 10V to 24V input**

Figure 3. Isolated high voltage capacitor charger charging waveform

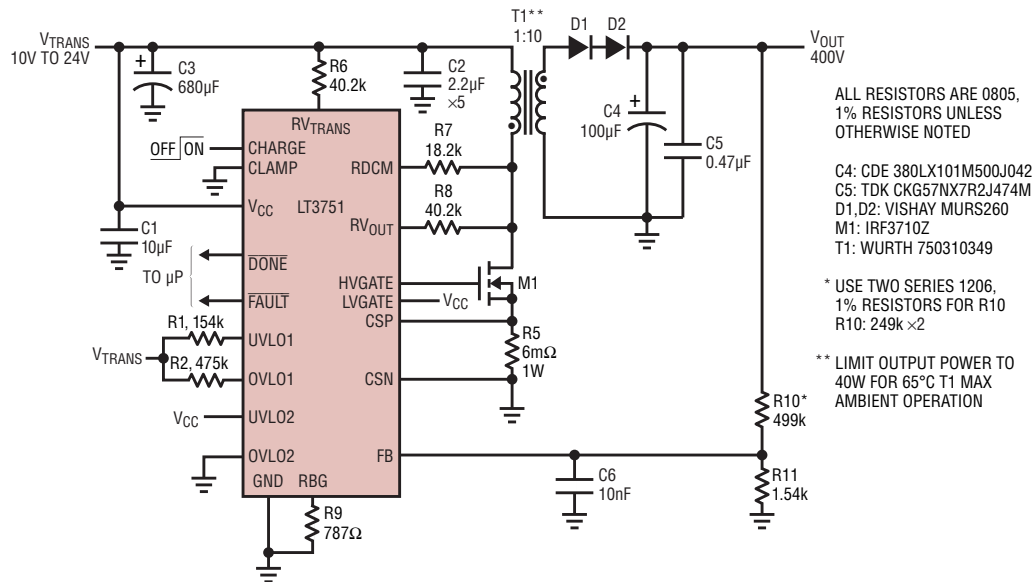DANGER HIGH VOLTAGE! OPERATION BY HIGH VOLTAGE TRAINED PERSONNEL ONLY

Figure 4. A 10V to 24V input, 400V regulated power supply

damage. To alleviate this issue, the LT3751 has an internal, selectable 5.6V or 10.5V gate driver clamp. No external components are needed, not even a capacitor. Simply tie the CLAMP pin to ground for 10.5V operation or tie to V_{CC} for 5.6V operation. Figure 1 shows the gate driver clamping at 10.5V with a V_{CC} voltage of 24V.

Not only does the internal clamp protect the FET from damage, it also reduces the amount of energy injected into the gate. This increases overall efficiency and reduces power consumption in the gate driver circuit. The gate driver overshoot is very minimal, as seen in Figure 1. Placing the external FET closer to the LT3751 HVGATE pin reduces overshoot.

High Voltage, Isolated Capacitor Charger from 10V to 24V Input

The LT3751 can be configured as a fully isolated stand-alone capacitor charger using a new differential discontinuous-conduction-mode (DCM) comparator—used to sense the boundary-mode condition—and a new differential output voltage (V_{OUT}) comparator. The differential operation of the DCM comparator and V_{OUT} comparator allow the LT3751 to accurately operate from high voltage input supplies of greater than 400V. Likewise, the LT3751's DCM comparator and V_{OUT} comparator can work with input supplies down to 4.75V. This accommodates an unmatched range of power sources.

Figure 2 shows a high voltage capacitor charger driven from an input supply ranging from 10V to 24V. Only five resistors are needed to operate the LT3751 as a capacitor charger. The output voltage trip point can be continuously adjusted from 50V to 450V by adjusting R9 given by:

$$R_9 = \left(\frac{0.98 \bullet N}{V_{OUT(TRIP)} + V_{DIODE}} \right) \bullet R_8$$

The LT3751 stops charging the output capacitor once the programmed output voltage trip point ($V_{OUT(TRIP)}$) is reached. The charge cycle is repeated by toggling the CHARGE pin. The maximum charge/discharge rate in
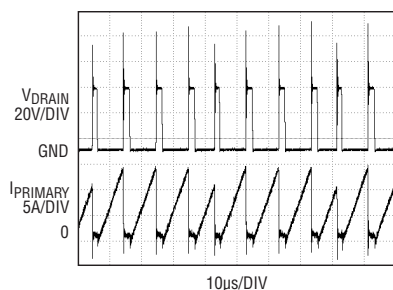
a. Switching waveform for I_{OUT} = 100mA
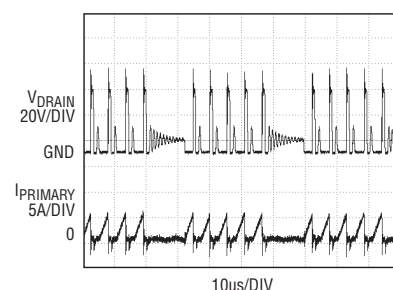
b. Switching waveform for I_{OUT} = 10mA
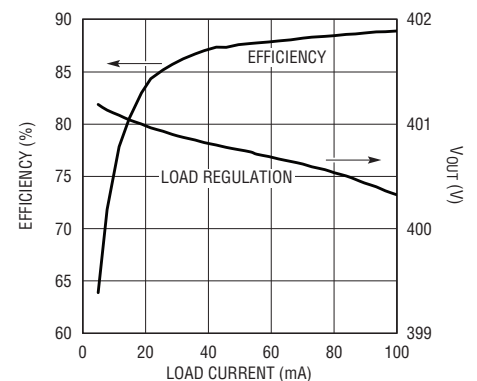
c. Efficiency and load regulation

Figure 5. High voltage regulator performance

the output capacitor is limited by the temperature rise in the transformer. Limiting the transformer surface temperature in Figure 2 to 65°C with no air flow requires the average output power to be ≤40W given by:

$$P_{AVG} = \frac{1}{2}C_{OUT} \bullet FREQUENCY \bullet$$

$$\left(2V_{OUT(TRIP)} \bullet V_{RIPPLE} - V_{RIPPLE}^2\right)$$

$$\leq 40W$$

where $V_{OUT(TRIP)}$ is the output trip voltage, V_{RIPPLE} is the ripple voltage on the output node, and frequency is the charge/discharge frequency. Two techniques are used to increase the available output power: increase the airflow across the transformer, or increase the size of the transformer itself. Figure 3 shows the charging waveform and average input current for a 100µF output capacitor charged to 400V in less than 100ms (R_9 = 976Ω).

For output voltages higher than 450V, the transformer in Figure 2 must be replaced with one having higher primary inductance and a higher turns ratio. Consult the LT3751 data
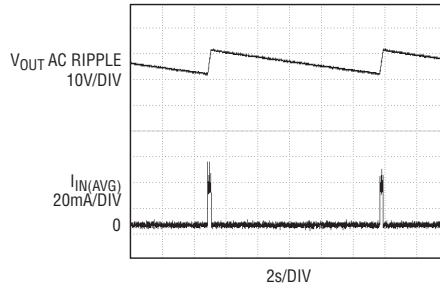

Figure 6. The LT3751 protecting the output during a no-load condition

sheet for proper transformer design procedures.

High Voltage Regulated Power Supply from 10V to 24V Input

The LT3751 can also be used to convert a low voltage supply to a much higher voltage. Placing a resistor divider from the output node to the FB pin and ground causes the LT3751 to operate as a voltage regulator. Figure 4 shows a 400V regulated power supply operating from an input supply range of 10V to 24V.

The LT3751 uses a regulation control scheme that drastically reduces audible noise in the transformer and the input and output ceramic bulk

capacitors. This is achieved by using an internal 26kHz clock to synchronize the primary winding switch cycles. Within the clock period, the LT3751 modulates both the peak primary current and the number of switching cycles. Figures 5a and 5b show heavy-load and light-load waveforms, respectively, while Figure 5c shows efficiency over most of the operating range for the application in Figure 4.

The clock forces at least one switch cycle every period which would overcharge the output capacitor during a no-load condition. The LT3751 handles no-load conditions and protects against over-charging the output node. Figure 6 shows the LT3751 protecting during a no-load condition.

Resistors can be added to RV_{OUT} and RBG to add a second layer of protection, or they can be omitted to reduce component count by tying RV_{OUT} and RBG to ground. The trip level for the V_{OUT} comparator is typically set 20% higher than the nominal regulation voltage. If the resistor divider were to fail, the V_{OUT} comparator would disable switching when the output climbed to 20% above nominal.

DANGER HIGH VOLTAGE! OPERATION BY HIGH VOLTAGE TRAINED PERSONNEL ONLY
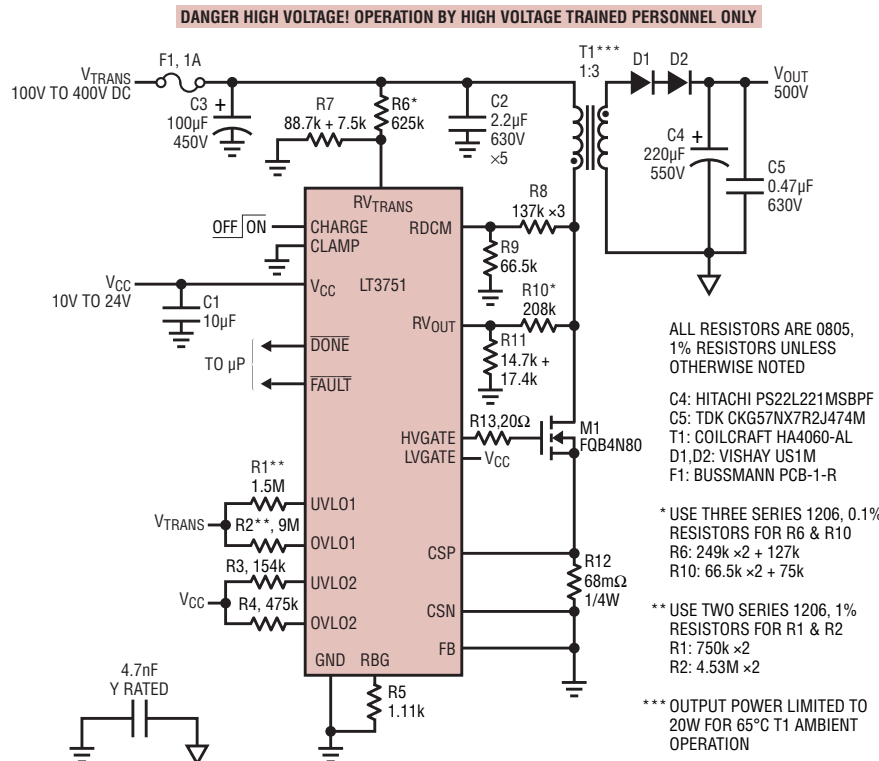
Figure 7. A 100V to 400V input, 500V output, isolated capacitor charger
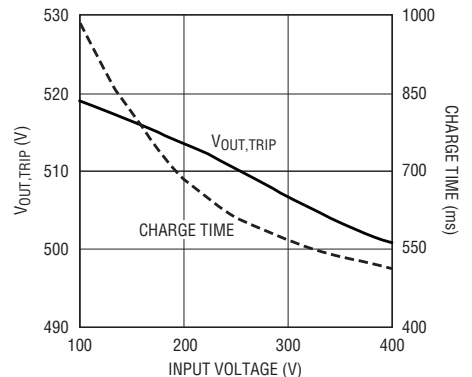
Figure 8. Isolated capacitor charger $V_{OUT(TRIP)}$ and charge time with respect to input voltage

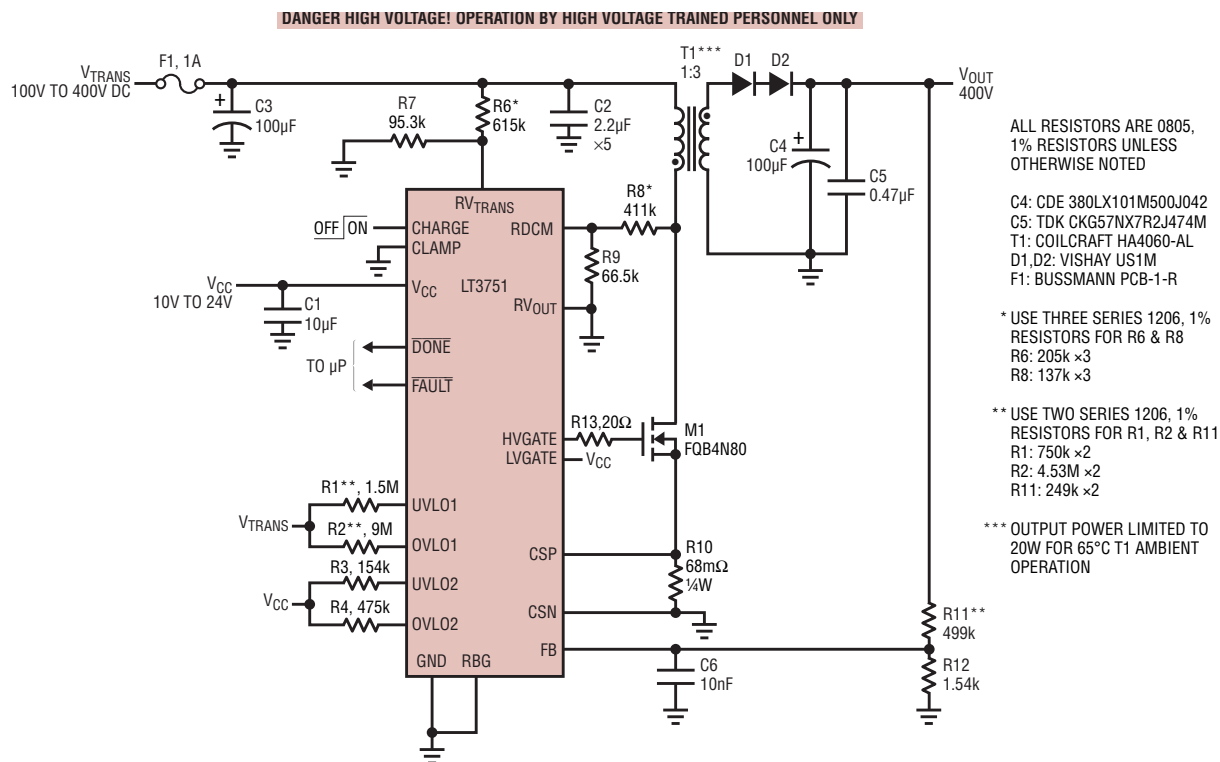**DANGER HIGH VOLTAGE! OPERATION BY HIGH VOLTAGE TRAINED PERSONNEL ONLY**

Figure 9. A 100V to 400V input, 400V output, capacitor charger and voltage regulator

Note that the FB pin of the LT3751 can also be used for a capacitor charger. The LT3751 operates as a capacitor charger until the FB pin reaches 1.225V, after which the LT3751 operates as a voltage regulator. This keeps the capacitor topped-off until the application needs to use its energy. The output resistor divider forms a leakage path from the output capacitor to ground. When the output voltage droops, the LT3751 feedback circuit will keep the capacitor topped-off with small, low current bursts of charge as shown in Figure 6.

High Input Supply Voltage, Isolated Capacitor Charger

As mentioned above, the LT3751 differential DCM and V_{OUT} comparators allow the part to accurately work from very high input supply voltages. An offline capacitor charger, shown in Figure 7, can operate with DC input voltages from 100V to 400V. The transformer provides galvanic isolation from the input supply to output node—no additional magnetics required.

Input voltages greater than 80V require the use of resistor dividers on the DCM and V_{OUT} comparators (charger mode only). The accuracy of the V_{OUT} trip threshold is heightened by increasing current I_Q through R_{10} and R_{11}; however, the ratio of R_6/R_7 should closely match R_{10}/R_{11} with tolerances approaching 0.1%. A trick is to use resistor arrays to yield the desired ratio. Achieving 0.1% ratio accuracy is not difficult and can reduce the overall cost compared to using individual 0.1% surface mount resistors. Note that the absolute value of the individual resistors is not critical, only the ratio of R_6/R_7 and R_{10}/R_{11}. The DCM comparator is less critical and can tolerate resistance variations greater than 1%.

The 100V to 400VDC input capacitor charger has an overall $V_{OUT(TRIP)}$ accuracy of better than 6% over the entire operating range using 0.1% resistor dividers. Figure 8 shows a typical performance for $V_{OUT(TRIP)}$ and charge time for the circuit in Figure 7.
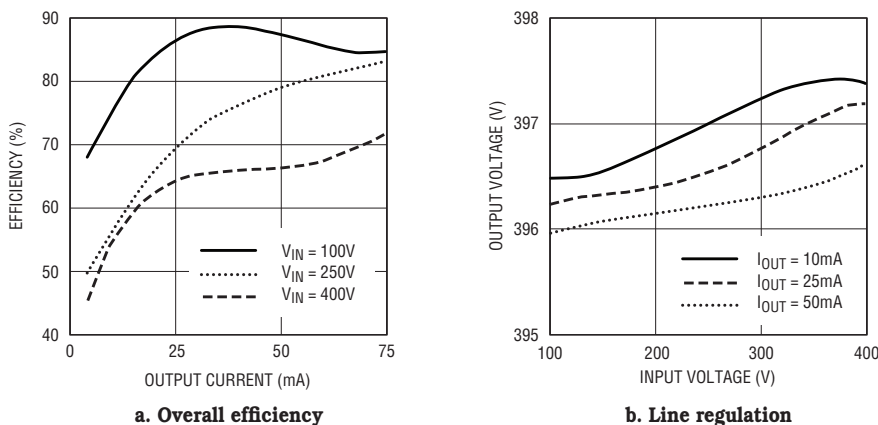

a. Overall efficiency

b. Line regulation

Figure 10. High voltage input and output regulator performance

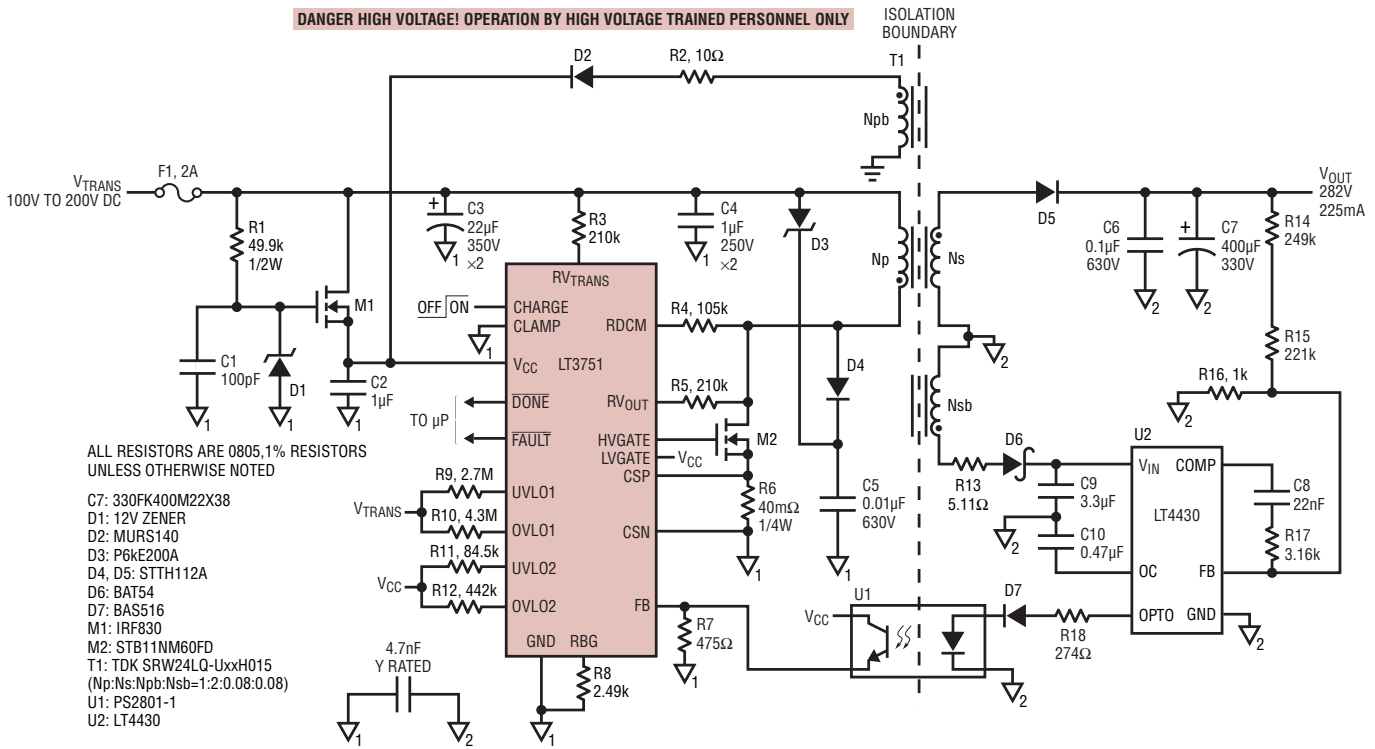**DANGER HIGH VOLTAGE! OPERATION BY HIGH VOLTAGE TRAINED PERSONNEL ONLY**

Figure 11. Fully isolated, high output voltage regulator

High Input Supply Voltage, Non-Isolated Capacitor Charger/Regulator

The FB pin of the LT3751 can also be configured for charging a capacitor from a high input supply voltage. Simply tie a resistor divider from the output node to the FB pin. The resistor dividers on the R_{VTRANS} and R_{DCM} pins can tolerate 5% resistors, and all the $R_{V(OUT)}$ and R_{BG} pin resistors are removed. This lowers the number and the tolerance of required components, reducing board real estate and overall design costs. With the output voltage resistor divider, the circuit in Figure 9 is also a fully functional, high-efficiency voltage regulator with load

and line regulation better than 1%. Efficiency and line regulation for the circuit in Figure 9 are shown in Figure 10a and Figure 10b, respectively.

