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## New Linear Regulators Solve Old Problems

Bob Dobkin

The architecture of linear regulators has remained virtually unchanged since the introduction of the 3-terminal floating voltage regulator in 1976. Architectures have settled on either the floating architecture (LT317) or an amplifier loop with feedback from the output to the amplifier. These architectures inherently limit regulator versatility, regulation and accuracy. In 2007, Linear Technology released the LT<sup>®</sup>3080, introducing a new linear regulator

architecture that expanded versatility, significantly improved performance, and allowed easy parallel operation. The LT3081 is one of this new class of linear regulator, specifically featuring a wide safe operating area for industrial applications.

### THE OLD WAY

In the old linear regulator architecture (Figure 1), feedback resistors set the output voltage and attenuate the feedback signal into the amplifier. In this scheme, regulation accuracy at the output is a function of the percentage of the output voltage, so absolute regulation accuracy (in volts) degrades as output voltage increases, even though percentage accuracy is maintained. The bandwidth of the regulator also changes with voltage—as the loop gain is decreased, the bandwidth decreases at higher output voltages, resulting in relatively slow transient response and high ripple.

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The LT3081 is one of a new breed of linear regulator, specifically featuring a wide safe operating area for industrial applications.

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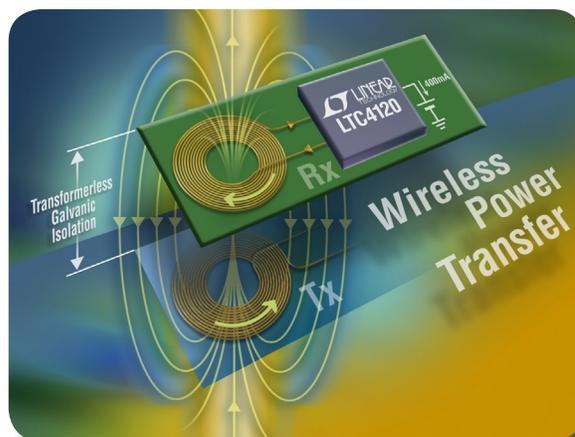
### NEW SOLUTION FOR WIRELESS CHARGING

Linear has released a wireless battery-charging solution that satisfies the demanding requirements of medical handheld devices, portable diagnostic equipment, lighting and signaling equipment, handheld industrial instruments, and industrial/military sensors. Linear's LTC<sup>®</sup>4120 combines a wireless power receiver with a constant-current/constant-voltage battery charger, functioning as the receive circuit component in a complete wireless power transfer system.

The LTC4120 works reliably with Linear Technology's simple discrete resonant transmitter reference design or with advanced off-the-shelf transmitters from PowerbyProxi, a leader in wireless power solutions. PowerbyProxi transmitters offer advanced features, including simultaneous charging of multiple receivers with a single transmitter and foreign object detection to prevent excessive heating of foreign objects.

The LTC4120 features dynamic harmonization control (DHC), a patented technique that enables optimal wireless power transfer over a variety of conditions while providing thermal management and overvoltage protection. This technique modulates the resonant frequency of the receiver tank to provide lossless adjustment of the power received as well as the power transmitted, enabling an efficient and robust solution for wireless battery charging.

The LTC4120 allows battery-powered devices to be wirelessly recharged without expensive, failure-prone connectors. Products incorporating the LTC4120 may be contained within sealed enclosures, in moving or rotating equipment or used where cleanliness or sanitation is critical. This provides battery charging for applications in which it is difficult or impossible to use a connector. For more info, see [www.linear.com/product/LTC4120](http://www.linear.com/product/LTC4120)



Linear's LTC4120 combines a wireless power receiver with a constant-current/constant-voltage battery charger, functioning as the receive circuit component in a complete wireless power transfer system.

## PARTNERSHIP TO ACCELERATE INTERNET OF THINGS

Linear Technology has announced a partnership to accelerate the commercial availability of ultralow power-connected products, promising to open a range of new applications for machine-to-machine connectivity. Linear has partnered with LogMein's Xively subsidiary, the creator of the first public cloud platform for the commercial Internet of Things (IoT) to accelerate the availability of ultralow power cloud-connected products. The partnership integrates Linear Technology's Dust Networks® SmartMesh IP™ wireless sensor networking Starter Kit with Xively Cloud Services.

This partnership opens up a range of potential applications—from smart parking and commercial building energy management to utility-scale solar plants and structural monitoring of bridges, tunnels and mines. Using Linear's Dust Networks SmartMesh IP Starter Kit with Xively's IoT platform provides users with a cloud-enabled wireless sensor network with carrier-class reliability and ultralow power consumption out-of-the-box, speeding time-to-pilot and reducing overall application development time.

The SmartMesh IP wireless sensor network is built for Internet Protocol (IP) compatibility, and is based on the 6LOWPAN and 802.15.4e standards. SmartMesh IP networks deliver >99.999% data reliability and >10 year battery life, making it practical to deploy wireless sensor networks in the most challenging environments. The DC9000 Starter Kit includes sample application code that allows a registered Xively user to connect the DC9000 kit to a web-connected computer, enabling the SmartMesh IP network to automatically be registered as a Xively “product,” and the nodes and sensors to securely send data to the user's Xively cloud as “devices.” For more info on SmartMesh IP kits, visit [www.linear.com/DC9000](http://www.linear.com/DC9000)



## ANALOG CIRCUIT DESIGN BOOK JAPANESE TRANSLATION

Linear Technology's well-received book, *Analog Circuit Design: A Tutorial Guide to Applications and Solutions*, edited by Bob Dobkin and Jim Williams, has just been published in a Japanese edition. CQ Publishing of Tokyo has published Part 1 Power Management section of the *Analog Circuit Design* book and it is now available via bookstores and online. For more info, visit [www.cqpub.co.jp/hanbai/books/42/42881.htm](http://www.cqpub.co.jp/hanbai/books/42/42881.htm).

The book debuted at the Analog Guru's Conference, held in Tokyo in October. At the conference, sponsored jointly by Linear Technology and Nikkei Electronics, 450 attendees heard presentations on analog technology by key industry figures, including Linear co-founder and Chief Technical Officer Bob Dobkin and Linear Vice President, Power Products Steve Pietkiewicz. Conference attendees received copies of the *Analog Circuit Design* book, signed by editor Bob Dobkin.

## LINEAR PRODUCT AWARDS

### EDN Hot 100 Products

**Power**—LT3081 40V, 1.5A Robust Linear Regulator with Current & Temperature Monitor Outputs

**Power**—LT4320 Ideal Diode Bridge Controller

**Analog**—LTC2378-20 20-Bit, 1MSPS, Low Power SAR ADC with 0.5ppm INL

**Wireless & Networking**—LTC4120 Wireless Power Receiver and 400mA Buck Battery Charger

**Selezione di Elettronica (Italy) Innovation Award 2013**

**Analog & Power ICs**—LTC5585 Wideband IQ Demodulator with IP2 and DC Offset Control

## CONFERENCES & EVENTS

**CAR-ELE Japan, 6th International Automotive Electronics Technology Expo, Tokyo Big Sight, West Hall, Tokyo, Japan, January 15–17**—Linear Technology will showcase its high performance analog IC solutions for automotive, including battery management systems. More info at [www.car-ele.jp/en/Home](http://www.car-ele.jp/en/Home)

**APEC 2014, Applied Power Electronics Conference, Fort Worth Convention Center, Fort Worth, Texas, March 16–20, Booth 532**—Presenting Linear's broad line of high performance power solutions. More info at [www.apec-conf.org/](http://www.apec-conf.org/)

**Electronica China 2014, Shanghai New International Expo Center, Shanghai, China, March 18–20, Hall W1, Booth 1218**—Linear will showcase its high performance analog and power management solutions. More info at [www.electronicachina.com/en/home.html](http://www.electronicachina.com/en/home.html) ■



The benefit of using an internal true current source as the reference, rather than a bootstrapped reference, as in prior regulators, is not obvious. A true reference current source allows the regulator to have gain and frequency response independent of the impedance on the positive input.

or set to zero, temperature and current can still be measured. Using a 1k resistor provides sufficient margin and ensures operation when the output is shorted.

#### One Resistor Sets the Output Voltage

The output is set with a resistor from the SET pin to ground and a 50 $\mu$ A precision current source set to the output. The internal follower amplifier forces the output voltage to be the same voltage as the SET pin. Unique to the LT3081, an output capacitor is optional. The regulator is stable with or without input and output capacitors. All the internal operating current flows through the output pin and minimum load is required to maintain regulation. Here, a 5mA load is required at all output voltages to maintain the device in full regulation.

The set resistor can add to the system temperature drift. Commercially available surface mount resistors have a wide range of temperature coefficients.

