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## Virtual Remote Sensing Improves Load Regulation by Compensating for Wiring Drops Without Remote Sense Lines

Tom Hack and Robert Dobkin

Accurately regulating a voltage at a load can be difficult when there are significant voltage drops between the power supply and the load. Even if a regulator produces a perfectly regulated voltage at its own output, variations in load current affect the IR drop along the wiring, resulting in significant voltage fluctuations at the load (see Figure 1).

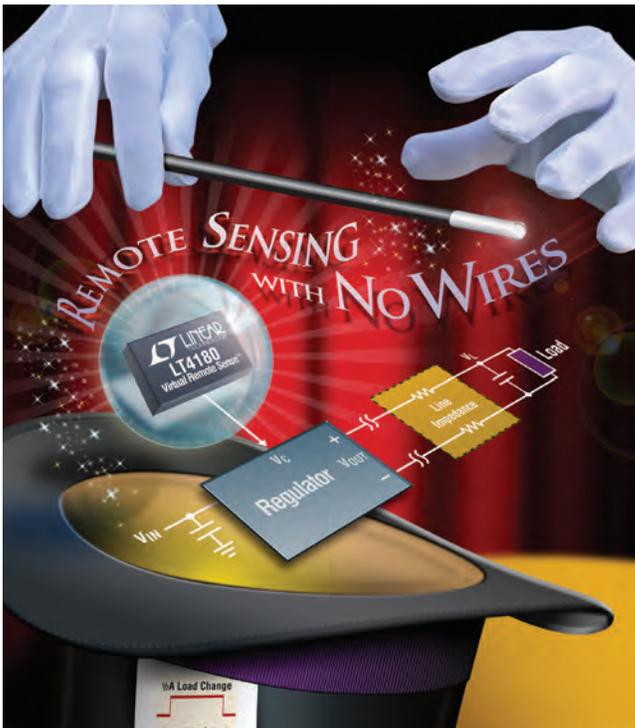
The conventional solution to improving regulation at the load involves adding extra wires for remote sensing (Figure 2a), but adding extra wires is not always desirable, or even possible. A new control method, Virtual Remote Sensing™ (VRS™), easily replaces and avoids the pitfalls of conventional solutions and in some instances solves the previously insoluble.

### LOAD-END REGULATION BEFORE VRS

Virtual Remote Sensing solves the problem of maintaining load regulation at the end of long wiring runs. VRS is easier to implement and generally performs better than conventional remote sensing techniques such as direct remote voltage sensing, voltage-drop compensation, and load-end regulation.

The first conventional technique, direct remote sensing (Figure 2a), produces excellent load-end regulation, but it requires two pairs of wires: one pair to provide the load current and a second pair to measure the voltage at the load for proper regulation. Traditionally, remote sensing requires foresight—it must be

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VRS avoids the limitations of conventional voltage drop compensation techniques while producing impressive load regulation over a wide range of conditions.

(LT4180, continued from the cover)

designed into the system. Unless an extra pair of sense wires is ready and waiting, remote sensing is impossible to implement after the fact.

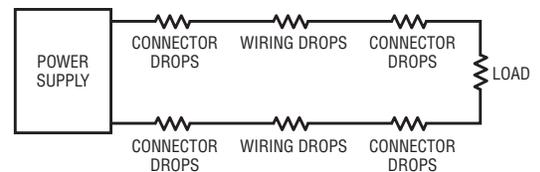
The second conventional technique, voltage drop compensation, doesn't require extra wires, but it does require careful estimation of the voltage drop of the load lines. The supply voltage is adjusted to make up for the estimated interconnection voltage drop. However, since the drop is only an estimated value and not measured, the accuracy of this method is questionable at best.

The third conventional technique involves placing a voltage regulator directly at the load (Figure 2b). This provides both accuracy and simplified wiring, but the regulator consumes valuable space at the load end, reduces overall power system efficiency and power dissipation near the load increases. In industrial and automotive systems, it may be impossible to place a regulator in the harsh environment at the load end.

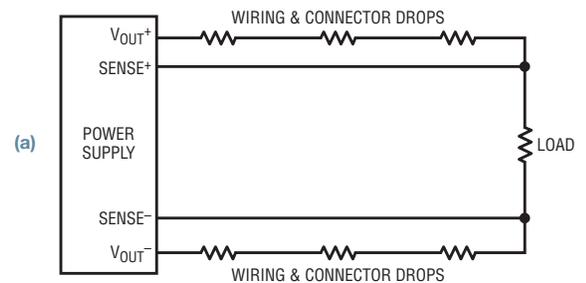
VRS avoids all of these limitations while producing impressive load regulation over a wide range of conditions.

(continued on page 4)

**Figure 1. A simple model shows the problem of load regulation over resistive interconnections.**



**Figure 2. Two conventional methods for solving the problem of wiring voltage drops: (a) Remote sensing solves the problem of load regulation, but adds wires across the divide, (b) local regulation stabilizes load voltage but is inefficient.**



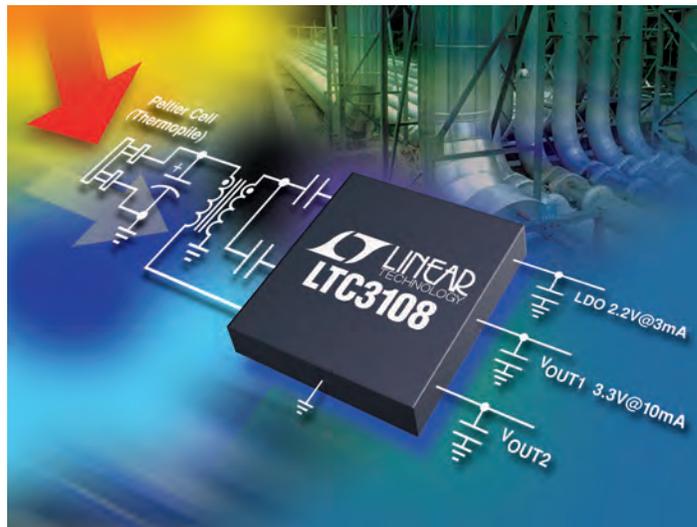
# Linear in the News

At the annual *EDN* Innovation Awards in April, *EDN* magazine honored Linear Technology's LTC<sup>®</sup>3108 ultralow step-up converter and power manager for energy harvesting with an award in the Power: Converters category. This is a device in Linear's new energy harvesting family—the industry's first ICs capable of taking extremely low levels of current from heat, vibration or other sources to power wireless transmitters or sensors.

The LTC3108 is a highly integrated DC/DC converter, ideal for harvesting and managing surplus energy from extremely low input voltage sources such as thermoelectric generators (TEG), thermopiles and small solar cells. Using a small step-up transformer, it provides a complete power management solution for wireless sensing and data acquisition. Extremely low quiescent current and high efficiency design ensure the fastest possible charge times of the output reservoir capacitor.

The LTC3108's self-resonant topology steps up from input voltages as low as 20mV. Energy harvesters are designed for applications using very low average power, but requiring periodic pulses of higher load current. For example, in many wireless sensor applications the circuitry is only powered to take measurements and transmit data periodically at low duty cycle.

Also honored with an *EDN* Innovation Award was Linear Design Manager Michael Kultgen, who received *EDN*'s award for Best Contributed Article for his article, "Managing High-Voltage Lithium-Ion Batteries in HEVs." The



## EDN INNOVATION AWARD

*EDN* magazine honored Linear Technology's LTC3108 ultralow step-up converter and power manager for energy harvesting with an award in the Power: Converters category. This is a device in Linear's new energy harvesting family—the industry's first ICs capable of taking extremely low levels of current from heat, vibration or other sources to power wireless transmitters or sensors.

article discusses an application based on Linear's LTC6802 battery stack monitor for lithium-ion batteries.

## VIRTUAL REMOTE SENSE DEBUT

Linear Technology has just introduced the LT<sup>®</sup>4180 Virtual Remote Sense™ (VRS) controller, which was covered in electronic design publications worldwide, including *EDN* and *Power Electronics Technology*. The LT4180 maintains a correctly regulated voltage at the load, regardless of load current or line impedance. VRS compensates for voltage drops in long cable, wire and circuit board trace runs without the remote sense wires used in conventional schemes.

The LT4180 is inserted into the feedback loop of just about any DC/DC regulator or module to continuously interrogate the line impedance and adjust the regulator's output voltage to produce the desired voltage at the load. The device's 3V to 50V input voltage range addresses a

variety of applications, including remote instrumentation, battery charging, wall adaptors, notebook power, surveillance equipment and halogen lighting.

## LINEAR VIDEO CHANNEL

Several new video design ideas from Linear Technology have just been posted. The following can be viewed on Linear's website at [video.linear.com](http://video.linear.com):

- "Low Power RF Mixers Enhance Receiver Performance" with James Wong
- "Synchronous PolyPhase<sup>®</sup> Boost Converter for Cool and Powerful Applications" with Goren Perica

On *EE Times China & Asia* websites:

- "Simple 2-Terminal Current Source" with Bob Dobkin

And on *EE Times Japan*'s site:

- "High Current LED Driver" with Walker Bai ■

## The LT4180 works with nearly any power supply or regulator: linear or switching, isolated or non-isolated.

(LT4180, continued from page 2)

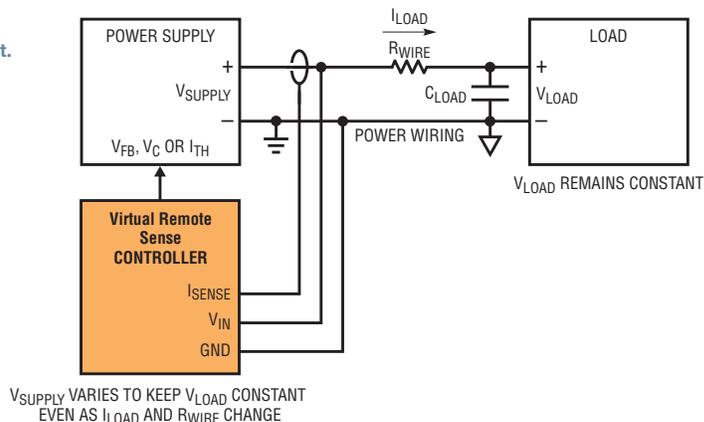
### WHAT IS VRS?

Figure 3 shows a simplified schematic of a Virtual Remote Sense system consisting of a power supply or regulator driving a load over a resistive interconnection (consisting of wiring plus connectors). Without VRS, supply voltage ( $V_{SUPPLY}$ ) and DC current ( $I_{LOAD}$ ) are known, but there is no way to determine how much voltage is delivered to the load and how much voltage is lost in the wiring, so proper load voltage regulation can't be achieved.

The LT4180 VRS solves this problem by interrogating the line impedance and dynamically correcting for the voltage drops. It works by alternating the output current between 95% of the required output current and 105% of the required output current. In other words, the LT4180 forces the supply to provide a DC current plus a current square wave with peak-to-peak amplitude equal to 10% of the DC current. Decoupling capacitor  $C$  (which normally insures low impedance for load transients in non-VRS systems) takes on an additional role by filtering out voltage transients from the VRS square wave.

Because  $C$  is sized to produce an “AC short” at the square wave frequency, the interrogating voltage square wave produced at the power supply is equal to  $V_{SUPPLY(AC)} = 0.1 \times I_{DC} \times R$ , measured in  $V_{P-P}$ . The voltage square wave measured at the power supply has a peak-to-peak amplitude equal to one tenth the DC wiring drop. This is not an estimate—it is a direct measurement of the voltage drop across the wiring over all load currents.

Figure 3. Virtual Remote Sensing is easy to implement.



Minor signal processing creates a DC voltage from this AC signal, which is introduced into the supply's feedback loop to provide accurate load regulation.

### SO HOW WELL DOES VRS WORK?

Static load regulation for the LT4180 is shown in Figure 4. In this case, load current was increased from zero until it produced a 2.5V drop in the wiring. The voltage at the load dropped only 73mV at maximum current from what it would be at no current. Even with an in-the-wire voltage drop equivalent to 50% of the nominal load voltage, the voltage at the load stayed within 1.5% of the no load current value. Less dramatic wiring drops produced even better results.

### VRS IS EXTREMELY FLEXIBLE

The LT4180 works with nearly any power supply or regulator: linear or switching, isolated or non-isolated. Power supplies can be synchronized to the LT4180 or not. To accommodate a variety of system and power supply requirements, VRS operating frequency can be adjusted over more than three decades. It also offers spread

spectrum operation to provide partial immunity from single-tone interference. Its large input voltage range simplifies design.

### SOLVING THE IMPOSSIBLE WITH VRS

Besides offering an alternative to conventional techniques, VRS opens up opportunities previously unavailable in battery charging, industrial and Ethernet, lighting, well logging and other applications.

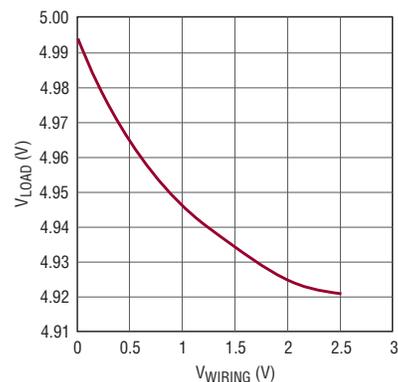


Figure 4. Static load regulation for the LT4180 is impressive over an impressive range of regulator-to-load wiring voltage drops.

## The LT4180 VRS solves the problem of line voltage drops by interrogating the line impedance and dynamically correcting for the voltage drops.

### Improve Battery Chargers

Figure 5 illustrates a poorly conceived power system for notebook computers, PDAs, cell phones or portable entertainment devices. An external power supply/battery charger is used to minimize the size of the portable electronic device. The

charger only works properly when the device is off and not drawing current. As the battery approaches full capacity, battery charging current ( $I_{BAT}$ ) is nearly zero. With  $I = 0$ , the battery charger voltage  $V_{SUPPLY}$  equals the battery float voltage and charge termination works properly.

But what happens if the system voltage regulator is drawing current? The battery voltage  $V_{BAT}$  can be less than the needed battery charger voltage,  $V_{SUPPLY}$ , thus slowing charging or even stopping it altogether. Interconnection resistance can't be lowered enough to solve this. The 1% Li-ion float voltage accuracy requirement translates into a 42mV float voltage error (for a one cell Li-ion battery). Because there are other float-voltage error sources, the wiring drop must be kept well below this.

Figure 5. A (flawed) battery charging architecture aims to reduce the device size with an external battery charger.

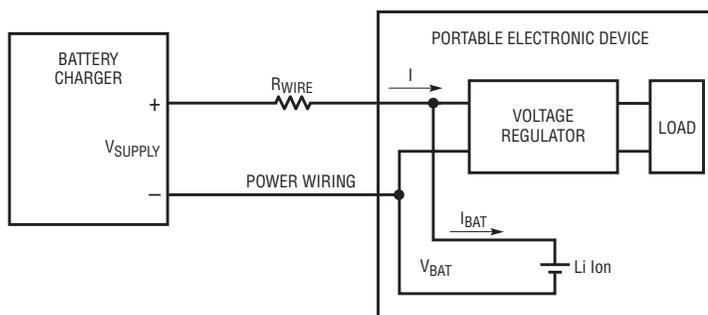


Figure 6. Typical battery charging architecture without VRS

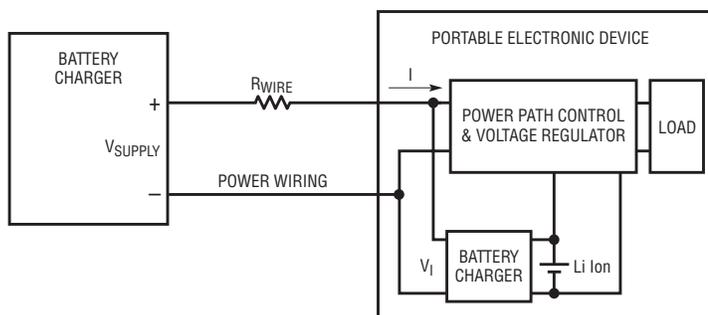
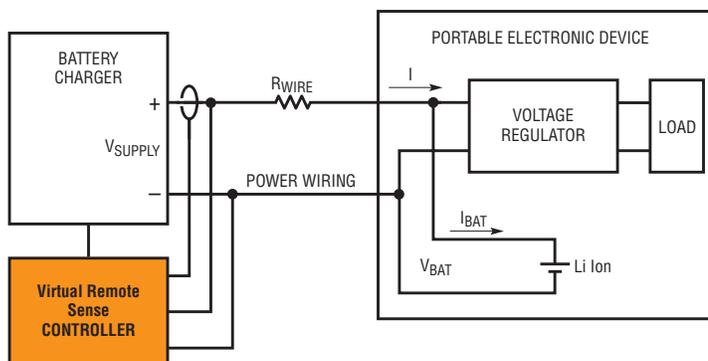


Figure 7. Simplified battery charging with VRS reduces the overall device size, achieving what the solution in Figure 5 could not.



The conventional solution uses a complex architecture like that shown in Figure 6, which incorporates the charger and a power path controller into the device. While this reduces wiring-related charging errors, it increases the size of the device and the power dissipation within the device because the charger and power path controller must be packed inside.

Figure 7 shows the no-compromise solution using VRS. Charger voltage is properly controlled at the device, independent of load current ( $I$ ), so an external battery charger supply can be used and a power path controller eliminated.

### Easily Compensate Line Drops in Power over Ethernet Applications

Power over Ethernet and industrial applications also benefit from VRS. VRS allows low voltage devices (with high operating current) to operate over CAT5 and CAT6 cable—without the drops caused by long runs. Even 10V-20V line drops can be compensated, allowing either no regulator or a simple linear regulator at the far end.

A VRS system can be used to improve lighting. For medium and large lighting systems, the improvement in energy efficiency easily pays for the upgrade from a standard transformer to a DC/DC converter. Additional benefits of using a VRS system include better color-temperature control and longer, more consistent bulb lifetimes.

### Retrofit Industrial Applications

VRS can also be used to simplify system retrofits for industrial applications. For example, a pair of power wires is available for new equipment, but regulation at the load-end is not up to the equipment spec. VRS can be easily dropped in to control the existing power supply or regulator. This is far easier and cheaper than adding another pair of wires for remote sensing or adding load-end regulation.