Alternatively, a resistor can be tied from V_{OUT} to the OVLO1 pin or OVLO2 pin. This mimics the V_{OUT} comparator, stopping charging once the target voltage is reached. The FB pin is tied to ground. The CHARGE pin must be toggled to initiate another charge sequence, thus the LT3751 operates as a capacitor charger only. Resistor R_{12} is omitted from Figure 9 and resistor R_{11} is tied from V_{OUT} directly to OVLO1 or OVLO2. R_{11} is calculated using the following equation:

$$R_{11} = \frac{V_{OUT(TRIP)} - 1.225}{50\mu A}$$

Note that OVLO1 or OVLO2 will cause the $\overline{FAULT}$ pin to indicate a fault when the target output voltage, $V_{OUT(TRIP)}$, is reached.

High Voltage Input/Output Regulator with Isolation

Using a resistor divider from the output node to the FB pin allows regulation but does not provide galvanic isolation. Two auxiliary windings are added to the transformer in circuit shown in Figure 11 to drive the FB pin, the

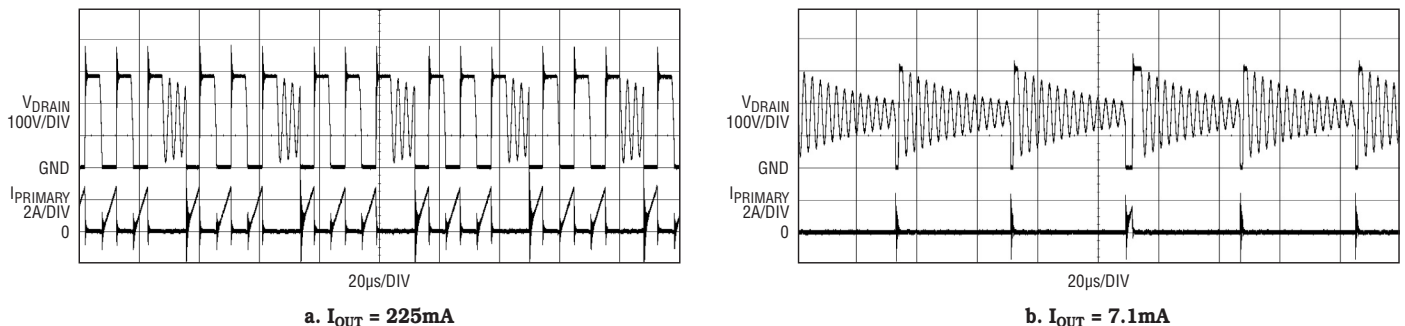a. I_{OUT} = 225mA

b. I_{OUT} = 7.1mA

Figure 12. Switching waveforms

How to Choose a Voltage Reference

by Brendan Whelan

Why Voltage References?

It is an analog world. All electronic devices must in some way interact with the "real" world, whether they are in an automobile, microwave oven or cell phone. To do that, electronics must be able to map real world measurements (speed, pressure, length, temperature) to a measurable quantity in the electronics world (voltage). Of course, to measure voltage, you need a standard to measure against. That standard is a voltage reference. The question for any system designer is not whether he needs a voltage reference, but rather, which one?

A voltage reference is simply that—a circuit or circuit element that provides a known potential for as long as the circuit requires it. This may be minutes, hours or years. If a product requires information about the world, such as battery voltage or current, power consumption, signal size or characteristics, or fault identification, then the signal in question must be compared to a standard. Each comparator, ADC, DAC, or detection circuit must have a voltage reference in order to do its job (Figure 1). By comparing the signal of interest to a known value, any signal may be quantified accurately.
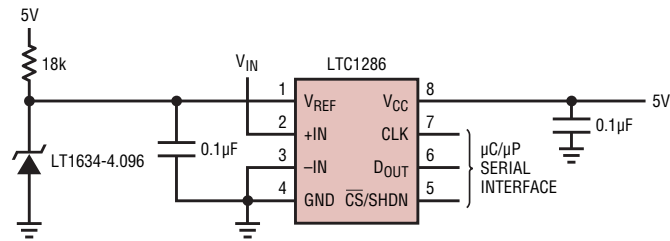

Figure 1. Typical use of a voltage reference for an ADC

Reference Specifications

Voltage references come in many forms and offer different features, but in the end, accuracy and stability are a voltage reference's most important features, as the main purpose of the reference is to provide a *known* output voltage. Variation from this known value is an error. Voltage reference specifications usually predict the uncertainty of the reference under

	Temperature Coefficient	Initial Accuracy	I_S	Architecture	V_{OUT}	Voltage Noise*	Long-Term Drift	Package
LT1031	5ppm/°C	0.05%	1.2mA	Buried Zener	10V	0.6ppm	15ppm/kHr	H
LT1019	5ppm/°C	0.05%	650µA	Bandgap	2.5V, 4.5V, 5V, 10V	2.5ppm		SO-8, PDIP
LT1027	5ppm/°C	0.05%	2.2mA	Buried Zener	5V	0.6ppm	20ppm/ month	SO-8, PDIP
LT1021	5ppm/°C	0.05%	800µA	Buried Zener	5V, 7V, 10V	0.6ppm	15ppm/kHr	SO-8, PDIP, H
LTC6652	5ppm/°C	0.05%	350µA	Bandgap	1.25V, 2.048V, 2.5V, 3V, 3.3V, 4.096V, 5V	2.1ppm	60ppm/√kHr	MSOP
LT1236	5ppm/°C	0.05%	800µA	Buried Zener	5V, 10V	0.6ppm	20ppm/kHr	SO-8, PDIP
LT1461	3ppm/°C	0.04%	35µA	Bandgap	2.5V, 3V, 3.3V, 4.096V, 5V	8ppm	60ppm/√kHr	SO-8
LT1009	15ppm/°C	0.2%	1.2mA	Bandgap	2.5V		20ppm/kHr	MSOP-8, SO-8, Z
LT1389	20ppm/°C	0.05%	700nA	Bandgap	1.25V, 2.5V, 4.096V, 5V	20ppm		SO-8
LT1634	10ppm/°C	0.05%	7µA	Bandgap	1.25V, 2.5V, 4.096V, 5V	6ppm		SO-8, MSOP-8, Z
LT1029	20ppm/°C	0.20%	700µA	Bandgap	5V		20ppm/kHr	Z
LM399	1ppm/°C	2%	15mA	Buried Zener	7V	1ppm	8ppm/√kHr	H
LTZ1000	0.05ppm/°C	4%		Buried Zener	7.2V	0.17ppm	2µV/√kHr	H

Table 1. Specifications for high performance voltage references

*0.1Hz–10Hz, Peak-to-Peak

certain conditions using the following definitions.

Initial Accuracy

The variance of output voltage as measured at a given temperature, usually 25°C. While the initial output voltage may vary from unit to unit, if it is constant for a given unit, then it can be easily calibrated.
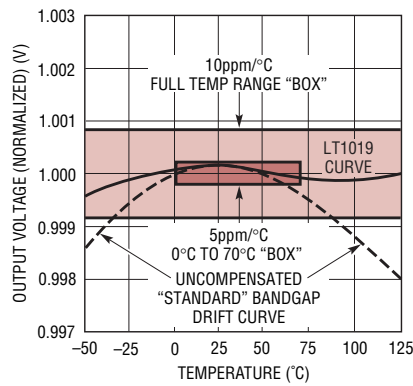
Temperature Drift

This specification is the most widely used to evaluate voltage reference performance, as it shows the change in output voltage over temperature. Temperature drift is caused by imperfections and nonlinearities in the circuit elements, and is often nonlinear as a result.

For many parts, the temperature drift, TC, specified in ppm/°C, is the dominant error source. For parts with consistent drift, calibration is possible. A common misconception regarding temperature drift is that it is linear. This leads to assumptions such as "the part will drift a lesser amount over a smaller temperature range." Often the opposite is true. TC is generally specified with a "box method" in order to give an understanding of the likely error over the entire operating temperature range. It is a calculated value based only on minimum and maximum values of voltage, and does not take into account the temperatures at which these extrema occur.

For voltage references that are very linear over the specified temperature range, or for those that are not carefully tuned, the worst-case error can be assumed to be proportional to the temperature range. This is because the maximum and minimum output voltages are very likely to be found at the maximum and minimum operating temperatures. However, for very carefully tuned references, often identified by their very low temperature drift, the nonlinear nature of the reference may dominate.

For example, a reference specified as 100ppm/°C tends to appear quite linear over any temperature range, as the drift due to component mismatches completely obscures the

Figure 2. Voltage reference temperature characteristics

inherent nonlinearity. In contrast, the temperature drift of a reference specified as 5ppm/°C will be dominated by the nonlinearities.

Voltage references come in many forms and offer different features, but in the end, accuracy and stability are a voltage reference's most important features, as the main purpose of the reference is to provide a known output voltage. Variation from this known value is an error. Voltage reference specifications usually predict the uncertainty of the reference under certain conditions.

This can be easily seen in the output voltage vs temperature characteristic of Figure 2. Note that there are two possible temperature characteristics represented. An uncompensated bandgap appears as a parabola, with minima at the temperature extrema and maximum in the middle. A temperature compensated bandgap, such as the LT1019, shown here, appears as an "S" shaped curve, with greatest slope near the center of the temperature range. In the latter case, nonlinearity is exacerbated so that the aggregate uncertainty over temperature is reduced.

The best use of the temperature drift specification is to calculate maximum total error over the specified temperature range. It is generally inadvisable to calculate errors over unspecified temperature ranges unless the temperature drift characteristics are well understood.

Long Term Stability

This is a measure of the tendency of a reference voltage to change over time, independent of other variables. Initial shifts are largely caused by changes in mechanical stress, usually from the difference in expansion rates of the lead frame, die and mold compound. This stress effect tends to have a large initial shift that reduces quickly with time. Initial drift also includes changes in electrical characteristics of the circuit elements, including settling of device characteristics at the atomic level. Longer-term shifts are caused by electrical changes in the circuit elements, often referred to as "aging." This drift tends to occur at a reduced rate as compared to initial drift, and to further reduce over time. It is therefore often specified as drift/$\sqrt{khr}$. Voltage references tend to age more quickly at higher temperatures.

Thermal Hysteresis

This often-overlooked specification can also be a dominant source of error. It is mechanical in nature, and is the result of changing die stress due to thermal cycling. Hysteresis can be observed as a change in output voltage at a given temperature after a large temperature cycle. It is independent of temperature coefficient and time drift, and reduces the effectiveness of initial voltage calibration.

Most references tend to vary around a nominal output voltage during subsequent temperature cycles, so thermal hysteresis is usually limited to a predictable maximum value. Each manufacturer has their own method for specifying this parameter, so typical values can be misleading. Distribution data, as provided in data sheets such as the LT1790 and LTC6652, is far more useful when estimating output voltage error.
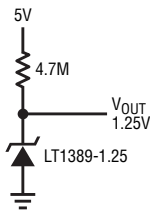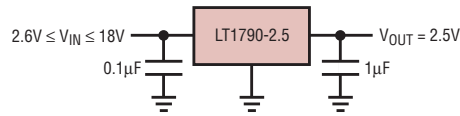
Figure 3. Shunt voltage reference

Figure 4. Series voltage reference

Other Specifications

Additional specifications that may be important, depending on application requirements include:
❏ Voltage Noise
❏ Line Regulation/PSRR
❏ Load Regulation
❏ Dropout Voltage
❏ Supply Range
❏ Supply Current

Reference Types

The two main types of voltage reference are shunt and series. See Table 2 for a list of Linear Technology series and shunt voltage references.

Shunt References

The shunt reference is a 2-terminal type, usually designed to work over a specified range of currents. Though most shunts are of the bandgap type and come in a variety of voltages, they can be thought of and are as simple to use as a Zener diode.

The most common circuit ties one terminal of the reference to ground and the other terminal to a resistor. The remaining terminal of the resistor is then tied to a supply. This becomes, in essence, a three terminal circuit. The shared reference/resistor terminal is the output. The resistor must be chosen such that the minimum and maximum currents through the reference are within the specified range over the entire supply range and load current range. These references are quite easy to design with, provided the supply voltage and load current do not vary much. If either, or both, may change substantially, then the resistor must be chosen to accommodate this variance, often forcing the circuit to dissipate significantly more power than required for the nominal case. It can be considered to function like a class A amplifier, in that sense.

Advantages of shunt references include simple design, small packages and good stability over wide current and load conditions. In addition, they are easily designed as negative voltage references and can be used with very high supply voltages, as the external resistor holds off most of the potential, or very low supplies, as the output can be as little as a few millivolts below the supply. Linear Technology offers shunt products including the LT1004, LT1009, LT1389, LT1634, LM399 and LTZ1000. A typical shunt circuit can be seen in Figure 3.

Series References

Series references are three (or more) terminal devices. They are more like low dropout (LDO) regulators, so they have many of the same advantages. Most notably, they consume a relatively fixed amount of supply current over a wide range of supply voltages, and they only conduct load current when the load demands it. This makes them ideal for circuits with large changes in supply voltage or load current. They are especially useful in circuits with very large load currents as there is no series resistor between the reference and supply.

Series products available from Linear Technology include the LT1460, LT1790, LT1461, LT1021, LT1236, LT1027, LTC6652, LT6660, and many others. Products such as the LT1021 and LT1019 may be operated either as a shunt or a series voltage reference. A series reference circuit is illustrated in Figure 4.
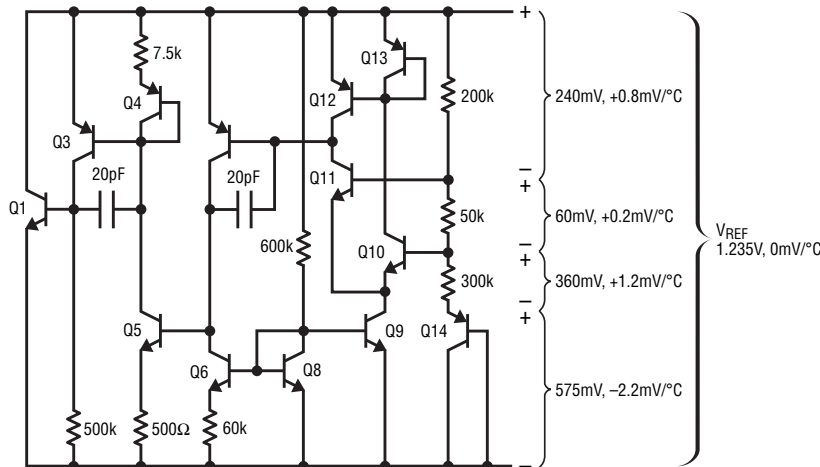

Figure 5. A bandgap circuit is designed for a theoretically zero temperature coefficient.
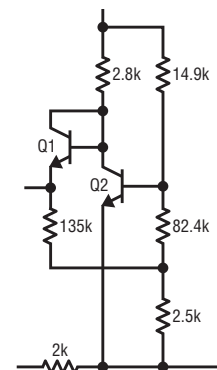

Figure 6. A 200mV reference circuit

Reference Circuits

There are many ways to design a voltage reference IC. Each has specific advantages and disadvantages.

Zener-Based References

The buried Zener type reference is a relatively simple design. A Zener (or avalanche) diode has a predictable reverse voltage that is fairly constant over temperature and very constant over time. These diodes are often very low noise and very stable over time if held within a small temperature range, making them useful in applications where changes in the reference voltage must be as small as possible.

This stability can be attributed to the relatively small number of components and die area as compared to other types of reference circuits, as well as the careful construction of the Zener element. However, relatively high variances in initial voltage and temperature drift are common. Additional circuitry may be added to compensate these imperfections, or to provide a range of output voltages. Both shunt and series references use Zener diodes.

Devices like the LT1021, LT1236 and LT1027 use internal current sources and amplifiers to regulate the Zener voltage and current to increase stability, as well as to provide various output voltages such as 5V, 7V and 10V. This additional circuitry makes the Zener diode more compatible with a wide variety of application circuits, but requires some additional supply headroom and may cause additional error.

Alternatively, the LM399 and LTZ1000 use internal heating elements and additional transistors to stabilize the temperature drift of the Zener diode, giving the best combination of temperature and time stability. In addition, these Zener-based products have extraordinarily low noise, giving the best possible performance. The LTZ1000 exhibits 0.05ppm/°C temperature drift, $2\mu V/\sqrt{kHr}$ long term stability and $1.2\mu V_{P-P}$ noise. To give some perspective, in a laboratory instrument, the total uncertainty in the LTZ1000's reference voltage due

to noise and temperature would be only about 1.7ppm plus a fraction of 1ppm per month due to aging.

Bandgap References

While Zener diodes can be used to make very high performance references, they lack flexibility. Specifically, they require supply voltages above 7V and they offer relatively few output voltages. In contrast, bandgap references can produce a wide variety of output voltages with little supply headroom—often less than 100mV. Bandgap references can be designed to provide very precise initial output voltages and low temperature drift, eliminating the need for time-consuming in-application calibration.

Bandgap operation is based on a basic characteristic of bipolar junction

transistors. Figure 5 shows a simplified version of the LT1004 circuit, a basic bandgap. It can be shown that a mismatched pair of bipolar junction transistors has a difference in V_{BE} that is proportional to temperature. This difference can be used to create a current that rises linearly with temperature. When this current is driven through a resistor and a transistor, the change over temperature of the base-emitter voltage of the transistor cancels the change in the voltage across the resistor if it is sized properly. While this cancellation is not completely linear, it can be compensated with additional circuitry to yield very low temperature drift.

The math behind the basic bandgap voltage reference is interesting in that it combines known temperature coef-

Table 2. Voltage references available from Linear Technology

| Type	Part	Description
Series	LT1019	Precision Bandgap
	LT1021	Precision Low Noise Buried Zener
	LT1027	Precision 5V Buried Zener
	LT1031	Precision Low Noise/Low Drift 10V Zener
	LT1236	Precision Low Noise Buried Zener
	LT1258	Micropower LDO Bandgap
	LT1460	Micropower Precision Bandgap
	LT1461	Micropower Ultra-Precision Bandgap
	LT1790	Micropower Low Dropout Bandgap
	LT1798	Micropower LDO Bandgap
	LT6650	Micropower 400mV/Adjustable Bandgap
	LTC6652	Precision Low Noise LDO Bandgap
Shunt	LM129	Precision 6.9V Buried Zener
	LM185	Micropower 1.2V/2.5V Zener
	LM399	Precision 7V Heated Zener
	LT1004	Micropower 1.2V/2.5V Bandgap
	LT1009	Precision 2.5V Bandgap
	LT1029	5V Bandgap
	LT1034	Micropower Dual (1.2V Bandgap/7V Zener)
	LT1389	Nanopower Precision Bandgap
	LT1634	Micropower Precision Bandgap
	LTZ1000	Ultra-Precision Heated Zener

ficients with unique resistor ratios to produce a voltage reference with theoretically zero temperature drift. Figure 5 shows two transistors scaled so that the emitter area of Q10 is 10-times that of Q11, while Q12 and Q13 hold their collector currents equal. This creates a known voltage between the bases of the two transistors of:

$$\Delta V_{BE} = \frac{kT}{q} \bullet \ln\left(\frac{AREA\ Q10}{AREA\ Q11}\right)$$

where k is the Boltzmann constant in J/kelvin (1.38×10^{-23}), T is temperature in kelvin (273 + T(°C)) and q is the charge of an electron in coulombs (1.6×10^{-19}). At 25°C, kT/q has a value of 25.7mV with a positive temperature coefficient of 86μV/°C. ΔV_{BE} is this voltage times ln(10), or 2.3, for a 25°C voltage of approximately 60mV with a tempco of 0.2mV/°C.