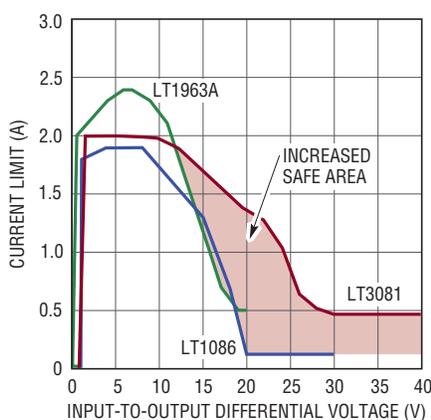


Figure 4. Comparative safe operating area performance

Depending on the manufacturer, these can go from 100ppm up to over 500ppm. While the resistor is not heated by power dissipation in the regulator, over a wide ambient temperature range its temperature coefficient can change the output by one to four percent. Lower temperature coefficient thin film resistors are available for precision applications.

#### Benefits of Internal Current Source as Reference

The benefit of using an internal true current source as the reference, rather than a bootstrapped reference, as in prior regulators, is not obvious. A true reference current source enables the regulator to have gain and frequency response independent of the impedance on the positive input. With all previous adjustable regulators, such as the LT1086, loop gain and bandwidth change with variations in output voltage. If the adjustment pin is bypassed to ground, bandwidth also changes. For the LT3081, the loop gain is unchanged with output voltage or bypassing. Output regulation is not a fixed percentage of output voltage, but is a fixed number of millivolts. Use of a true current source allows all of the gain in the buffer amplifier to provide regulation, and none of that gain is needed to amplify the reference to a higher output voltage.

Table 1. Linear regulators featuring updated architecture

DEVICE	OUTPUT CURRENT	I <sub>SET</sub>	ADJUSTABLE CURRENT LIMIT/CURRENT MONITOR	TEMPERATURE MONITOR	LDO	COMMENTS
LT3080	1.1A	10 $\mu$ A	No/No	No	Yes	
LT3081	1.5A	50 $\mu$ A	Yes/Yes	Yes	No	Output CAP Optional
LT3082	200mA	10 $\mu$ A	No/No	No	No	
LT3083	3A	50 $\mu$ A	No/No	No	Yes	
LT3085	600mA	10 $\mu$ A	No/No	No	Yes	
LT3086	2.1A		Yes/Yes	Yes + Temp Limit	Yes	
LT3090	600mA	-50 $\mu$ A	Yes/Yes	Yes	Yes	Negative Regulator
LT3092	200mA	10 $\mu$ A	No/No	No	No	Current Source Operation Needs No Output CAP

The LT3081 extends the safe operating area, offering nearly 1A of output current at 25V of differential. Even above 25V, 500mA is still available. This allows the regulator to be used in applications where widely varying input voltages can be applied during operation.

### Wide Safe Operating Area

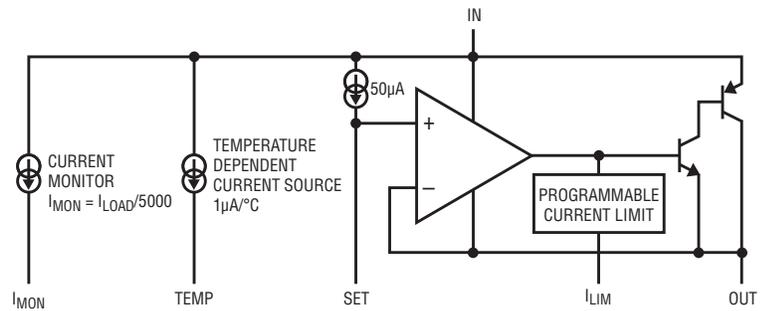
Industrial applications require a wide safe operating area. Safe operating area (SOA) is the ability to carry large currents at high input-output differentials. The safe operating area for several regulators is compared in Figure 4.

The LT1086, introduced in the mid-1980s, is a 1.5A regulator in which output current drops very low above 20V input/output differential. Above 20V only about 100mA of output current is available. This causes output voltage to go unregulated if the load current is above 100mA and transients on the input cause the high voltage current limit to be exceeded. The LT1963A is a low dropout regulator that also has a limited safe operating area.

The LT3081 extends the safe operating area, offering nearly 1A of output current at 25V of differential. Even above 25V, 500mA is still available. This allows the regulator to be used in applications where widely varying input voltages can be applied during operation. Wide operating safe area is obtained by using a large structure for the PNP pass device. Also, the LT3081 is protected (along with the load) for reverse input voltage.

Figure 5 shows a block diagram of the LT3081. There are three current sources: two that report output current and temperature and a third that supplies the 50µA reference current. The LT3081, while not a low dropout regulator, operates down to 1.2V across the device — slightly better than older devices such as the LT1086.

Figure 5. Block diagram of the LT3081



### No Input or Output Capacitors Needed

The internal amplifier configuration, in conjunction with well regulated internal bias supplies, allows the LT3081 to be stable with no external capacitors. One caveat: it cannot be designed to tolerate all possible impedances in the input and load, so it is important to test the stability in the actual system used. If instability is found, external capacitors will ensure that the device is stable at all output currents. External capacitors also improve the transient response since it is no longer limited by the bandwidth of the internal amplifier.

### Paralleling Devices for More Current

Paralleling devices—typically forbidden with prior regulators since they do not share current—is easy with these new current source reference regulators. Paralleling is useful for increasing output current or spreading the heat. Since it is set up as a voltage follower, tying all the SET pins together makes the outputs the same voltage. If the outputs are at the same voltage, only a few milliohms of ballast are needed to allow them to share current.

Figure 6 shows a distribution of the offset voltage for the LT3081. The distribution is within 1mV so ensuring 10% sharing requires no more than 10mΩ of ballast resistance. The ballast resistor can be less than an inch of a trace on a PC board or a small piece of wire, and provides good current balance from parallel devices. Even at 1V output, this degrades the regulation by only about 1.5%. Table 2 shows PC board resistance.

Figure 7 shows a schematic of two LT3081s paralleled to obtain 3A output. The set resistor now has twice the set current flowing through it, so the

Figure 6. Offset voltage distribution

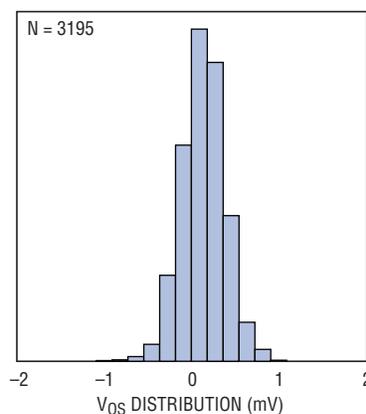


Table 2. PC board trace resistance

WEIGHT (oz)	10MIL WIDTH	20MIL WIDTH
1	54.3	27.1
2	27.1	13.6

Trace resistance is measured in mΩ/in.

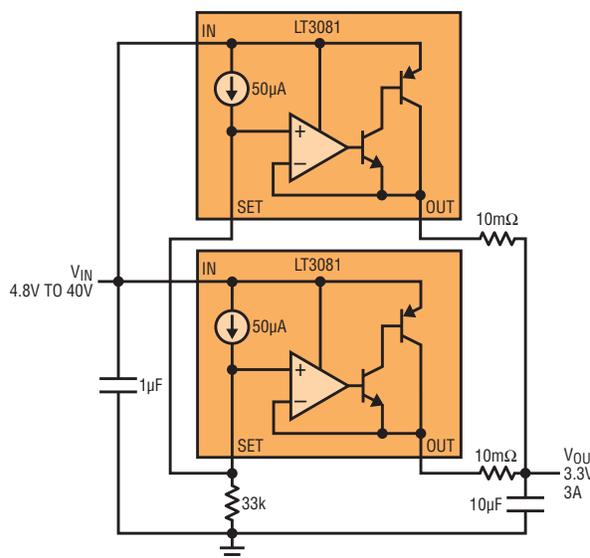
output is  $100\mu\text{A}$  times  $R_{\text{SET}}$  and the  $10\text{m}\Omega$  output resistors ensure ballasting at full current. Any number of devices can be paralleled for higher current. The I pins can be paralleled (if used) so one resistor sets the current limit.

Figure 8 shows the LT3081 paralleled with a fixed regulator. This is useful when a system that has been designed has insufficient output current available. It provides a quick fix for higher output current. The output voltage of the fixed device is divided down by just a few millivolts by the divider. The SET pin of the LT3081 is tied about  $4\text{mV}$  below the fixed output. This ensures no current flows from the LT3081 under a no-load condition. Then the  $20\text{m}\Omega$  resistors provide sufficient ballast to overcome this offset and ensure current matching at higher output currents.