### Increase the Efficiency and Light Output of High Intensity Lighting Applications

While incandescent lighting is on the decline, high intensity halogen lights continue to be popular. The operating voltage of halogens directly affects their light output, efficiency,

lifetime and color temperature as shown in Figure 8, and as follows:

- Light output is approximately proportional to  $v^{3.4}$
- Power consumption is approximately proportional to  $v^{1.6}$
- Lifetime is approximately *inversely* proportional to  $v^{16}$
- Color temperature is approximately proportional to  $v^{0.42}$

Normally these devices operate at 12V, but their operating current is relatively high, so line drops between the regulator and the light can be high. In this case, the load-end discrepancy can easily reach 1V or more. A 12V halogen operated at 11V produces 25% less light than when operated at 12V, with only a 13% power

savings. So to produce light at 11V that is equivalent to that produced at 12V would require 25% more bulbs running relatively less efficiently. Simply put, running

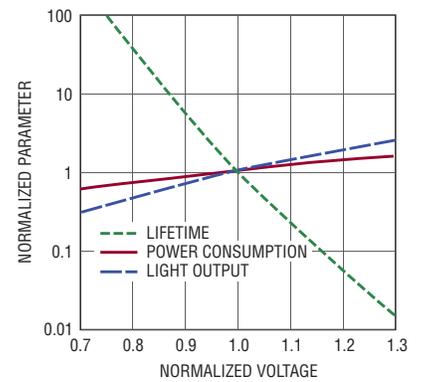
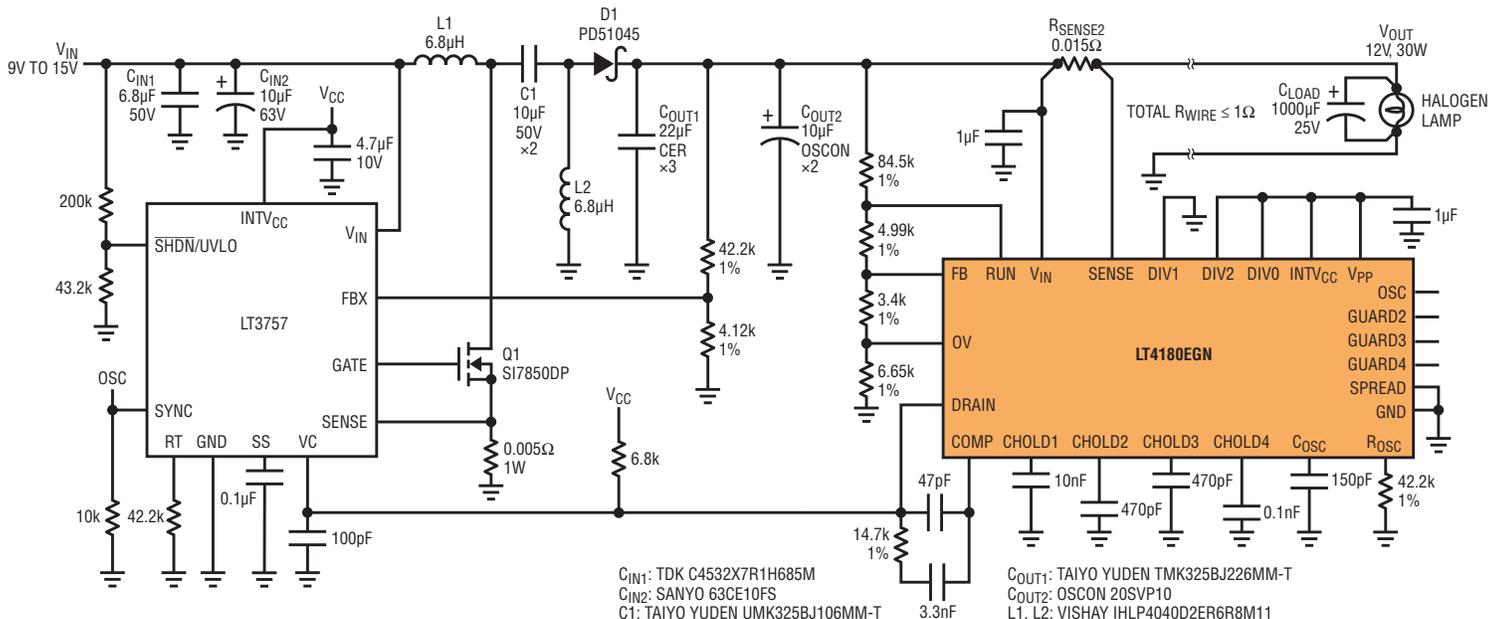


Figure 8. Lamp parameters vs normalized lamp voltage show that better voltage regulation at the lamp improves output, saves energy and prolongs lamp life.

Figure 9. An automotive halogen headlamp power supply



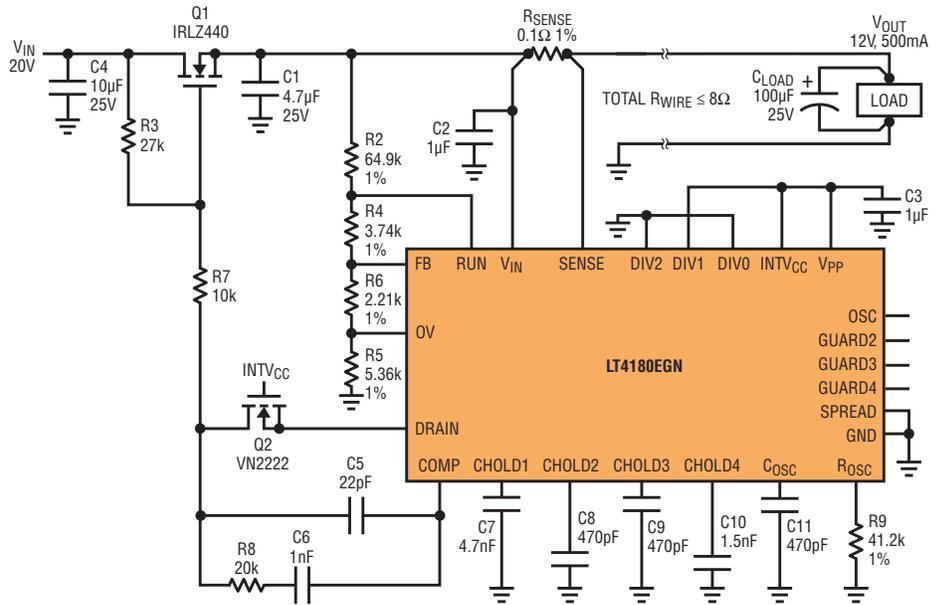


Figure 10. A full featured VRS linear supply

halogens at the correct voltage offers more precise lighting control, more predictable color temperature and better efficiency.

A VRS system can be used to accurately maintain correct bulb intensity. A capacitor is placed in the vicinity of the bulbs, and the voltage is controlled at that point. For medium and large lighting systems, the improvement in energy efficiency easily pays for the upgrade from a standard transformer to a DC/DC converter. Additional benefits of using a VRS system include better color-temperature control and longer, more consistent bulb lifetimes.

A SEPIC-based automotive halogen headlight power supply (Figure 9) improves bulb reliability while also ensuring optimum illumination. The design maintains 12V at headlight voltage over a 9V to 15V input voltage range. It works well up to 1Ω interconnection resistance. Using VRS allows the SEPIC converter to be placed far from the load—say in the passenger compartment, away from extreme under-the-hood environments, thus improving reliability.

Residential and commercial track-style lighting also benefit. The cost of properly regulating lamp voltage is quickly recouped in the form of lower power consumption and higher efficiency. Two to three kilowatt-hours can be saved per day on a 250W string while maintaining the same amount of light. Color temperature (while not as dependent on voltage as other lamp parameters) also benefits. VRS allows remote voltage regulation of a single lamp, or provides

first-order regulation of several lamps distributed over a single power rail.

**VRS Might be the Only Solution When the Line Lengths Are in Miles**

VRS can be used in oil and gas well logging applications where instrumentation is often connected by cables from thousands, to tens of thousands of feet long.

**A COLLECTION OF APPLICATIONS**

The LT4180 includes all components needed for a linear power supply (except for the pass transistor). Undervoltage lockout, overvoltage lockout and soft-correct are also available, so a full featured linear

VRS power supply can be built with few components (Figure 10). The linear supply in Figure 10 provides 12V at 500mA with an 18V input. Pass transistor Q1 is driven via R3, R7 and Q2 via the DRAIN pin. Q2 serves to keep DRAIN pin voltage below the absolute maximum rating. C5, R8, and C6 provide compensation. R2, R4, R5, and R6 set output voltage and lockout thresholds. R1 is the current sense resistor. C7–C10 are hold capacitors used by the VRS, while C11 and R9 set the square wave frequency. Typical load-step response is shown in Figures 11 and 12 with 4Ω wiring resistance and 100μF and 1100μF load-end

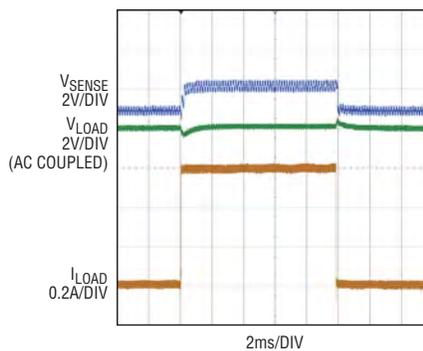


Figure 11. Load step response of the linear supply shown in Figure 10 with 100μF decoupling capacitance.

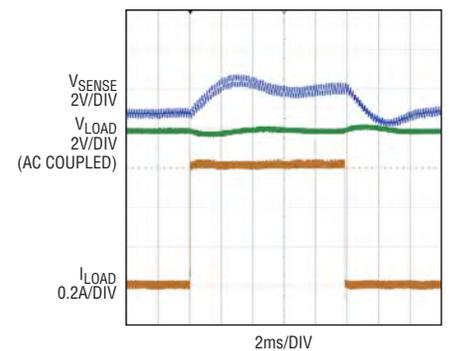
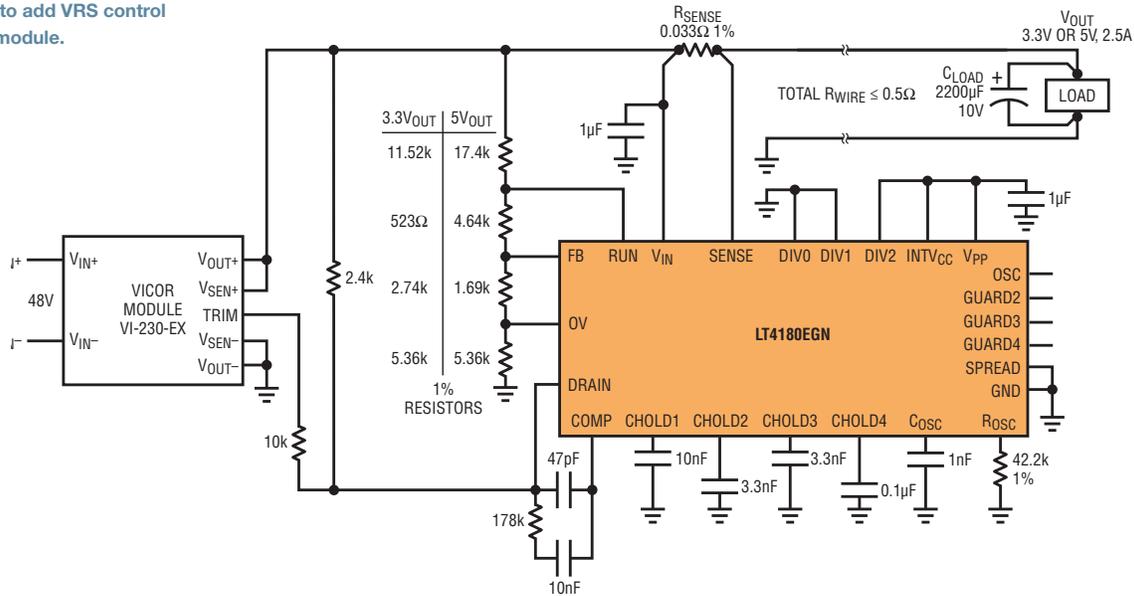


Figure 12. Load step response of the linear supply shown in Figure 10 with 1100μF decoupling capacitance.

Figure 13. It's easy to add VRS control to a power supply module.



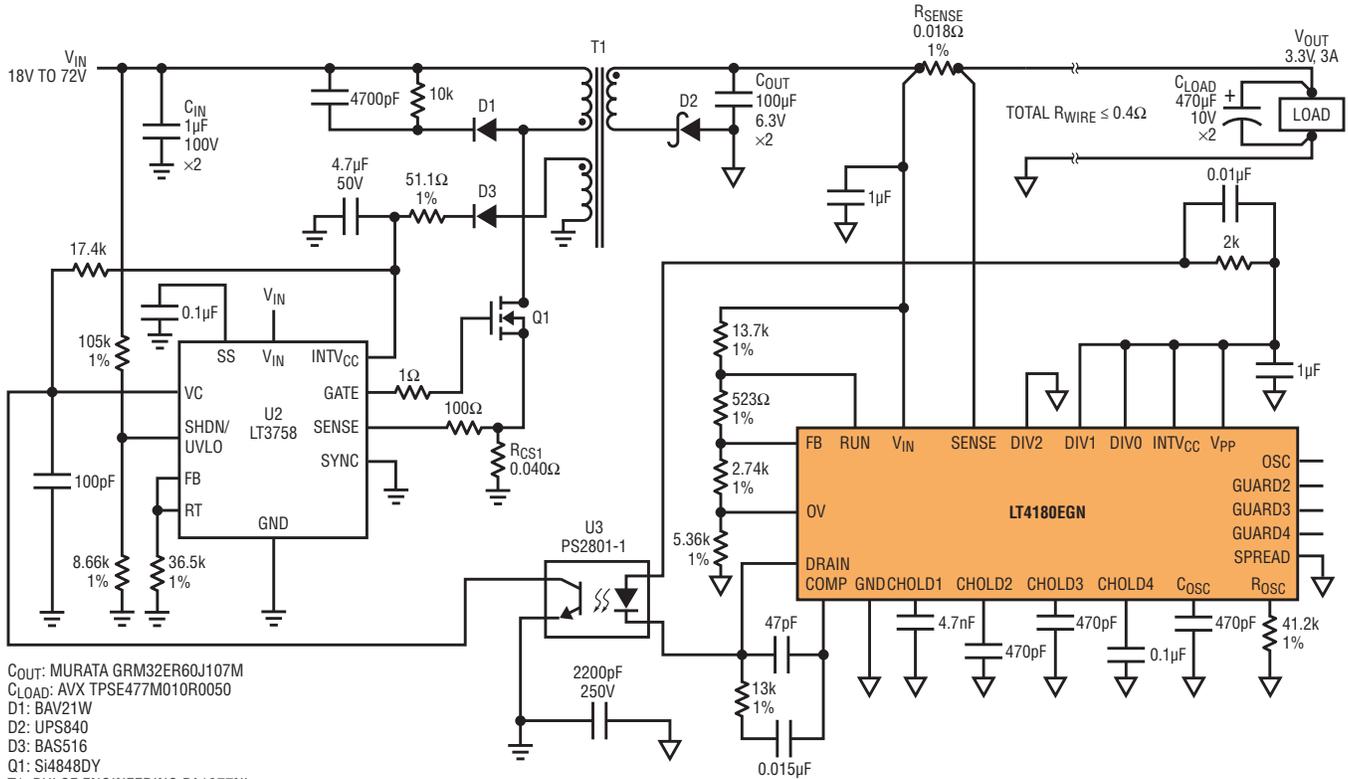
capacitances. VRS transient response is well controlled with widely varying  $C_{LOAD}$ .

Figure 13 shows how the LT4180 interfaces to a Vicor power module, providing Virtual Remote Sensing for a 3.3V/2.5A or 5V, 2.5A load through 0.5Ω wiring resistance. Output voltage is adjusted by

changing the values of the feedback and overvoltage resistors. Nominal input voltage is 48V. VRS is produced via the module's trim pin. This design works with 0.5Ω wiring resistance and 2200μF decoupling capacitance.

A fully isolated flyback converter capable of supplying 3.3V at 3A from an 18V to 72V input is shown in Figure 14. It is designed to correct for up to 0.4Ω of wiring resistance. Recommended load-decoupling capacitance is 940μF. Isolation is achieved through T1 and

Figure 14. A fully isolated VRS flyback converter



The LT4180 gives power supply designers a valuable new tool, enabling use of Virtual Remote Sensing for accurate load voltage regulation over highly resistive interconnections. Virtual Remote Sensing provides alternatives previously unavailable for simplifying or improving designs.

opto-isolator U3. While not shown in this design, it is also possible to provide an opto-isolated OSC signal from the LT4180 to a power supply for synchronization.

Figure 15 shows a buck regulator capable of supplying 12V at 1.5A to a load with up to 2.5Ω of wiring resistance. 470μF load decoupling capacitance is recommended. Input voltage range is 22V to 36V.

### CONCLUSION

While conventional 2-wire remote sensing gives proper voltage at the load, there are many drawbacks. The sense wires are an additional cost in the system as well as consuming connector space for the system. Reliability issues can occur if the sense wires are disconnected or broken.

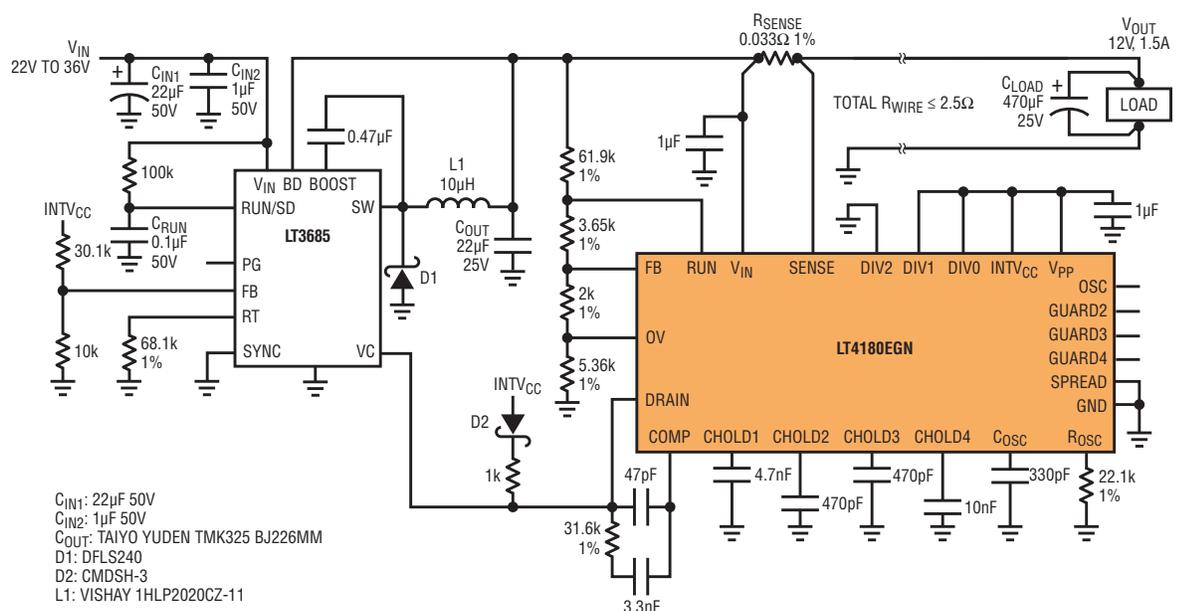
In contrast, a Virtual Remote Sense system produces excellent regulation at the load, with none of the drawbacks of wired remote sense. Unlike other compensation schemes such as negative resistance, Virtual Remote Sensing continuously corrects the output—even if the line-drop resistance changes—by determining real-time wire drops and connector drops. The additional noise on the power supply lines from the Virtual Remote Sense circuitry is easily removed by the capacitor at the load, which is always included in remote sense systems anyway.

The LT4180 can interface with IC regulators as well as preconfigured purchased offline supplies. In most cases,

the cost of adding a VRS IC to a power supply system is much less than laying wires for traditional remote sensing.