Applying this voltage to the 50k resistor tied between the bases creates a current that is proportional to temperature. This current biases a diode, Q14 with a 25°C voltage of 575mV with a –2.2mV/°C temperature coefficient. Resistors are used to create voltage drops with positive tempcos, which are added to the Q14 diode voltage, thus producing a reference voltage potential of approximately 1.235V with theoretically 0mV/°C temperature coefficient. These voltage drops are shown in Figure 5. The balance of the circuit provides bias currents and output drive.

Linear Technology produces a wide variety of bandgap references, including the LT1460, a small and inexpensive precision series reference, the LT1389, an ultralow power shunt reference, and the LT1461 and LTC6652, which are very high precision, low drift references. Available output voltages include 1.2V, 1.25V, 2.048V, 2.5V, 3.0V, 3.3V, 4.096V, 4.5V, 5V and 10V. These reference voltages can be provided over a wide range of supplies and load conditions with minimal voltage and current overhead. Products may be very precise, as with the LT1461, LT1019, LTC6652 and LT1790; very small, as with the LT1790 and LT1460 (SOT23), or
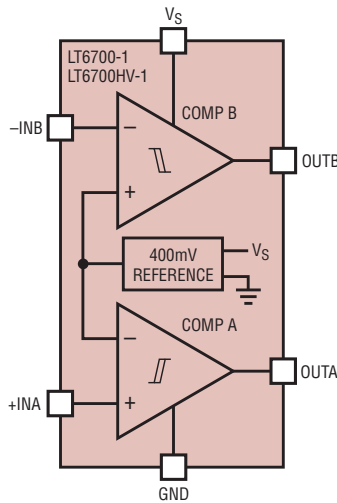
Figure 7. The LT6700 allows comparisons with thresholds as low as 400mV.

LT6660 in a 2mm × 2mm DFN package; or very low power, such as the LT1389, which requires only 800nA. While Zener references often have better performance in terms of noise and long term stability, new bandgap references such as the LTC6652, with 2ppm peak-to-peak noise (0.1Hz to 10Hz) are narrowing the gap.

Fractional Bandgap References

These are references based on the temperature characteristics of bipolar transistors, but with output voltages that may be as low as a few millivolts. They are useful for very low voltage circuits, especially in comparator applications where the threshold must

be less than a conventional bandgap voltage (approximately 1.2V).

Figure 6 shows the core circuit from the LM10, which combines elements that are proportional and inversely proportional to temperature in a similar fashion to the normal bandgap reference to obtain a constant 200mV reference. A fractional bandgap usually uses a ΔV_{BE} to generate a current that is proportional to temperature, and a V_{BE} to generate a current that is inversely proportional. These are combined in the proper ratio in a resistor element to generate a temperature-invariant voltage. The size of the resistor may be varied to alter the reference voltage without affecting the temperature characteristic. This differs from a traditional bandgap circuit in that the fractional bandgap circuit combines currents, while the traditional circuits tend to combine voltages, usually a base-emitter voltage and an I•R with opposite TC.

Fractional bandgaps like the LM10 circuit are based in part on a subtraction as well. The LT6650 has a 400mV reference of this type, combined with an amplifier. This allows the reference voltage to be altered by changing the gain of the amplifier, and gives a buffered output. Any output voltage from 0.4V to a few millivolts below the supply voltage can be generated with this simple circuit. In a more integrated solution, the LT6700 (Figure 7) and LT6703 combine a
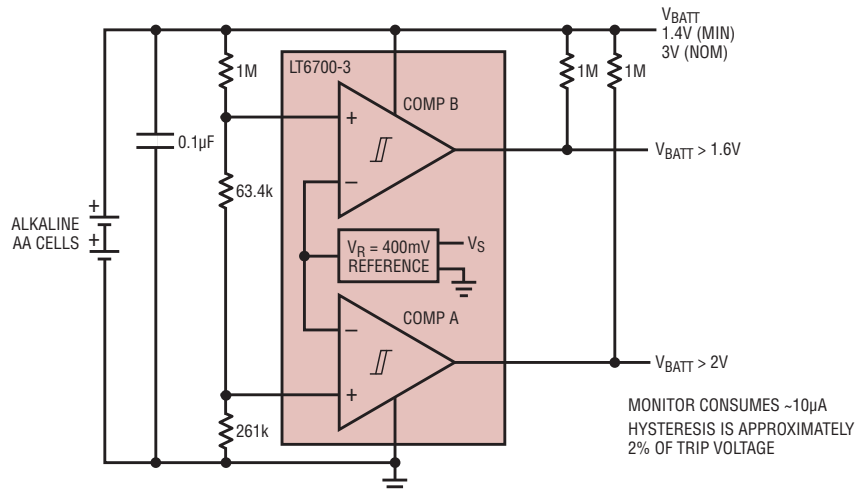
Figure 8. Higher thresholds are set by dividing the input voltage.

400mV reference with comparators, and can be used as voltage monitors or window comparators. The 400mV reference allows monitoring of small input signals, which decreases the complexity of monitor circuits and enables monitoring of circuit elements working on very low supplies as well. For larger thresholds, a simple resistor divider may be added (Figure 8). Each of these products is available in a small footprint package (SOT23), consumes low power (less than 10μA) and works on a wide supply range (1.4V to 18V). In addition, the LT6700 is available in a 2mm × 3mm DFN package and the LT6703 is available in a 2mm × 2mm DFN package.

Choosing a Reference

So, now, with all those options, how do you choose the right reference for your application? Here are a few hints that can narrow the range of options:

❑ **Is the supply voltage very high?** Choose a shunt.

❑ **Does the supply voltage or load current vary widely?** Choose a series.

❑ **Require high power efficiency?** Choose a series.

❑ **Figure your real-world temperature range.** Linear Technology provides guaranteed specifications and operation over various temperature ranges including 0°C to 70°C, –40°C to 85°C and –40°C to 125°C.

❑ **Be realistic about required accuracy.** It is important to understand the precision required by the application. This will help identify critical specifications. With the requirement in mind, multiply temperature drift by the specified temperature range. Add initial accuracy error, thermal hysteresis, and long term drift over the intended product life. Remove any terms that will be factory calibrated or periodically recalibrated. This gives an idea of total accuracy. For the most demanding applications, noise, line regulation and load regulation errors may also be added. As an example, a

reference with 0.1% (1000ppm) initial accuracy error, 25ppm/°C temperature drift over –40°C to 85°C, 200ppm thermal hysteresis, 2ppm peak-to-peak noise and 50ppm/$\sqrt{kHr}$ time drift would have a total uncertainty of over 4300ppm at the time the circuit is built. This uncertainty increases by 50ppm in the first 1000 hours the circuit is powered. The initial accuracy may be calibrated, reducing the error to 3300ppm + 50ppm • $\sqrt{(t/1000hours)}$.

Linear Technology offers a wide variety of voltage reference products. These include both series and shunt references—using Zeners, bandgaps and other schemes. References are available in multiple performance and temperature grades, as well as in nearly every conceivable package type.

❑ **What is the real supply range?** What is the maximum expected supply voltage? Will there be fault conditions such as battery load dump or hot-swap inductive supply spikes that the reference IC must withstand? This may significantly reduce the number of viable choices.

❑ **How much power can the reference consume?** References tend to fall into a few categories: more than 1mA, ~500μA, <300μA, <50μA, <10μA, <1μA.

❑ **How much load current?** Will the load draw substantial current or produce current that the reference must sink? Many references can provide only small currents to the load and few can absorb substantial current. The load regulation specification is a good guide.

❑ **How much room do you have?** References come in a wide variety of packages, including metal cans, plastic packages (DIP, SOIC, SOT) and very small packages, including the LT6660 in a 2mm × 2mm DFN. There is a widely held view that references in larger package sizes have less error due to mechanical stress than smaller packages. While it is true that some references may give better performance in larger packages, there is evidence that suggests performance difference has little to do directly with the package size. It is more likely that because smaller dice are used for products that are offered in smaller packages, some performance tradeoffs must be made to fit the circuit on the die. Usually, the package's mounting method makes a more significant performance difference than the actual package—careful attention to mounting methods and locations can maximize performance. Also, devices with smaller footprints can show reduced stress when a PCB bends compared to devices with larger footprints. This is discussed in detail in application note AN82, "Understanding and Applying Voltage References," available from Linear Technology.

Conclusion

Linear Technology offers a wide variety of voltage reference products. These include both series and shunt references designed with Zeners, bandgaps and other types. References are available in multiple performance and temperature grades and nearly every conceivable package type. Products range from the highest precision available to small and inexpensive alternatives. With a vast arsenal of voltage reference products, Linear Technology's voltage references meet the needs of almost any application.

See also Linear Technology's application note AN82 "Understanding and Applying Voltage References," available at www.linear.com. ∠T

1.2A Monolithic Buck Regulator Shrinks Supply Size and Cost with Programmable Output Current Limit

by Tom Sheehan

Introduction

Power supplies are often overqualified for their job. This is because power ICs often specify a current limit that is more than twice the rated output current of the device. The power supply components are sized to handle the maximum current that the IC can deliver, even if loads are unlikely to draw that current during normal operation. The components are bigger and more expensive than they need to be.

There is, however, an alternative: set an *accurate* maximum output current on the supply once the real world load is known. Accurately setting the maximum output current reduces the required current rating of the regulator's power path components, thus replacing big, expensive components with smaller, less expensive ones. A limit on the regulator's maximum output limits the maximum power dissipation of both the supply and the load, thus reducing the potential for localized heating. Monitoring and controlling the output current also makes for a robust solution, which is able to withstand harsh overload and short circuit conditions.
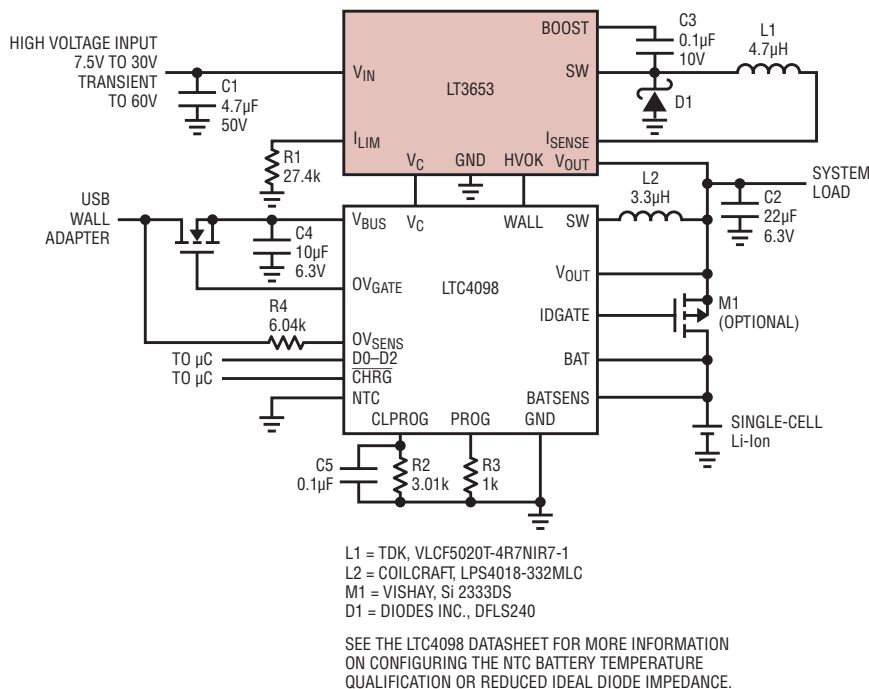
The LT3653 and LT3663 are monolithic step-down switching regulators that have an accurate output current limit programmable from 400mA to 1.2A. The LT3663 is a general purpose high voltage step-down regulator while the LT3653 is designed for use with Linear Technology Bat-Track™ enabled battery chargers and power management ICs (PMICs). The maximum input voltages of 30V (LT3653) or 36V (LT3663) with 60V transient ride through capability are well suited to automotive, industrial, distributed supply, and wall transformer applications.

Programmable Output Current Limit

Monolithic switching regulators typically limit the peak switch current to protect the internal switch from being damaged during an overload or short circuit event. The maximum switch peak current limit is typically more than two times the maximum output current rating of the part. While the peak switch current limit prevents overstressing the IC, it does not keep the entire application from overheating during an overload condition. For example, a regulator with an output current rating of 1A is typically capable of providing over 2A at the output. During an output overload condition, the power dissipation of the regulator could more than double, making thermal management more difficult. The LT3653 and LT3663 reduce localized hot spots by controlling the total power dissipation of the application with a programmable, accurate current limit.

Conservative design principles call for power path components that are rated for worst-case currents. In the above example, where a 1A part is capable of delivering 2A, the power path components must be sized for greater than 2A, because during an output short circuit or overload the inductor and diode can conduct up to 2A. In contrast, the PowerPath components in LT3653 and LT3663 applications are sized based on the programmed maximum output current limit. Therefore, an application with a 750mA output current limit requires only 750mA rated components. This allows for smaller, lower

L1 = TDK, VLCF5020T-4R7NIR7-1
L2 = COILCRAFT, LPS4018-332MLC
M1 = VISHAY, Si 2333DS
D1 = DIODES INC., DFLS240

SEE THE LTC4098 DATASHEET FOR MORE INFORMATION ON CONFIGURING THE NTC BATTERY TEMPERATURE QUALIFICATION OR REDUCED IDEAL DIODE IMPEDANCE.

Figure 1. Charging a single cell Li-ion battery from either a USB input or high voltage input. This solution offers a seamless, highly efficient, low part count approach to dual input charging and PowerPath™ control of a Li-ion battery-powered application. If additional integration is required for more system supplies, the LT3653 can be used in a similar fashion with the LTC3576 PMIC.

cost devices and a smaller overall application footprint.

In early product development, system designers usually don't know how much current their load will draw. Once they choose a power supply, they are committed. However, with the programmable current limit of the LT3653 and LT3663, once the load has been fully characterized, they can change the output current limit by changing an inexpensive 1% resistor.

The output current limit is implemented by monitoring and controlling the average inductor current. When an overcurrent event occurs, the regulator disables the power switch. This robust solution withstands short circuit and overload conditions throughout the entire input voltage range.

The LT3653 Plays Well with Bat-Track Battery Chargers

The LT3653 is a 1.5MHz constant frequency, current mode control, step-down regulator designed for use with Linear Technology's Bat-Track-enabled battery charger PowerPath power managers. The LT3653 steps down a high voltage input to power the system load and charge a single-cell Li-ion battery charger.

Minimizing the voltage across a linear battery charger increases efficiency. To accomplish this, a Bat-Track battery charger controls the LT3653's V_C Pin, overriding the error amplifier. In this way, the output voltage of the LT3653 is regulated by the battery charger to a potential slightly above the battery voltage, typically 300mV.
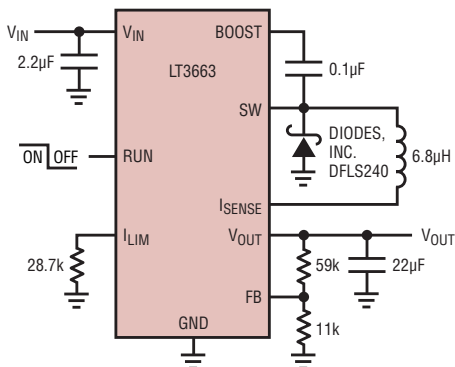
Figure 3. A LT3663 application producing 5V at 1.2A from an input of 7.5V to 36V. The input is capable of handling 60V transients.
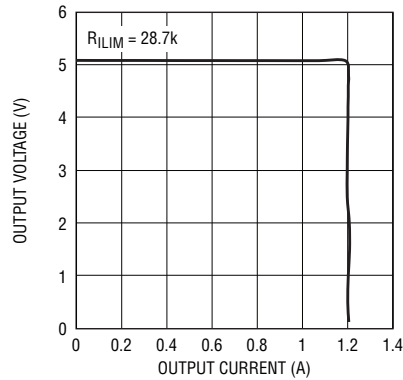
Figure 2. The LT3663 output current limit at 1.2A

Input overvoltage protection allows the LT3653 to handle 60V input transients. The HVOK pin indicates that the internal bias supplies are present and no faults have occurred (i.e., over-temperature and input overvoltage and undervoltage). The LT3653 includes internal compensation, and an internal boost diode to minimize the number of external components. The LT3653 is available in an 8-lead 2mm × 3mm DFN package with an exposed pad.

Charging a Single Cell Li-Ion Battery from Either a USB or High Voltage Input

Figure 1 shows a LT3653 and LTC4098 application charging a single cell Li-ion battery from either a USB input or high voltage input. This solution offers a seamless, highly efficient, low part count approach to dual input charging and power path control of a Li-ion battery-powered application. If additional integration is required for more system supplies, the LT3653 can be used in a similar fashion with the LTC3576 PMIC.

When a high voltage input is applied, the LT3653 HVOK pin signals the LTC4098 that it is capable of delivering power. The LTC4098 takes control of the LT3653's V_C pin and regulates the output voltage to just above the battery voltage. This Bat-Track function optimizes the battery charger efficiency.

When present, the high voltage input supplies the battery charge current and the system load current. If the total current increases beyond the

LT3653 programmed current limit, the regulator's output voltage decreases to reduce charge current as the battery charger enters dropout. If the system load continues to increase, the battery charge current first decreases to zero and then reverses direction to deliver power to the system load, supplementing the LT3653. The transitions between these modes of operation are seamless to the system load. The output current from the LT3653 regulator never exceeds the programmed output current limit.

The LT3663 Directly Accepts 36V Inputs

The LT3663 is a 1.5MHz constant frequency, current mode control, general purpose, monolithic switching regulator suited for automotive batteries, industrial power supplies, distributed supplies, and wall transformers. The LT3663 includes a low current shutdown mode, input overvoltage and undervoltage lockout, and thermal shutdown. The LT3663 is available in 8-lead (2mm × 3mm) DFN package with exposed pad. An 8-lead MSOP package with exposed pad will be available soon.

The LT3663 can also function as a constant current, constant voltage (CC/CV) source to charge a supercapacitor or other energy storage device. The IC operates in constant current mode at the programmed current limit until the capacitor reaches the programmed output voltage. It then operates in a constant voltage mode to maintain that voltage.

Figure 2 shows the LT3663 output current limit at 1.2A. For output currents below 1.2A the regulator is in constant voltage mode. When the output current is increased to 1.2A it goes into constant current mode. The output current is maintained at 1.2A from V_{OUT} nominal down to 0V.

7.5V–36V to 5V Buck Regulator with 1.2A Output Current Limit

Figure 3 shows a LT3663 application producing 5V at 1.2A from an input of 7.5V to 36V. The input is capable

Boost Converters for Keep-Alive Circuits Draw Only 8.5μA of Quiescent Current

by Xiaohua Su

Introduction

Industrial remote monitoring systems and keep-alive circuits spend most of their time idle. Many of these systems use batteries, so to maximize run time power losses, even during low power idle modes, must be minimized. Even at no load, power supplies draw some current to produce a regulated voltage for keep-alive circuits.

The LT8410/-1 DC/DC boost converter features ultralow quiescent current and integrated high value feedback resistors to minimize the draw on the battery when electronics are idle.

An entire boost converter takes very little space, as shown in Figure 1.

Ultralow Quiescent Current Low Noise Boost Converter with Output Disconnect

When a micropower boost converter is in regulation with no load, the input current depends mainly on two things—the quiescent current (required to keep regulation) and the output feedback resistor value. When the output voltage is high, the output feedback resistor can easily dissipate more power than the quiescent current of the IC. The quiescent current of the LT8410/-1 is a low 8.5μA, while the integrated output feedback resistors

have very high values (12.4M/0.4M). This enables the LT8410/-1 to dissipate very little power in regulation at no load. In fact, the LT8410/-1 can regulate a 16V output at no load from 3.6V input with about 30μA of average input current. Figures 2, 3 and 4 show the typical quiescent and input current in regulation with no load.

The LT8410/-1 controls power delivery by varying both the peak inductor current and switch off time. This control scheme results in low output voltage ripple as well as high efficiency over a wide load range. As shown in Figure 5, even with a small 0.1μF output capacitor, the output ripple is typically less than 10mV. The part also features output disconnect, which disconnects the output voltage from the input during shutdown. This output disconnect circuit also sets a maximum output current limit, allowing the chip survive output shorts.