LAYOUT CONSIDERATIONS

With the  $50\mu\text{A}$  current source used to generate the reference voltage, leakage paths to or from the SET pin can create errors in the reference and

Figure 7. Paralleling devices

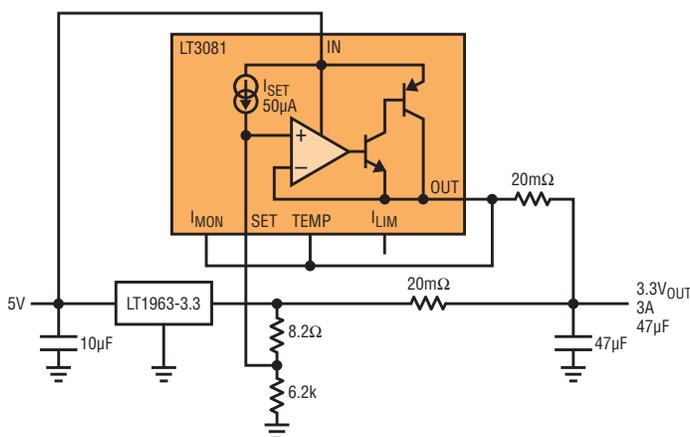


output voltages. Cleaning of all insulating surfaces to remove fluxes and other residues is required. Surface coating may be necessary to provide a moisture barrier in high humidity environments. Minimize board leakage by encircling the SET pin and circuitry with a guard ring tied to the OUT pin. Increasing the set current as shown also decreases the effects of spurious leakages.

The low  $50\mu\text{A}$  SET current can cause problems in some applications. High value film potentiometers are not as stable as lower value wire wounds. Board leakage can also introduce instabilities in the output. Problems can be minimized by increasing the set current above the nominal  $50\mu\text{A}$ .

Figure 9 shows a solution using lower value set resistors. Here an increased current is generated through  $R_2$  and summed with the SET pin current, giving a much larger current for adjusting the output. SET current flows through a  $4\text{k}$  resistor, generating  $200\text{mV}$  across  $R_1$ . Then the current through  $R_2$  adds to the SET current, giving a total of  $1.05\text{mA}$  flowing through  $I_{\text{SET}}$  to ground. This makes the voltage less sensitive to leakage currents around  $R_{\text{SET}}$ . Care should be taken to Kelvin connect  $R_2$  directly to the output. Voltage drops from the output to  $R_2$  affect regulation.

Figure 8. Increasing the output current of a fixed regulator



The LT3081 is one of a new family of linear regulators that yields an order-of-magnitude better regulation against load and line changes compared to traditional devices. Regulation and transient response, measured in millivolts, are now maintained regardless of output voltage.

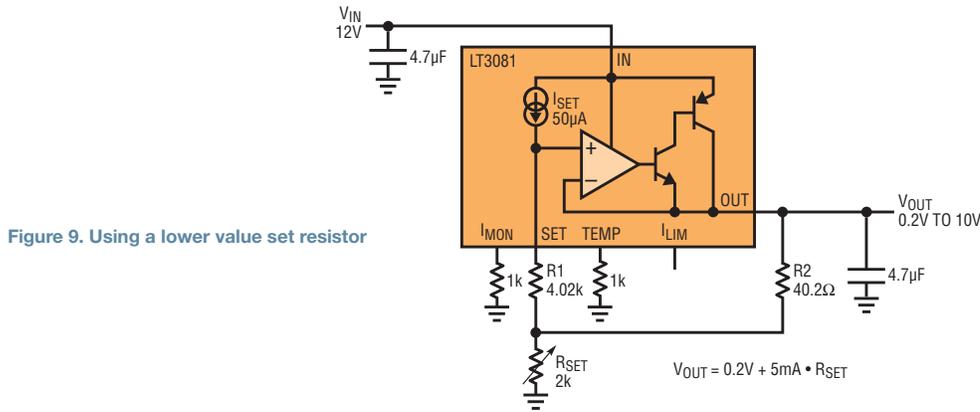


Figure 9. Using a lower value set resistor

Another configuration uses an LT3092 as an external current source of 1mA. This provides increased set current and allows the output to be adjusted down to zero. Figure 10 shows an LT3092 current source used to provide the current reference to an LT3081. The 1mA generated reference current allows the adjustment set resistor to be much lower in value while enabling the device to be adjusted down to zero.

The current monitor output can be used to compensate for line drops, as shown in Figure 11. Feeding the current monitor through a portion of the set resistor generates a voltage at the SET pin that raises the output as a function of current. The value of the comp resistor is  $R_2 = 5000 \cdot R_{CABLE(TOTAL)}$  and  $V_{OUT} = 50\mu A \cdot (R_{SET} + R_{COMP})$ . Several volts of line drop can be compensated this way.

Figure 10. Using an external reference current

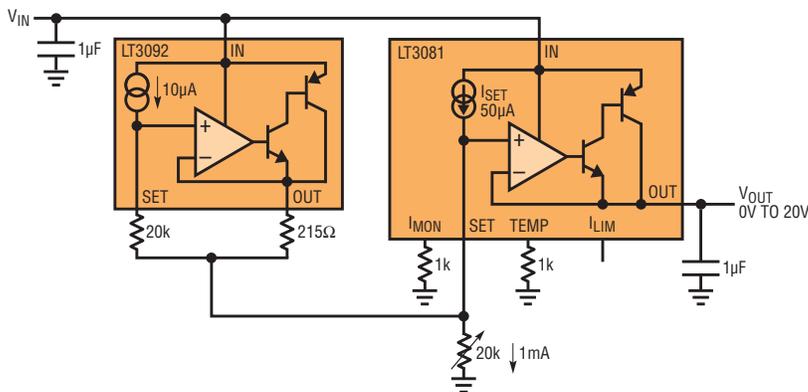
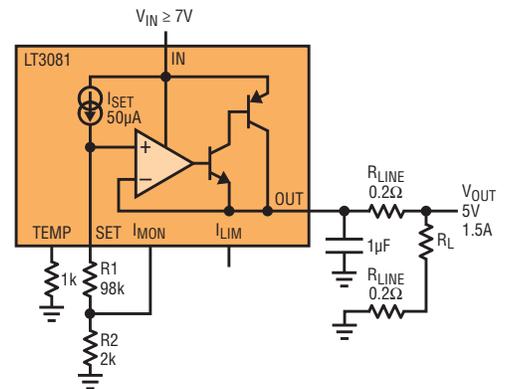


Figure 11. Using current monitor output to compensate for line drops



## CONCLUSION

The LT3081 is one of a new family of linear regulators that yields an order-of-magnitude better regulation against load and line changes compared to traditional devices. Regulation and transient response, measured in millivolts, are now maintained regardless of output voltage.

These new regulators are far more robust and versatile than previous generations, offering features that were previously unavailable. Temperature and current monitoring and adjustable current limiting are now added. Paralleling these new regulators no longer requires external current balance circuitry to prevent current hogging. Line drops can be easily compensated. Current limit thresholds are now user-defined, as opposed to fixed in the regulator, and outputs are adjustable to zero. The LT3801, in particular, features a wide safe operating area to support load currents in the face of wide input swings. ■

# 2.7V to 38V $V_{IN}$ Range, Low Noise, 250mA Buck-Boost Charge Pump Converter

George H. Barbehenn

The LTC3245 is a buck-boost regulator that dispenses with the traditional inductor, and instead uses a capacitor charge pump. The LTC3245's input voltage range is 2.7V to 38V, and it can be used without a feedback divider to produce one of two fixed output voltages, 3.3V and 5V, or programmed via a feedback divider to any output voltage from 2.5V to 5.5V. Maximum output current is 250mA. As a result of its buck-boost topology, the LTC3245 is capable of regulating a voltage above or below the input voltage, allowing it to satisfy automobile cold crank requirements.

The LTC3245 achieves efficiency of 80% when delivering 5V, 100mA from a 12V source, significantly higher efficiency than an LDO, making it possible to avoid the space and cost requirements of an LDO with a heat sink. The LTC3245 is available in an exposed pad MSOP12 or 3mm × 4mm DFN12.

## CHARGE PUMP OPERATION

Figure 3 shows a simplified block diagram of the LTC3245 converter. Charge pumps can operate as  $N/M \times V_{IN}$  converter, where  $N$  and  $M$  are integers.  $1/2$ , 1, and 2 are the simplest forms and only require one flying capacitor. Higher order  $N$  and  $M$  require more flying capacitors and switches.

Because  $N$  and  $M$  are integers, a straight charge pump cannot be used to produce an arbitrary output. Instead the controller modifies  $V_{IN}$  to produce  $V_{IN}'$ , which is then fed to the charge pump. The charge pump can operate in one of three modes, buck, LDO or boost, resulting in  $1/2 V_{IN}'$ ,  $V_{IN}'$  or  $2V_{IN}'$ , respectively.

By properly controlling both  $V_{IN}'$  and the operating mode of the charge pump any arbitrary voltage can be achieved. When operating in buck mode, the input current is approximately half that of an equivalent LDO, offering a significant efficiency improvement.

## INPUT RIPPLE AND EMI

The LTC3245 charges the flying capacitor each switching cycle, so  $V_{IN}$  must be sufficiently decoupled to minimize EMI.