The LT4180 gives power supply designers a valuable new tool to accurately regulate load voltage over highly resistive interconnections. Virtual Remote Sensing provides alternatives previously unavailable for simplifying or improving designs. The LT4180 VRS works with virtually any power supply or regulator: switching or linear, isolated or non-isolated, synchronized or unsynchronized. It contains a VRS regulation circuit and a variety of features such as undervoltage and overvoltage lockout, and opto-isolator drivers. ■

Figure 15. A VRS buck converter



# Unique Analog Multiplier Continuously Monitors Instantaneous Power and Simplifies Design of Power Control Loops

Mitchell Lee and Thomas DiGiacomo

As energy consumption in electronics is increasingly scrutinized, the ability to accurately monitor and control power becomes an important part of any system design. To measure instantaneous power, one must simultaneously measure current and voltage and multiply the results. While traditional analog multiplier ICs can perform continuous multiplication, they typically lack the operating range and input sensitivity required for power monitoring and control. The high price of the multiplier itself and the necessary additional signal conditioning circuits make such a solution costly both in dollars and in board area.

Digital solutions, on the other hand, can provide sensitivity and dynamic range, but lack the ability to *continuously* monitor power. For instance, the LTC4151 combines a 7V to 80V operating voltage range, a current sense amplifier, a MUX, and an I<sup>2</sup>C interface with a 12-bit ADC to measure current and voltage. Multiplication is performed in a host processor. This makes for an accurate power monitoring solution, but the 7.5Hz conversion rate of the ADC limits its utility in closed loop applications, where it is unable to respond to rapid changes.

The LT2940 power monitor solves the problems of creating power monitor and control systems by combining all of the necessary features in a single IC (see Figure 1). Here are a few of its features:

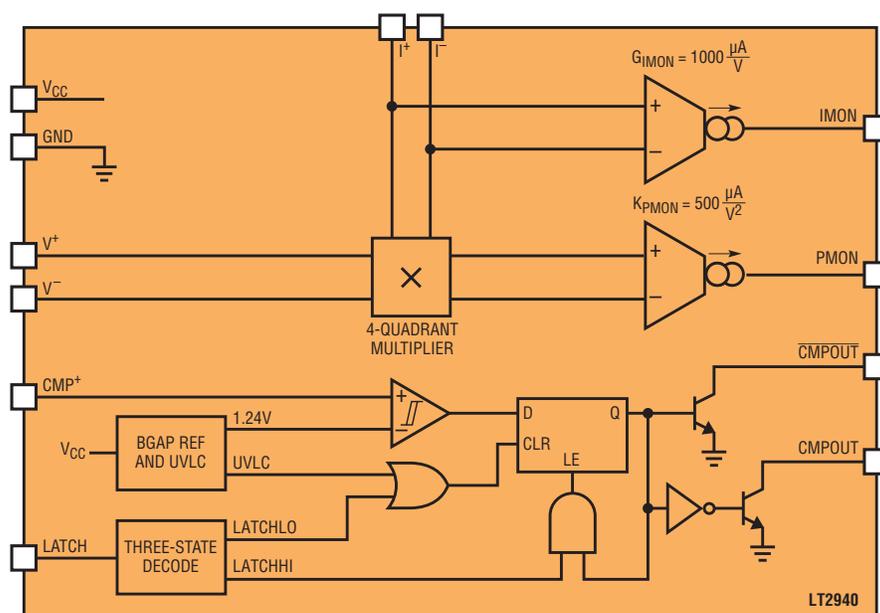
- Measures Power of a Supply or Load
- 4V–80V High Side Current Sense, 100V Max
- Full 4-Quadrant Operation
- 500kHz Bandwidth
- Current Mode Power and Current Monitor Outputs

A wide, 4V to 80V operating range makes the LT2940's current sense suitable for 48V telecom power as well as intermediate bus voltages in the 5V to 2.4V range.

4-quadrant capability allows the LT2940 to monitor bidirectional power flow such as in battery applications, or to measure power flow in reversible and regenerative motor drives. In AC applications where 4-quadrant operation is necessary, there is plenty of bandwidth to accurately track the results of a chopped sinusoid at common line frequencies of 50Hz, 60Hz and 400Hz.

The LT2940 also includes a current monitor output that allows the load current to be examined directly. The output signals for both the power and current are current mode, a feature readily appreciated when using the LT2940 in a servo loop, or when simply filtering the output. An integrated comparator with complementary open-collector outputs and selectable latching allows the LT2940 to be used for direct control, so an entire control loop can be implemented with a single IC.

Figure 1. Block diagram of the LT2940



The LT2940 has been designed specifically for measuring the power flowing into or out of sources such as regulators, converters and batteries, as well as input power to loads ranging from telecom cards to motors to RF amplifiers.

### CURRENT IS NOT (ALWAYS) POWER

In systems supplied by a well-regulated fixed voltage, there is no need to directly measure power. In these systems, power is simply inferred by measuring current, and scaling the gain of the current sense amplifier to represent multiplication by the fixed supply voltage. In systems where the supply voltage is not regulated to a fixed value, or is not regulated at all, monitoring current as a means of inferring power is not feasible. Instead, both voltage and current must be measured simultaneously and multiplied together to determine the power. Central office telecom systems are good examples of wide-range, unregulated supplies. These systems are battery operated and commonly designed to operate over a range of 36v to 72v or more.

Other 48v-based systems such as servers and mass storage are powered by regulated supplies, but it is not unusual for the supply bus to be set to a regulation point higher than 48v to reduce backplane

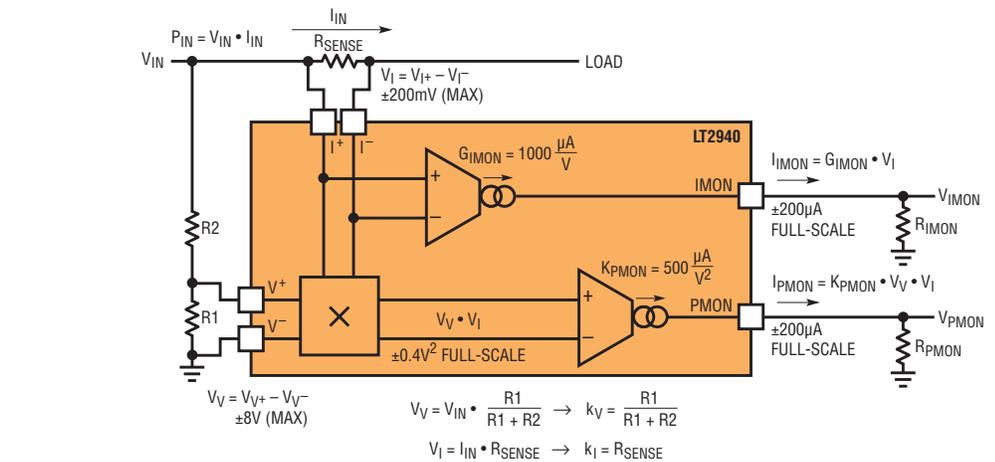


Figure 2. The LT2940 signal path and typical external components

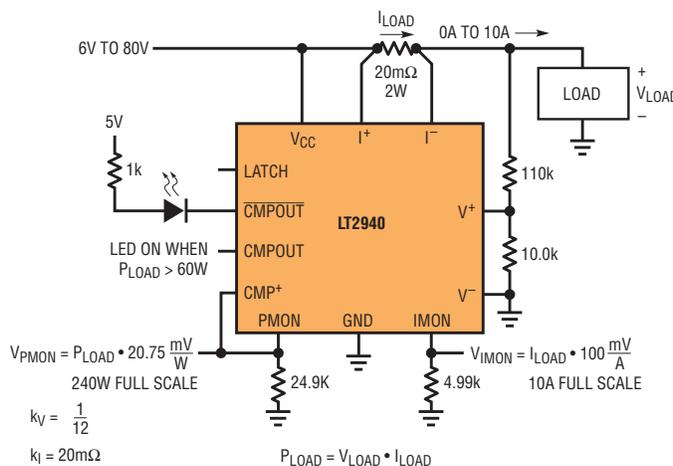
distribution current, or to achieve longer hold-up times in case of supply drop-outs or outright loss of power. A product designed into such a system may encounter one application regulated to 48v, while another might be adjusted for 57v, and yet another set to 62v. Such a wide operating range precludes ascertaining power from a simple current measurement without customized gain calibration.

### CORE OPERATION

The LT2940 has been designed specifically for measuring the power flowing into or out of sources such as regulators, converters, and batteries, as well as input power to loads ranging from telecom cards to motors to RF amplifiers. Unlike traditional analog multipliers that conjure up images of split supplies and narrow input ranges, the LT2940 is designed to bolt up to DC supplies ranging from 6v to 80v with little more than a current sense resistor and a voltage divider.

Figure 2 illustrates the signal path block diagram along with typical external components common to almost all applications. The current sense input pins I<sup>+</sup> and I<sup>-</sup> measure up to ±200mV differentially over a common mode range of 4v to 80v, independent of the supply pin V<sub>CC</sub>. The voltage sense input pins V<sup>+</sup> and V<sup>-</sup> measure up to ±8v with a common mode range limited by the V<sub>CC</sub> and GND pins.

Figure 3. A load monitor that alarms above 60W





### APPLICATIONS

The LT2940 translates power into a simple analog control signal, making it possible to easily produce a variety of applications that were heretofore difficult, or nearly impossible, to realize: power monitors, power servo controls and regulated heaters, just to name a few. What follows are just a few of the possible applications.

#### Simple Power Monitoring

Figure 3 shows a core use of the LT2940, a simple power monitoring application. The circuit operates over a 6V to 80V range, measuring up to 240W. Below 24V the measurement range is limited to 10A maximum. The comparator output lights an alarm LED when the load power exceeds 60W. Owing to the bandwidth of the LT2940, even relatively short over-power excursions in the 2μs–3μs range are easily detected by the comparator. By adding a MOSFET disconnect switch and controlling the LATCH pin, it is possible to form a 60W overpower circuit breaker (see the application in Figure 4, for example).

The circuit breaker application in Figure 4 extends simple monitoring into the realm of control. The LT2940 is configured to measure motor current and power, and to protect the field magnets in the event the current exceeds the motor's 8A rating. This circuit also highlights the use of the LATCH pin to keep the motor off after a triggering event until the RESET signal is pulled low.

Addition of positive voltage bias to the PMON and IMON output networks, and a rectifier between IMON and CMP+ allows the monitor and circuit breaker to work in two quadrants, covering both "motor" and "generator" modes of operation. Other applications below employ these techniques.

The LT2940 current sense inputs, I+ and I- are designed to operate over a range of 4V to 80V. Nevertheless, it is possible to translate bottom-side (ground)

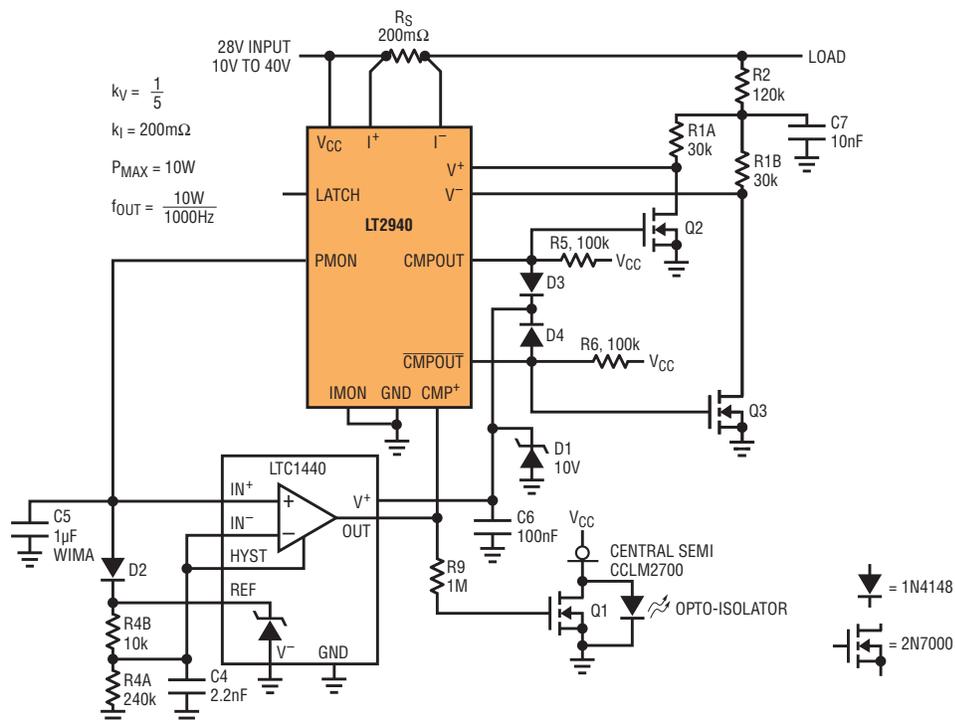
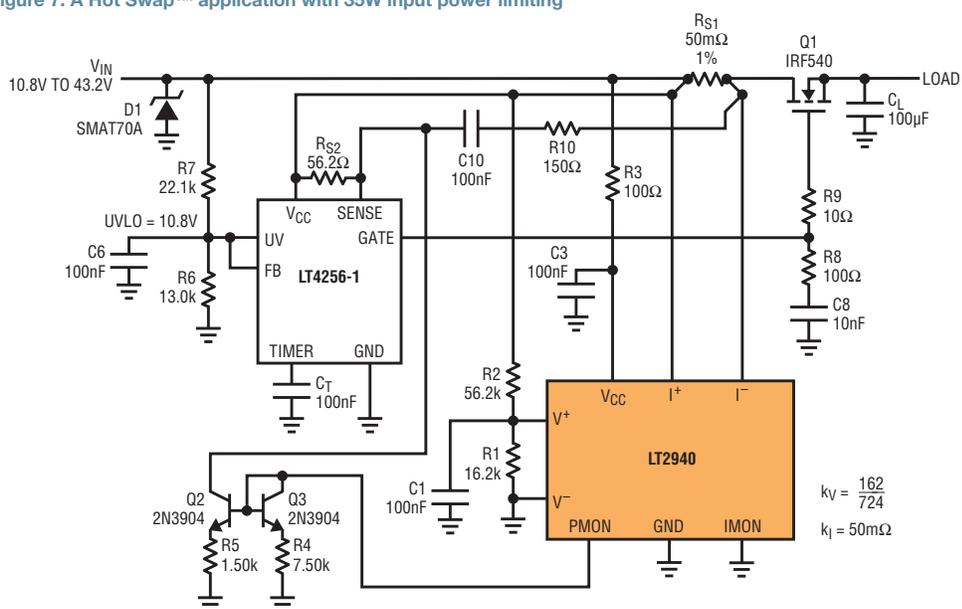


Figure 6. A 28V power-to-frequency converter

current sense information to I+ and I- using the circuit shown in Figure 5. An LT1635 performs the necessary translation for both positive and negative current flow. The PMON and IMON outputs are biased with a Zener diode so that positive and negative power and current measurements are available.

In some environments isolation is necessary for safety or noise reasons. Figure 6 shows a power-to-frequency converter using the LT2940. The PMON output alternately charges and discharges a film capacitor, C5. When the voltage on C5 charges to the upper threshold on hysteretic comparator LTC1440, its output flips the phase of

Figure 7. A Hot Swap™ application with 35W input power limiting



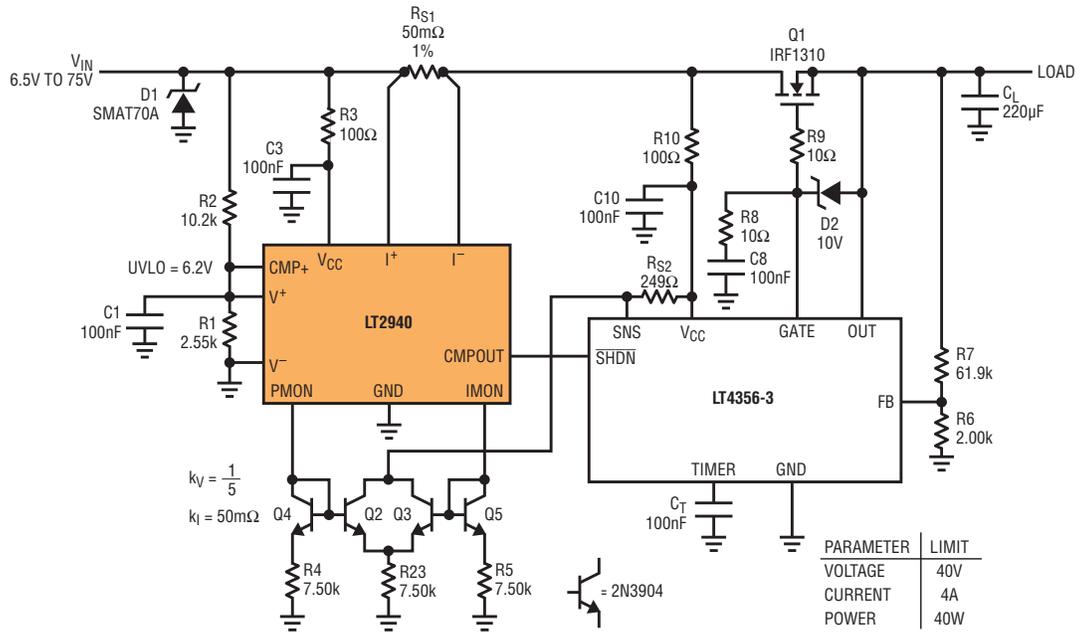


Figure 8. An overvoltage protection regulator with power and current limiting

the voltage sense inputs  $v^+$  and  $v^-$  using a pair of MOSFETS (Q2 and Q3). The LT2940's comparator serves as a phase splitter to develop complementary signals with which to drive Q2 and Q3. When the phasing of the voltage sense inputs is reversed, PMON discharges C5 to the LTC1440's lower threshold, whereupon the action is repeated. The frequency is proportional to the current at PMON and, ultimately, the power consumed by the load. An optoisolator conveniently communicates the frequency across an isolation barrier.

### Advanced Power Monitoring

Power limiting is crucial to applications such as running off a backup generator or supplying multiple line cards in an enclosure with low air flow. The LT2940 meshes well with Hot Swap and Surge Stopper circuits, which control current or voltage to provide important power control capability.

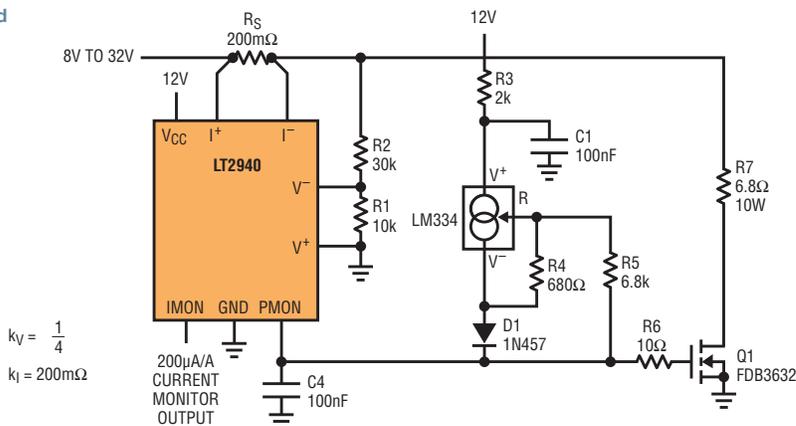
To limit the input power in the LT4256-1 Hot Swap application shown in Figure 7, the LT2940 controls its current sense input. The Hot Swap controller serves the

GATE pin to achieve a known drop across a sense resistor in a standard application. The inrush current is set by gate capacitor, C8. In this application circuit, the LT2940's output signal substitutes power for current at the LT4256's current sense, so that the load current is limited in inverse proportion to the input supply voltage; load power is thus regulated. The LT4256's current circuit breaker behaves as a power circuit breaker with a regulated limit of 35W.