An Excellent Choice for High Impedance Batteries

A power source with high internal impedance, such as a coin cell battery, may show normal output voltage on a voltmeter, but its voltage can collapse under heavy current demands. This makes it incompatible with high
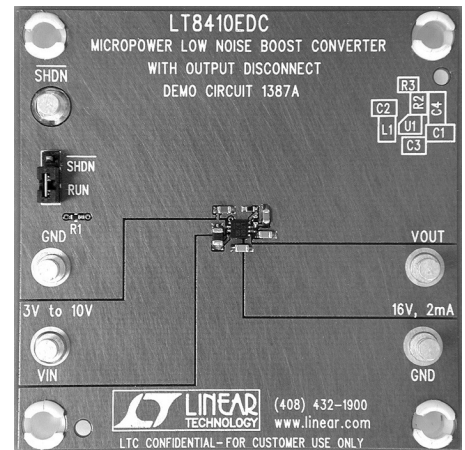

Figure 1. The LT8410/-1 is designed to facilitate compact board layout.

switch-current DC/DC converters. The LT8410/-1 has an integrated power switch and Schottky diode, and the switch current limits are very low (25mA for the LT8410 and 8mA for the LT8410-1). This low switch current limit enables the LT8410/-1 to operate very efficiently from high impedance sources, such as coin cell batteries, without causing inrush current problems. Figure 6 shows the LT8410-1 charging an electrolytic capacitor. Without any additional external circuitry, the input current for
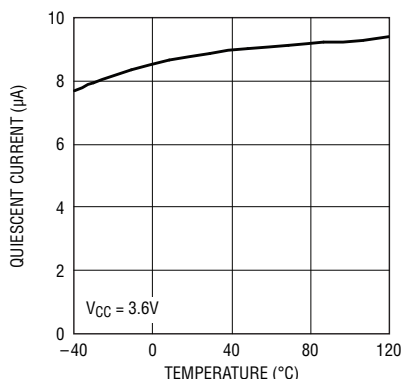

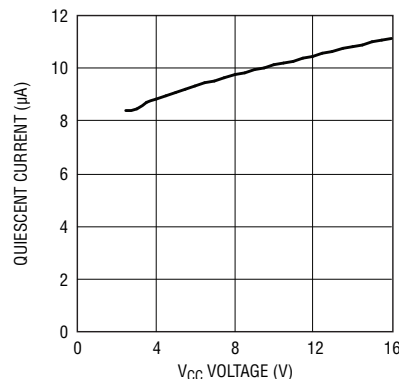**Figure 2. Quiescent current vs temperature—not switching**

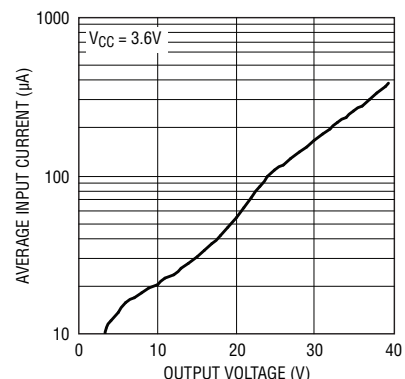**Figure 3. Quiescent current vs V_{CC} voltage—not switching**

Figure 4. Average input current in regulation with no load

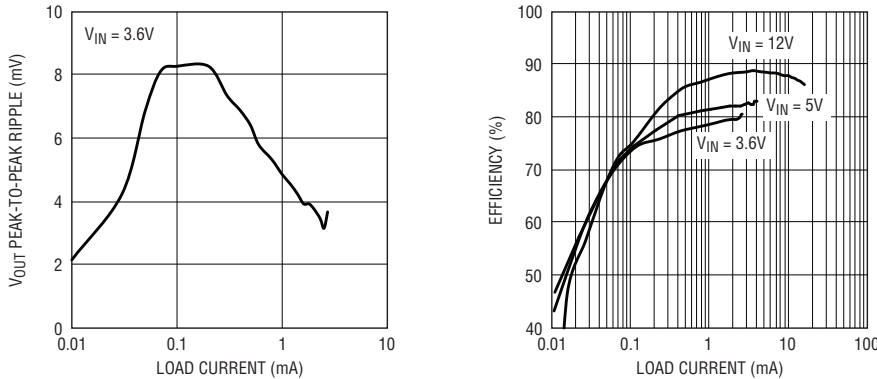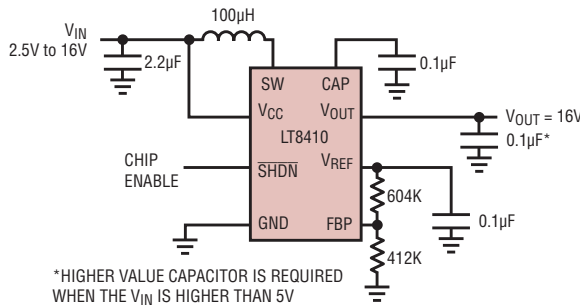*HIGHER VALUE CAPACITOR IS REQUIRED WHEN THE V_{IN} IS HIGHER THAN 5V*

Figure 5. General purpose bias with wide input voltage and low output voltage ripple

the entire charging cycle is less than 8mA.

Tiny Footprint with Small Ceramic Capacitors

Available in a tiny 8-pin 2mm × 2mm DFN package, the LT8410/-1 is internally compensated and stable for a wide range of output capacitors. For most applications, using 0.1µF output capacitor and 1µF input capacitor is sufficient. An optional 0.1µF capacitor at the V_{REF} pin implements a soft-start feature. The combination of small package size and the ability to use small ceramic capacitors enable the

LT8410/-1 to fit almost anywhere. Figure 1 shows the size of a circuit similar to that shown in Figure 4, illustrating how little board space is required to build a full featured LT8410/-1 application.

$\overline{SHDN}$ Pin Comparator and Soft-Start Reset Feature

An internal comparator compares the $\overline{SHDN}$ pin voltage to an internal voltage reference of 1.3V, giving the part a precise turn-on voltage level. The $\overline{SHDN}$ pin has built-in programmable hysteresis to reject noise and tolerate slowly varying input voltages. Driving

the $\overline{SHDN}$ pin below 0.3V shuts down the part and reduces input current to less than 1µA. When the part is on, and the $\overline{SHDN}$ pin voltage is close to 1.3V, 0.1µA current flows out of the $\overline{SHDN}$ pin. A programmable enable voltage can be set up by connecting external resistors as shown in Figure 7.

The turn-on voltage for the configuration is:

$$1.30 \bullet \left(1 + \frac{R1}{R2}\right)$$

and the turn-off voltage is:

$$(1.24 - R3 \bullet 10^{-7}) \bullet \left(1 + \frac{R1}{R2}\right) - (R1 \bullet 10^{-7})$$

where R1, R2 and R3 are resistance in Ω. Programming the turn-on/turn-off voltage is particularly useful for applications where high source impedance power sources are used, such as energy harvesting applications.

By connecting an external capacitor (typically 47nF to 220nF) to the V_{REF} pin, a soft-start feature can be implemented. When the part is brought

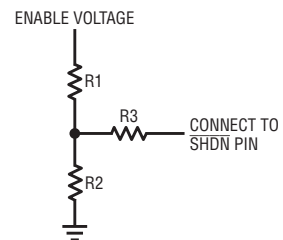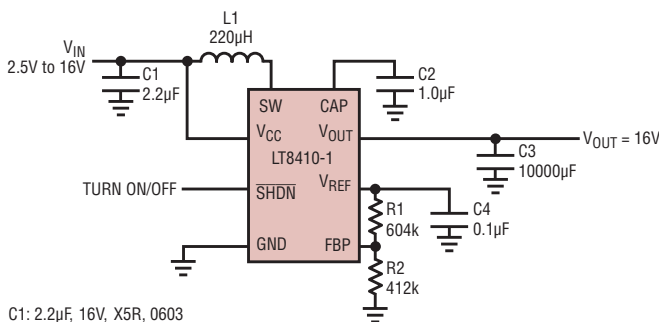**Figure 7. Programming the enable voltage by using external resistors**

C1: 2.2µF, 16V, X5R, 0603
C2: 1.0µF, 25V, X5R, 0603*
C3: 10000µF, Electrolytic Capacitor
C4: 0.1µF, 16V, X7R, 0402
L1: COILCRAFT LPS3008-224ML
* HIGHER CAPACITANCE VALUE IS REQUIRED FOR C2 WHEN THE V_{IN} IS HIGHER THAN 12V
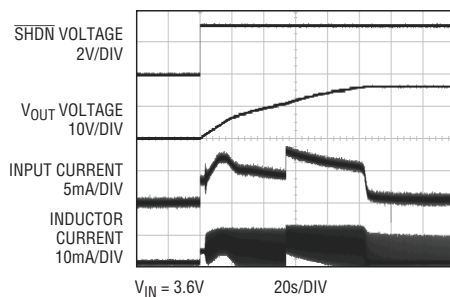

Figure 6. Capacitor charger with the LT8410-1 and charging waveforms

Industrial/Automotive Step-Down Regulator Accepts 3.6V to 36V and Includes Power-On Reset and Watchdog Timer in 3mm × 3mm QFN

by Ramanjot Singh

Introduction

As the number of microprocessors in automotive and industrial applications continues to expand, so does the need for rugged step-down regulators that can operate over a wide input voltage range and withstand high voltage transients and output shorts. Microprocessor-based applications also require supervisory functions, such as power-on reset (POR) and watchdog timing, to ensure high system reliability. The regulator must have high efficiency at light loads to increase battery life. The LT3689 delivers all of these features in tiny 16-pin 3mm × 3mm QFN and 16-pin MSOP packages.

Features of the LT3689 Step-Down Regulator

The LT3689 employs a constant frequency, current mode architecture to provide 800mA of continuous output current. The part operates from a wide 3.6V to 36V input range and can protect itself from input transients up to 60V. It is internally compensated, which helps to lower the external component count. The switching frequency can be set anywhere between 350kHz and 2.2MHz by tying a resistor from the R_T pin to ground, allowing the designer to optimize component size and efficiency. The switching frequency can also be synchronized to an external clock for noise sensitive applications. An external resistor divider programs the output voltage to any value above the part's 0.8V reference. Also, the boost diode is integrated into the IC to minimize solution size and cost. Figure 1 shows a typical application of LT3689.

Soft-Start and Output Short Circuit Protection

The LT3689 includes a soft-start feature that limits the maximum inrush current during start-up and recovery from fault conditions. The soft-start circuit ramps up the peak switch current limit in approximately 150µs, reducing the peak input current.

The DA pin is used to monitor the current in the catch diode. If the catch diode current at the end of switch cycle is higher than the DA current limit then the part delays the switch turn-on until the catch diode current drops below the DA current limit. This protects the LT3689 in the face of inductor current runaway situations,

especially during output overload or short at high switching frequencies with high input voltages and small inductor values. Other protection features such as frequency foldback, cycle-by-cycle current limit, and thermal shutdown together ensure that the part is not damaged by excessive switch currents during startup, overload or short circuit.

Pin Selectable Modes of Operation: Low Ripple Burst Mode Operation and Pulse-Skipping Mode

Two modes of operation can be selected through the SYNC pin. Applying a logic low to the SYNC pin enables the low ripple Burst Mode® operation, which maintains high efficiency at light loads while keeping output ripple low. In Burst Mode operation, the LT3689 delivers single cycle bursts of current to the output capacitor followed by sleep periods. Between bursts, all circuitry associated with controlling the output switch is shutdown, reducing the V_{IN} pin and OUT pin currents in a typical application to a mere 50µA and 75µA, respectively. As the load current decreases to a no load condition, the percentage of sleep time increases, thus decreasing average input current.

A logic high on SYNC disables Burst Mode operation, allowing the part to skip pulses at light loads. The advantage of this pulse-skipping mode over Burst Mode operation is that the part continues to switch at the programmed frequency (set by R_T) down to very low load currents, above 15mA at $12V_{IN}$ in a typical application.
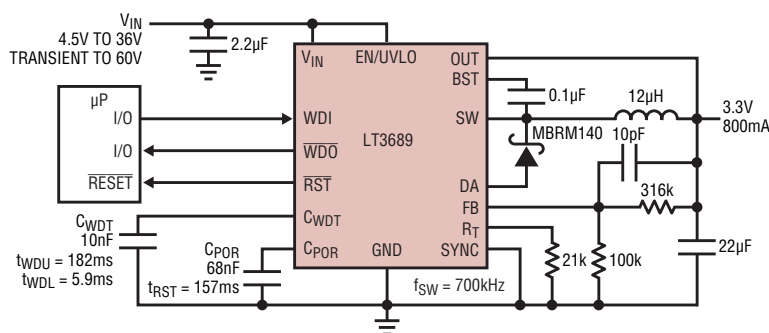
Figure 1. LT3689 typical application circuit with reset time set to 157ms and watchdog timeout period set to 182ms

Programmable Undervoltage Lockout

The LT3689 can be shutdown by pulling the EN/UVLO pin below 0.3V. In shutdown, quiescent current is less than 0.5µA. The EN/UVLO pin can also be used to perform an accurate undervoltage lockout (UVLO) function. A resistor divider from V_{IN} pin can be used to program the UVLO threshold of the circuit using the 1.26V accurate threshold of the EN/UVLO pin. A 4µA current hysteresis on this pin is also provided to allow the user to program desired voltage hysteresis. The LT3689 also has an internal UVLO that prevents the part from switching if V_{IN} pin ever goes below 3.3V (typical). The part only starts switching when V_{IN} is higher than 3.4V and EN/UVLO pin is above the 1.26V threshold.

Low Dropout

The LT3689 features low dropout for output voltages above 3V. The minimum operating voltage of the device is determined either by the LT3689's internal undervoltage lockout or by its maximum duty cycle. Unlike many buck regulators, the LT3689 can extend its duty cycle by staying on for multiple cycles, provided that the boost capacitor is charged above the minimum voltage of 2.5V. Eventually, after several switching cycles, the boost capacitor discharges. Internal circuitry detects this condition and charges the boost capacitor only when needed. Also, a bigger boost capacitor allows even higher duty cycle, allowing extremely low dropout operation. The dropout voltage for a 5V typical application is about 400mV at 200mA load and 900mV at 800mA load.
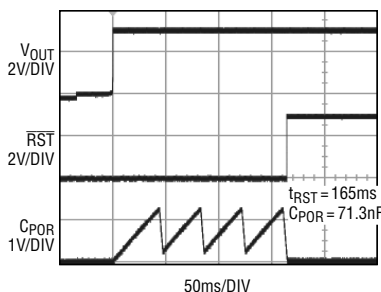

Figure 2. Power-on reset feature of LT3689

Power-On Reset (POR)

Many microprocessor-based applications powered by the output of a switching regulator must know when the regulator output is ready and stable before the microprocessor starts operating. Likewise, once running, the electronic system must be warned when the regulator output has dropped below a minimum tolerable threshold, such as during overload or shutdown conditions. This is required to prevent unreliable operation and to allow the microprocessor to perform housekeeping operations before power is completely lost.

The LT3689's accurate internal voltage reference and glitch immune precision POR comparator and timer circuit feed these specific needs of microprocessor-based applications. The switcher's output voltage must be above 90% of programmed value for its $\overline{RST}$ pin to remain high (refer to Figure 2). The LT3689 asserts $\overline{RST}$ during power-up, power-down and brown-out conditions. Once the output voltage rises above the $\overline{RST}$ threshold, the adjustable reset timer is started and $\overline{RST}$ is released after the reset timeout period. On power-down, once output voltage drops below $\overline{RST}$ threshold, $\overline{RST}$ is held at a logic low. The reset timer is adjustable using an external capacitor. The $\overline{RST}$ pin has a weak pull up to the OUT pin.

The POR comparator is designed to avoid false triggering. High frequency noise on the FB pin can falsely trip $\overline{RST}$, particularly when the monitored output is already near the reset threshold. This can cause oscillatory behavior at the $\overline{RST}$ pin. The traditional way of tackling this problem is to add some DC hysteresis in the comparator input, which changes the threshold point once the output flips. The problem is that the addition of DC hysteresis makes the trip voltage less accurate, since the trip point changes once the output changes. The LT3689 does not use hysteresis. Instead, it performs an integration-like function on transient events at the comparator. In this way the magnitude and duration of the event are both important to the comparator threshold. Figure 3 illustrates the typical transient duration versus comparator overdrive (as a percentage of trip threshold) required to trip the comparator.

Selecting the Reset Timing Capacitor

The reset timeout period is adjustable in order to accommodate a variety of microprocessor applications. The reset timeout period, t_{RST}, is adjusted by connecting a capacitor between the C_{POR} pin and ground. The value of this capacitor is determined by:

$$C_{POR} = t_{RST} \cdot 432 \cdot 10^{-9}$$

with C_{POR} in Farads and t_{RST} in seconds. The C_{POR} value per millisecond of delay can also be expressed as $C_{POR}/ms = 432$ (pF/ms).

Leaving the C_{POR} pin unconnected generates a minimum reset timeout of approximately 25µs with 10kΩ pull-up to 5V on $\overline{RST}$ pin. Maximum reset timeout is limited by the largest available low leakage capacitor. The accuracy of the timeout period will be affected by capacitor leakage (the nominal charging current is 2µA) and capacitor tolerance. A low leakage ceramic capacitor is recommended.

Watchdog Modes: Timeout or Window

The LT3689 also includes an adjustable watchdog timer that monitors a microprocessor's activity. If a code execution error occurs in a µP, the watchdog detects the error and sets the $\overline{WDO}$ pin low. This signal can be used to interrupt a routine or to reset a µP.
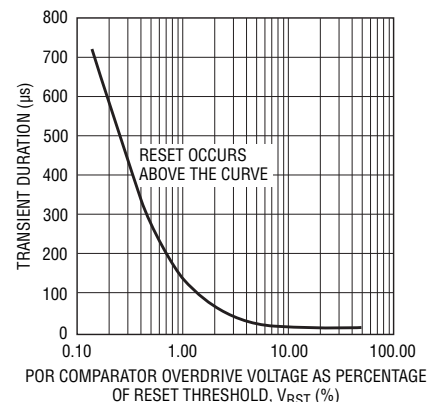

Figure 3. Typical transient duration vs POR comparator overdrive

The watchdog is operated either in timeout or window mode (refer to Figure 4). In timeout mode, the microprocessor needs to toggle the WDI pin before the watchdog timer expires to keep the $\overline{WDO}$ pin high. If the voltage on the WDI pin does not transition during the programmed timeout period then the circuitry pulls $\overline{WDO}$ low.

In window mode, the WDI pin's negative-going pulses must appear inside a programmed time window to prevent $\overline{WDO}$ from going low. If more than two falling pulses are registered in the lower boundary period (t_{WDL}), the $\overline{WDO}$ pin is forced low. The $\overline{WDO}$ pin also goes low if no negative edge is supplied to the WDI pin within the upper boundary period (t_{WDU}).

During a code execution error, the microprocessor outputs WDI pulses that are either too fast or too slow. This condition asserts $\overline{WDO}$ low and forces the microprocessor to reset the program.

In window mode, the WDI signal is bounded by an upper and lower boundary periods for normal operation. The period of the WDI input signal should be longer than the window mode's lower boundary period and shorter than the upper boundary period to keep $\overline{WDO}$ high under normal conditions. The window mode's lower and upper boundary periods have a fixed ratio of 31. These times can be increased or decreased by adjusting an external capacitor on the C_{WDT} pin.

In both watchdog modes, when $\overline{WDO}$ is asserted, the reset timer is enabled. Any WDI pulses that appear while the reset timer is running are ignored. When the reset timer expires, the $\overline{WDO}$ is allowed to go back high again. Therefore, if no input is applied to the WDI pin then the watchdog circuitry produces a train of pulses on the $\overline{WDO}$ pin. The high time of this pulse train is equal to the upper boundary period and low time is equal to the reset period. Also, $\overline{WDO}$ and $\overline{RST}$ cannot be logic low simultaneously. If $\overline{WDO}$ is low and there is an undervoltage lockout fault, $\overline{RST}$ goes low and $\overline{WDO}$ will go high.

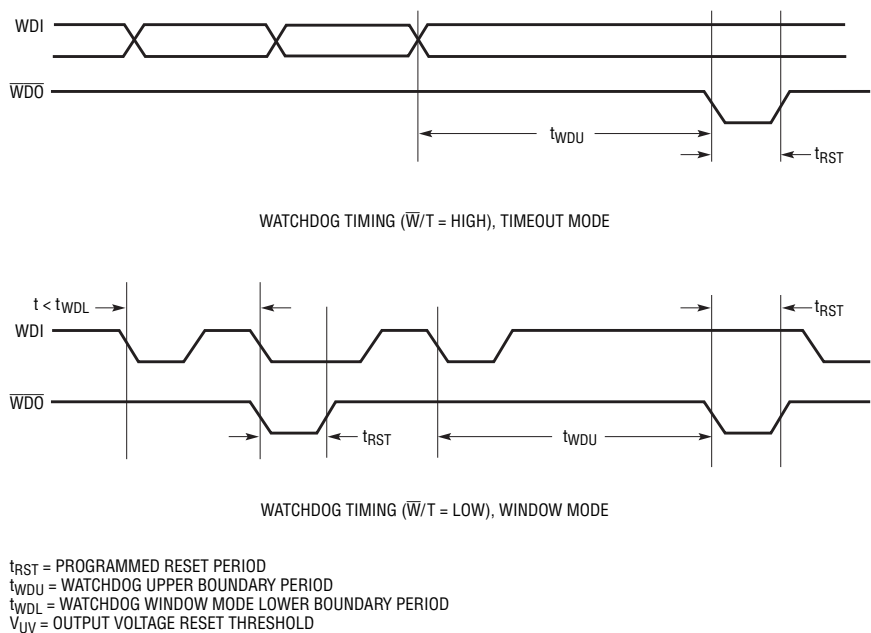The $\overline{WDE}$ pin allows the user to turn on or off the watchdog function. This feature can be used to reliably program the connected microprocessor in the factory. During factory programming of the microprocessor, $\overline{WDE}$ pin can be kept high to prevent $\overline{WDO}$ from toggling and thus prevents $\overline{WDO}$ from interfering with the microprocessor's programming procedure.