To decouple the LTC3245, place a  $3.3\mu\text{F} \sim 10\mu\text{F}$  MLCC capacitor as close to the  $V_{IN}$  pin as possible. One way to move it closer is to limit the voltage rating on the capacitor, which helps minimize the size of the cap, and the smaller it is, the nearer the  $V_{IN}$  pin it can be placed. For instance, although LTC3245 is rated to operate up to 38V input, for an automotive supply, an MLCC with 16V rating should be sufficient.

A decoupling capacitor with a short, low inductance supply connection, but a high inductance ground connection,

Figure 1. A 5V output buck-boost converter

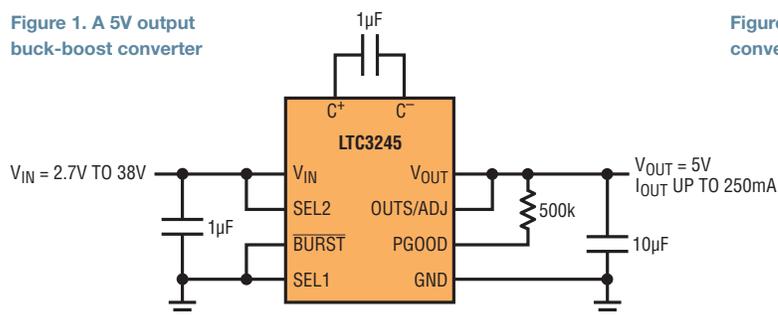
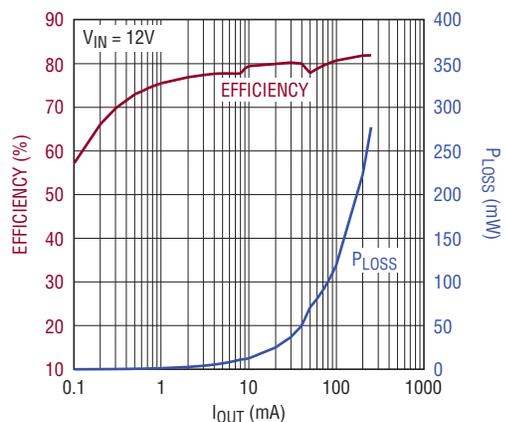
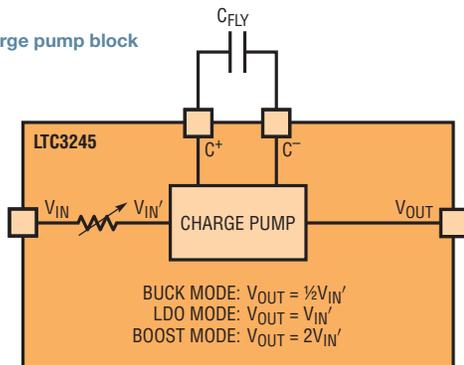


Figure 2. Efficiency of the converter in Figure 1



The LTC3245 achieves efficiency of 80% when delivering 5V, 100mA from a 12V source, significantly higher efficiency than an LDO, making it possible to avoid the space and cost requirements of an LDO with a heat sink.

Figure 3. Detail of the charge pump block



is not very effective. The ideal situation is when the supply connection is short and wide, and the ground connection is an area fill with a very wide connection to the exposed pad on the LTC3245.

The assumption is made that  $V_{IN}$  does not have a very long connection back to a low impedance supply. If the input supply is high impedance, or the connection to the input supply is longer than 5cm, it is recommended that the supply be decoupled with additional bulk capacitance, as needed. In many cases, 33 $\mu$ F is adequate.

The LTC3245 can be optimized for light load efficiency or low output ripple by choosing high efficiency Burst Mode<sup>®</sup> operation or low noise mode. Burst Mode operation features low quiescent current and hence higher efficiency at low load currents. Low noise mode trades off light load efficiency for lower output ripple at light loads.

Figure 4 shows measured radiated and conducted signatures of the LTC3245, tested in a microchamber in accordance with CISPR25. As can be seen here,

when properly decoupled, the LTC3245 does not present any issue when striving to meet government regulations for radiated or conducted emissions.

### CHOOSING THE FLYING CAPACITOR

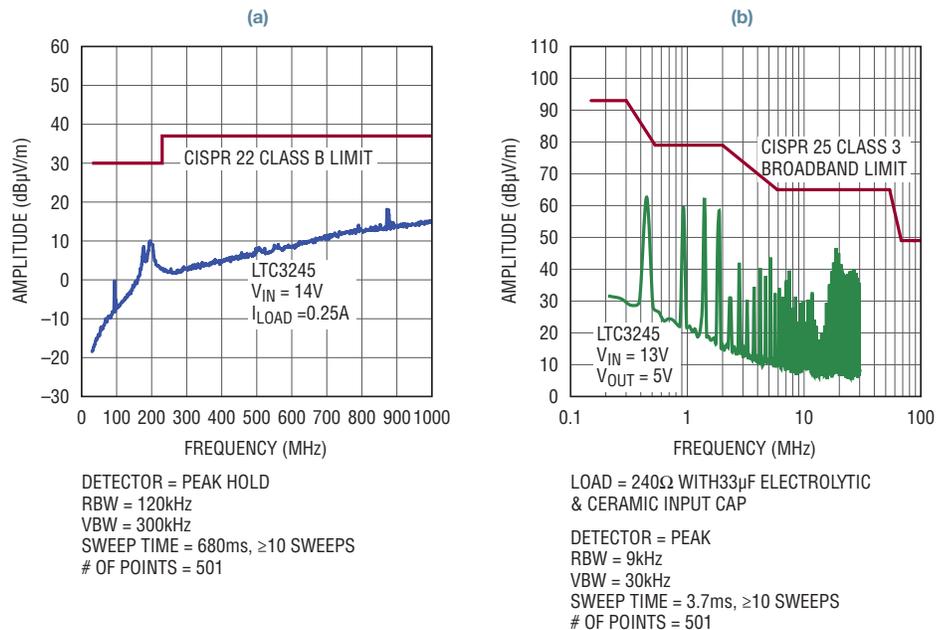
The detail of the charge pump block (Figure 3) suggests that the flying capacitor is only involved in the charge pump

itself. However, the flying capacitor is also involved in the variable attenuator that generates  $V_{IN}'$ . Consequently, the capacitor should not be chosen based on straightforward calculation, but instead by observing a few constraints.

The flying capacitor cannot be polarized, such as an electrolytic or tantalum capacitor. The voltage rating of the flying capacitor should be about 1V more than the output voltage, such as using a 6.3V flying capacitor for a 5V output.

The minimum capacitance of the flying capacitor must be 0.4 $\mu$ F. Since polarized capacitors are not allowed, the most appropriate capacitor is MLCC. MLCC capacitors with enough capacitance to meet the 0.4 $\mu$ F are likely Class II dielectric capacitors, with strong voltage

Figure 4. Radiated (a) and conducted (b) emissions



The OUTS/ADJ pin is used either for sensing  $V_{OUT}$  for fixed 3.3V and 5V outputs or as the feedback pin for an adjustable output voltage. It is connected directly to the output when using the fixed values. An adjustable output can be set anywhere between 2.5V and 5V through the choice of suitable feedback resistors.

coefficients on their capacitance. The voltage coefficient of the capacitance is a function of the maximum voltage, so a capacitor of maximum voltage of 16V operating at 5V will have much more in-circuit capacitance than a 6.3V capacitor of the same nominal capacitance and size, operating at 5V.

So, a 0.47 $\mu$ F, 6.3V, Class II dielectric capacitor operating at 5V will likely not meet the minimum capacitance, while a 0.47 $\mu$ F, 50V, Class II dielectric capacitor likely will. A capacitor such as the TDK C1005X5R1C105K 1 $\mu$ F, 16V, 0402 is suitable for most applications.

### OUTPUT CAPACITOR

The choice of output capacitor value is a trade-off between ripple and step response. As the output capacitance is increased, the ripple decreases but the step response is also increasingly overdamped.

The required voltage rating of the output capacitor is the output voltage of the regulator, so a 6.3V capacitor would suffice for a 5V output. Nevertheless, as discussed above, Class II dielectric capacitors lose more than half their nominal capacitance at their rated voltage. Consequently, it may be necessary to choose a larger capacitor when operating close to the rated voltage of the capacitor, to minimize ripple.

A good compromise between ripple and response is a capacitor with a capacitance, at bias, of  $10\times \sim 20\times$  the flying capacitor. This means 10 $\mu$ F to 20 $\mu$ F for

the recommended flying capacitor value of 1 $\mu$ F. Since Class II capacitors lose a little more than half their capacitance at rated voltage, this indicates a 47 $\mu$ F nominal capacitance capacitor.

### ADJUSTABLE OUTPUT

Besides the two fixed output voltage values of 3.3V and 5V, it is possible to program the output voltage of the LTC3245 using feedback resistor as shown in Figure 5.