The 5:1 multiplying mirror brings the 200μA full scale PMON output current up to 1mA, which makes the SENSE pin input current error negligible, in addition to avoiding the LT2940's PMON pin compliance limit. The 100nF and 150Ω between SENSE and  $\Gamma^-$  provide necessary feed-forward compensation.

The application in Figure 8 marries the power and current sensing of the LT2940 with the surge voltage protection of the LT4356-3 to put a lid on excess voltage, current, and power. The LT4356 serves the GATE pin to limit the output voltage normally. The LT2940 limits power and current by feeding a control signal into the SNS input. Transistors Q2 through

Figure 9. An 8W load for an 8V to 32V supply bus





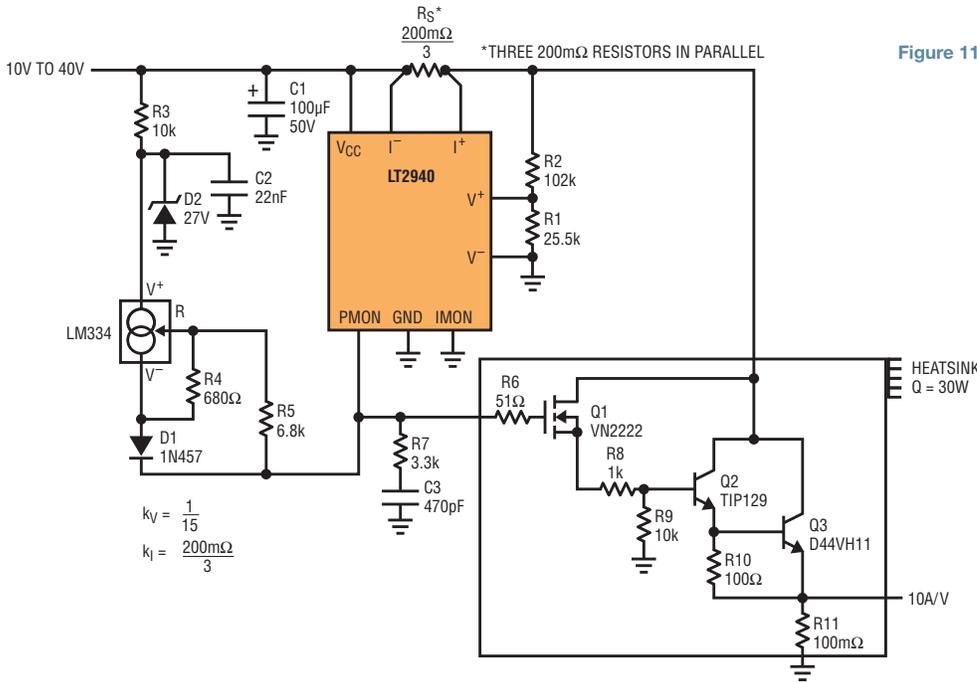


Figure 11. A 30W linear heat source

additional thermal hysteresis is provided by R13 and the complementary comparator output,  $\overline{CMPOUT}$ . One last feature of the load box circuit is reverse polarity protection, courtesy of diode D1.

Figure 11 shows a 30w linear heat source using a bipolar transistor in a TO-220 package as the primary dissipater (Q3). Components Q2, Q3 and R11 are mounted on the body to be heated, such as a heat sink, an enclosure, or a reaction vessel. For testing thermal performance, thermal resistance is given by  $T_{RISE}/30W$ . Here the  $I^+$  and  $I^-$  pins are reverse connected so that the PMON pin sinks current. The PMON output is balanced against a 200μA current source, the result driving the gate of Q1 to servo the power dissipation to 30w.

By using pulse width modulation, heat can be dissipated efficiently in one or more resistors. Figure 12 illustrates a

200μA is generated by an LT1635 op amp and reference, and controlled by a 10-turn potentiometer. This current balances against the measured power at PMON and regulates Q1 to a predictable power.

The LT2940's integrated comparator is used to shut down the circuit in case of undervoltage or overtemperature conditions. At reduced input voltage, a constant power servo attempts to draw ever more current, leading to a theoretical result

of infinite current at zero input voltage. Undervoltage lockout prevents the servo loop from shorting the supply at low input voltages. The input voltage is monitored by the comparator and if less than 10V, CMPOUT shorts the MOSFET gate to ground.

Overtemperature is sensed by Q2, which pulls down on CMP+ and shuts the MOSFET off in case of excessive temperature. Although the comparator includes its own small hysteresis, generous

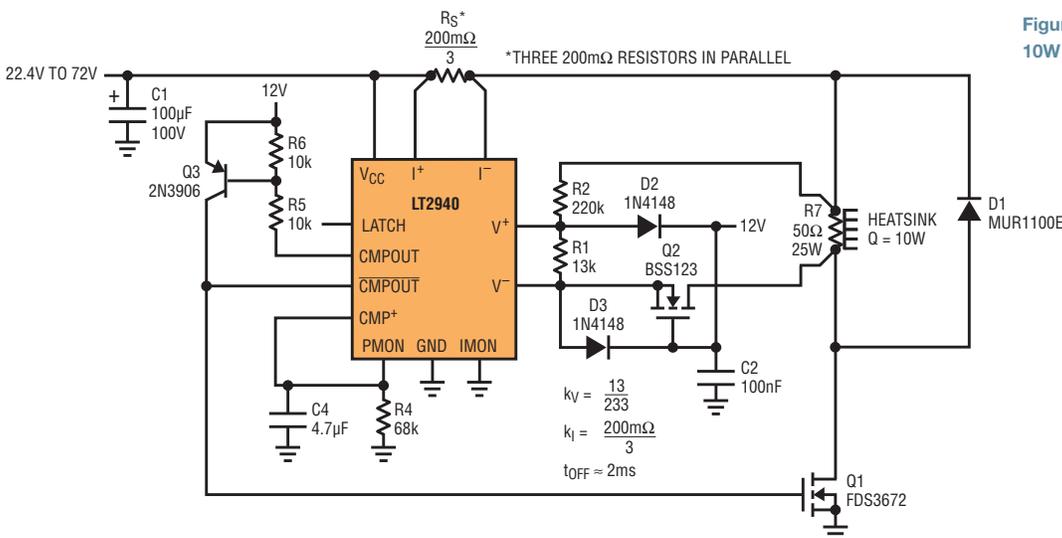
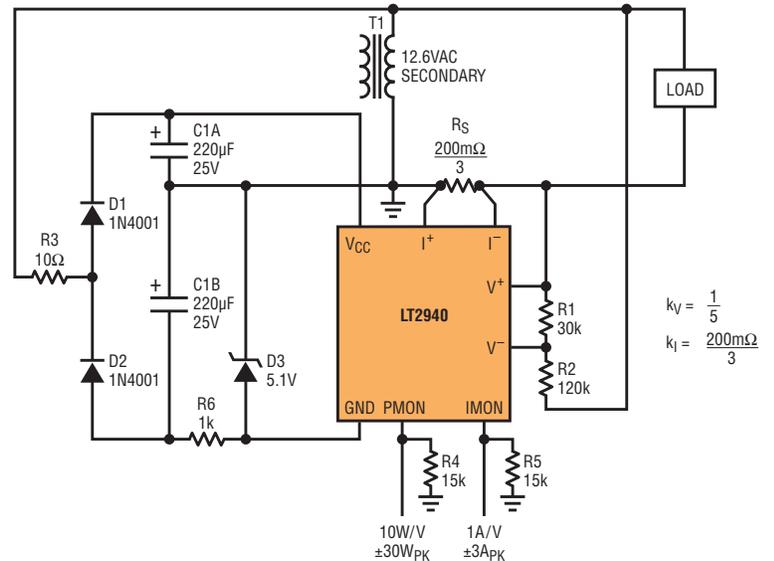


Figure 12. A wide input range 10W PWM heat source

Figure 13. An AC power and current monitor



low pulse-width modulated heat source that operates over a wide, telecom-like supply range. To distribute the heat across a circuit board, around an enclosure, or at various key points on a heat sink (to create a physical model of actual thermal conditions), multiple resistors connected in series and/or parallel may be substituted for the single 50Ω unit shown.

The integrated comparator is used as a PWM engine. When the CMP+ pin is low, Q1 is turned on, which connects the load resistor to the input. The PMON output sources current into C4 and R4, charging the CMP+ pin to its threshold of 1.24V. Q1 turns off, the power (and PMON's output current) falls to zero, and R4 discharges C4 slightly until the comparator trips and again drives Q1 on. The typical 35mV hysteresis in the comparator assures oscillation. Constant average power is maintained, with CMP+ maintained around 1.24V,

and R4 sinking a roughly constant current away from the CMP+ node, balanced by an equivalent average from PMON.

The simplest PWM scheme is employed here, with an N-channel MOSFET driven directly from one of the comparator's outputs, made push-pull with help from Q3. One side of the load is connected to the supply, and this means that during off times the voltage sense inputs, v+ and v- would be pulled up to as high as 72V, in violation of their 36V absolute maximum rating. Q2 and D2 clamp the inputs during the off time, protecting

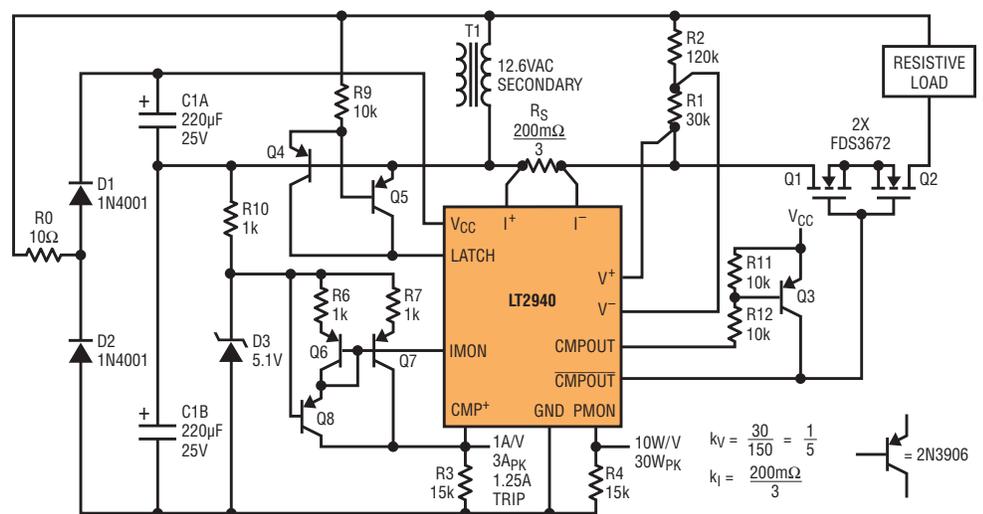
them from harm. See the LT2940 data sheet for a simpler PWM heater application for lower supply voltages.

#### AC Power Monitoring

4-quadrant operation allows the LT2940 to be used in AC applications, as shown in Figures 13, 14 and 15. Note that the LT2940 outputs a current proportional to instantaneous power, which is different from RMS-to-DC power metering such as the LTC1968 provides.

The LT2940 monitors the load power and current on a 12.6V secondary winding in Figure 13. A split supply is derived

Figure 14. A secondary side AC circuit breaker



## The LT2940 brings together the features that make power monitoring and control not only possible, but easy.

from the same winding so that bidirectional measurement of instantaneous real power and instantaneous current is possible. Note that averaging the power output results in average real power. The load can be any combination of resistance, reactance, or nonlinear devices including chopped or rectified circuits.

In Figure 14 the LT2940's comparator is used in conjunction with two MOSFETS to form a circuit breaker with cycle-by-cycle limiting, again operating on a 12.6V secondary. Devices Q4 and Q5 form a window comparator that resets

the circuit breaker after each half cycle, and Q6, Q7 and Q8 form a full-wave current rectifier to drive the CMP+ input. Thus only an absolute value current measurement is available at CMP+.

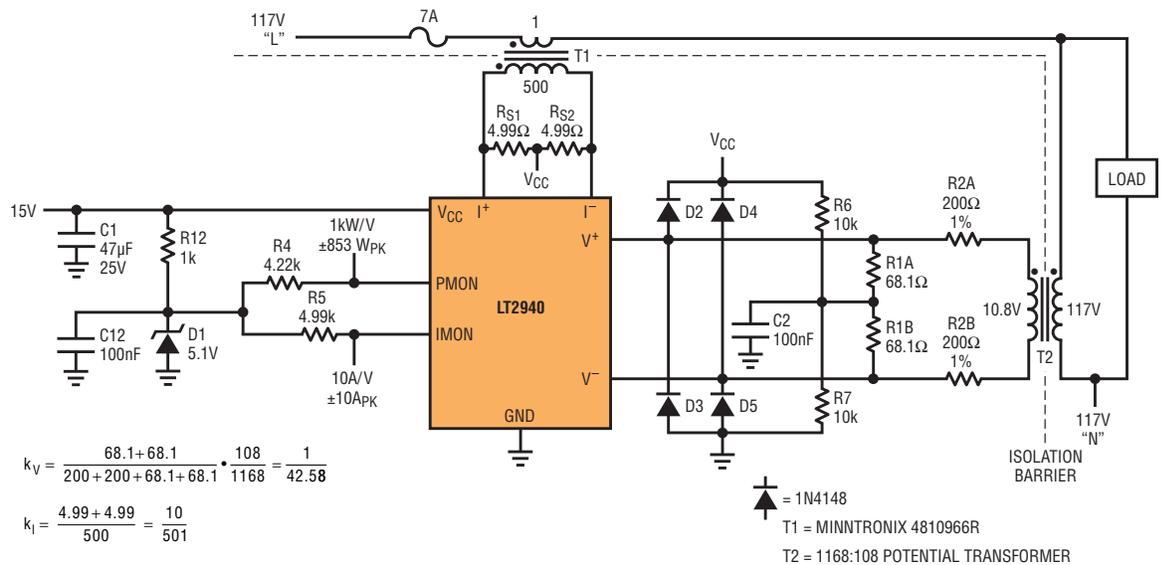
With isolation, the LT2940 is capable of monitoring the AC power of line operated loads, as shown in Figure 15. *Care must be exercised when working with AC line connected circuits.* To preserve output accuracy, a phase-accurate current transformer is essential—the component shown is capable of less than 1° phase error. The phase accuracy of

the potential transformer is equally critical, but accuracy is easily achieved using an off-the-shelf line transformer. Note that the constants  $k_V$  and  $k_I$  are inclusive of transformer ratios.

### CONCLUSION

The LT2940 brings together the features that make power monitoring and control not only possible, but easy. It is available in both a leadless 12-pin DFN (3mm × 3mm) and a 12-lead MSOP package, and is featured in the DC1495A evaluation kit. ■

Figure 15. A fully isolated AC power and current monitor



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# 2-Phase Synchronous Step-Down DC/DC Controller with Programmable Stage Shedding Mode and Active Voltage Positioning for High Efficiency and Fast Transient Response

Jian Li and Charlie Zhao

The LTC3856 is a versatile and feature-rich single-output 2-phase synchronous buck controller with on-chip drivers, remote output voltage sensing, inductor DCR temperature compensation, Stage Shedding™ mode and active voltage positioning (AVP). It is suitable for converting inputs of 4.5V–38V to outputs from 0.6V up to 5V. The LTC3856 facilitates the design of high efficiency, high power density solutions for telecom and datacom systems, industrial and medical instruments, DC power distribution systems and computer systems. The controller is available in 32-pin 5mm × 5mm QFN and 38-pin TSSOP packages.

## MAJOR FEATURES

The LTC3856's constant-frequency peak current-mode control architecture allows a phase-lockable frequency of up to 770kHz. For high frequency applications, the LTC3856 can operate at low duty cycles due to its small minimum on-time (90ns),

making it possible to produce a large step-down ratio applications in very little space.

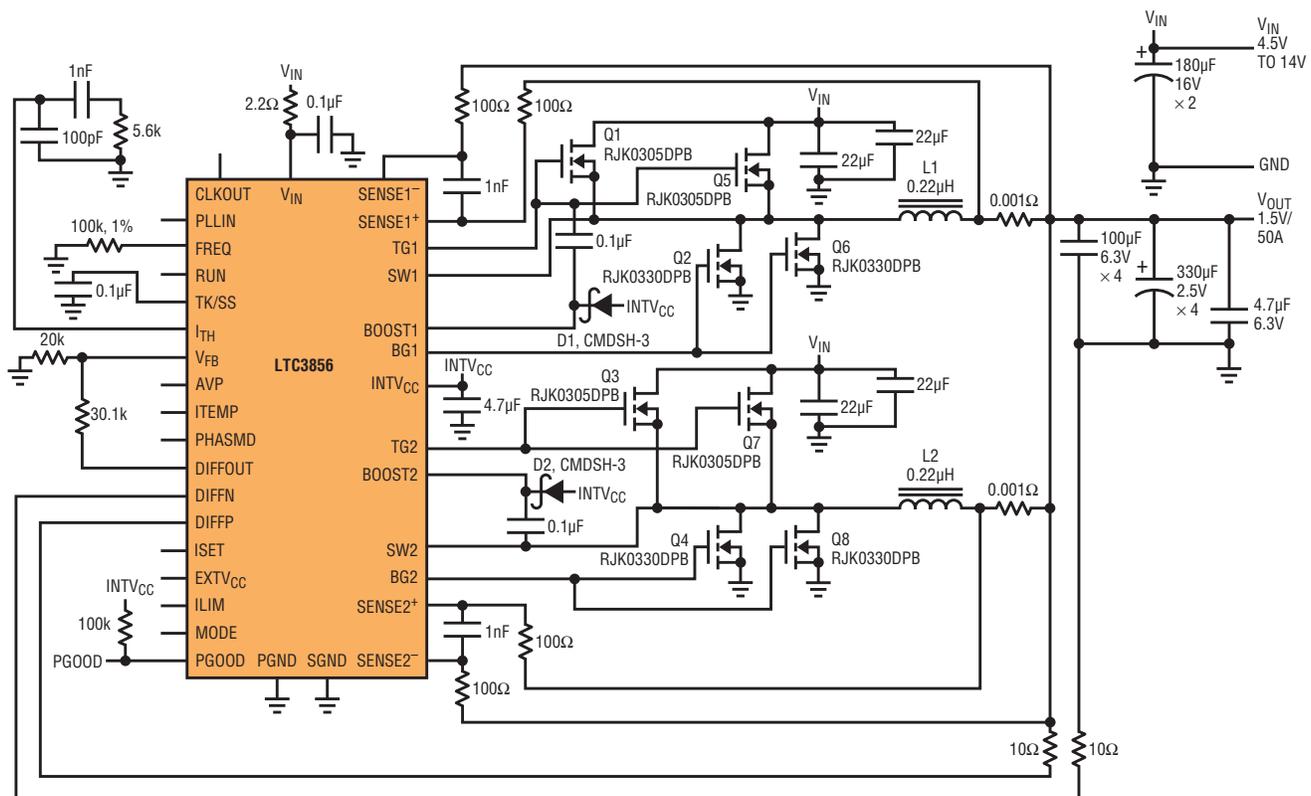
The LTC3856 includes a high speed differential amplifier for remote output voltage sensing, which can eliminate the regulation error due to PCB voltage

drops during heavy load conditions.