Tying the $\overline{WDO}$ and $\overline{RST}$ pins together will generate a reset signal when either the output voltage falls 10% below the regulation value or if there is a watchdog error.

Selecting the Watchdog Timing Capacitor

The watchdog upper boundary period is adjustable and can be optimized for software execution. The watchdog upper boundary period is adjusted by connecting a capacitor between the C_{WDT} pin and ground. Given a specified watchdog upper boundary period, the capacitor is determined by:

$$C_{WDT} = t_{WDU} \cdot 55 \text{ (pF/ms)}$$

The window mode lower boundary period has a fixed relationship to upper boundary period for a given capacitor. The lower boundary period is related to the upper boundary period by the following:

$$t_{WDL} = 1/31 \cdot t_{WDU}$$

Leaving the C_{WDT} pin unconnected generates a minimum watchdog upper boundary period of approximately 200µs with 10kΩ pull-up to 5V on $\overline{WDO}$ pin. Maximum timeout is limited by the largest available low leakage capacitor. The accuracy of the upper and lower boundary periods is affected by capacitor leakage (the nominal charging current is 2µA) and capacitor tolerance. A low leakage ceramic capacitor is recommended.

Conclusion

The wide input range, low quiescent current, supervisory features, robustness and small size of the LT3689 makes it an ideal candidate to power automotive and industrial applications. The part withstands 60V V_{IN} transients and normal operation is guaranteed for max V_{IN} of 36V, and the part is robust against inrush and short circuit conditions. The Burst Mode circuitry provides high efficiencies at light loads. Programmable switching frequency allows the designer to trade off between component size and efficiency. The accurate POR and Watchdog circuitry of LT3689 allows complete supervisory control of a microprocessor connected to the output of the LT3689 switching regulator. *LT*

t_{RST} = PROGRAMMED RESET PERIOD
t_{WDU} = WATCHDOG UPPER BOUNDARY PERIOD
t_{WDL} = WATCHDOG WINDOW MODE LOWER BOUNDARY PERIOD
V_{UV} = OUTPUT VOLTAGE RESET THRESHOLD

Figure 4. Watchdog timing diagram

Complete APD Bias Solution in 60mm² with On-the-Fly Adjustable Current Limit and Adjustable V$_{APD}$

By Xin (Shin) Qi

Introduction

The overriding factor limiting functionality in fiber-optic communication systems is available space. A compact APD (avalanche photo diode) bias solution with a high degree of feature integration is the key to breaking new ground in system size and performance. The LT3571 offers such a solution in a tiny 3mm × 3mm QFN package.

The LT3571 combines a current mode step-up DC/DC converter and a high side fixed voltage drop APD current monitor with an integrated 75V power switch and Schottky diode. The combination of a traditional voltage loop and a unique current loop allows customers to set an accurate APD current limit at any given bias voltage. The integrated high side current monitor provides an 8% accurate current that is proportional to the load current, making it possible to adjust the APD bias voltage via the CTRL pin. This feature-rich device makes it possible to produce a single stage boost converter to bias high voltage APDs in only 60mm².

Low Noise APD Bias Supply

The gain of the APD is dependent on the bias voltage, so the bias supply must minimize the noise contamination from switching regulators and other sources. Figure 1 shows the LT3571 configured to produce an ultralow noise power supply for a 45V APD with 2.5mA of load current capability. The MONIN voltage is regulated by the internal voltage reference and the resistor divider made up of R1 and R2. Resistor R$_{SENSE}$ is selected to set the APD current limit at 200mV/1.2R$_{SENSE}$ – 0.2mA.

The CTRL pin can override the internal reference, making it possible to optimize the APD bias on the fly to maximize receiver performance.
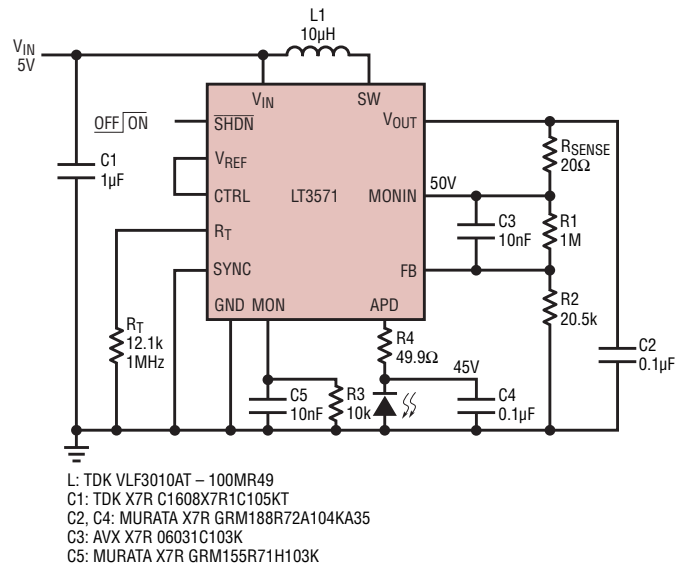

L: TDK VLF3010AT – 100MR49
C1: TDK X7R C1608X7R1C105KT
C2, C4: MURATA X7R GRM188R72A104KA35
C3: AVX X7R 06031C103K
C5: MURATA X7R GRM155R71H103K

Figure 1. Low noise APD bias supply

When the CTRL pin is connected to a supply above 1V, the output voltage is regulated with feedback at 1V. When driven below 1V, the feedback and the output voltage follow accordingly.

The APD pin, the output of the current monitor, provides a voltage to the APD load that is fixed 5V below the MONIN pin. The LT3571 includes a precise current mirror with a factor-of-five attenuation. The proportional current output signal at the MON pin can be used to accurately indicate the APD signal strength. The voltage variance of APD pin voltage is only ±200mV over the entire input current range and the whole temperature range. Figure 2 shows the evaluation board for this topology.

The topology uses several filter capacitors to achieve ultralow noise performance. The capacitor at V$_{OUT}$ pin and the 0.1µF capacitor at the APD pin suppress switching noise. The 10nF feedforward capacitor across the MONIN and FB pins filters out high frequency internal reference and error amplifier noise. Figure 3 shows the measured switching noise is less than 500µV$_{P–P}$ at 1mA load current. This exceptionally low noise bias voltage
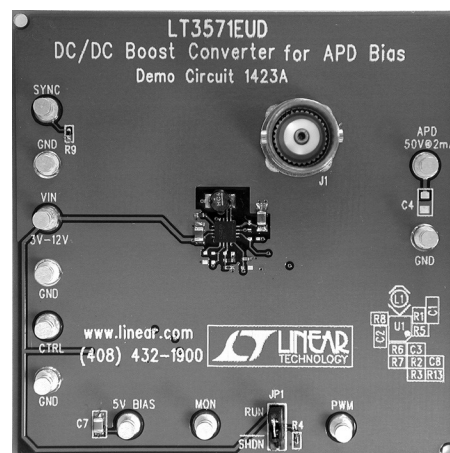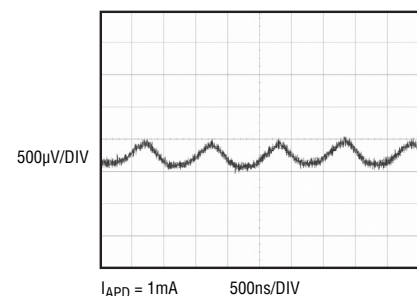

Figure 2. The LT3571 evaluation board

Figure 3. AC-coupled noise ripple at APD pin

gives the APD greater sensitivity and dynamic range.

Fast APD Current Monitor Transient Response

Design efforts in modern communications systems increasingly focus on 10Gbits/s GPON systems, which demand that the transient response of the APD current monitor is less than 100ns for a two-decades-of-magnitude input current step. To meet this challenging requirement, many designers rely on a simple discrete current mirror topology to reduce parasitic capacitance on the signal path, sacrificing monitor accuracy and board space. In contrast, the LT3571's APD current monitor is carefully designed to provide not only a fixed voltage drop and high accuracy, but also the required fast transient response.

Figure 4 shows a compact circuit that responds quickly to current transients. Unlike the ultralow noise topology shown in Figure 1, the filter capacitor at the APD pin is moved to the MONIN pin. C2, C3 and R_{SENSE} form a π filter to isolate the APD current monitor from high frequency switching noise. The capacitor at the MON pin is also removed to reduce the transient delay on the measurement path.

The transient speed is measured using the same technique described in the Linear Technology Design Note 447 "A Complete Compact APD Bias Solution for a 10GBit/s GPON System." Figures 5 and 6 show the measured input signal falling transient response and input signal rising transient response, respectively, where the input current levels are 10μA and 1mA. Note that there is an inversion and DC offset present in the measurement. The measurements show a transient response time of less than 100ns, well within the stringent speed demands of the 10Gbits/s GPON system.

APD Bias Voltage Temperature Compensation

Typically, the APD reverse bias voltage is designed with a compensatory positive temperature coefficient. This can be easily implemented via the CTRL pin of the LT3571—a less complex
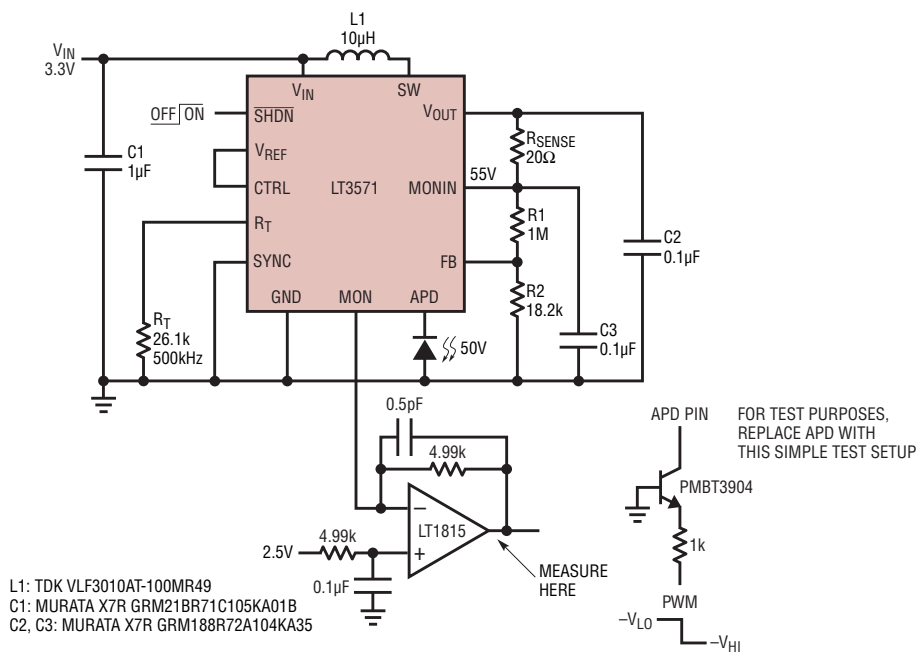

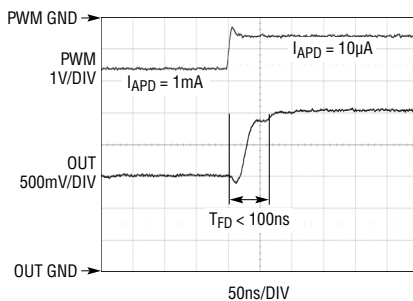**Figure 4. APD bias supply with ultrafast current monitor transient speed**

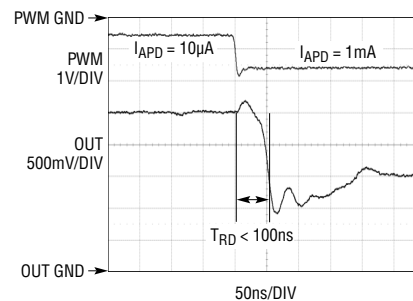**Figure 5. Transient response on input signal falling edge (1mA to 10μA)**

Figure 6. Transient response on input signal rising edge (10μA to 1mA)
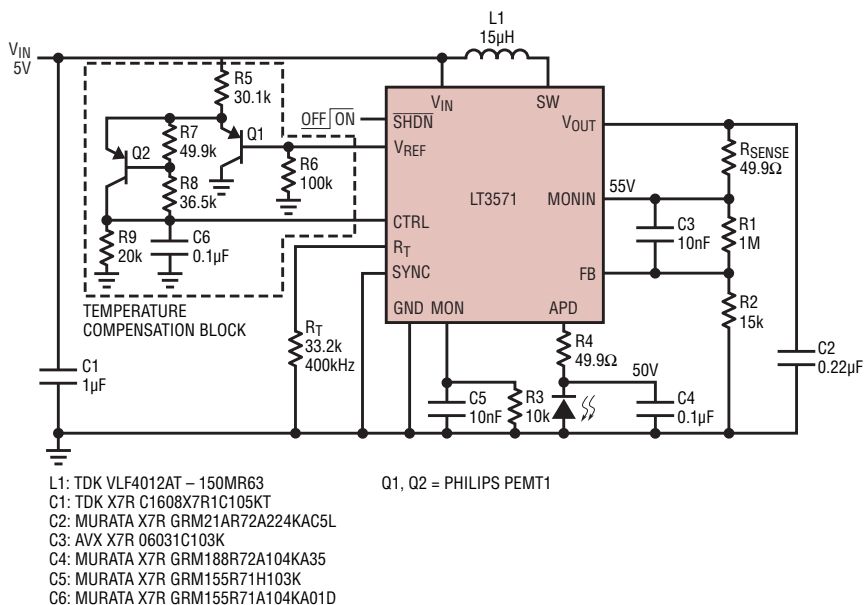

Figure 7. Temperature-compensated APD power supply

and expensive solution than typical microprocessor-controlled methods.

The simplest scheme uses a resistor divider from the V_{REF} pin to the CTRL pin, where the top resistor in the divider is an NTC (negative temperature coefficient) resistor. While simple, this method suffers from nonlinear temperature coefficient of the NTC resistor. A more precise method uses a transistor network as shown in Figure 7. The PTC (Positive Temperature Coefficient) of the CTRL pin voltage is
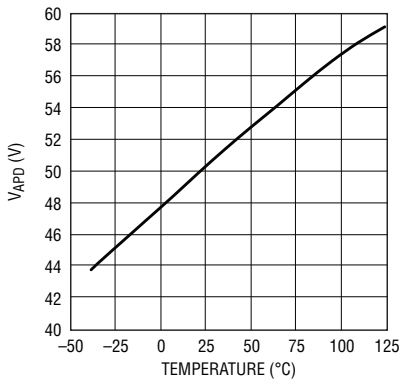

Figure 8. Temperature response of the circuit shown in Figure 7

realized by an emitter follower of Q1 and a V_{BE} multiplier of Q2.

Assuming:

$$V_{BE(Q1)} = V_{BE(Q2)} = V_{BE}$$

and

$$\frac{dV_{BE(Q1)}}{dT} = \frac{dV_{BE(Q2)}}{dT} = \frac{2mV}{°C}$$

then the CTRL pin voltage is

$$V_{CTRL} = V_{REF} - \frac{R8}{R7} V_{BE}$$

with

$$PTC = \frac{dV_{CTRL}}{dT} = \frac{R8}{R7} \cdot \frac{2mV}{°C}$$

Given V_{OUT} at room and dV_{OUT}/DT, the R1/R2 and R8/R7 can be calculated as follows

$$\frac{R8}{R7} = \frac{V_{REF}}{V_{BE} + \dfrac{2mV}{°C} \cdot \dfrac{V_{OUT}}{dV_{OUT}/dT}}$$

$$\frac{R1}{R2} = \frac{V_{BE} \cdot \dfrac{dV_{OUT}}{dT} + \dfrac{2mV}{°C} \cdot V_{OUT}}{\dfrac{2mV}{°C} \cdot V_{REF}} - 1$$

Resistors R5–R9 are selected to make I(Q1) = I(Q2) ≈ 10µA, and

$$\frac{dV_{BE(Q1)}}{dT} = \frac{dV_{BE(Q2)}}{dT} = \frac{2mV}{°C}$$

Simulation using LTspice always gives a good starting point. The circuit shown in Figure 7 is designed to have V_{APD} = 50V (V_{OUT} = 55V) at room and dV_{APD}/dT = 100mV/°C (dV_{OUT}/dT = 100mV/°C). The measured temperature response is shown in Figure 8, which is very close to the design target.

Conclusion

The LT3571 is a highly integrated, compact solution to APD bias supply design. It provides a useful feature set and the flexibility to meet a variety of challenging requirements, such as low noise, fast transient response speed, and temperature compensation. With a high level of integration and superior performance, the LT3571 is the natural choice for APD bias supply design. ⊿⊤

LT8410, continued from page 23

out of shutdown, the V_{REF} pin is first discharged for 70µs with a strong pull down current, and then charged with 10µA to 1.235V. This achieves soft start since the output is proportional to V_{REF}. Full soft-start cycles occur even with short $\overline{SHDN}$ low pulses since V_{REF} is discharged when the part is enabled.

In addition, the LT8410/-1 features a 2.5V to 16V input voltage range, up

to 40V output voltage and overvoltage protection for CAP and V_{OUT}.

Conclusion

The LT8410/-1 is a smart choice for applications which require low quiescent current and low input current. The ultralow quiescent current, combined with high value integrated feedback resistors, keeps the average input current very low, significantly

extending battery operating time. Low current limit internal switches (8mA for the LT8410-1, 25mA for the LT8410) make the part ideal for high impedance sources such as coin cell batteries. The LT8410/-1 is packed with features without compromising performance or ease of use and is available in a tiny 8-pin 2mm × 2mm package. ⊿⊤

LT3653/63, continued from page 21

of handling 60V transients. Figure 4 shows the circuit efficiency at multiple input voltages.

The current limit of the application is set to 1.2A, therefore, the power path components are sized to handle 1.2A maximum. To reduce the application footprint, the LT3663 includes internal compensation and a boost diode. The RUN pin, when low, puts the LT3663 into a low current shutdown mode.
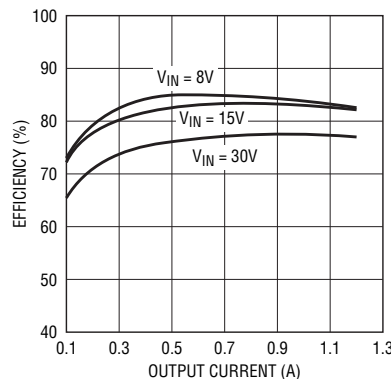

Figure 4. Efficiency of the circuit in Figure 3

Conclusion

The accurate programmable output current limit of the LT3653 and LT3663 eliminates localized heating from an output overload, reduces the maximum current requirements on the power components, and makes for a robust power supply solutions. ⊿⊤

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Don't Want to Hear It? Avoid the Audio Band with PWM LED Dimming at Frequencies Above 20kHz by Eric Young

Introduction

The requirements of LED drivers become more demanding as application designers exploit the unique characteristics of LEDs. Linear Technology offers a complete portfolio of LED drivers with the performance levels required to meet even the most challenging design requirements. One area where these LED drivers especially excel is in the performance and flexibility of their PWM dimming capabilities. LEDs can be turned on and off rapidly—it takes only nanoseconds to illuminate or extinguish the source. PWM dimming exploits this characteristic to achieve orders of magnitude dimming, even while maintaining a constant output spectrum over the entire dynamic light intensity range.