Adjustable output mode is achieved by setting SEL2 low and SEL1 high. The OUTS/ADJ pin is used either for sensing the output for fixed output voltages or as the feedback pin for an adjustable output voltage. It is connected directly to the output when using the fixed values. For adjustable output, the feedback reference voltage is  $1.200V \pm 2\%$ . The output can be set anywhere between 2.5 and 5V, through the choice of suitable feedback resistors.

### SHUTDOWN

The LTC3245 can also be placed in shutdown to reduce the quiescent current to just 4 $\mu$ A. Pull both SEL1 and SEL2 low to shutdown the LTC3245.

### PGOOD

PGOOD is an active high, open drain signal that indicates the output of the LTC3245 is in regulation. The threshold for the PGOOD indication is 90% of the desired feedback or sense voltage.

### CONCLUSION

The LTC3245 is a switched capacitor buck-boost DC/DC converter that produces a regulated output (3.3V, 5V or adjustable) from a 2.7V to 38V input. No inductors are required. Low operating current (20 $\mu$ A with no load, 4 $\mu$ A in shutdown) and low external parts count (three small ceramic capacitors) make the LTC3245 ideal for low power, space-constrained automotive and industrial applications. ■

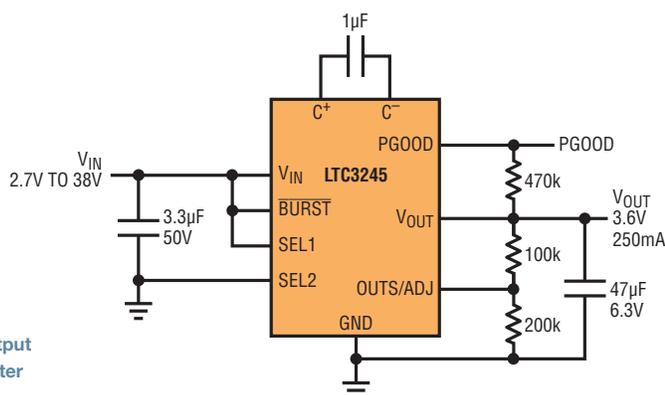


Figure 5. A 3.6V output buck-boost converter



is 5A with an input supply of 12V. The choice of a 500mV differential is based on a combination of parameters: the desired maximum output current, dropout margin when accounting for part-to-part variation, effect of operating temperature and transient response performance.

### DUAL OUTPUT REGULATOR OPTIMIZED FOR LOW NOISE

In Figure 2, the input supply is also 12V, but it produces two LDO outputs at 5V and 3.3V. The 5V LDO output has the same low 500mV input-to-output differential voltage as in Figure 1, so LDOs 1 and 2 are operated near their rated currents. The input-to-output differential voltage for the 3.3V output is higher so the output currents for LDOs 3, 4 and 5 each require derating to 330mA. Nevertheless, the three LDOs in parallel perform heat-spreading, so the available combined output current of the derated LDOs increases to 1A. The thermal images in Figure 3 show comparable LDO temperatures for the solutions in Figures 1 and 2.

#### Setting the BIAS

The BIAS inputs to the LDOs must be 1.6V higher (worst case) than the highest corresponding LDO output voltage. Consequently, in Figure 1, BIAS is taken from  $V_{IN}$ , but in Figure 2, BIAS for the 5V-output LDOs comes from  $V_{IN}$  and

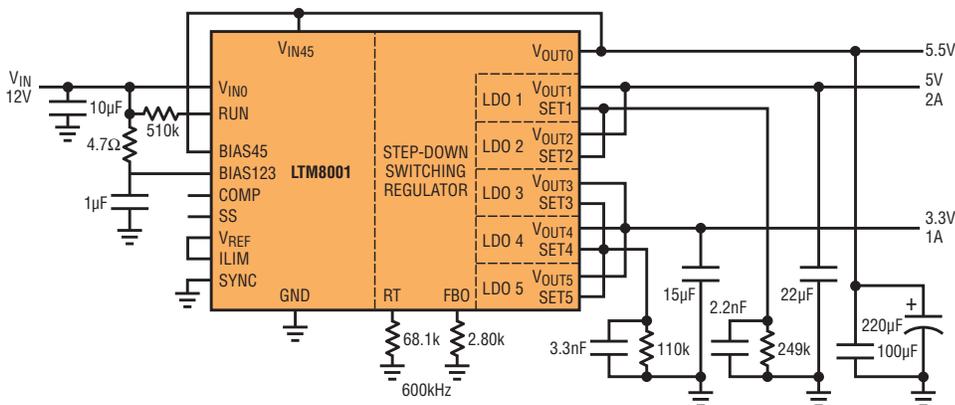


Figure 2. A solution optimized for dual low noise outputs. The 5V LDO output features the same low 500mV input-to-output differential voltage as in Figure 1, so LDOs 1 and 2 are operated near their rated currents. The input-to-output differential voltage for the 3.3V output is higher, so the available current from each of the three paralleled LDOs (3, 4 and 5) is derated to 330mA, for a combined 1A.

BIAS for the 3.3V-output LDOs comes from the 5.5V switcher output ( $V_{OUT0}$ ).

#### Optimizing Output Noise

LDOs can be used to post-regulate the outputs of switching converters to produce low noise voltages because LDOs reject voltage ripple on both the switching converter output and the BIAS inputs to the LDO. The solution in Figure 2 is optimized for low noise outputs. The capacitance on the switching converter output  $V_{OUT0}$  and LDO outputs  $V_{OUT1}$ – $V_{OUT5}$  are optimized to minimize voltage ripple. The bias input BIAS45 from  $V_{IN}$  is filtered. The LDO reference is a source of noise so capacitors are added to

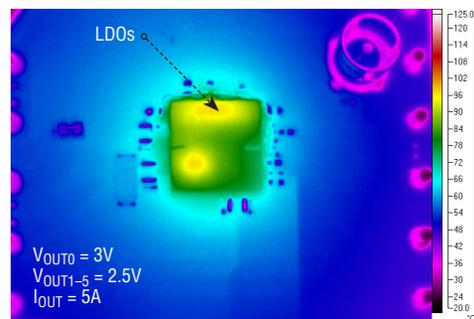
SET1–SET5 to bypass the LDO references. Figure 4 shows that the noise spectrum first harmonic peak is  $-63\text{dBm}$  for the 5V LDO1-LDO2 output in Figure 2.

#### CONCLUSION

The LTM8001 is a high efficiency step-down converter that is followed by five in-package low noise LDOs for a total output current capability of 5A. The outputs of the LDOs can be used to produce five different output voltages, or combined in a variety of parallel arrangements for high current outputs and heat spreading. ■

Figure 3. Thermal imaging shows comparable LDO temperatures for the circuits in Figures 1 (a) and 2 (b) when operated at maximum current, proving the versatility of the LTM8001 in spreading heat, even when applied in very different solutions.

(a) Thermal image of solution in Figure 1



(b) Thermal image of solution in Figure 2

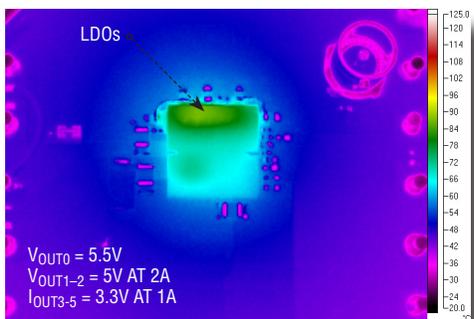
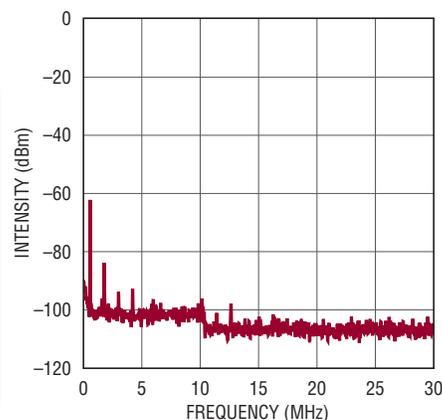


Figure 4. The solution in Figure 2 is optimized for a low noise output. The 5V output produces a noise spectrum first harmonic peak of  $-63\text{dBm}$ .



# Precision Fully Differential Op Amp Drives High Resolution ADCs at Low Power

Kris Lokere

The LTC6362 op amp produces differential outputs, making it ideal for processing fully differential analog signals or taking a single-ended signal and converting it to fully differential. Many alternative op amps of this fully differential nature are optimized for very high speed operation, resulting in high power consumption and lack DC accuracy. The LTC6362 is unique in that it features differential outputs, low power consumption and accurate DC offset voltage (see Table 1).

## HOW DOES IT WORK?

Let's take a closer look at how a differential op amp works. Just like a regular op amp, it has two inputs, but unlike a regular op amp, it also has two outputs, labeled  $-OUT$  and  $+OUT$ . A regular op amp features high open-loop gain between the differential input and the one output; a fully differential op amp features high open-loop gain between the differential input and the differential output.