Figure 1 shows a typical 4.5V~14V input, 1.5V/50A output application schematic.

The LTC3856's two channels operate anti-phase, which reduces the input RMS current ripple and thus the input

Figure 1. A 1.5V/50A, 2-phase converter featuring the LTC3856



At light load, switching-related power losses dominate the total loss. With Stage Shedding mode, the LTC3856 can shut down one channel at light loads to reduce switching related losses.

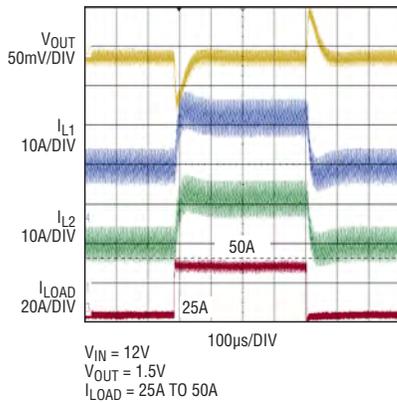


Figure 2. Load transient performance

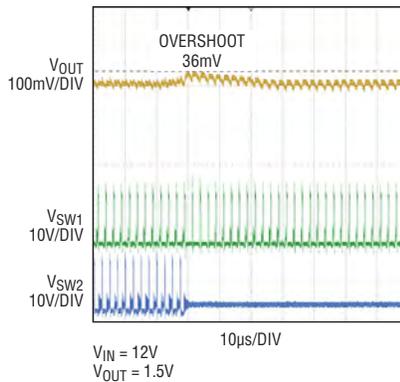


Figure 3. Stage Shedding mode: 2-phase to 1-phase transition

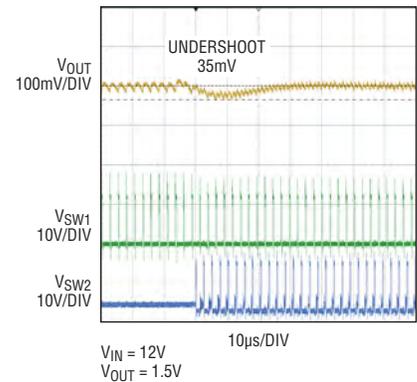


Figure 4. Stage Shedding mode: 1-phase to 2-phase transition

capacitance. Up to six LTC3856s can be combined for 12-phase operation by using the CLKOUT, PLLIN and PHASMD pins. Due to its peak current mode control architecture, the LTC3856 provides fast cycle-by-cycle dynamic current sharing plus tight DC current sharing, as shown in Figure 2.

The LTC3856's maximum current sense voltage is selectable for either 30mV, 50mV or 75mV, allowing the use of either the inductor DCR or a discrete sense

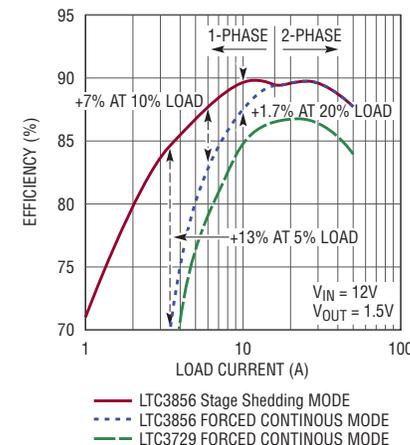


Figure 5. Efficiency comparison

resistor as the sensing element. Inductor winding resistance (DCR) changes over temperature, so the LTC3856 senses the inductor temperature via the ITEMP pin and maintains a constant current limit over a broad temperature range. It makes high efficient inductor DCR sensing more reliable for high current applications.

At heavy load, the LTC3856 operates in constant frequency PWM mode. At light loads, it can operate in any of three modes: Burst Mode® operation, forced continuous mode and Stage Shedding™ mode. Burst Mode operation switches in pulse trains of one to several cycles, with the output capacitors supplying energy during internal sleep periods. This provides the highest possible efficiency at very light load. Forced continuous conduction mode (CCM) offers continuous PWM operation from no load to full load, providing the lowest possible output voltage ripple. Programmable Stage Shedding mode is unique to the LTC3856. In Stage Shedding mode, one channel can be shut down at

light load to reduce switching related losses, thus improving efficiency in the load range up to 20% of full load.

The programmable active voltage positioning (AVP) is another unique design feature of the LTC3856. AVP modifies the regulated output voltage depending on its current loading. AVP can improve overall transient response and save output capacitors.

#### STAGE SHEDDING MODE

At light load, switching-related power losses dominate the total loss. With Stage Shedding mode, the LTC3856 can shut down one channel at light loads to reduce switching related losses. When the MODE pin is tied to INTVCC, the

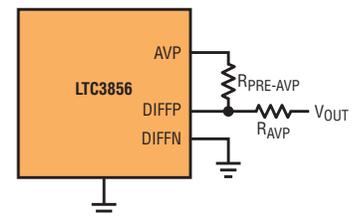


Figure 6. Programmable AVP

The LTC3856's two channels operate anti-phase, which reduces the input RMS current ripple and thus the input capacitance. Up to six LTC3856s can be combined for 12-phase operation. Due to its peak current mode architecture, the LTC3856 provides fast cycle-by-cycle dynamic current sharing, plus tight DC current sharing.

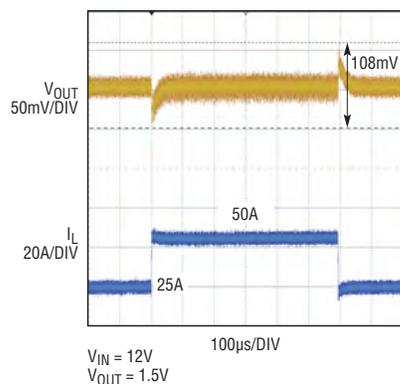


Figure 7. Transient performance without AVP

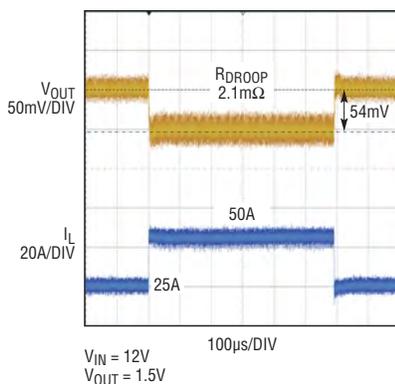


Figure 8. Transient performance with AVP

LTC3856 enters Stage Shedding mode. This means that the second channel stops switching when  $I_{TH}$  is below a certain programmed threshold. The threshold voltage  $V_{SHED}$  on  $I_{TH}$  is programmed according to the following formula:

$$V_{SHED} = 0.5 + \frac{5}{3}(0.5 - V_{ISET})$$

There is a precision 7.5µA flowing out of the ISET pin. Connecting a resistor to SGND sets the  $V_{ISET}$  voltage.

Current mode control allows the LTC3856 to transition smoothly from 2-phase to 1-phase operation and vice versa, as shown in Figures 3 and 4. A voltage mode, multi-phase supply cannot transition between 1- and 2-phase operation as smoothly.

The efficiency improvements brought on by Stage Shedding mode are shown in Figure 5. Due to stronger gate driver and shorter dead-time, the LTC3856 can achieve around 4%~5% higher efficiency than the LTC3729, a comparable single-output, 2-phase controller, over the whole load range. With Stage Shedding

mode, significant efficiency improvement is further achieved at light load. At 5% load, the efficiency is improved by 13%.

#### ACTIVE VOLTAGE POSITIONING

Transient performance is an important parameter in the requirements for the latest power supplies. To minimize the voltage excursions during a load step, the LTC3856 uses AVP to lower peak-to-peak output voltage deviations caused by load steps without having to increase the output filter capacitance. Likewise, the output filter capacitance can be reduced in applications while maintaining peak-to-peak transient response.

The LTC3856 senses inductor current information by monitoring the voltage across the sense resistors  $R_{SENSE}$  or the DCR sensing network of the two channels. The voltage drops are added together and applied as  $V_{PRE-AVP}$  between the AVP and DIFFP pins, which are connected through resistor  $R_{PRE-AVP}$ . Then,  $V_{PRE-AVP}$  is scaled through  $R_{AVP}$  and added to the output voltage as the compensation

for the load voltage drop. As shown in Figure 6, the load slope ( $R_{DR00P}$ ) is set to:

$$R_{DR00P} = \frac{R_{SENSE} \cdot R_{AVP}}{R_{PRE-AVP}} \cdot \frac{V}{A}$$

With proper design, AVP can reduce load transient-induced peak-to-peak voltage spikes by 50%, as shown in Figures 7 and 8.

#### CONCLUSION

The LTC3856 delivers an oversized set of features for its small 5mm × 5mm 32-pin QFN package. It can run at high efficiency using temperature compensated DCR sensing with Stage Shedding mode/Burst Mode operation. AVP can improve the transient response even as output capacitance is reduced. Tracking, strong on-chip drivers, multichip operation and external sync capability fill out its menu of features. The LTC3856 is ideal for high current applications, such as telecom and datacom systems, industrial and computer systems applications. ■

# Choose a Regulator with an Accurate Input Current Limit to Safely Extract Maximum Power from USB

Albert Lee

Most of us would not use a paintbrush to sign our names on our checks. It's simply the wrong tool for the job—wasting paint where a thin line of ink offers better results. Likewise, using a power supply IC with a broad-brush  $\pm 40\%$  output current limit is wasteful, requiring a designer to leave 80% of the input supply on the table, or worse, ignore the tolerance and risk collapse of the input supply. To design an input-current-limited supply that maximizes input power usage requires a detailed approach, ideally incorporating a regulator with tight tolerances.

The LTC3619 and LTC3619B are 400mA and 800mA dual monolithic synchronous buck regulators with tight  $\pm 5\%$  programmable average input current limits. The LTC3619 uses Burst Mode operation to improve efficiency at light loads, while the LTC3619B uses pulse-skipping to improve efficiency and also reduce noise. The accurate input current limit allows utilization of 90% of the maximum input current for fast supercap charging, strong signal and lag-free operation without risk of collapsing the input supply.

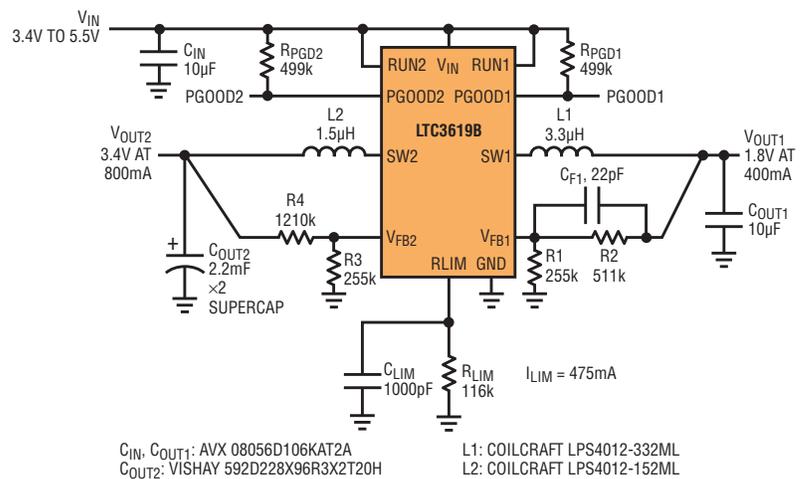


Figure 1. GSM receiver power supply operates from the USB. Although GSM receivers require bursts of power beyond the USB current limit, this design complies with USB current limits by incorporating a supercap to provide short bursts of current at  $V_{OUT2}$ . Only  $V_{OUT2}$  is actively input current limited. The current at  $V_{OUT1}$  must operate within the power constraints of the USB input, but it is not actively input current limited, so the voltage at  $V_{OUT1}$  remains stable for important circuits.

An increasing number of portable electronics are powered from the USB, which is current limited to 500mA. Surges in current on the USB commonly occur from plugging a device into a port. If the surge current is high, non-current-limited load dumps at the USB port can glitch the source power supply, which can affect other systems that depend on the supply. Plugging in an improperly current-limited USB device into a laptop can glitch the laptop CPU, causing it to lock up or reboot. High peak current pulses during GSM wireless data transfers can also cause supply glitches. The accurate current limits of the LTC3619 and LTC3619B protect the supply while maximizing usage of available input current.

## GSM APPLICATION

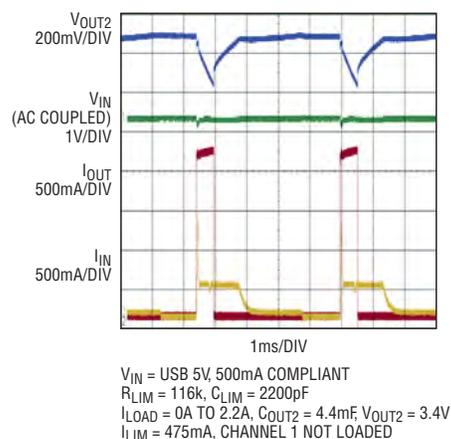
Users expect a high level of mobile functionality in their electronic devices—they want their GSM modems to detect a strong signal at all corners of the city. As more current is required for better transmission and reception, the importance of an accurate input current limit cannot be ignored.

Programming the LTC3619B's accurate input current limit to maximize available current usage is simple with an external resistor  $R_{LIM}$  and capacitor  $C_{LIM}$ , sized using the following expression:

$$R_{LIM} = \frac{55k\Omega \cdot A}{I_{DC}}$$

Determine  $R_{LIM}$  for the average input current being limited ( $I_{DC}$ ), and choose

Users expect a high level of mobile functionality in their electronic devices—they want their GSM modems to detect a strong signal at all corners of the city. As more current is required for better transmission and reception, the importance of an accurate input current limit cannot be ignored.



**Figure 2. Operation of the GSM modem power supply showing bursts of current beyond the input current limit and quick charging of the supercap on  $V_{OUT2}$ . Note the stability of  $V_{IN}$ .**

$C_{LIM}$  greater than 100pF for averaging  $I_{LIM}$  current. When the sum of both channels' current exceeds the input current limit, only channel 2 is current limited while channel 1 remains regulated. Input current limit starts when  $V_{RLIM} = 1V \pm 5\%$ .

The LTC3619B's input current limit is trimmed to less than 2% at room temperature and deviates no more than a few percent over temperature. (See the LTC3619B data sheet for detailed explanation in selecting  $R_{LIM}$  and  $C_{LIM}$ .) Figure 1 shows the LTC3619B in a solution that converts USB input ( $V_{IN} = 5V$ ) to  $V_{OUT2} = 3.4V$  to deliver GSM pulsed current load. Figure 2 shows the GSM output waveforms.

GSM modems demand high bursts of current, up to 2A, to the RF power amplifier, which well exceeds the maximum 500mA input current available via USB. In Figure 1, LTC3619B quickly charges a reservoir cap or supercap, which in turn can adequately provide bursts of current.

Because channel 1 is not input current limited, its output voltage does not collapse even as the GSM modem draws high current bursts. Thus, it is safe to use this channel to power up base-band chips, power management ICs or

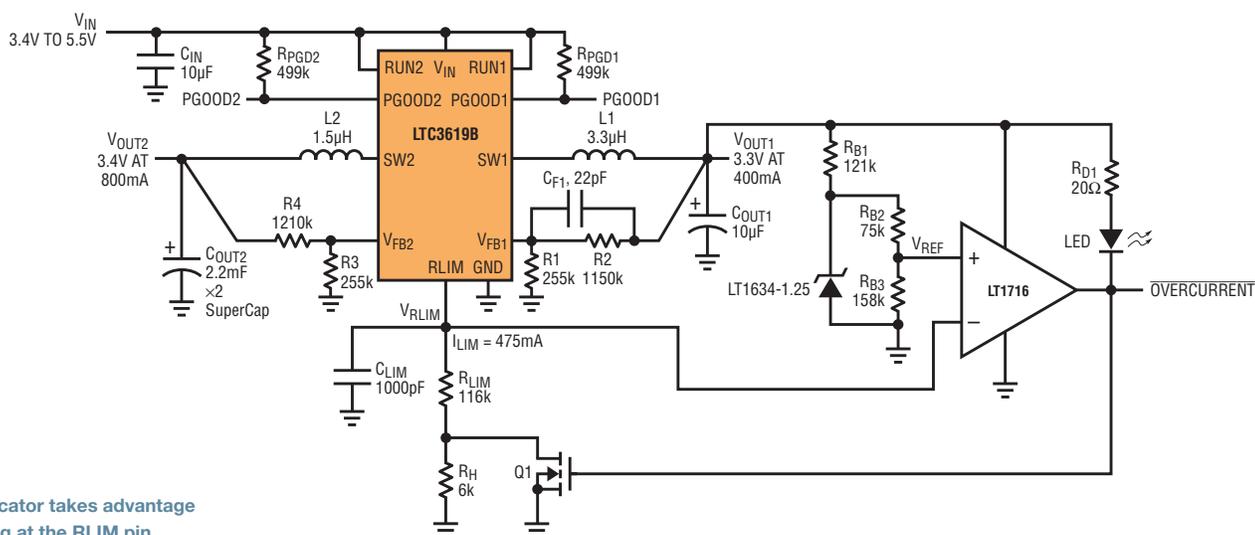
keep-alive circuitry while maximizing the available current for channel 2.

### $V_{OUT}$ POWER GOOD

The PGOOD indicator pins are useful for monitoring the regulation status of the two outputs. The PGOOD pins are open-drain outputs—pulled high with an external pull-up resistor when in regulation. In Figure 1, if the supercap is not fully charged or the output is not in regulation, the PGOOD2 pin is pulled low by the internal NFET. A red LED indicator can be used instead of a pull-up resistor. Of course, PGOOD can be used to handshake with other circuitry, providing a ready signal for the next load dump.

### OVERCURRENT INDICATOR

In applications that require additional system control, the LTC3619B provides accurate current sense information for both channels. This information can be used for protection schemes, feedback control and other features. Figure 3 shows a scheme for an LED overcurrent indicator.



**Figure 3. Overcurrent indicator takes advantage of input current monitoring at the RLIM pin.**

Figure 4 shows the relationship between the  $V_{RLIM}$  voltage and the input current. Set  $V_{REF}$  equal to  $V_{RLIM}$  voltage at the overcurrent limit of interest. For example, if the overcurrent limit were to be set at 85% of max current or 400mA, then set  $V_{REF} = (400\text{mA}/55\text{k}\Omega\text{-A}) \times 116\text{k}\Omega = 0.85\text{V}$ . The 0.85V reference is created via a resistor divider off of the LT1634 precision shunt voltage reference and is compared to  $V_{RLIM}$  using the LT1716 precision comparator to create the overcurrent signal.