The broad field of available LED drivers narrows quite a bit when one considers PWM dimming at frequencies above 20kHz. Why 20kHz? Although most LED light designers worry about perceptible flicker at PWM frequencies below about 100Hz, in some applications the human eye is not the limiting factor; it is the human ear. The human ear perceives vibrations up to about 20kHz, which in some applications can become the important factor in determining PWM frequency. The versatile LT3755 and LT3756 are members of an elite group
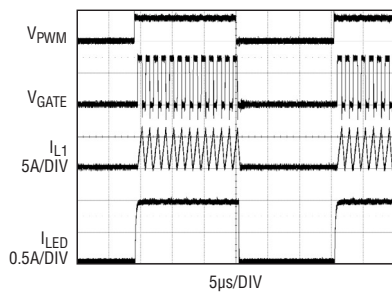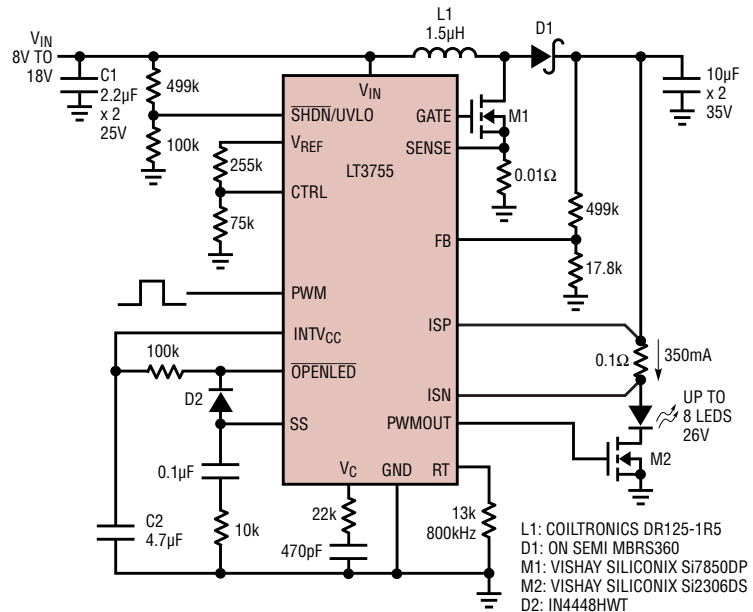

Figure 1. This 10W boost LED driver stays out of the audio band by achieving 50:1 PWM dimming at 20kHz. Lower PWM frequencies can result in an audible hum as ceramic capacitors vibrate.

Figure 2. DCM operation of the boost LED driver in Figure 1

of LED controllers that can support very high PWM dimming ratios, as much as 50:1, at 20kHz. These controllers support a variety of topologies, including buck mode, boost and buck-boost at various power levels.

High Performance PWM Dimming

The PWM dimming method is straightforward; the LED is driven by a tightly regulated current for a fixed interval in every PWM period. During the off-phase, the current in the LED is zero. During the on-phase, the current is carefully regulated. It is important that the "on" current is consistent, since an LED's output spectrum is a function of forward current. The duty cycle of the PWM signal corresponds to the dimming value.

Although the concept is simple, designing a controller that can achieve this at a high PWM frequency is any-

thing but simple. The rise and fall times of the pulsed current should be fast, less than 100ns. Generating a suitable PWM current pulse from an arbitrary input voltage can prove a challenge. This usually requires a high bandwidth DC/DC converter to regulate the current, a storage/filter capacitor across the LED to provide current during PWM on/off transitions, and a disconnect switch to ensure that the current waveform has sharp turn-on and off edges.

Hysteretic converters, while simple to use from the standpoint of closed loop stability, have problems. The slow LED current rise and fall times are one consequence of using a large value inductor to smooth the current through the LED because there is no output capacitor. And since the average current in the LED is related to the ripple current in the inductor, which is in turn sensitive to input volt-

age transients, the LED light output changes with input supply. In most cases, this method cannot provide acceptable PWM performance.

What determines PWM performance? The PWM interval or frequency is determined by the application, and there are several considerations to bear in mind. First, the human eye generally does not perceive flicker if the PWM frequency is greater than 120Hz, thus a lower bound on the interval is typically taken to be 8ms.

The achievable dimming ratio is a function of the minimum on- and off-times of the current pulse provided by the driver circuit. So an 8µs minimum pulse yields a 1000:1 dimming capability at 120Hz. The 20kHz audible requirement comes about because audible physical vibrations can be introduced to the PC board by the ceramic capacitors, and these caps are ubiquitous in high bandwidth converter circuits because of their low ESR, ruggedness, and long-term reliability. Ceramic capacitors physically change dimension (as well as value) with a change in applied voltage, and rapid voltage transients during the PWM transients cause rapid changes in dimensions that couple vibrations into the boards. If you ever noticed
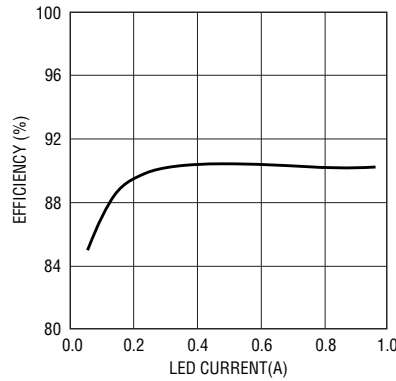

Figure 3. The efficiency of the boost LED driver in Figure 1 is greater than 90%.

an annoying buzz or hum next to a handheld device containing one of these circuits, then you have observed this effect.

The use of a disconnect switch in series with the LED greatly reduces the voltage transient and therefore the hum from the output capacitor. While good design techniques can greatly minimize audible noise for lower PWM frequencies, the elimination of audible emission is not assured so long as PWM frequency is below 20kHz. Many application designers don't want to tinker with acoustics, preferring instead quiet running circuits that do a reasonable job of PWM dimming. The LT3755 and LT3756 current-mode switching con-

trollers can be configured into several different converter circuits to provide a high bandwidth, well regulated output current that can be pulsed at intervals as short as 1µs.

Discontinuous Conduction Mode Is the Secret to Maximizing PWM Performance

The key to short on/off times is for the switching regulator to operate in discontinuous conduction mode (DCM). In this mode, the inductor current always starts from zero at the beginning of each switching period and the peak inductor current is determined by the load and adjusted through the switch duty cycle. In contrast, continuous conduction mode (CCM) maintains a relatively constant switch duty cycle and adjusts the average inductor current to meet the demands of the load.

DCM is superior for high performance PWM dimming because it delivers the required energy to the output in a single switching period. This allows the controller to bypass the typical minimum PWM period of 3-4 switching cycles to reach steady state, a familiar requirement of CCM. Operation in DCM places greater demands
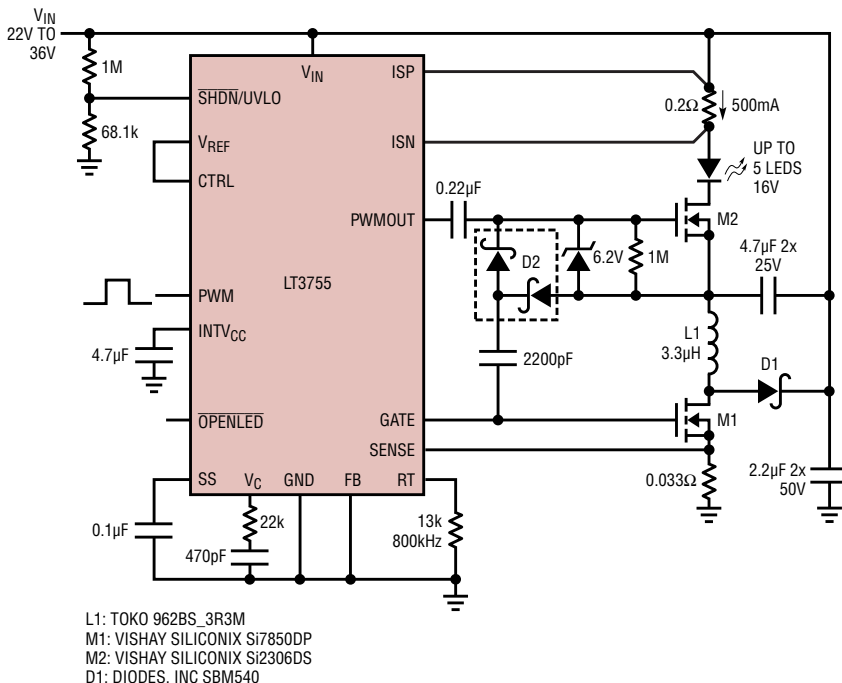

L1: TOKO 962BS_3R3M
M1: VISHAY SILICONIX Si7850DP
M2: VISHAY SILICONIX Si2306DS
D1: DIODES, INC SBM540

Figure 4. An 8W buck-mode LED driver with 50:1 PWM dimming at 20kHz and 90% efficiency
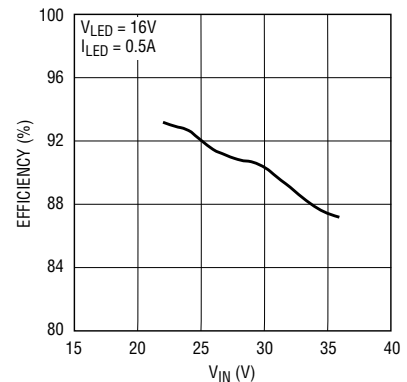
on switching components because the switching components see higher peak currents for a given load. Because of this, a controller is easier to use than a monolithic converter because its maximum switching current can be programmed to the needs of the application, without having to change the application's features.

Operating in DCM does come at a price when compared to CCM: efficiency, input supply range and analog dimming range all suffer some reduction. The ratio of maximum-to-minimum input supply range is slightly less than the ratio of the minimum PWM pulse width to minimum switch on-time. Likewise, provided the input supply is fixed, the maximum analog dimming ratio is the same ratio of minimum PWM pulse to minimum switch on-time. Nevertheless, the benefit of this technique is that minimum PWM period is four to five times shorter compared with continuous conduction mode. If the application calls for high PWM dimming ratio, DCM mode provides a sure path to achieve that objective. Three application circuits built with LT3755 and shown here demonstrate this technique.
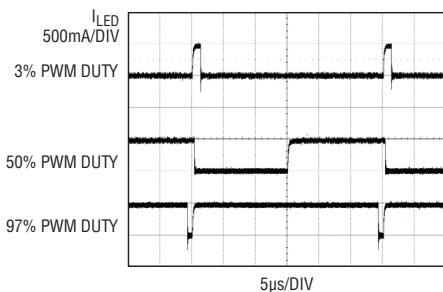
Figure 5. Three PWM dimming settings for the buck mode driver in Figure 4. Even at 33kHz there is no perceptible change in the LED current from minimum to maximum duty cycle.
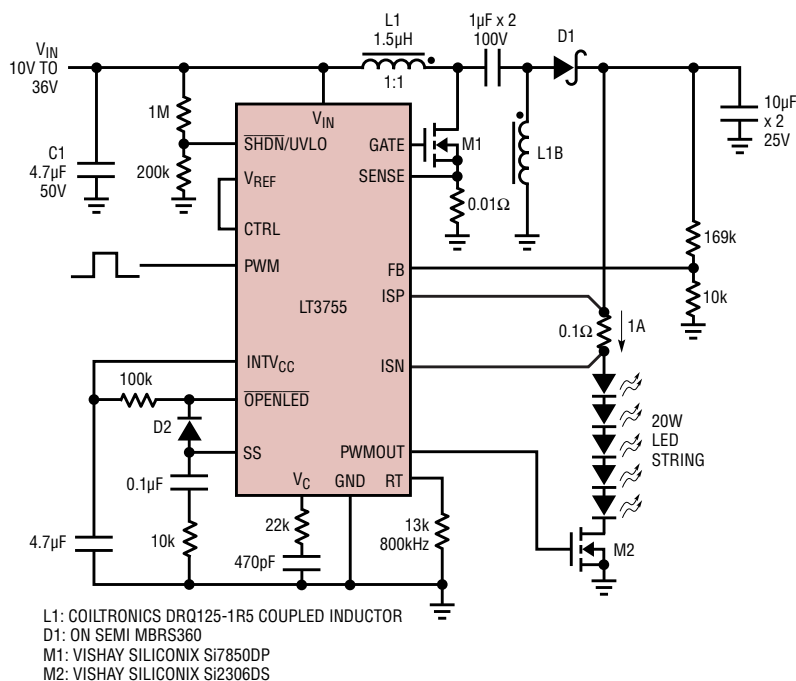
Figure 1 shows a 9W boost converter that regulates 26V of LEDs at a steady 350mA from a supply ranging between 8V and 18V. If the supply is fixed at 12V, the regulator operates at constant switching frequency for LED currents programmed by the CTRL pin between 125mA and 1A (2.4W to 27W). The minimum on-time is 1μs, as is the minimum off-time. The switching waveforms in Figure 2 show the operation at 50% duty cycle, 27V/1A load and 12V supply. Notice the fast rise and fall times of the LED current signal, even at 1A. At maximum load, the GATE pin is 7V for almost 1μs (same as the minimum pulse width)

and the inductor current reaches zero before the start of the each GATE pulse, a characteristic of DCM operation. Figure 3 shows the efficiency versus LED current at 12V input, which peaks at just over 90%.
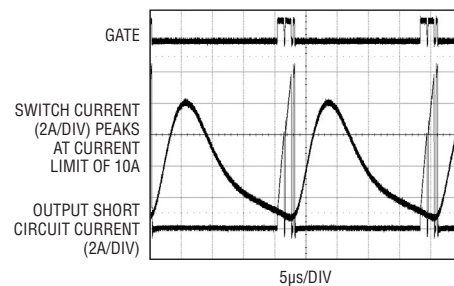
Figure 4 shows a buck-mode converter that regulates a 16V LED string at 500mA from a 22V to 36V supply. This circuit has an external charge-pump and level shift to drive the gate of an LED disconnect NMOS. This level shift provides much faster rise and fall times than the familiar resistor level shift driving a PMOS, and uses much less current. The scope trace in Figure 5 shows PWM dimming at several duty cycles—it is clear that the output LED current has no perceptible variation as pulse width is smoothly adjusted between the minimum on-time and the minimum off-time. The efficiency of this 8W circuit exceeds 90%.

Figure 6 shows a SEPIC converter driving a 1A, 20V LED string from a 12V-to-36V supply. In addition to providing step-up and step-down capability, this circuit is handy because it provides input-output isolation and built in protection from a short to GND on the output. The efficiency of this circuit exceeds 87%. The minimum

L1: COILTRONICS DRQ125-1R5 COUPLED INDUCTOR
D1: ON SEMI MBRS360
M1: VISHAY SILICONIX Si7850DP
M2: VISHAY SILICONIX Si2306DS

Figure 6. A 20W SEPIC LED Driver with 50:1 PWM dimming at 20kHz and output fault protection

Figure 7. The SEPIC converter in Figure 6 maintains control during an output fault to GND.

Eliminate EMI Worries with 2A, 15mm × 9mm × 2.82mm µModule Step-Down Regulator

by David Ng

Introduction

 "We failed EMI." Those three dreaded words strike fear into the hearts and minds of electronics design engineers. There are four words that are even worse: "We failed EMI again." The psyche of many a seasoned engineer is scarred with dark memories of long days and nights in an EMI lab, struggling with aluminum foil, copper tape, clamp-on filter beads and finger cuts to fix a design that just won't keep quiet. A big part of the problem is the necessary profusion of switching power supplies, which contribute significantly to the radiated system EMI.

The LTM8032 is a DC/DC switching step-down µModule regulator built specifically for low EMI. It is rated for up to $36V_{IN}$, $10V_{OUT}$ at 2A, and features adjustable frequency, synchronization, a power good status flag and soft-start. It is small, measuring only 15mm × 9mm × 2.82mm, integrating the inductor, power stage and controller in a ROHS e3-compliant molded LGA package.

10V/2A Supply Is EN55022 and CSIPR 22 Class B Compliant

Like most other µModule regulators, the LTM8032 is easy to use. As shown in Figure 1, all that is needed for a complete power design are the resistors to set the output voltage and operating frequency, and the input and output caps.

The LTM8032 is test-proven EN55022 and CSIPR 22 class B complaint, tested in an NRTL 5-meter chamber, set up as shown in the photo given in Figure 2. The LTM8032 is mounted on a circuit board with no bulk capacitance installed. The input and output capacitance are the minimum ceramic values specified in the data sheet for proper operation.
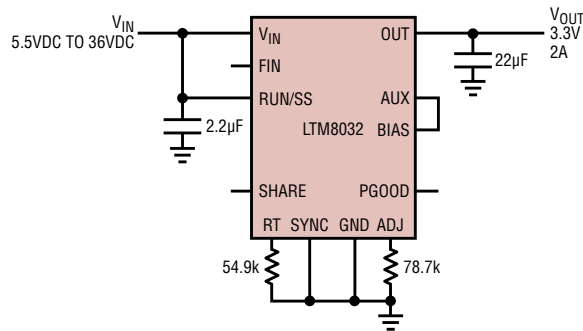

Figure 1. Just two resistors, input and output caps are needed to complete a power supply design with the LTM8032.

> *The LTM8032 is a DC/DC switching step-down µModule regulator built for low EMI. It is rated for up to $36V_{IN}$, $10V_{OUT}$ at 2A, integrating the inductor, power stage and controller in a ROHS e3-compliant molded LGA package.*

The assembled unit is placed atop an all-wood table. The all-wood construction ensures that the test setup does not shield or shadow noise emanating from the device under test (DUT). The power source, a linear lab grade power supply, is on the floor. The load for the LTM8032, with its heat sink, is also on the table top.

Before measuring the emissions from the LTM8032, a baseline measurement is taken to establish the
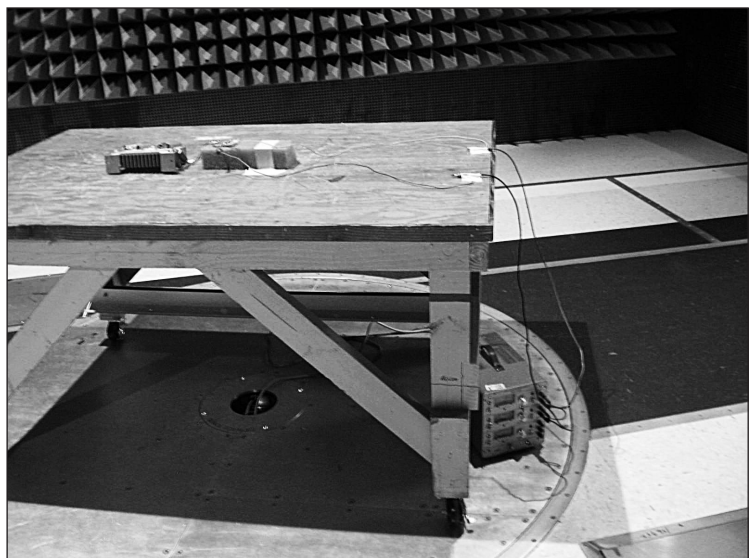
Figure 2. For EMI testing, the DUT is mounted on a circuit board and placed on a wooden table. The power source is on the floor.

Diode Turn-On Time Induced Failures in Switching Regulators
Never Has So Much Trouble Been Had by So Many with So Few Terminals

by Jim Williams and David Beebe

This article is excerpted from the Linear Technology Application Note AN122 with the same title.

Introduction

Most circuit designers are familiar with diode dynamic characteristics such as charge storage, voltage dependent capacitance and reverse recovery time. Less commonly acknowledged and manufacturer specified is diode forward turn-on time. This parameter describes the time required for a diode to turn on and clamp at its forward voltage drop. Historically, this extremely short time, units of nanoseconds, has been so small that user and vendor alike have essentially ignored it. It is rarely discussed and almost never specified. Recently, switching regulator clock rate and transition time have become faster, making diode turn-on time a critical issue. Increased clock rates are mandated to achieve smaller magnetics size; decreased transition times somewhat aid overall efficiency but are principally needed to minimize IC heat rise. At clock speeds beyond about 1MHz, transition time losses are the primary source of die heating.

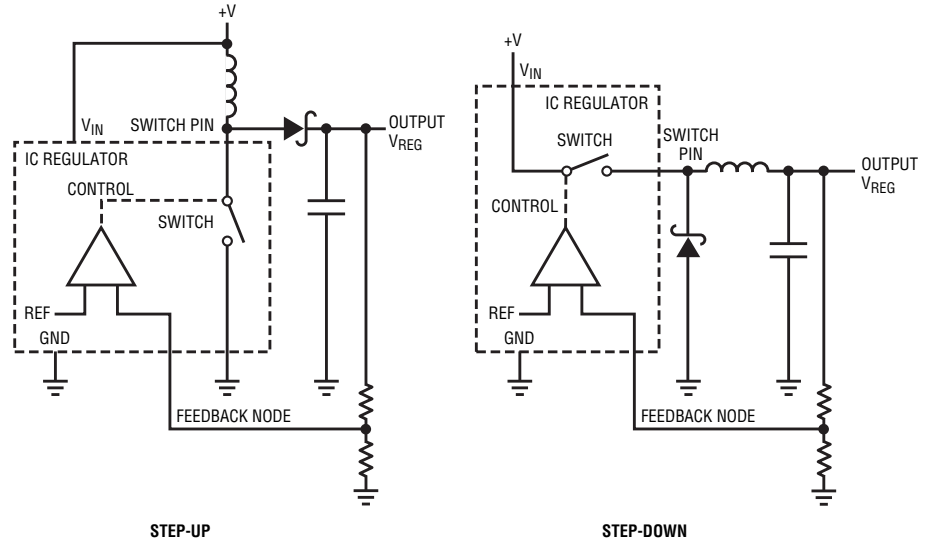A potential difficulty due to diode turn-on time is that the resultant

Figure 1. Typical voltage step-up/step-down converters. Assumption is diode clamps switch pin voltage excursion to safe limits.

transitory "overshoot" voltage across the diode, even when restricted to nanoseconds, can induce overvoltage stress, causing switching regulator IC failure. As such, careful testing is required to qualify a given diode for a particular application to insure reliability. This testing, which assumes low loss surrounding components and layout in the final application, measures turn-on overshoot voltage due to diode parasitics only. Improper

associated component selection and layout will contribute additional over-stress terms.