Feedback should also be applied differentially. Figure 1 shows four external resistors feeding a portion of the differential output back into the differential input. Just like in a regular

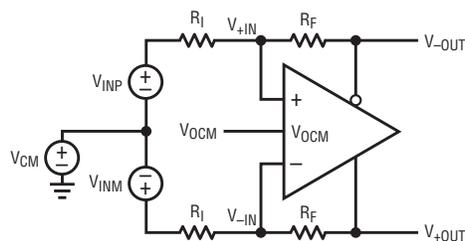


Figure 1. Four external resistors apply feedback around the differential op amp.

op amp, the high open-loop gain combined with the feedback effectively forces the two inputs to bias at nearly identical voltages, often called “virtual ground.” Circuit analysis is based on the concept of virtual ground. The differential gain of the circuit is equal to

$$V_{OUTDIFF} = V_{+OUT} - V_{-OUT} \approx \left( \frac{R_F}{R_I} \right) \cdot (V_{INP} - V_{INM}) \quad (1)$$

Equation 1 shows that the differential output voltage depends only on the difference between the two inputs, regardless of the absolute voltage on each input. To convert a single-ended input to a differential output, simply connect one of the inputs to ground.

Although Equation 1 explains how to determine the differential output voltage, it neither reveals the voltages of each output, nor the average voltage of the two output nodes. For the LTC6362,

## Why Fully Differential Analog Signals?

An analog signal is usually represented as one signal measured with respect to a fixed potential such as ground, also known as a single-ended signal. But there are times when it is better, or necessary, to make the analog signal fully differential. Fully differential means that two nets each vary with signal. Whenever one voltage goes higher, the other voltage goes lower by the same amount. The analog signal is defined as the voltage difference between these two nets.

One benefit of fully differential signal processing is that it can reduce sensitivity to external interference, such as power supply noise, ground bounce or electromagnetic interference (EMI). For example, if power

supply noise couples equally to both conductors that carry your fully differential signal, then the difference signal may remain undisturbed.

Another benefit of fully differential signal processing is that you can squeeze more signal into a given supply voltage range. For example, in a system that is powered from a single 5V supply, a traditional single-ended signal can vary at most 5V. But a fully differential signal can vary from  $-5V$  to  $5V$ , for  $10V_{p-p}$ . That is because either of the two nets can be higher or lower than the other, which effectively doubles the signal swing. For a given noise floor, doubling the maximum signal swing

results in a 6dB improvement in the signal-to-noise ratio (SNR).

Finally, some semiconductor components mandate, per the data sheet, that you provide a fully differential signal into the input. This fully differential input requirement is near universal for ADCs that convert at a high sample rate (e.g., pipeline ADCs at  $>10M$ sp/s) as well as for ADCs that achieve very high resolution, high linearity and low noise (e.g., SAR ADCs at  $\geq 18$  Bit and  $\geq 100$ dB SNR). Therefore, to use those components, you have no choice but to convert your analog signal to fully differential at some point in your signal chain.

Once you understand that the differential gain of the circuit is simply set by the resistor ratio  $R_F/R_I$ , and the output common mode is independently set by the voltage on  $V_{OCM}$ , it is easy to apply the LTC6362 for a variety of signal translations.

the instantaneous average of the two outputs (also called the output common mode voltage) is independent of the input voltage, but instead is user-determined by the voltage on the  $V_{OCM}$  pin.

$$V_{OUTCM} = V_{OCM} = \left( \frac{V_{+OUT} + V_{-OUT}}{2} \right) \quad (2)$$

If both the average of the two outputs and the difference of the two outputs are known, equations 1 and 2 can be used in as a linear system of two equations and two variables to determine the value of each output.

Figure 2 shows how the LTC6362's differential output can be governed by Equation 1 while the common mode output is held at  $v_{OCM}$ . The LTC6362 features an additional feedback loop with a separate error amplifier. Two internal resistors measure the instantaneous average of the two outputs, and feed that into the error amplifier with its other input tied to the  $v_{OCM}$  pin. The output of the error amplifier is connected to the main amplifier in

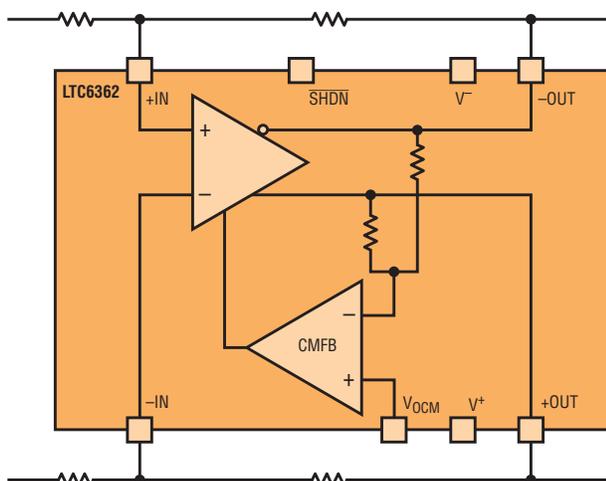


Figure 2. Inside the LTC6362 differential op amp is an additional feedback loop and error amplifier, providing common mode feedback (CMFB).

such a way that it attempts to move each of the op amp outputs higher or lower depending on how the error amplifier is driven. (You can think of it as extra current injected into a mirror driving the main op amp's output stages). When this common mode feedback loop closes to a stable operating point, it ensures that the average of the two outputs is equal to  $v_{OCM}$ , while the difference between the

two outputs is controlled by the differential feedback around the main op amp.

### HOW DO YOU HOOK IT UP?

Once you understand that the differential gain of the circuit is simply set by the resistor ratio  $R_F/R_I$ , and the output common mode is independently set by the voltage on  $v_{OCM}$ , it is easy to apply the LTC6362 for a variety of signal translations. Figures 3 and 4 show some typical examples.

Table 1. LTC6362 key specifications

Supply Current	1mA
Maximum Offset Voltage	200µV
-3dB Bandwidth	34MHz
Input Noise Density	3.9nV/√Hz
Input Voltage Range	Rail-to-Rail
Output Voltage Range	Rail-to-Rail
Supply Voltage	2.8V – 5.25V

Figure 3. Translating a 0V-5V input signal to 9V<sub>p-p</sub> differential output

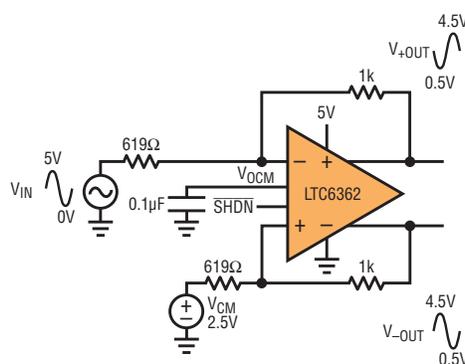
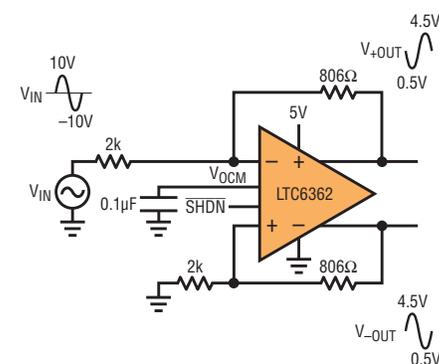
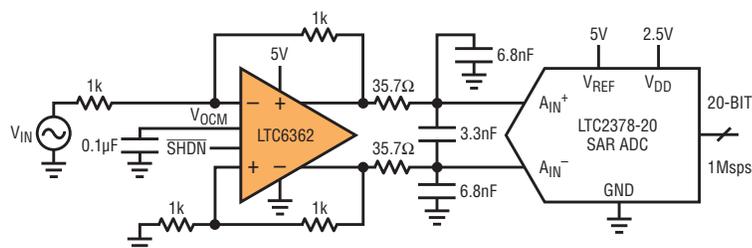


Figure 4. Translating a ±10V input signal to a 9V<sub>p-p</sub> differential output



The input signal can swing below ground, even though the op amp is not powered from a negative supply rail. This works thanks to the divider action of the feedback resistors and the LTC6362's rail-to-rail input stage. The input pins of the op amp are fully specified to operate for any voltage up to and including either of the supply rails.

Figure 5. The LTC6362 driving the 20-bit 1Msps LTC2378-20 ADC



You may notice that in the example of Figure 4, we let the input signal swing below ground, even though the op amp is not powered from a negative supply rail. This works thanks to the divider action of the feedback resistors and the LTC6362's rail-to-rail input stage. The input pins of the op amp itself are fully specified to operate for any voltage up to and including either of the supply rails. You can analyze or simulate any of the example circuits and see that the voltage at the op amp inputs does not need to go below ground even though the voltage at the signal input does.