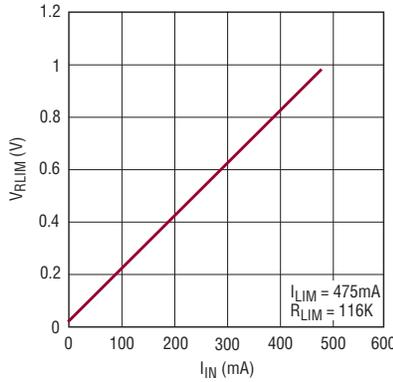


Figure 4.  $V_{RLIM}$  vs  $I_{IN}$  for circuit in Figure 3

In this example, when  $V_{RLIM}$  rises above  $V_{REF}$ , the LT1716 pulls  $\overline{\text{OVERCURRENT}}$  low and turns on LED1 while turning off Q1, allowing  $R_H$  to provide about 5% of hysteresis for  $V_{RLIM}$ .  $V_{RLIM}$  would have to be reduced by 5% in order to clear the  $\overline{\text{OVERCURRENT}}$  condition.

### POWERING AN LED LIGHT AT INPUT CURRENT LIMIT

If an application requires independent monitoring of only one channel, the LTC3606B pulse-skipping, single channel monolithic buck with input current limit is a good choice. The LTC3606B can be used to power an LED driver chip or to directly drive a large LED light, such as LED1 in Figure 5.

LED1 has a nominal on-voltage of 3.2V and its current is limited by the input current limit of 400mA set by  $R_{LIM}$ . In this case, the current through LED1 is limited to 625mA, as calculated from  $(V_{IN}/V_{OUT}) \times I_{LIM} = (5\text{V}/3.2\text{V}) \times 400\text{mA}$ , assuming  $V_{IN} = 5\text{V}$ . The circuit is configured so that the  $\overline{\text{LAMPGOOD}}$  indicator goes high when LED1 turns off or burns out. When turned on, the current through LED1 is at the input current limit and the voltage at  $V_{OUT}$  is 3.2V instead of the regulated 4V. Because the operating LED forces  $V_{OUT}$  more than 11% out of regulation, PGOOD (a.k.a.  $\overline{\text{LAMPGOOD}}$ ) falls low, indicating the lamp is on. If LED1 turns off or burns out (no current

through it),  $V_{OUT}$  returns to the regulated 4V and  $\overline{\text{LAMPGOOD}}$  is pulled high via  $R_{PGD}$ , indicating that the lamp is not operating.

If the LTC3619B were used in the previous example, the available input current to channel 2 would be dependent on the input current of channel 1. Using the expression below, the current out of  $R_{LIM}$  pin can be calculated. This is the summed representation of the inductor currents from both channels and illustrates how the currents are distributed.

$$I_{R(LIM)} = I_{OUT1} \times D1 \times K1 + I_{OUT2} \times D2 \times K2,$$

where  $D1 = V_{OUT1}/V_{IN}$  and  $D2 = V_{OUT2}/V_{IN}$  are the duty cycles of channels 1 and 2, respectively.

$K1$  and  $K2$  are the ratio

$$\frac{R_{DS(ON)}(\text{POWER PFET})}{R_{DS(ON)}(\text{SENSE PFET})}$$

of channels 1 and 2, respectively and are internally trimmed to better than 2% accuracy at 1/55kΩ-A. Assuming  $K = K1 = K2$ , and dividing both sides by

$K$ , the input current is derived. Setting input current to the input current limit, we get the following expression.

$$I_{LIM} = I_{OUT1} \times D1 + I_{OUT2} \times D2,$$

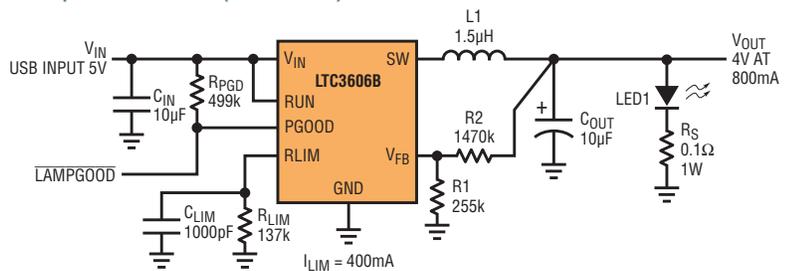
Channel 2 is configured to power LED1 at  $V_{OUT2} = 3.2\text{V}$ . If channel 1 is loaded at 400mA with  $V_{OUT1} = 1.8\text{V}$ , which translates to 144mA of input current ( $I_{OUT1} \times D1$ ), this subsequently leaves 400mA through LED2 instead of the original 625mA, which translates to 256mA of input current ( $I_{OUT2} \times D2$ ) available for channel 2.

According to the above  $I_{LIM}$  expression, a higher LED1 turn-on voltage, for instance due to looser manufacturing tolerance, would result in reduced current through LED1 and vice versa. This is an appealing, self-adjusting feature in this application to keep the intensity constant over manufacturing differences. Using an LED driver IC to power multiple LEDs on channel 2 is preferred although the LTC3606B may be used to drive an LED light bulb efficiently.

### CONCLUSION

The LTC3619 and the LTC3606B are buck regulators that combine average input current limit and current sense information in a 10-lead MSOP and DD package. The LTC3619's accurate input current limit is ideal for USB powered applications, where the USB port's output current is limited. At the same time, the current sense information simplifies designs in applications that require detection and monitoring of current in a single-chip solution requiring no additional board area. ■

Figure 5. LED driver with lamp status indicator ( $\overline{\text{LAMPGOOD}}$ )



# Fast Time Division Duplex (TDD) Transmission Using an Upconverting Mixer with a High Side Switch

Vladimir Dvorkin

Many wireless infrastructure time division duplex (TDD) transmit applications require fast on/off switching of the transmitter, typically within one to five microseconds. There are several different ways to implement fast Tx on/off switching, including the use of RF switches in the signal path, or on/off switching of the supply voltage for different stages of the transmitter chain. The advantages of the latter method are low cost, very good performance and power saving during the Tx off-time. In particular, a good place to apply supply switching is at the transmit upconverting mixer because this removes both the transmit signal and all other mixing products from the mixer RF output.

The LT5579 high performance upconverting mixer fits various TDD and Burst Mode transmitter applications with output frequencies up to 3.8GHz. Fast on/off supply voltage ( $V_{CC}$ ) switching for the LT5579 is as simple as adding an external high side power supply switch (note that this technique is equally effective for the lower frequency upconverting mixer, LT5578).

## HIGH SIDE $V_{CC}$ SWITCH FOR A BURST MODE TRANSMITTER USING THE LT5579 MIXER

The high side  $V_{CC}$  switch circuit in Figure 1 uses a P-channel MOSFET (IRLML6401) with an  $R_{DS(ON)}$  of less than  $0.1\Omega$ . An N-channel enhancement mode FET (2N7002), connected from the drain of IRLML6401 to ground, further improves fall time.

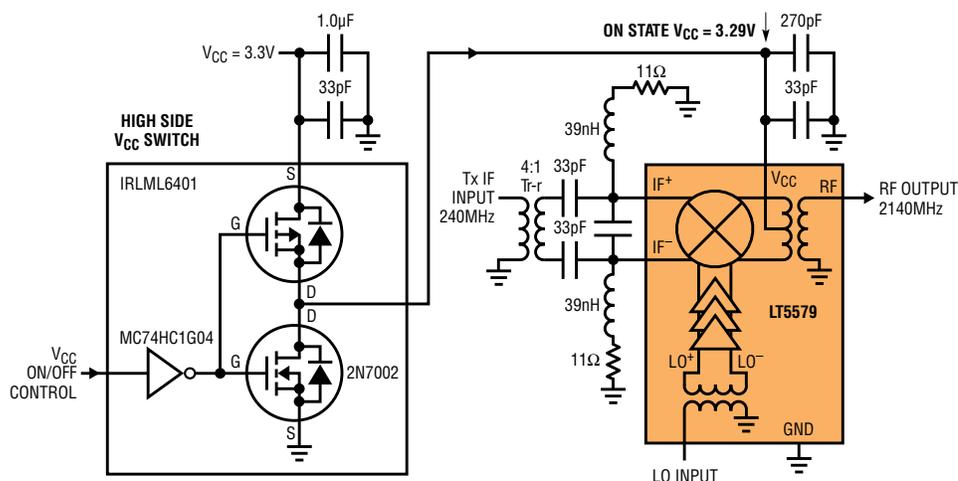


Figure 1. Upconverting mixer with high side  $V_{CC}$  switch

The 2N7002's  $R_{DS(ON)}$  is less than  $4\Omega$ , which is sufficient for this application.

The input driver for the high side  $V_{CC}$  switch is a high speed CMOS inverter (MC74HC1G04) capable of driving capacitive loads. The IRLML6401 input capacitance is typically  $830pF$  and the 2N7002 input capacitance is under  $50pF$ . For faster rise times, two high speed CMOS drivers can be used in parallel. Likewise, for faster fall times, a different N-channel MOSFET with lower on-resistance can be used.

With the LT5579 supply current of  $220mA$ , the power supply voltage drop across the MOSFET is only  $11mV$ . The response time of the high side  $V_{CC}$  switch is shown in Figure 2. Total turn-on time is only  $650ns$  and total turn-off time is  $500ns$ . These measurements were performed using two RF bypass capacitors at the mixer  $V_{CC}$  pin ( $33pF$  and  $270pF$ ). Higher value RF bypass

(continued on page 27)

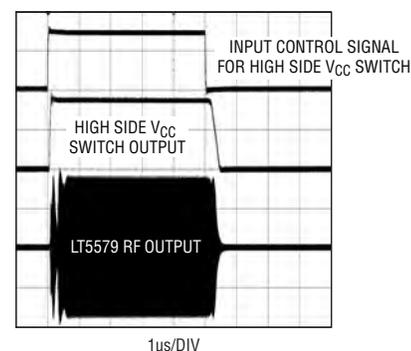


Figure 2.  $V_{CC}$  turn-on and turn-off waveforms

# Driving Lessons for a Low Noise, Low Distortion, 16-Bit, 1MSPS SAR ADC

Guy Hoover

It is challenging to design an ADC driving topology that delivers uncompromising performance, especially when designing around an ultralow noise SAR ADC such as the 1MSPS LTC2393-16. For both single-ended and differential applications, a well thought out driving topology fully realizes the ultralow noise and low distortion performance required in your data acquisition system.

The LTC2393-16 is the first in a family of high performance SAR ADCs from Linear Technology that utilizes a fully differential architecture to achieve an excellent SNR of 94.2dB and THD of  $-105$ dB. In order to take full advantage of the ADC performance, we present driving solutions for both single-ended and differential applications. Both topologies fully demonstrate the ultralow noise and low distortion capabilities of the LTC2393-16.

## SINGLE-ENDED TO DIFFERENTIAL CONVERTER

The circuit of Figure 1 converts a single-ended 0V to 4.096V signal to a differential  $\pm 4.096$ V signal. This circuit is useful for sensors that do not produce a differential signal. Resistors R1, R2 and capacitor C2 limit the input bandwidth to approximately 100kHz.

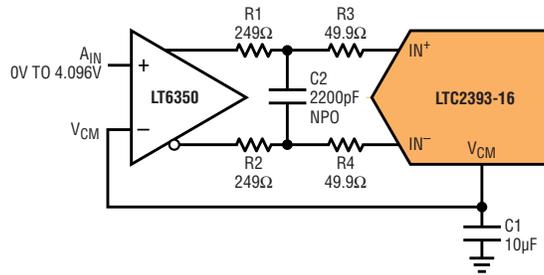


Figure 1. Single-ended to differential converter

When driving a low noise, low distortion ADC such as the LTC2393-16, component choice is essential for maintaining performance. All of the resistors used in this circuit are relatively low values. This keeps the noise and settling time low. Metal film resistors are recommended to reduce distortion caused by self-heating. An NPO capacitor is used for C2 because of its low voltage coefficient, which minimizes distortion. The excellent linearity characteristics of NPO and silver mica capacitors make these good choices for low distortion applications. Finally, the LT6350 features low noise, low distortion and a fast settling time.

The 16k-point FFT in Figure 2 shows the performance of the LTC2393-16 in the circuit of Figure 1. The measured SNR of 94dB and THD of  $-103$ dB match closely with the typical data sheet specs for the LTC2393-16, showing that little, if any, degradation of the ADC's specifications result from inserting the single-ended to differential converter into the signal path.

## FULLY DIFFERENTIAL DRIVE

The circuit of Figure 3 AC-couples and level shifts the sensor output to match the common mode voltage of the ADC. The lower frequency limit of this circuit is about 10kHz. The lower frequency limit can be extended by increasing the values of C3 and C4. This circuit is useful for sensors with low impedance differential outputs.

The circuit of Figure 1 could be AC-coupled in a similar manner. Simply bias AIN to VCM through a 1k resistor and couple the signal to AIN through a 10μF capacitor.

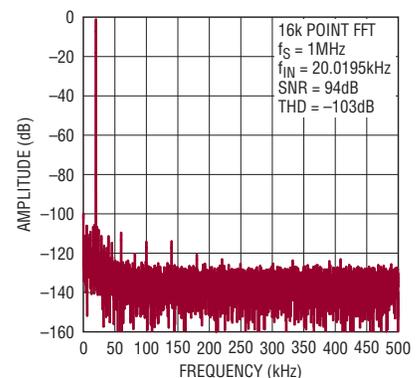


Figure 2. LTC2393-16 16k point FFT using circuit of Figure 1

The LTC2393-16 with its fully differential inputs can improve SNR by as much as 6dB over conventional differential input ADCs. This ADC is well suited for applications that require low distortion and a large dynamic range.

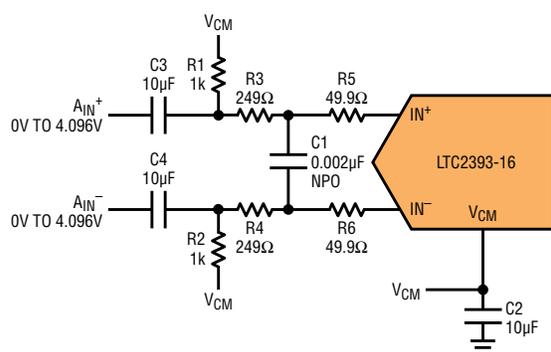


Figure 3. AC-coupled differential input

### PCB LAYOUT

The circuits shown are quite simple in concept. However, when dealing with a high speed 16-bit ADC, PCB layout must also be considered. Always use a ground plane. Keep traces as short as possible. If a long trace is required for a bias node such as  $V_{CM}$ , use additional bypass capacitors for each component attached to the node and make the trace as wide as possible. Keep bypass capacitors as close to the supply pins as possible. Each bypass capacitor should have its own low impedance

return to ground. The analog input traces should be screened by ground. The layout involving the analog inputs should be as symmetrical as possible so that parasitic elements cancel each other out.

Figure 4 shows a sample layout for the LTC2393-16. Figure 4 is a composite of the top metal, ground plane and silk-screen layers. See the DC1500A Quick Start Guide at [www.linear.com](http://www.linear.com) for a complete LTC2393-16 layout example.

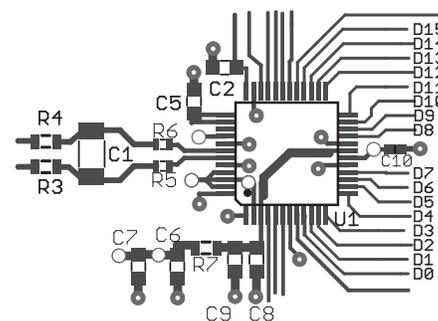


Figure 4. Sample layout for LTC2393-16

### CONCLUSION

The LTC2393-16 with its fully differential inputs can improve SNR by as much as 6dB over conventional differential input ADCs. This ADC is well suited for applications that require low distortion and a large dynamic range. Realizing the potential low noise, low distortion performance of the LTC2393-16 requires combining simple driver circuits with proper component selection and good layout practices. ■

(LT5579 continued from page 25)

capacitors can be used, which would result in correspondingly slower rise/fall times.

The LT5579 upconverting mixer circuit shown in Figure 1 was optimized and tested at an RF output frequency of 2140MHz. The RF output envelope in Figure 2 shows a dip about 300ns after the  $V_{CC}$  switch turns on, followed by another, smaller dip at about the 500ns point. Both dips represent the mixer's internal feedback circuit reaction to the ramping supply voltage.

LO leakage to the RF output of the LT5579 was measured at  $-40\text{dBm}$  when  $V_{CC}$  is on and  $-46\text{dBm}$  for  $V_{CC}$  off. The LO port of the LT5579 is internally matched and has a return loss of 10dB to 18dB over a frequency range of 1100MHz to 3200MHz.

When the LT5579 mixer is in the off state, the return loss of the LO port is about 3dB to 5dB across the same frequency range of 1100MHz to 3200MHz. It is advisable to use an LO injection VCO with a buffered output for better reverse isolation,

and to avoid any VCO pulling while the LO port impedance changes when switching between the on and off states.

### CONCLUSION

LT5579 and LT5578 mixers without an ENABLE pin can be used in TDD applications with external  $V_{CC}$  switching. Using only three parts (IRLML6401, 2N7002 and an MC74HC1G04), a high performance high side  $V_{CC}$  switch allows turn-on and turn-off in under 1µs. ■

# Ultrafast, Low Noise, Low Dropout Linear Regulators Running in Parallel Produce Clean, Efficient, High Current Point-of-Load Power for FPGA and Server Backplanes

Kelly Consoer

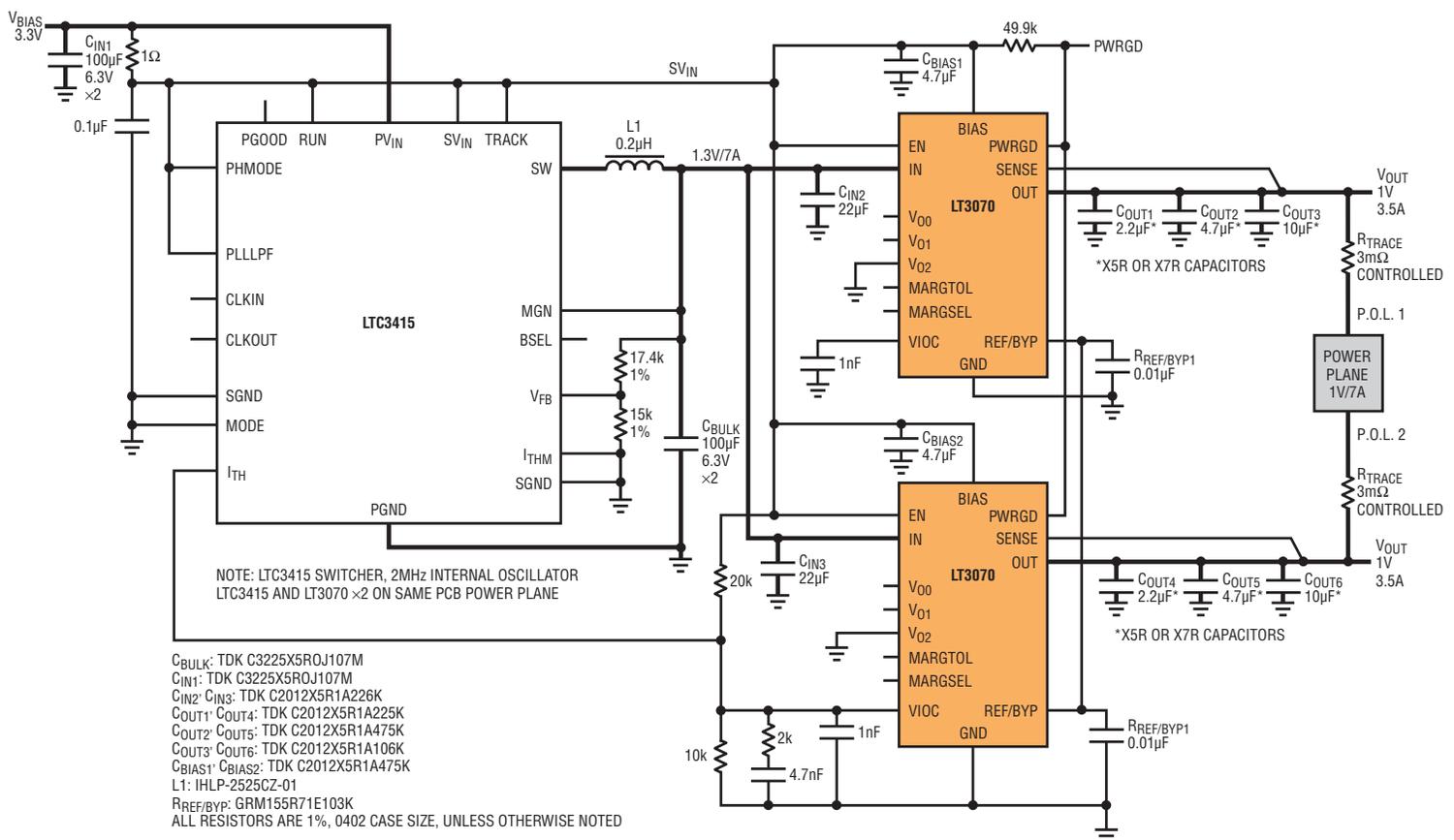
The latest FPGAs and processors have migrated to deep submicron geometries with gigahertz+ data rates and integrated telemetry channels that operate from 0.9V to 1.8V supply rails. The transient power demand for these processors can reach over 10A peak current in nanoseconds. At these processor speeds and high currents even the local PCB backplane impedances around the processor can contribute to supply droop, where millivolts of supply droop or supply noise can degrade data integrity.