Diode Turn-On Time Perspectives

Figure 1 shows typical step-up and step-down voltage converters. In both cases, the assumption is that the diode clamps switch pin voltage excursions to safe limits. In the step-up case, this limit is defined by the switch pins
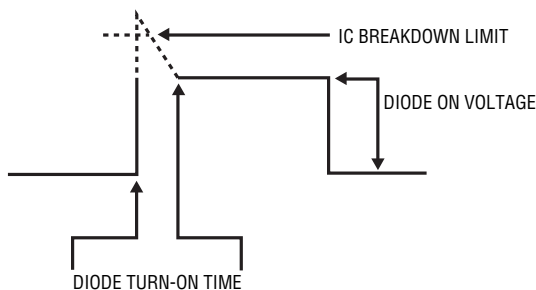

Figure 2. Diode forward turn-on time permits transient excursion above nominal diode clamp voltage, potentially exceeding IC breakdown limit.
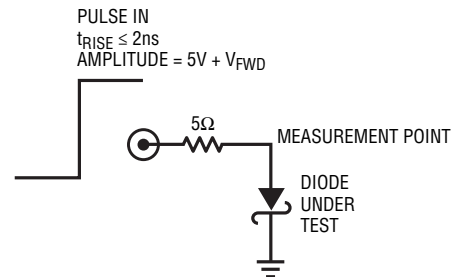

Figure 3. Conceptual method tests diode turn-on time at 1A. Input step must have exceptionally fast, high fidelity transition.
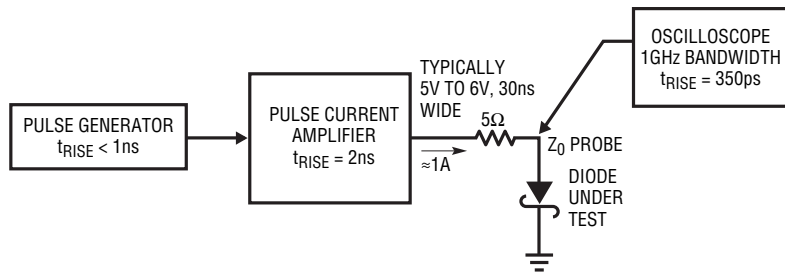
AN122 F04

Figure 4. Detailed measurement scheme indicates necessary performance parameters for various elements. Subnanosecond rise time pulse generator, 1A, 2ns rise time amplifier and 1GHz oscilloscope are required.
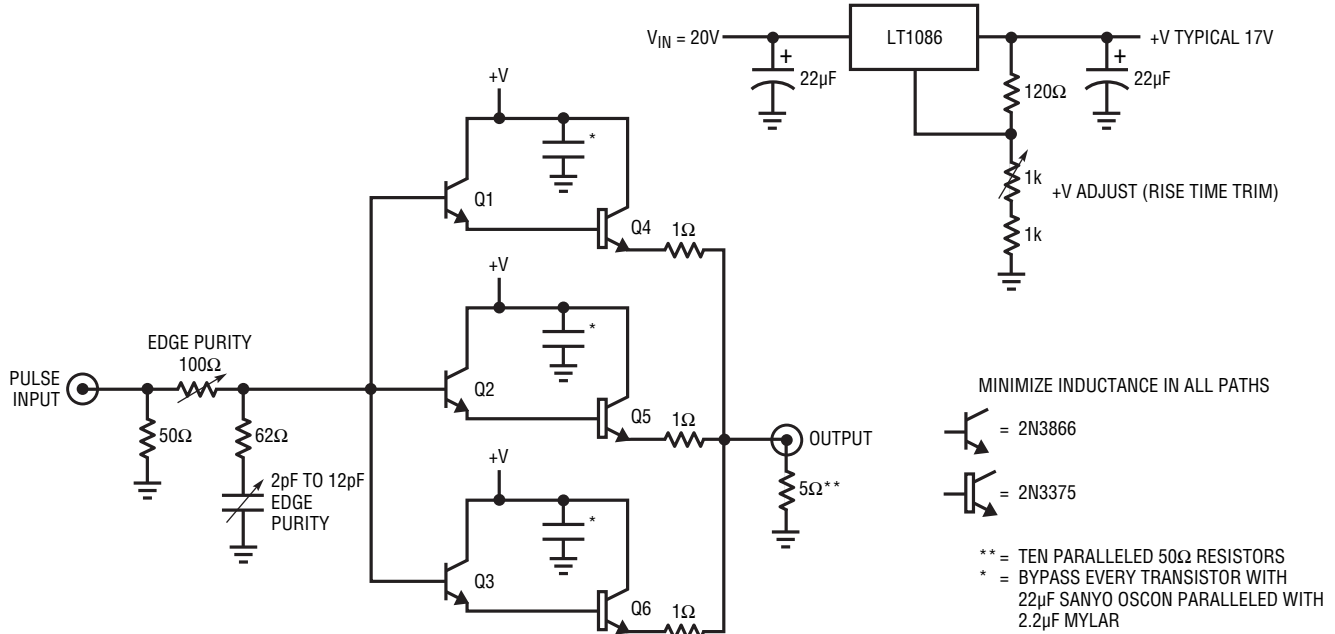

AN122 F05

Figure 5. Pulse amplifier includes paralleled, darlington driven RF transistor output stage. Collector voltage adjustment ("rise time trim") peaks Q4 to Q6 F_T, input RC network optimizes output pulse purity. Low inductance layout is mandatory.

maximum allowable forward voltage. The step-down case limit is set by the switch pins maximum allowable reverse voltage.

Figure 2 indicates the diode requires a finite length of time to clamp at its forward voltage. This forward turn-on time permits transient excursions above the nominal diode clamp voltage, potentially exceeding the IC's breakdown limit. The turn-on time is typically measured in nanoseconds, making observation difficult. A further complication is that the turn-on overshoot occurs at the amplitude extreme of a pulse waveform, precluding high resolution amplitude measurement. These factors must be considered when designing a diode turn-on test method.

Figure 3 shows a conceptual method for testing diode turn-on time. Here, the test is performed at 1A although other currents could be used. A pulse steps 1A into the diode under test via the 5Ω resistor. Turn-on time voltage excursion is measured directly at the diode under test. The figure
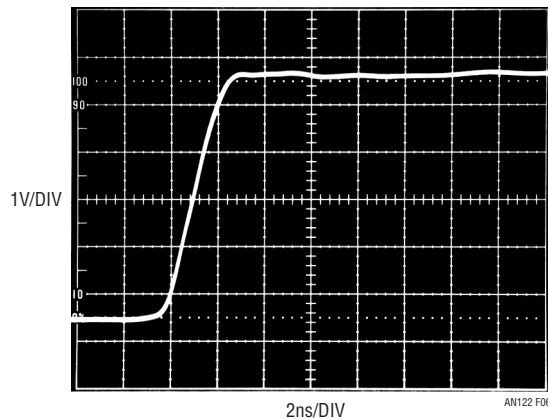

AN122 F06

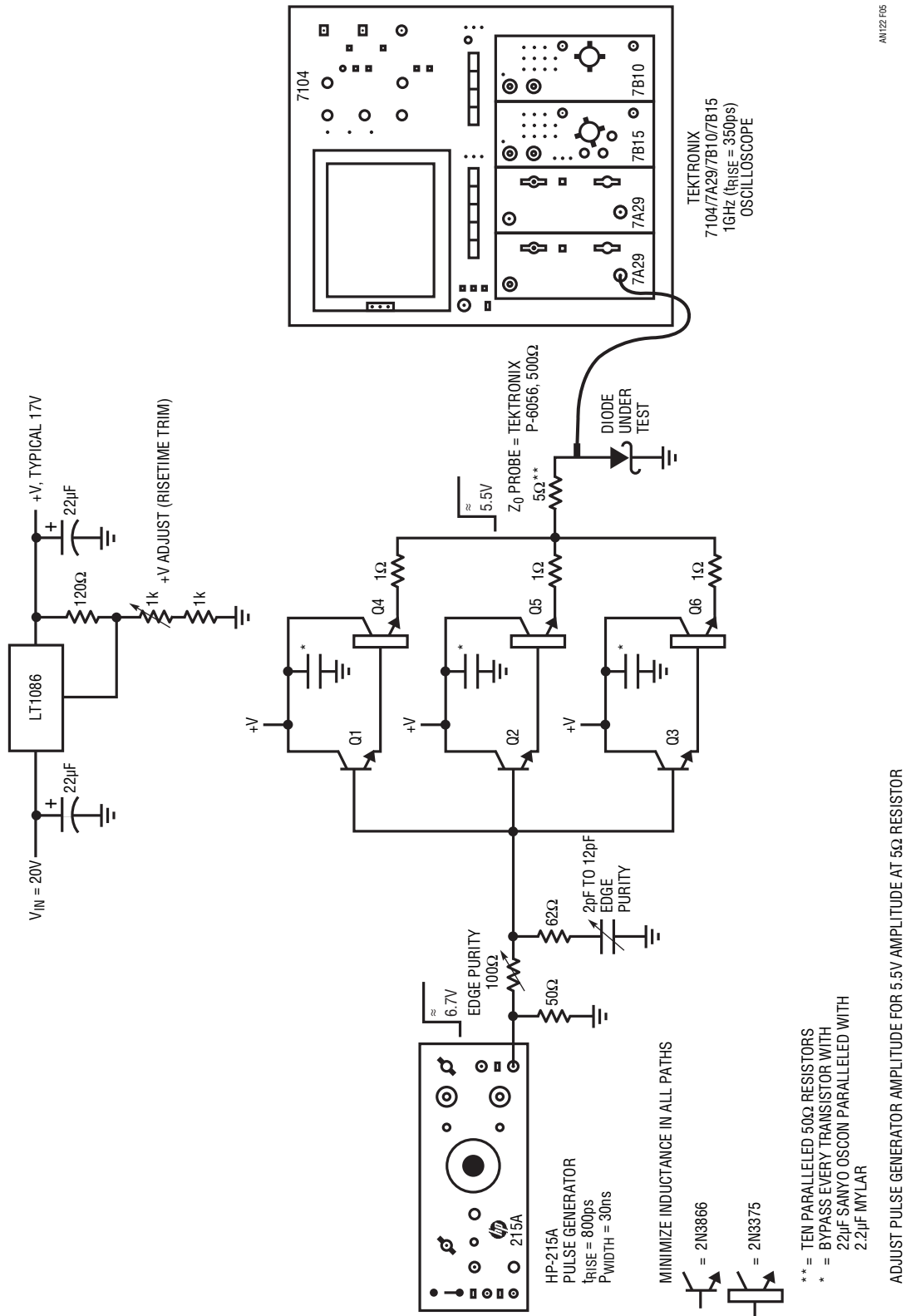**Figure 6. Pulse amplifier output into 5Ω. Rise time is 2ns with minimal pulse-top aberrations.**

Figure 7. Complete diode forward turn-on time measurement arrangement includes subnanosecond rise time pulse generator, pulse amplifier, Z0 probe and 1GHz oscilloscope.
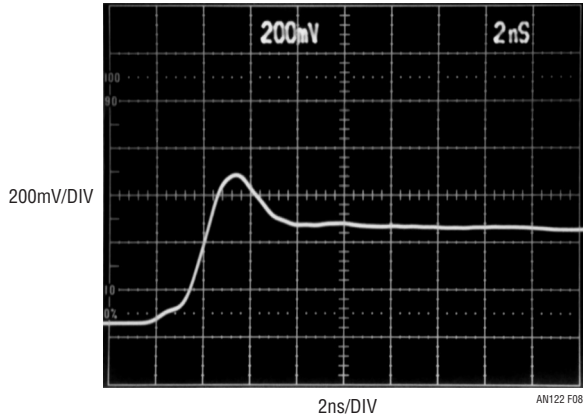
Figure 8. "Diode Number 1" overshoots steady state forward voltage for ≈3.6ns, peaking 200mV.
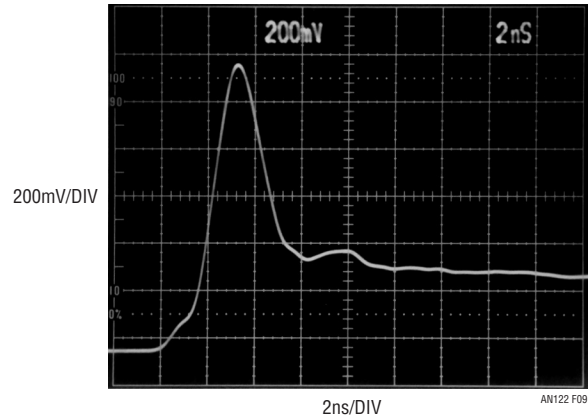
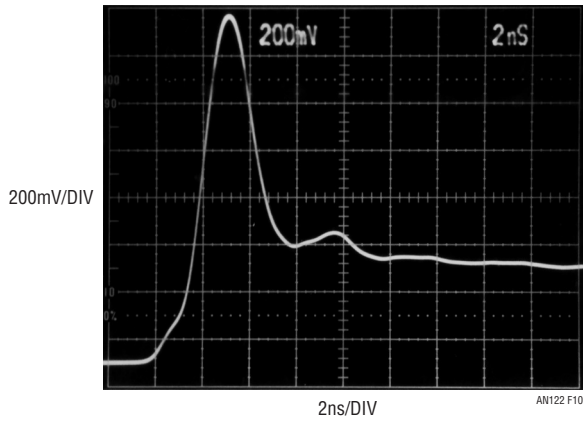Figure 9. "Diode Number 2" peaks ≈750mV before settling in 6ns... > 2x steady state forward voltage.

Figure 10. "Diode Number 3" peaks 1V above nominal 400mV VFWD, a 2.5x error.

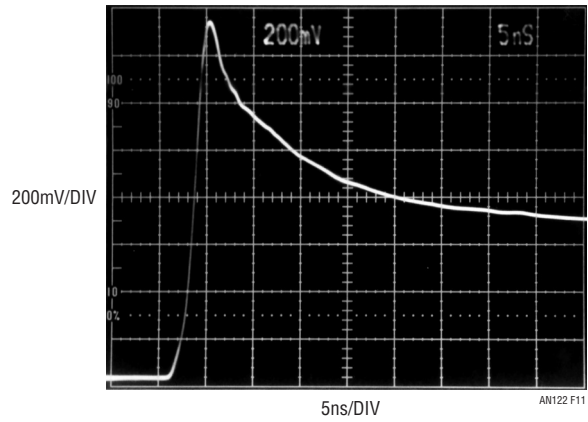Figure 11. "Diode Number 4" peaks ≈750mV with lengthy (note horizontal 2.5x scale change) tailing towards VFWD value.
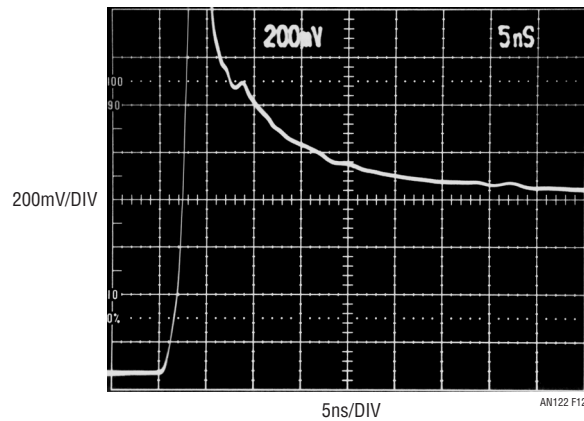

Figure 12. "Diode Number 5" peaks offscale with extended tailing (note horizontal slower scale compared to Figures 8 thru 10).

is deceptively simple in appearance. In particular, the current step must have an exceptionally fast, high-fidelity transition and faithful turn-on time determination requires substantial measurement bandwidth.

Detailed Measurement Scheme

A more detailed measurement scheme appears in Figure 4. Necessary performance parameters for various elements are called out. A subnanosecond rise time pulse generator, 1A, 2ns rise time amplifier and a 1GHz oscilloscope are required. These specifications represent realistic operating conditions; other currents and rise times can be selected by altering appropriate parameters.

The pulse amplifier necessitates careful attention to circuit configuration and layout. Figure 5 shows the amplifier includes a paralleled, Darlington driven RF transistor output stage. The collector voltage adjustment ("rise time trim") peaks Q4 to Q6 FT; an input RC network optimizes output pulse purity by slightly retarding input pulse rise time to within amplifier passband. Paralleling allows Q4 to Q6

to operate at favorable individual currents, maintaining bandwidth. When the (mildly interactive) edge purity and rise time trims are optimized, Figure 6 indicates the amplifier produces a transcendently clean 2ns rise time output pulse devoid of ringing, alien components or post-transition excursions. Such performance makes diode turn-on time testing practical.[1]

Figure 7 depicts the complete diode forward turn-on time measurement arrangement. The pulse amplifier, driven by a sub-nanosecond pulse generator, drives the diode under test. A Z0 probe monitors the measurement point and feeds a 1GHz oscilloscope.[2,3,4]

Diode Testing and Interpreting Results

The measurement test fixture, properly equipped and constructed, permits diode turn-on time testing with excellent time and amplitude resolution.[5] Figures 8 through 12 show results for five different diodes from various manufacturers. Figure 8 (Diode Number 1) overshoots steady state forward voltage for 3.6ns, peaking 200mV. This is the best performance of the five. Figures 9 through 12 show

increasing turn-on amplitude and time which are detailed in the figure captions. In the worst cases, turn-on amplitudes exceed nominal clamp voltage by more than 1V while turn-on times extend for tens of nanoseconds. Figure 12 culminates this unfortunate parade with huge time and amplitude errors. Such errant excursions can and will cause IC regulator breakdown and failure. The lesson here is clear. Diode turn-on time must be characterized and measured in any given application to insure reliability. *LT*

Notes

[1] An alternate pulse generation approach appears in Linear Technology *Application Note 122*, Appendix F, "Another Way to Do It."

[2] Z0 probes are described in Linear Technology *Application Note 122* Appendix C, "About Z0 Probes." See also References 27 thru 34.

[3] The subnanosecond pulse generator requirement is not trivial. See Linear Technology *Application Note 122* Appendix B, "Subnanosecond Rise Time Pulse Generators For The Rich and Poor."

[4] See Linear Linear Technology *Application Note 122* Appendix E, "Connections, Cables, Adapters, Attenuators, Probes and Picoseconds" for relevant commentary.

[5] See Linear Technology *Application Note 122* Appendix A, "How Much Bandwidth is Enough?" for discussion on determining necessary measurement bandwidth.

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amount of ambient noise in the room. Figure 3 shows the noise spectrum in the chamber without any devices running. This can be used to determine the actual noise produced by the DUT.

Figure 4 shows the worst case LTM8032 emissions plot, which occurs at maximum power out, 10V at

2A, from the maximum input voltage, 36V. There are two traces in the plot, one for the vertical and horizontal orientations of the test lab's receiver antenna. As shown in the figure, the LTM8032 easily meets the CISPR 22 class B limits, with 20db of margin for most of the frequency spectrum, with either antenna orientation.