#### LOW NOISE AND HOW DO YOU CALCULATE IT?

One key advantage of differential signal processing is the potential to double the signal-to-noise ratio of the system. So let's take a closer look at how to analyze the noise performance of a differential op amp such as the LTC6362.

The op amp specifies an input-referred noise density of just under  $4\text{nV}/\sqrt{\text{Hz}}$ . This can be modeled as a noise voltage source in series with one of the inputs. The effect of that noise on the output of the circuit is the inverse of the feedback

ratio. For example, if all resistors are equal value, then half of the output voltage is fed back into the op amp inputs, and the op amp input noise appears times a factor of two at the output.

In addition, each of the four feedback resistors contributes noise at  $\sqrt{4kTR}$ . You need to combine the contributions of these noise sources. For the example with  $R_F = R_I = 1\text{k}$ , this results in a total output noise of  $12\text{nV}/\sqrt{\text{Hz}}$ .

Once you know the total output noise density of the circuit, you can calculate RMS noise by integrating over the bandwidth of interest. As usual, the lower the bandwidth, the more you time-average the noise, and therefore the lower the total observed noise becomes. For example, the noise bandwidth of a 1MHz single-pole filter is 1.57MHz. Integrating the  $12\text{nV}/\sqrt{\text{Hz}}$  noise density over that bandwidth results in approximately  $15\mu\text{V}_{\text{RMS}}$  of total noise.

You can calculate the maximum SNR of the circuit by dividing the maximum signal by the noise. The LTC6362 features rail-to-rail outputs which, on a single 5V supply, result in an almost  $10\text{V}_{\text{P-P}}$  differential output swing. If you convert that to RMS ( $3.5\text{V}_{\text{RMS}}$ ) and divide by the noise

( $15\mu\text{V}_{\text{RMS}}$ ), you get an SNR of more than 233,000 or 107dB in this 1MHz of noise bandwidth. The 107dB figure makes the LTC6362 a good match for the LTC2378-20 20-bit SAR ADC, which features 104dB SNR, and requires fully differential input drive.

#### DRIVING A 20-BIT SAR ADC

Figure 5 shows how to use the LTC6362 to drive the LTC2378-20, a 20-bit SAR ADC sampling at 1Msps. The LTC6362 takes a single-ended input signal and converts it to a fully differential output, just as the ADC specifies.

The RC filter network in between the amplifier and the ADC serves several purposes. First, the filter network reduces the amount of wideband noise that would enter the ADC. For an ADC sampling at 1Msps, the Nyquist criterion says that any signals above 500kHz would alias and become indistinguishable from lower frequency signals. That is true for wideband noise as well, so there is no reason to let the wideband noise enter into the ADC. Second, the capacitors serve as a charge reservoir to absorb charge kick-back from the ADC's internal sampling capacitors. Every time that the ADC finishes the previous conversion, it reconnects

The combined performance of the 20-bit SAR ADC and LTC6362 achieves an SNR of 103dB and THD of 107dB. This is breakthrough performance for a data acquisition system at 1Msps. Best of all, the LTC6362 eases the burden of providing a fully differential input into the ADC with a precise common mode.

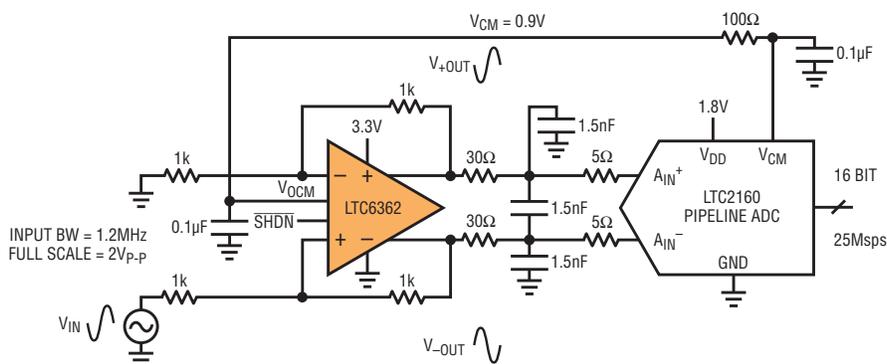


Figure 6. LTC6362 drives the LTC2160 16-bit high speed pipeline ADC

MEASURED PERFORMANCE FOR LTC6362 DRIVING LTC2160:  
 INPUT:  $f_{IN} = 2\text{kHz}$ ,  $-1\text{dBFS}$   
 SNR: 77.0dB  
 HD2:  $-98.9\text{dBc}$   
 HD3:  $-102.3\text{dBc}$   
 THD:  $-96.3\text{dB}$

discharged sample capacitors (about 45pF) to the amplifier circuit. By placing a much larger reservoir capacitor at the ADC inputs, you reduce the voltage excursion caused by these sample capacitors.

The period of time after the ADC has finished a conversion and before it starts the next conversion is called the acquisition time. This is the period of time during which the sample capacitors remain connected to the amplifier circuit.

Ideally, during this time, the RC network fully settles to within the resolution of the ADC. In practice, you may make a trade-off between wideband noise and settling time. Fortunately, the LTC2378-20 ADC charge kickback is relatively linear, ensuring excellent linearity, even when the sample capacitors have not yet fully settled to their final value.

The combined performance of the circuit shown in Figure 5 achieves an

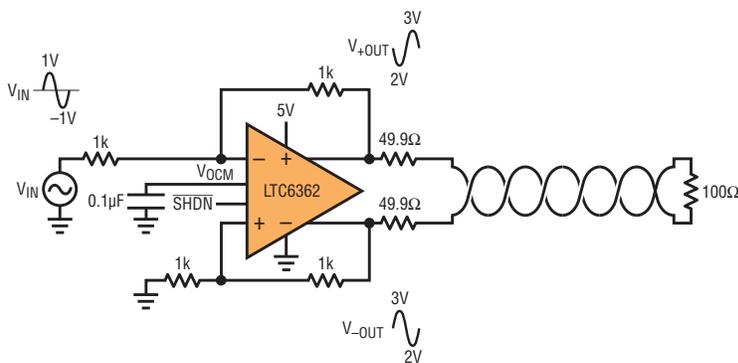
SNR of 103dB and THD of 107dB. This is breakthrough performance for a data acquisition system at 1Msps. Best of all, the LTC6362 eases the burden of providing a fully differential input into the ADC with a precise common mode. Indeed, the input to this circuit is a traditional single-ended analog signal. To configure the circuit for other input ranges, simply choose appropriate external resistors, as shown in the earlier figures.

#### DRIVING A 25Msps PIPELINE ADC

High speed pipeline ADCs typically require that their inputs be driven fully differential. Figure 6 shows how to take a single-ended analog signal with frequency content from DC to about 1MHz and oversample it at 25Msps.

Here, the LTC6362 converts the signal to differential outputs. The LTC2160 ADC wants the common mode of its inputs to be at 0.9V. You achieve this by connecting the VCM pin of the ADC to the

Figure 7. LTC6362 conditions a signal to drive across a differential twisted pair.



The LTC6362 is a capable and versatile differential op amp. Accurate DC specifications, low power and rail-to-rail operation enable it to drive a variety of high performance ADCs that require differential signals, as well as to perform active filtering or drive differential cables.

v<sub>OCM</sub> pin of the amplifier. Few differential op amps are capable of supporting this low of a common mode voltage this low while achieving the performance of the LTC6362. The SNR of the circuit is 77dB, equal to the SNR spec of the ADC, and quite impressive given the low total power consumption of 45mW for the ADC and 3mW for the amplifier.

#### DIFFERENTIAL LINE DRIVER

Sometimes you need to transport an analog signal over a relatively large distance from one PCB to another. A good way to do that is by using a differential twisted pair, because it provides immunity to noise coupling and other interference. As previously discussed, the LTC6362 can convert a traditional single-ended signal to fully differential, in this case driving it over the differential line, as shown in Figure 7.

#### DIFFERENTIAL ACTIVE FILTER

An op amp-RC active filter can be used to produce a lowpass filter with multiple poles, and a cutoff frequency that is relatively well established. Circuit examples that do this with a traditional op amp are readily available. The LTC6362 is can be used to fully implement such filters differentially.