To manage the system power distribution, switching regulators may be used locally to downconvert from a higher voltage rail. These regulators are traditionally bypassed with a large quantity of bulk caps to reduce ripple and buffer the load

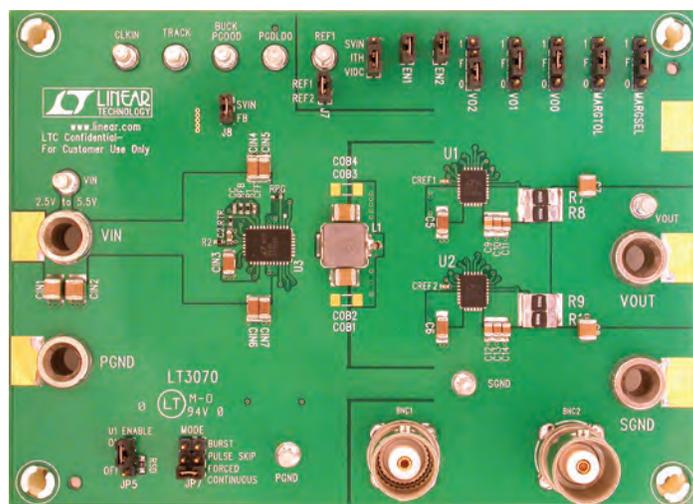
transients. Unfortunately, bulk tantalum and electrolytic caps have parasitic ESL and ESR that limits their ability to bypass fast high current load transients. To a lesser extent, even large ceramic caps are bandwidth limited by their inherent ESL.

The LT3070 and LT3071 family of UltraFast™ transient response, low noise, low dropout linear regulators addresses this challenge. The high bandwidths of the LT3070 and LT3071 reduce the supply impedance at the point-of-load using

Figure 1. Paralleled LT3070s with VIIOC control of upstream LTC3415 switcher implementation



**Figure 2. Two paralleled LT3070s clean up the power produced by an LTC3415 switching regulator with minimal impact on efficiency and board real estate**



only a few small, low ESR, ceramic capacitors, saving bulk capacitance and cost. These linear regulators supply up to 5A of output current with a typical dropout voltage of 85mV. A 0.01 $\mu$ F reference bypass capacitor decreases output voltage noise to 25 $\mu$ V<sub>RMS</sub>, yielding an output that is not only responsive but also very quiet. The LT3070 and LT3071 incorporate a unique VIOC tracking function to control the switching regulator powering the input.

The LT3070 has digital controls that allow the user to margin the system output voltage in increments of  $\pm 1\%$ ,  $\pm 3\%$  or  $\pm 5\%$ . The LT3071 has an analog margining pin that allows the user to margin the system output voltage  $\pm 10\%$ . The LT3071 also has a IMON output current monitor to support system load current diagnostics.

The features included in the LT3070 and LT3071 make them ideal for high performance FPGAs, microprocessors or sensitive communication supply applications.

#### A TIME FOR SHARING

Multiple LT3070/71s can be paralleled to increase available output current with minimal ballasting. In fact, eight LT3070s have successfully been paralleled to share a common 30A load.

Simply tie the REF/BYP pins of the paralleled regulators together. This effectively gives an averaged value of multiple 600mV reference voltage sources. Tie the OUT pins of the paralleled regulators to the common load plane through a small piece of PC trace ballast or an actual surface mount sense resistor beyond the primary output capacitors of each regulator.

The required ballast is dependent on the application output voltage and peak load current. The recommended ballast is the value that contributes 1% to load regulation. For example, two LT3070/71 regulators configured to output 1V, sharing a 10A load require 2m $\Omega$  of ballast at each output. The Kelvin SENSE pins connect to the regulator side of the ballast resistors to keep the individual control loops from conflicting with each other, as shown in Figure 1. Keep this ballast trace area free of solder to maintain a controlled resistance.

#### SIMPLIFIED INTEGRATION WITH SWITCHING REGULATORS YIELDS HIGH EFFICIENCY

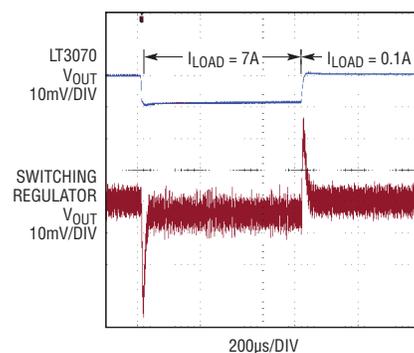
Figure 1 illustrates the use of the LT3070's VIOC function. This feature transfers control of the switcher output (LDO input) to one of the LDOs such that the digital output control of the LDO sets the output of the LTC3415 to maintain 300mV headroom  $V_{IN}$  to  $V_{OUT}$ , across the LDOs, optimizing power dissipation on the fly.

The demo board in Figure 2 demonstrates a single LTC3415 switching regulator supplying two LT3070 LDOs with co-joined outputs ballasted by sense resistors. Close examination of this board layout reveals a pair of small routes from the switcher PGND to the LDO SGND. Likewise, similar routes are incorporated

in a buried layer between the switcher output and the LDO inputs, terminated by the LDO input decoupling capacitors. These serve as  $\pi$ -filters to help isolate the LDO reference from the switching noise.

Figure 3 illustrates the quiet performance the two LT3070s sharing a pulsed 0.1A to 7A load while exhibiting less than 10mV of excursion on the output load. This is in sharp contrast to the relatively noisy load regulation waveform of the of the adjoining switching regulator.

In conclusion, the LT3070 and LT3071 provide power efficient, area efficient, UltraFast transient response, low noise, point-of-load regulation for the most demanding server applications. ■



**Figure 3. Waveform of paralleled LT3070s contrasted to the LTC3415 switcher output**

# Tiny Digital Predistortion Receiver Integrates RF, Filter and ADC

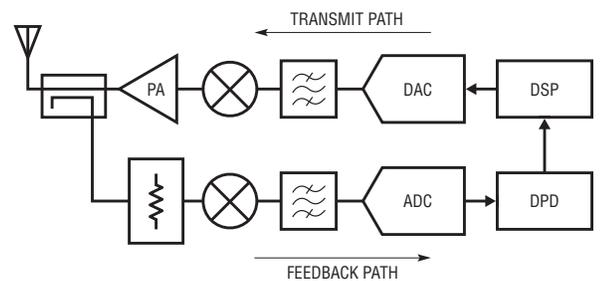
Todd Nelson

The power amplifier (PA) consumes more electrical power than any other block in a cellular basestation and is therefore a significant factor in the operating expense for the service provider. Complex digital modulation requires extremely high linearity from the PA, so it must be driven well below saturation where it is most efficient. To improve PA efficiency, designers use digital techniques to reduce the crest factor and improve PA linearity, allowing it to run closer to saturation. Digital predistortion (DPD) has emerged as the preferred method of PA linearization. A great deal of focus is paid to the DPD algorithm but another critical element is the RF feedback receiver.

## DPD RECEIVER REQUIREMENTS

The DPD receiver converts the PA output from RF back to digital as part of a feedback loop (see Figure 1). Key design requirements are the input frequency range and power level, the intermediate frequency and the bandwidth to be digitized. Some of these are derived

Figure 1. Digital predistortion signal chain



directly from the PA specifications and some are optimized at design time.

The baseband transmit signal is upconverted to the carrier frequency and is defined in frequency by the air interface standard: WCDMA, TD-SCDMA, CDMA2000, LTE, etc. Since the purpose of the DPD loop is to measure the PA transfer function, it is not necessary to separate the carriers or demodulate the digital data. PA nonlinearity produces odd order intermodulation products which constitute spectral regrowth in the adjacent and alternate channels. Third-order products appear within a range of three times the bandwidth of the desired channel (see Figure 2). Likewise, fifth-order products appear within a range of five times the bandwidth and seventh-order products within seven times the bandwidth. Therefore, the DPD receiver must acquire a multiple of the transmit bandwidth equivalent to the order of the intermodulation products being linearized.

The trend in current development is to mix the desired channel to an intermediate frequency (IF) and capture the full bandwidth of all the intermodulation products. The exact IF is chosen to ease

filtering and avoid other frequencies that are already fixed based on specification requirements. The sample rate is similarly chosen as a multiple of the digital modulation chip rate, for example, 3.84MHz in WCDMA. Finally, the Nyquist theorem dictates that the sample rate must be at least twice the sampled bandwidth. Although many configurations are acceptable, one that meets these constraints is an IF of 184.32MHz, an ADC sample rate of 245.76MHz and a bandwidth of 122.88MHz.

In the case of a 20W PA, the average output power is 43dBm. The peak to average ratio (PAR) is about 15dBm. To set the average input power to the mixer of the receive chain at -15dBm, the combination of the coupler and attenuator insertion loss needs to be 58dB (refer to Figure 1). The in-band noise of the PA is specified by the WCDMA standard at a maximum of -13dBm/MHz (-73dBm/Hz). Therefore, the combination of the coupler and attenuation (-58dB) and the PA noise limit (-13dBm/MHz) yields a receiver sensitivity level that must be below -71dBm/MHz (-131dBm/Hz). For sufficient margin, a number at least 6dB to 10dB better than this is desirable. This sets the frequency

At the board level, the  $\mu$ Module packaging integrates all of the key components into a small area including the passive filter and decoupling components. This can save significant board space, simplify layout and improve performance.

plan, power level and sensitivity requirements for the DPD receiver.

### INTEGRATED DPD RECEIVER

Once the system requirements are defined, the task turns to the circuit implementation using a mixer, IF amplifier, ADC, passive filtering, matching networks and supply bypassing. While calculations and simulations are helpful, there is no substitute for evaluation of real hardware, which generally leads to multiple printed circuit board (PCB) iterations. However, a new class of integrated receivers based on Linear Technology's  $\mu$ Module<sup>®</sup> packaging technology greatly simplifies this task. The LTM<sup>®</sup>9003 digital predistortion  $\mu$ Module receiver is a fully integrated DPD receiver—essentially RF-to-bits in a single device.

The LTM9003 consists of a high linearity active mixer, an IF amplifier, an L-C bandpass filter and a high speed ADC (see Figure 3). The wire-bonded bare die assembly ensures that the overall form factor is highly compact, but also allows the reference and supply bypass capacitors to be placed closer to the die than possible with traditional packaging. This reduces the potential for noise to degrade the fidelity of the ADC. This idea extends to the high frequency layout techniques used throughout the receiver chain of the LTM9003.

The integration eliminates many challenges of driving high speed ADCs. Linear circuit analysis cannot account for the current pulses resulting from the sample-and-hold switching action of the ADC. Traditional circuit layout requires multiple iterations to define an input network that

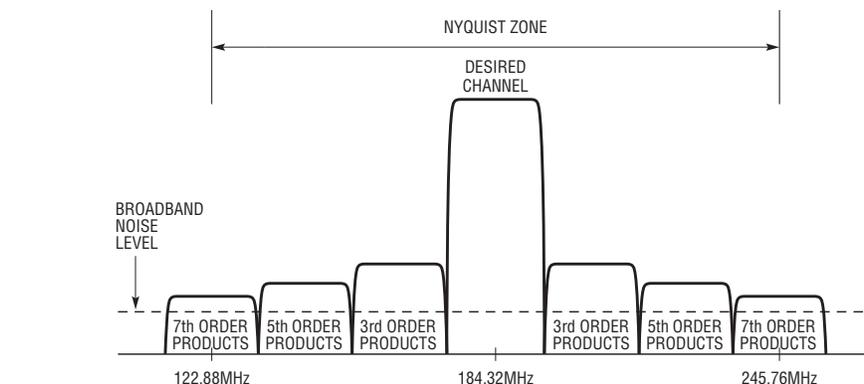


Figure 2. Intermodulation products

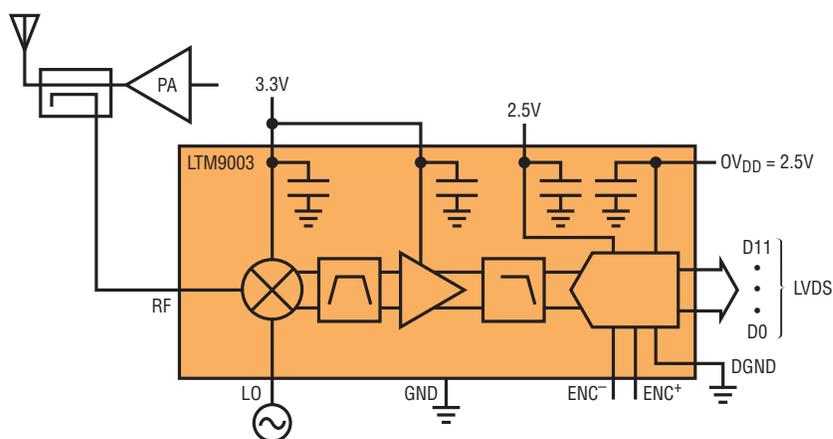
absorbs these pulses, is absorptive out of band, and yet works seamlessly with the preceding amplifier. The IF amplifier must also be capable of driving this network without adding distortion. Solving these challenges may be the greatest hidden attribute of the LTM9003  $\mu$ Module receiver.

The passive bandpass filter is a third order filter with an extremely flat pass-band. The center 25MHz of the band

exhibits less than 0.1dB ripple, and over the entire 125MHz the passband ripple is only 0.5dB. The third order configuration ensures that the shoulders of the frequency response are monotonic which is important for many DPD algorithms.

The overall performance of the LTM9003 greatly exceeds the system requirements described above. With a single tone at  $-2.5\text{dBm}$ , which is equivalent to  $-1\text{dBFS}$  at

Figure 3. LTM9003 integrated digital predistortion receiver



At the engineering level, the LTM9003 saves time. Filter design and component matching require PCB iteration to get it right. It is particularly challenging to design a filter that is undisturbed by the switching action of the ADC sample and hold circuitry. Even the placement of capacitors for supply decoupling affects overall performance and can cause board layout revisions.

the ADC, the signal to noise ratio (SNR) is typically  $-145\text{dBm/Hz}$ . This figure is well below the target value of  $-131\text{dBm/Hz}$  defined by the WCDMA standard. The worst-case harmonics are  $60\text{dBc}$ . The  $\text{IP}_3$  figure of  $25.7\text{dBm}$  means that the LTM9003 could support an ACPR of  $87\text{dBc}$  if the PA were linear enough. Relative to the system requirements and the capability of the best power amplifiers available, the LTM9003 greatly exceeds the requirements. The entire chain consumes about  $1.5\text{W}$  from a  $3.3\text{V}$  and a  $2.5\text{V}$  supply, yet requires a circuit board area of only  $11.25\text{mm} \times 15\text{mm}$ .

#### ALTERNATE CONFIGURATIONS

$\mu\text{Module}$  technology also offers an unexpected level of flexibility. By changing the values of the passive components or substituting ICs that are optimized as a group, the LTM9003 can be made available in application-specific versions, with no loss of performance or increased complexity.

For example, the LTM9003-AA utilizes a low power, silicon germanium active

mixer operating from a  $3.3\text{V}$  supply. The  $2 \times \text{RF} - 2 \times \text{LO}$  product gives a  $60\text{dBc}$  second harmonic, which is the worst spur in the spectrum. This can be improved at the expense of power consumption by replacing the mixer with a similar  $5\text{V}$  part. The second harmonic is then improved by  $4\text{dB}$  in the LTM9003-AB. Similarly, the sample rate can be reduced by substituting a  $210\text{MSps}$  ADC which consumes less power and the L-C filter values can be changed to realize a different filter bandwidth yet still achieve excellent passband flatness.

#### BIG BENEFITS IN SMALL PACKAGE

The benefits of using the LTM9003 for PA linearization occur at several levels. At a high level, DPD allows you to run the PA with less back-off. The result is that the PA is more efficient and therefore consumes less power for the same output power level.

At the board level, the  $\mu\text{Module}$  packaging integrates all of the key components into a small area including the passive filter and decoupling components. This

saves significant board space, simplifies layout and improves performance. The integration may enable a high performance remote radio head (RRH).

At the engineering level, the LTM9003 saves time. Filter design and component matching require PCB iteration to get it right. It is particularly challenging to design a filter that is undisturbed by the switching action of the ADC sample and hold circuitry. Even the placement of capacitors for supply decoupling affects overall performance and can cause board layout revisions. These tasks can consume months of engineering time to debug each revision and evaluate the changes. With the LTM9003, this work has already been done.