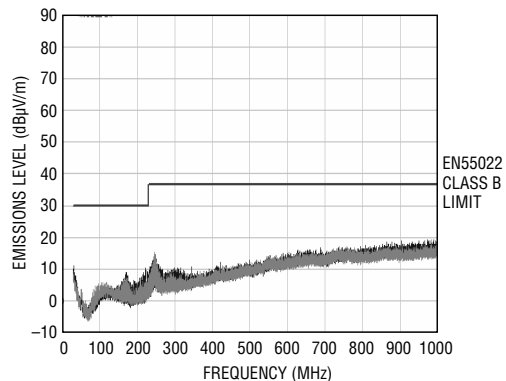## Conclusion

The LTM8032 switching step-down regulator is both easy to use and quiet, meeting the radiated emissions requirements of CISPR22 and EN55022 class B by a wide margin. *LT*

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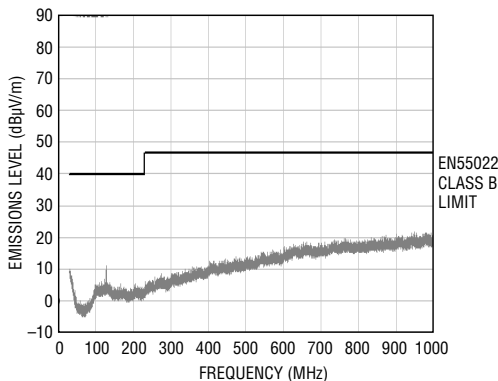
Figure 3. The baseline measurement of ambient noise in the 5-meter chamber (no devices operating)

Figure 4. The LTM8032 emissions for 20W out, 36V$_{IN}$

µModule Regulator Fits a (Nearly) Complete Buck-Boost Solution in 15mm × 15mm × 2.8mm for 4.5V–36V V_{IN} to 0.8V–34V V_{OUT}

by Judy Sun, Sam Young and Henry Zhang

Introduction

Linear Technology offers a number of high efficiency synchronous 4-switch buck-boost DC/DC converter solutions for applications where V_{OUT} falls within the range of V_{IN}. The LTM4605, LTM4607 and LTM4609 µModule regulators are nearly self-contained buck-boost solutions that share pin-compatible 15mm × 15mm × 2.8mm packages. The package includes the controller, four power FETs and a number of other discrete components. Only an external inductor, a sensing resistor, a voltage setting resistor and a few input and output capacitors are needed to complete a high efficiency buck-boost converter.

Table 1 shows the input voltage, output voltage and current specifications of these three buck-boost µModule regulators. The LTM4609 is the latest addition to this family. It satisfies the needs of high output voltage applications with an output range of 0.8V–34V.

## High Performance with Minimum Component Count

As with all Linear Technology µModule regulators, the LTM4609 requires only
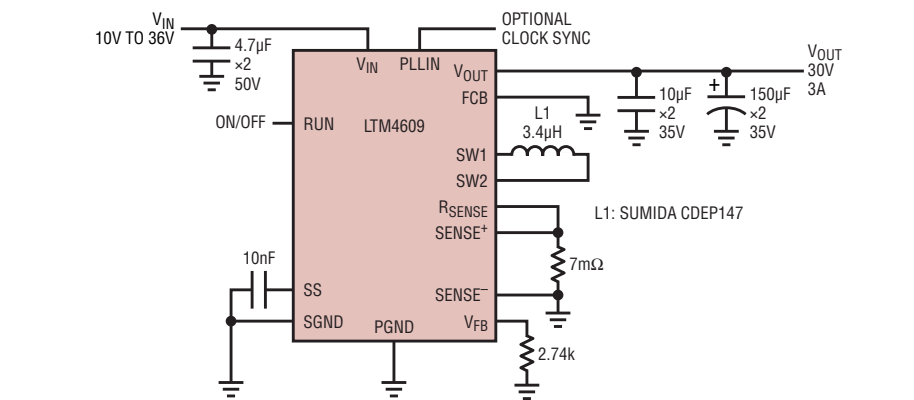
Figure 1. Just a few components form a complete 10V to 36V input, 30V/3A output converter using the LTM4609.

a few external components to complete a wide input range buck-boost converter. Figure 1 shows a 10V to 36V input, 30V output design. The output current capability is 3A at 10V V_{IN}, and 8A with 36V input.

Figure 2 shows the efficiency of this converter, up to 98% in buck mode and 95% in boost mode. The low profile LGA package features low thermal resistance from junction to pin, thus maintaining an acceptable junction temperature even at high output power. The LTM4609's high
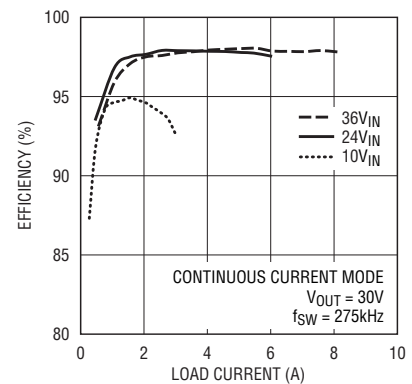
Figure 2. Efficiency of the 30V buck-boost converter
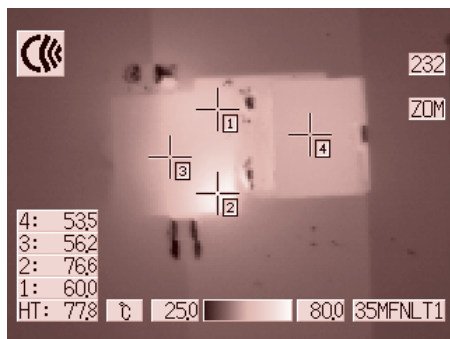
V_{IN} = 36V, V_{OUT} = 30V, I_{OUT} = 8A
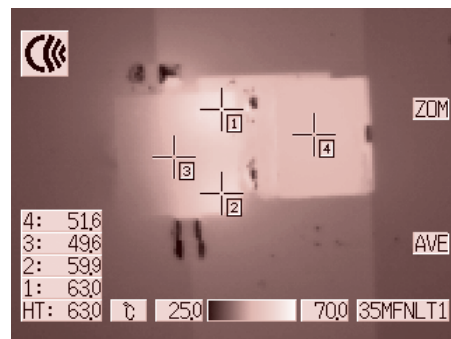
V_{IN} = 24V, V_{OUT} = 30V, I_{OUT} = 6A
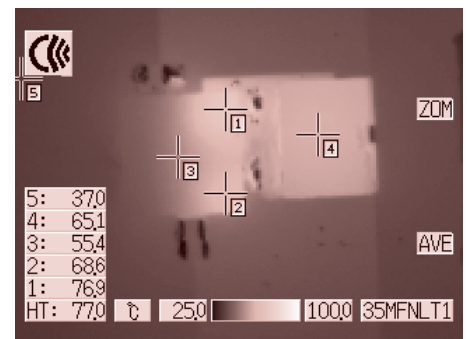
V_{IN} = 12V, V_{OUT} = 30V, I_{OUT} = 3A

Figure 3. Thermal-graph taken with the LTM4609 running at different input voltages. The LTM4609 is on the left, the inductor (Sumida CDEP147) is on the right. No heat sink or forced air flow. Ambient temperature = 25°C.

efficiency combined with its excellent thermal management capability enables it to deliver up to 240W output power without a heat sink or forced airflow. Figure 3 shows the thermal-graphs taken with three different input voltages and loads at 25°C ambient temperature. With 240W output and 36V input, the maximum temperature rise of the LTM4609 is only 52.8°C.

Input Ripple Reduction

One way to improve efficiency in a switching DC/DC converter is to minimize the turn-on and turn-off times of the MOSFET—shorter transitions correspond to lower switch losses. However, fast transitions also lead to high frequency switching n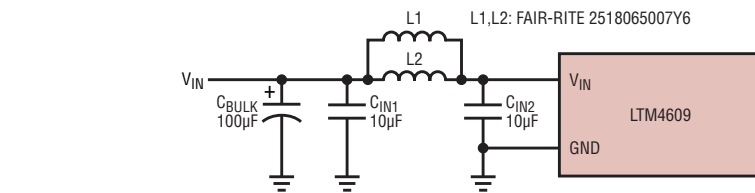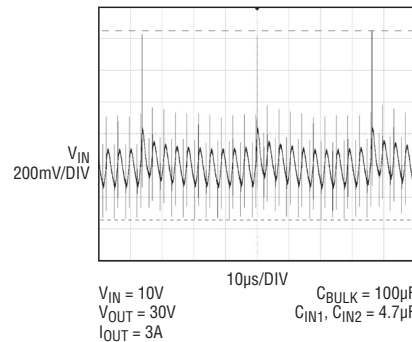oise, which can pollute the input power source. For the applications where the input voltage ripple must be limited, a simple LC π filter can be inserted at the input side to attenuate the high frequency input voltage noise. Figure 4 shows the LTM4609 with an input π filter. The filter includes two 10µF low ESR ceramic capacitors and two very small magnetic beads. For lower output power applications, only one magnetic bead is necessary.
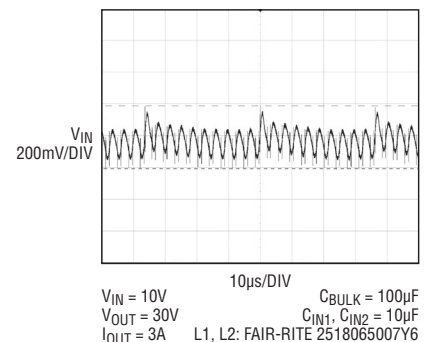
Figure 5 shows the input ripple reduction with the π filter. Figure 5a shows the input ripple with 100µF aluminum electrolytic plus 2 × 4.7µF

Figure 4. The LTM4609 µModule regulator with an input π filter.

V_IN = 10V
V_OUT = 30V
I_OUT = 3A
C_BULK = 100µF
C_IN1, C_IN2 = 4.7µF

5a. Input voltage waveform without the input π filter shown in Figure 4

V_IN = 10V
V_OUT = 30V
I_OUT = 3A
C_BULK = 100µF
C_IN1, C_IN2 = 10µF
L1, L2: FAIR-RITE 2518065007Y6

5b. Input voltage waveform with input π filter as shown in Figure 4

Figure 5. The input π filter shown in Figure 4 effectively reduces the input voltage spike caused by switching action of the MOSFETs.

ceramic input capacitors. Figure 5b shows the input ripple with the filter shown in Figure 4. Both waveforms are measured across the 100µF aluminum capacitor. A 67% reduction in input ripple is obtained with the input π filter, which requires only two small additional magnetic beads.

Conclusion

Buck-boost µModule regulators are easy-to-use, high performance solutions for applications where a regulated output voltage sits within the range of the input voltage. The 15mm × 15mm × 2.8mm LTM4609 widens the input/output voltage range of the pin compatible LTM4605 and LTM4607. The advanced package technology, as well as the high efficiency design of the LTM4609, allows it to deliver up to 240W of output power without heat sinks or forced airflow. For applications that require low input voltage ripple, a simple π filter can be added by inserting one or two small magnetic beads to significantly reduce the high frequency input noise. *LT*

Table 1. Specification comparison of the LTM4605, LTM4607 and LTM4609

	LTM4605	LTM4607	LTM4609
V_IN	4.5V ~ 20V	4.5V ~ 36V	4.5V ~ 36V
V_OUT	0.8V ~ 16V	0.8V ~ 24V	0.8V ~ 34V
I_OUT	5A (12A in buck mode)	5A (10A in buck mode)	4A (10A in buck mode)
Package	15mm × 15mm × 2.8mm LGA		

PWM on- and off-times are 1µs as with the other circuits. Figure 7 shows the waveforms during a short circuit fault on the output. The input current remains in control as the switch current ramps up to the set limit of 10A, then skips the next few cycles while the current sensed by the LED

resistor ramps down to 1.5A. This faulted mode of circuit operation can continue indefinitely without damage to the components.

Conclusion

The LT3755 and LT3756 offer unparalleled performance for an LED

controller generating PWM pulse widths as narrow as 1µs, which enables 50:1 PWM dimming at frequencies above the audible range. Other features include open LED protection, an open LED status indicator, and programmability of the LED current via an analog input. *LT*

New Device Cameos

Micropower Low Noise Boost Converter with Output Disconnect

The LT3495/LT3495B/LT3495-1/LT3495B-1 is a low noise boost converter with integrated power switch, feedback resistor and output disconnect circuitry. The part controls power delivery by varying both the peak inductor current and switch off-time. This new control scheme results in low output voltage ripple as well as high efficiency over a wide load range.

For the LT3495/LT3495-1, the off-time of the switch is not allowed to exceed a fixed level, guaranteeing the switching frequency stays above the audio band for the entire load range. This feature is disabled for the LT3495B/LT3495B-1, which leads to higher efficiency at light load. The difference between the LT3495/LT3495B and LT3495-1/LT3495B-1 is the level of the switch current limit. The LT3495/LT3495B has a typical peak current limit of 650mA while the LT3495-1/LT3495B-1 has a typical peak current limit of 350mA.

The LT3495 series has an output disconnect PMOS that blocks the load from the input during shutdown. During normal operation, the maximum current through this PMOS is limited by circuitry inside the chip, which helps the chip survive output shorts.

The input voltage range of the LT3495 series is a wide 2.5V to 16V and the output voltage can be up to 40V. In addition, the part is well compensated internally, and can be stable with very small ceramic output capacitors.

Other features include low quiescent current (60µA in active mode and 0.1µA in shutdown mode), integrated output dimming, maximum switching on-time and undervoltage lockout. Combining the small ceramic capacitors and space saving 10-pin 3mm × 2mm DFN packages, the LT3495 enables compact solutions for many applications.

Dual 550mA, 1MHz Synchronous Boost Regulator with Output Disconnect in a 3mm × 3mm DFN

The LTC3535 is a dual-channel 1MHz, current mode synchronous boost DC/DC converter with integrated output disconnect and soft-start. The LTC3535's internal 550mA switches deliver output voltages as high as 5.25V from an input voltage range of 0.7V at start-up/0.5V when running to 5V, making it ideal for single- or multicell alkaline/NiMH as well as Li-ion/polymer applications. Each of the LTC3535's channels has it own power input and is completely independent, offering maximum design flexibility. For example, one channel can deliver up to 50mA of continuous output current at 3.3V while the other channel delivers up to 100mA at 1.8V to power a microcontroller from a single alkaline cell. The 1MHz switching frequency minimizes external component sizes while providing up to 94% efficiency. Combined with a compact 3mm × 3mm DFN-12 package, the LTC3535 dual channel boost provides the tiny and efficient solution footprint required in handheld applications.

Burst Mode® operation lowers quiescent current to only 18µA (both channels), providing extended battery run time in handheld applications. The LTC3535 is an ideal part for handheld dual boost applications where small solution size and maximum battery run time are defining factors.

Ideal Diode Equipped High Power Battery Charger Handles All Chemistries

The LTC4012, LTC4012-1, LTC4012-2 and LTC4012-3 are a family of high power buck battery chargers all in a 20-lead 4mm × 4mm QFN package. Compared to the LTC4009 family of chargers, the 4012 family adds Ideal Diode™ input reverse current input protection and extends the high efficiency to higher current levels. Combined with just a few external components and external termination control, the LTC4012 family facilitates construction of chargers capable of delivering up to 4A to batteries with output power levels approaching 66W in a very small footprint.

The LTC4012 family builds upon the proven quasi-constant frequency, constant off-time PWM buck control architecture as found in Linear Technology's LTC4008. This unique buck topology provides continuous switching with synchronous rectification even with no load current, critical to preventing audible noise in constant voltage charge termination applications. However, LTC4012 family uses switching NFETs along with an adaptive gate drive to avoid overlap conduction losses. The higher 550kHz switching frequency reduces both the inductor size and output capacitance requirements while offering efficiencies up to 95% or more. If the duty cycle goes below 20% or above 80%, the LTC4012 lowers the switching frequency to avoid pulse skipping that might otherwise begin to occur at 550kHz. Under low dropout conditions requiring high duty cycle operation, the internal watchdog timer prevents the LTC4012 from switching below 25kHz, achieving a maximum duty cycle of 98% without producing audible noise. There is also an input current monitor function that prevents input power overload when the input power is shared with a load.

There are four versions of the LTC4012. The LTC4012 and LTC4012-3 offer a user programmable voltage set point using an external resistor divider allowing for multi-chemistry support. The Li-ion optimized LTC4012-1 and LTC4012-2 support one to four series cells via pin selection. The LTC4012-1 provides 4.1V/cell charging while the LTC4012-2 produces 4.2V/cell. Output voltage accuracy is typically ±0.5% and a maximum of ±0.8% over temperature. These ICs contain a switch that in shutdown removes voltage divider current drained from the

battery whether external or internal. Programming the charge current only requires a single external resistor.

The fault management system of the LTC4012 family suspends charging immediately for various conditions. First is battery overvoltage protection, which can occur with the sudden loss of battery load during bulk charge. Second, each IC features internal over-temperature protection to prevent silicon damage during elevated thermal operation.

The LTC4012 family has a logic-level shutdown control input and three open-drain status outputs. First is an input current limit (ICL) status flag to tell the system when V_{IN} is running at over 95% of its current capacity. The input current limit accuracy is typically ±3% and a maximum of ±4% over the full operating temperature range. Next is the AC present status, which indicates when V_{IN} is within a valid range for charging under all modes of operation. The last is a charge status output can indicate bulk or C/10 charge states. The control input and status outputs of the LTC4012, along with the analog current monitor output, can be used by the host system to perform necessary preconditioning, charge termination and safety timing functions.

4MHz Synchronous Step-Down DC/DC Converter Delivers up to 1.25A from a 3mm × 3mm DFN

The LTC3565 is a high efficiency synchronous step-down regulator that can deliver up to 1.25A of continuous output current from a 3mm × 3mm DFN (or MSOP-10E) package. Using a constant frequency of (up to 4MHz) and current mode architecture, the LTC3565 operates from an input voltage range of 2.5V to 5.5V making it ideal for single cell Li-Ion, or multicell Alkaline/NiCad/NiMH applications. It can generate output voltages as low as 0.6V, enabling it to power the latest generation of low voltage DSPs and microcontrollers. An independent RUN pin enables simple turn-on and shutdown. Its switching frequency is user programmable from 400kHz to 4MHz, enabling the designer to optimize efficiency while avoiding critical noise-sensitive frequency bands. The combination of its 3mm × 3mm DFN-10 (or MSOP-10) package and high switching frequency keeps external inductors and capacitors small, providing a very compact, thermally efficient footprint.

The LTC3565 uses internal switches with an $R_{DS(ON)}$ of only 0.13Ω (N-Channel lower FET) and 0.15Ω (P-Channel upper FET) to deliver efficiencies as high as 95%. It also utilizes low dropout 100% duty cycle operation to allow output voltages equal to V_{IN}, further extending battery run time. The LTC3565 utilizes Automatic Low Ripple (< 25mV$_{P–P}$) Burst Mode® operation to offer only 40µA no load quiescent current. If the application is noise sensitive, Burst Mode operation can be disabled using a lower noise pulse-skipping mode, which still offers only 330µA of quiescent current. The LTC3565 can be synchronized to an external clock throughout its entire frequency range. Other features include ±2% output voltage accuracy and over-temperature protection. *LT*


LT3751 controller, and the optocoupler on the feedback resistor divider. The auxiliary windings provide the desired galvanic isolation boundary while maintaining an isolated feedback path from the output node to the LT3751 FB pin. Figures 12 and 13 show the regulator's performance.

The fully isolated, high voltage input/output regulator yields over 90% efficiency. Load regulation is excellent as shown in Figure 13b, due mainly to the added gain of the optocoupler circuit.

Conclusion

The ability to run from any input supply voltage ranging from 4.75V to greater than 400V and the abundance of safety features make the LT3751 an excellent choice for high voltage capacitor chargers or high voltage regulated power supplies. In fact, the LT3751 is, for now, the only

a. Efficiency

b. Load regulation
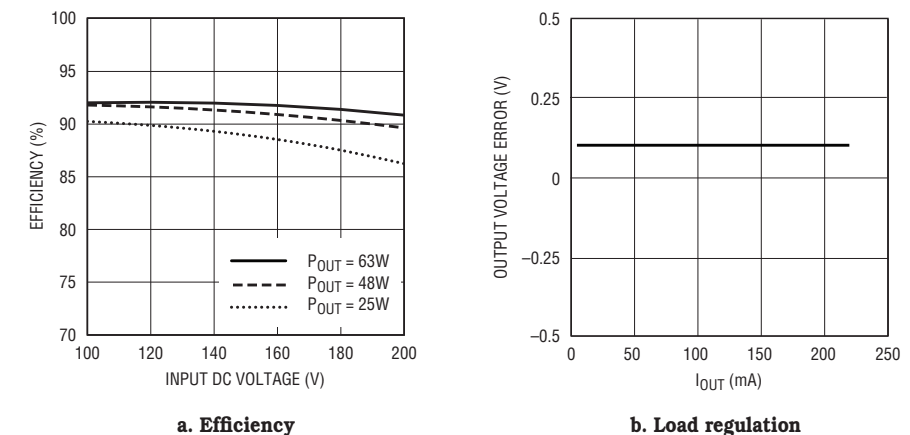
Figure 13. Fully isolated, high voltage regulator performance

boundary-mode capacitor charger controller that can accurately operate from extremely high input voltages. The LT3751 simplifies design by integrating many functions that—due to cost and board real-estate—would otherwise not be realizable. Although several designs are shown here, the LT3751 includes many more features than we can show in one article. We recommended consulting the data sheet or calling the Linear Technology applications engineering department for more in-depth coverage of all available features. *LT*

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