#### CONCLUSION

While the digital algorithms for DPD garner much attention, the analog receiver design is similarly demanding. The LTM9003  $\mu\text{Module}$  receiver simplifies this design by integrating the entire receiver in a single tiny package. ■

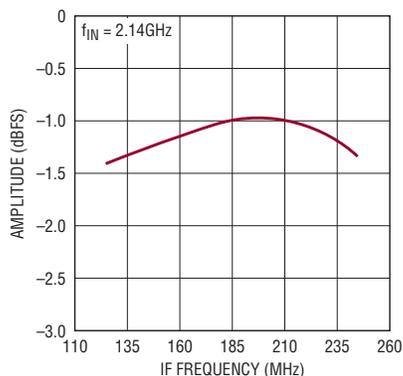


Figure 4. IF frequency response

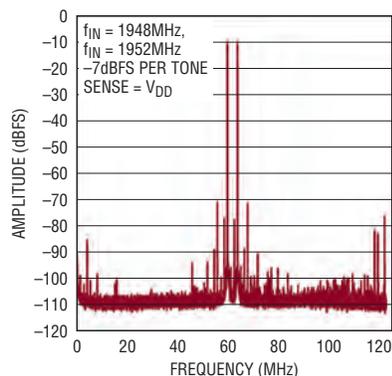


Figure 5. 64k point 2-tone FFT

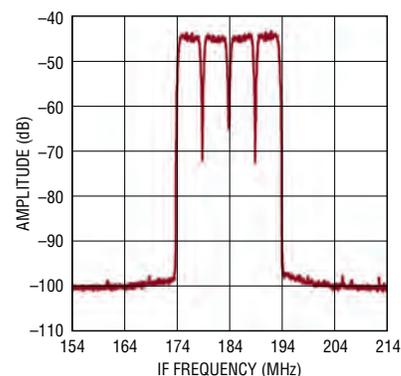


Figure 6. FFT of 4-channel WCDMA input at 2.14GHz

# Dual Output High Efficiency Converter Produces 3.3V and 8.5V Outputs from a 9V to 60V Rail

Victor Khasiev

The LTC3890 dual output DC/DC controller brings a unique combination of high performance features to applications that require low voltage outputs from high voltage inputs. It can produce two output voltages ranging from 0.8V to 24V from an input voltage of 4V to 60V. It is also very efficient, with a no-load quiescent current of only 50 $\mu$ A.

Many high-input-voltage step-down DC/DC converter designs use a transformer-based topology or external high side drivers to operate from up to 60V<sub>IN</sub>. Others use an intermediate bus converter requiring an additional power stage. However, the LTC3890 simplifies design, with its smaller solution size, reduced cost and shorter development time compared to other design alternatives.

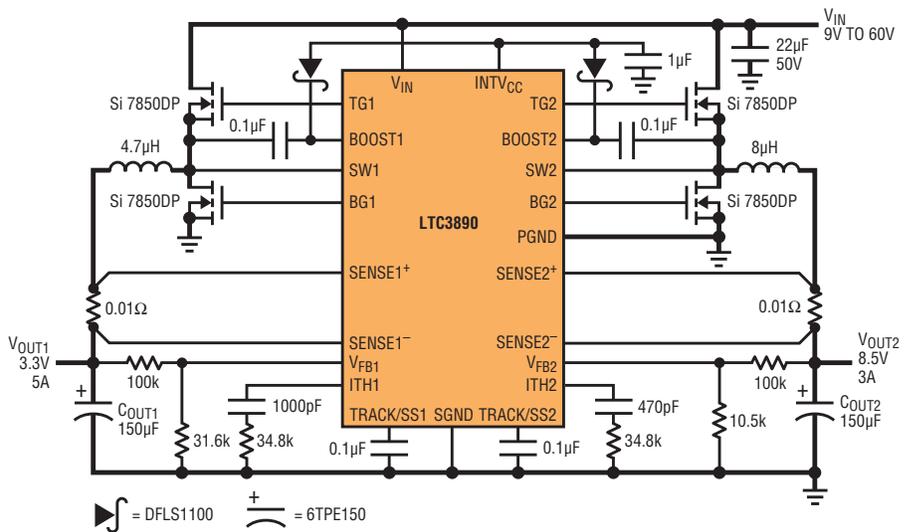


Figure 1. High efficiency dual 8.5V/3.3V output step-down converter

## FEATURE RICH

The LTC3890 is a high performance synchronous buck DC/DC controller with integrated N-channel MOSFET drivers. It uses a current mode architecture and operates from a phase-lockable fixed frequency from 50kHz to 900kHz. The device features up to 99% duty cycle capability for low voltage dropout

applications, adjustable soft-start or voltage tracking and selectable continuous, pulse-skipping or Burst Mode operation with a no-load quiescent current of only 50 $\mu$ A. These features, combined with a minimum on-time of just 95ns, make this controller an ideal choice for high step-down ratio applications. Power loss and

(continued on page 35)

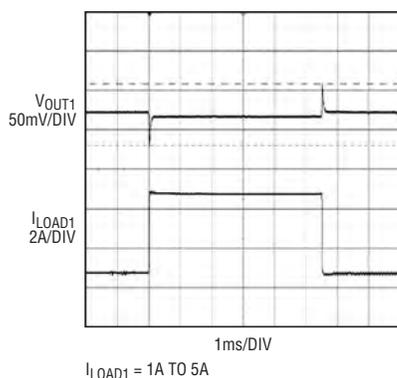


Figure 2. Transient response of 3.3V channel

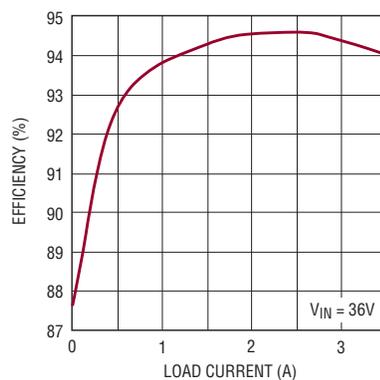


Figure 3. Efficiency of the converter in Figure 1 for the V<sub>OUT2</sub> 8.5V channel

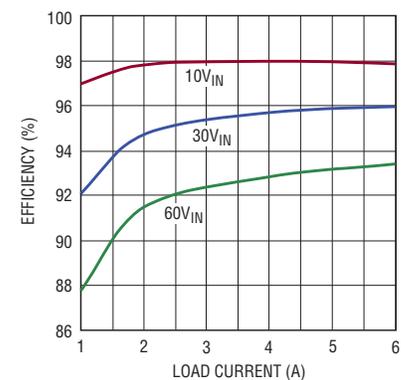


Figure 4. Efficiency of the LTC3890 configured as a 2-phase single output of 8.5V at up to 6A

# Product Briefs

## LOW VOLTAGE HOT SWAP CONTROLLER PROVIDES ADJUSTABLE FAULT TIMER

The LTC4280 is a low voltage Hot Swap controller with an onboard ADC and an I<sup>2</sup>C compatible interface. In many ways, the LTC4280 is the successor to the LTC4215. The FILTER pin of the LTC4280 replaces the soft-start (SS) pin on the LTC4215, allowing the pads for the SS capacitor to be used for an overcurrent filter capacitor without layout changes. In other respects it is pin compatible with LTC4215 and LTC4215-2 layouts.

In systems that are subject to fast input voltage steps and output current surges, the LTC4215 family of parts can have trouble meeting fault filtering requirements because of those parts' fixed overcurrent fault timers. The LTC4280 allows the filter time to be set to any value with the capacitor on the FILTER pin, which provides 123ms/μF of fault filtering.

After start-up, the LTC4280 has a current limit of 26mV with foldback down to 10mV and a 25mV circuit breaker, while the LTC4215 family has a current limit of 75mV without foldback and a 25mV circuit breaker. Reducing the current limit in the LTC4280 reduces power dissipation during overcurrent transients, which allows for transients 45 times longer than the LTC4215 while maintaining the same MOSFET safe operating area (SOA) performance.

The LTC4280 works in applications from 12V (with transients to 24V) down to 3.3V and is available in a 4mm × 5mm QFN package.

## 2.5μA QUIESCENT CURRENT, 62V, 350mA, 2.2MHZ STEP-DOWN DC/DC CONVERTER

The LT3990 is a 350mA, 62V ultralow quiescent current step-down switching regulator. Burst Mode operation keeps quiescent current under 2.5μA at no load. The LT3990's 4.2V to 62V input voltage range makes it ideal for automotive and industrial applications that need continuous output with ultralow power drain. Its internal 550mA switch can deliver up to 350mA of continuous output current at voltages as low as 1.21V. The ultralow quiescent current improves the performance of automotive or industrial systems, which demand always-on operation and optimum battery life. Switching frequency is user programmable from 200kHz to 2.2MHz, enabling the designer to optimize efficiency while avoiding critical noise-sensitive frequency bands. The combination of its 10-lead 3mm × 3mm DFN-10 (or 16-lead MSOP) package and high switching frequency keeps external inductors and capacitors small, providing a very compact, thermally efficient footprint.

The LT3990 utilizes a high efficiency 550mA, 300mV switch, with the necessary boost and catch Schottky diodes, oscillator, control and logic circuitry integrated into a single die. Low ripple Burst Mode operation ensures high efficiency at low output currents while keeping output ripple below 5mV<sub>p-p</sub>. Special design techniques and a new high voltage process enable high efficiency over a wide input voltage range, and the LT3990's current mode topology

enables fast transient response and excellent loop stability. Other features include a power good flag, soft-start capability and output short protection.

## OVERVOLTAGE/OVERCURRENT PROTECTION CONTROLLER SAFEGUARDS SENSITIVE LOW VOLTAGE ELECTRONICS FROM INPUT POWER SURGES

The LTC4361 is a 2.5V to 5.5V overvoltage and overcurrent protection controller designed to safeguard low voltage, portable electronics from damaging input voltage transients and current surges. Overvoltage events can occur due to power adapter failure or faults, or when hot-plugging an AC adapter into the power input of the device. The wrong power adapter can also inadvertently be plugged into a device, potentially causing damage from overvoltage or negative supply voltage.

The LTC4361 utilizes a 2% accurate 5.8V overvoltage threshold to detect an overvoltage event and responds quickly within 1μs to isolate the downstream components from the input. Overvoltage protection of up to 80V can be achieved with this simple IC/MOSFET solution without the need of additional external components such as capacitors or transistors at the input. In addition, the LTC4361 monitors voltage drop across a current sense resistor at the input of the circuit to protect against overcurrent faults. The LTC4361 is designed for mobile electronics with multiple power supply inputs, such as cell phones, MP3/MP4 players and digital cameras charged via USB ports, wall or car battery adapters.

The LTC4361 overvoltage and overcurrent protection controller utilizes a 2% accurate 5.8V overvoltage threshold to detect an overvoltage event and responds quickly within 1 $\mu$ s (max) to isolate the downstream components from the input.

The LTC4361 controls a low cost external N-channel MOSFET so that under normal operation it provides a low loss path from the input to the load. Inrush current limiting is achieved by controlling the voltage slew rate of the gate. If the voltage at the input exceeds the overvoltage threshold of 5.8V, the GATE is pulled low within 1 $\mu$ s to protect the load. While the IC operates from supplies between 2.5V and 5.5V, the input pins can withstand 80V transients or DC overvoltages. The LTC4361 features a soft shutdown controlled by the ON pin and provides a gate drive output for an optional external P-channel MOSFET for reverse voltage protection. A power good output pin indicates gate turn-on. Following an overvoltage condition, the LTC4361 automatically restarts with a start-up delay. The LTC4361 is available in two options; the LTC4361-1 latches off after an overcurrent event, where as the LTC4361-2 performs an auto-retry following a 130ms delay.

The new LTC4360 overvoltage protection controller is recommended for applications that do not require overcurrent protection. While offering many of the same features as the LTC4361, the two LTC4360 versions are differentiated by pin functions. The LTC4360-1 features soft shutdown control with low shutdown current of 1.5 $\mu$ A, while the LTC4360-2 can drive an optional external P-channel MOSFET for negative voltage protection.

The LTC4361 is offered in 8-lead (2mm  $\times$  2mm) DFN and SOT-23 packages, and the LTC4360 is offered in a tiny 8-lead SC70 package.

### 180MHZ, 1mA POWER EFFICIENT RAIL-TO-RAIL I/O OP AMPS

The LTC6246/LTC6247/LTC6248 are single/dual/quad low power, high speed unity gain stable rail-to-rail input/output operational amplifiers. On only 1mA of supply current, they feature an impressive 180MHz gain-bandwidth product, 90V/ $\mu$ s slew rate and a low 4.2nV/ $\sqrt{\text{Hz}}$  of input-referred noise. The combination of high bandwidth, high slew rate, low power consumption and low broadband noise makes these amplifiers unique among rail-to-rail input/output op amps with similar supply

currents. They are ideal for lower supply voltage high speed signal conditioning systems. The LTC6246 family maintains high efficiency performance from supply voltage levels of 2.5V to 5.25V and is fully specified at supplies of 2.7V and 5.0V. For applications that require power-down, the LTC6246 and the LTC6247 in MS10 offer a shutdown pin, which disables the amplifier and reduces current consumption to 42 $\mu$ A. The LTC6246 family can be used as a plug-in replacement for many commercially available op amps. ■

(LTC3890 continued from page 33)

supply noise are minimized by operating the two output stages out-of-phase.

### DUAL OUTPUT APPLICATION

Figure 1 shows the LTC3890 operating in an application that converts a 9V to 60V input into 3.5V/5A and 8.5V/3A outputs. The transient response for the 3.3V output with a 4A load step is less than 50mV (as shown in Figure 2).

Figure 3 shows the efficiency of the 8.5V channel with a 36V input voltage.

### SINGLE OUTPUT APPLICATION

The LTC3890 can also be configured as a 2-phase single output converter by simply connecting the two channels together. For example, a 9V to 60V input can be converted to an 8.5V output at 6A. Figure 4 shows the efficiency of this configuration at input voltages of 10V, 30V and 60V.

Current mode control provides good current balance between the phases.

Less than 10% mismatch can be achieved, as shown in Figure 5.

### CONCLUSION

Although there are many choices in dual-output controllers, the LTC3890 brings a new level of performance with its high voltage operation, high efficiency conversion and ease of design. ■

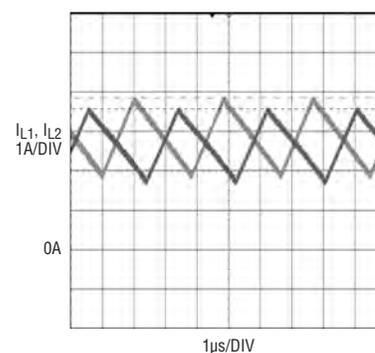
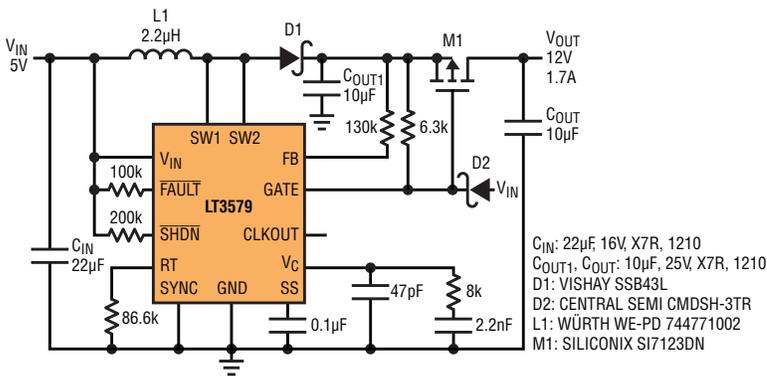
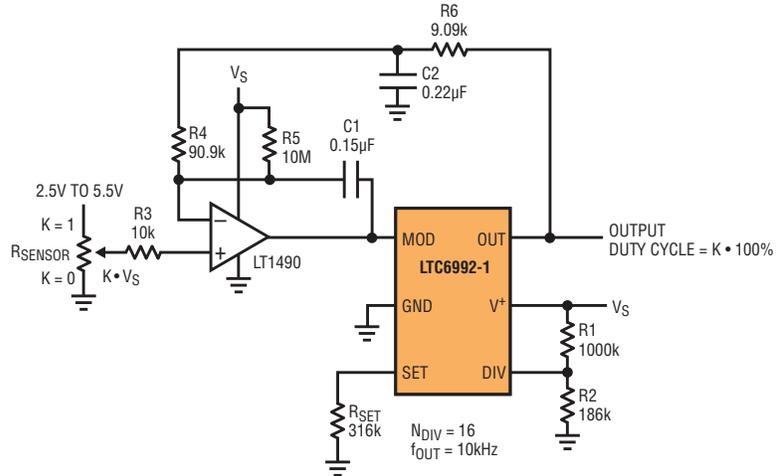


Figure 5. The inductor current in a 2-phase single output converter. Currents in both inductors shown with a 24V Input and 8.5V at 6A output.

**RATIOMETRIC SENSOR TO PULSE WIDTH**

The LTC6992 is a silicon oscillator with an easy-to-use analog voltage-controlled pulse width modulation (PWM) capability. A single resistor,  $R_{SET}$ , sets the oscillation frequency. Applying a voltage between 0V and 1V on the MOD pin sets the duty cycle. In this example, the MOD pin is driven by the output of a ratiometric sensor, creating an output waveform of constant frequency and a duty cycle proportional to the sensor output. The LTC6992 is part of the TimerBlox™ family of versatile silicon timing devices. [www.linear.com/6992](http://www.linear.com/6992)

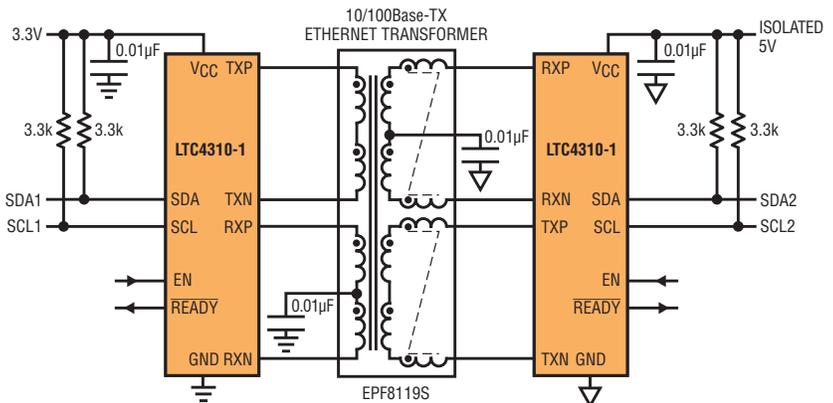


**1MHZ, 5V TO 12V BOOST CONVERTER WITH OUTPUT SHORT CIRCUIT PROTECTION**

The LT3579 is a flexible DC/DC converter that can easily be configured in boost, SEPIC, inverting and flyback topologies. It incorporates two integrated 42V switches, a 3.4A master switch and a 2.6A slave switch, which can be tied together for a total switch current of 6A. It also offers a range of integrated fault protection features for output shorts, input/output overvoltages and overtemperature conditions. In this example, the GATE pin of the LT3579 is tied to an external PMOS FET to implement robust output short-circuit protection. [www.linear.com/3579](http://www.linear.com/3579)

**1500V ISOLATED I<sup>2</sup>C SYSTEM**

The LTC4310 provides bidirectional I<sup>2</sup>C communications between two I<sup>2</sup>C buses whose grounds are isolated from one another. Each LTC4310 encodes I<sup>2</sup>C bus logic states into signals that are transmitted across an isolation barrier to another LTC4310. The receiving LTC4310 decodes the transmission and drives its I<sup>2</sup>C bus to the appropriate logic state. The isolation barrier can be bridged by an inexpensive Ethernet, or other transformer, to achieve communications across voltage differences reaching thousands of volts, or it can be bridged by capacitors for lower voltage isolation. [www.linear.com/4310](http://www.linear.com/4310)



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