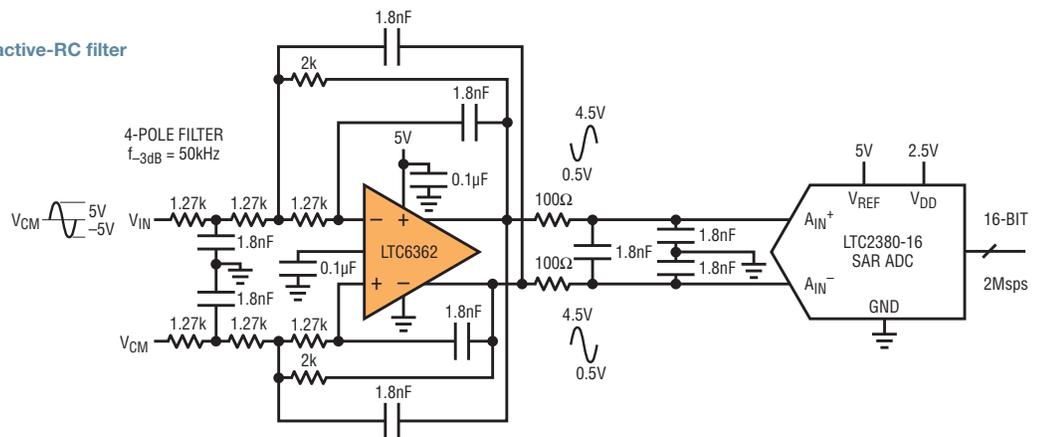
Figure 8 shows an example of a 4-pole 50kHz lowpass filter. In this example, the LTC6362 performs three functions at once: it converts a single-ended input signal to fully differential, it forms a 4-pole lowpass filter and it drives a high performance ADC (in this example, the 16-Bit 20Msps LTC2380-16).

Not all differential op amps are capable of being used in this way. The feedback capacitors create a high frequency short directly from the op amp output to the op amp input, meaning that the feedback factor at high frequencies is much stronger than in a circuit with feedback resistors. If the op amp does not have sufficient phase margin, oscillations or ringing occur. In contrast to a traditional op amp, LTC6362 performs admirably in this configuration.

#### CONCLUSION

The LTC6362 is a capable and versatile differential op amp. Accurate DC specifications, low power and rail-to-rail operation enable it to drive a variety of high performance ADCs that require differential signals, as well as perform active filtering or drive differential cables. ■

Figure 8. Configuring the LTC6362 as a differential active-RC filter





One potential problem of driving two parallel strings from a single output is that twice the current could run in one of the strings if the other string becomes open or nonconducting. The dual current regulation loops in the LT3796-1 can be used to prevent this current hogging situation.

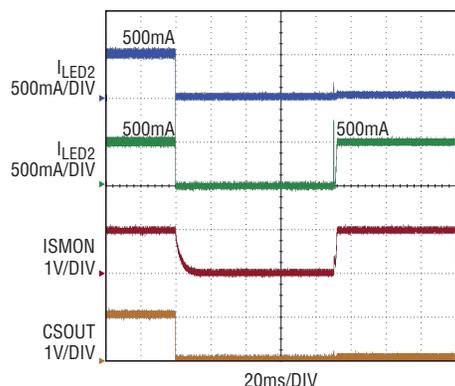


Figure 2. Shorting one LED in string 1 in Figure 1

One potential problem of driving two parallel strings from a single output is that twice the current could run in one of the strings if the other string becomes open or nonconducting. The dual current regulation loops in the LT3796-1 can be used to prevent this current hogging situation.

Figure 1 shows how to configure the ISP/ISN and CSP/CSN current sense amplifiers in a boost LED driver. In normal operation, the ISP/ISN current loop dominates and sets the LED current. The current through LED string 1 is reported at ISMON. The CSP/CSN current loop normally provides monitoring at CSOUT, since it is set 25% below the regulation point of FB2. The CSOUT pin reports the current of LED string 2 as

$$V_{CSOUT} = 1V \cdot \frac{I_{LED2}}{500mA}$$

If the two strings become unbalanced, the reporting pins, ISMON and CSOUT, will show the relative LED current difference in voltage. By externally comparing the

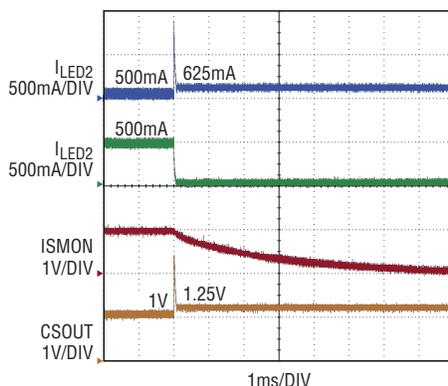


Figure 3. Shorting one LED in string 2 in Figure 1

two analog signals, a fault signal can be triggered by an external controller.

Figure 2 shows the oscilloscope waveform when one LED from string 1 is shorted. The ISP/ISN current sense amplifier instantly senses the overcurrent event, and disconnects both PMOS switches. After one soft-start cycle, the ISP/ISN current loop starts to regulate the LED string at a new output voltage, which corresponds to the forward voltage drop of eight LEDs. Since the new output voltage can't drive nine LEDs, LED string 2 stops conducting current and CSOUT reports 0V. Similarly, if one LED is shorted in string 2, the CSP/CSN current loop takes control and regulates the output current to 625mA through the FB2 pin, as shown in Figure 3. LED string 1 now stops conducting current and ISMON reports 0V.

This converter also provides high performance PWM dimming and robust short-circuit protection for both strings through the high side PMOS disconnect switch driver pin TG. The built-in overcurrent

comparator inside ISP/ISN current sense amplifier protects from a short circuit of string 1, whereas the circuit formed by D4, Q1 and R9–R11 detects the short circuit of string 2, drives FB1 pin high and turns off PMOS switches, M2 and M3.

### LED DRIVER WITH 100:1 ANALOG DIMMING RATIO

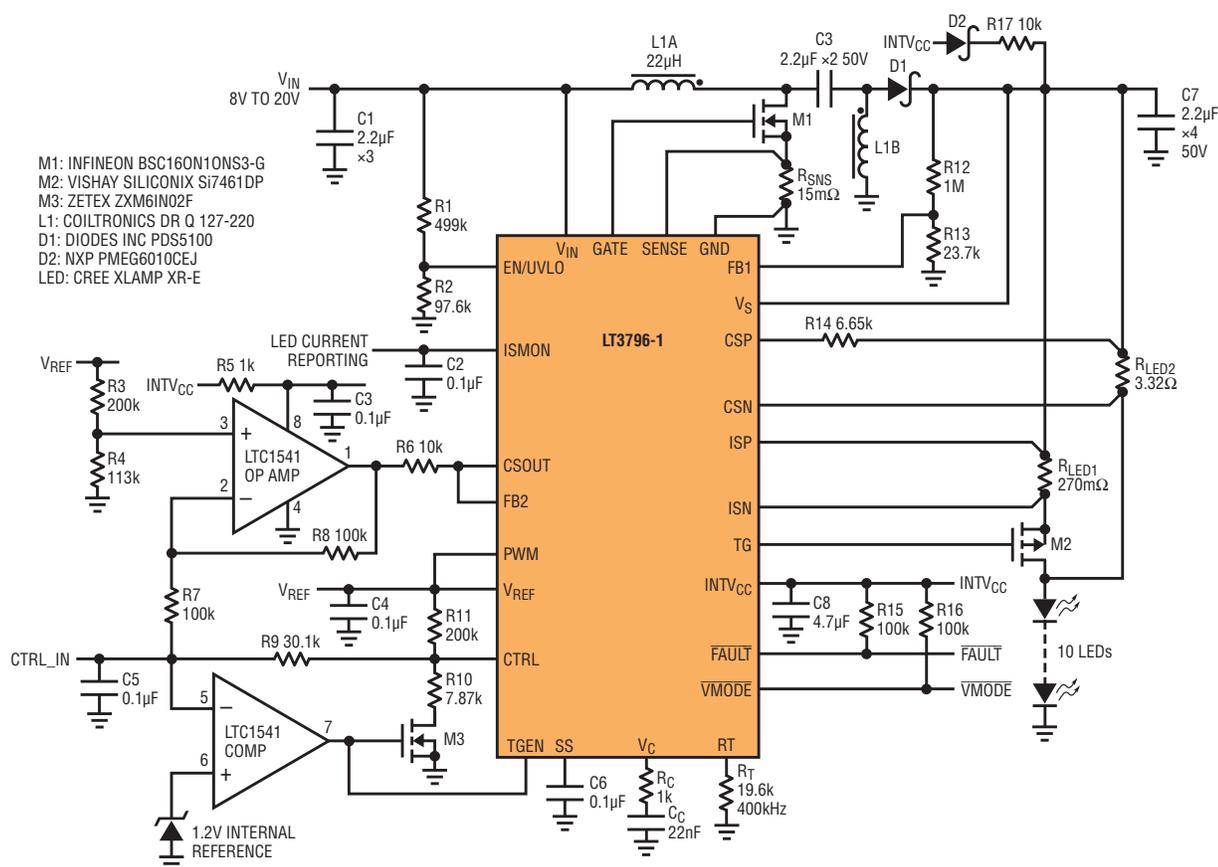
Many high power LED applications require a high analog dimming ratio, which is difficult to achieve with a single current sense path. The problem is dynamic range: at high currents, a low differential voltage is needed to limit power dissipation in the sense resistor, typically 250mV or less, but with so little signal to work with, the several mV accuracy of the current sense amplifier becomes an appreciable contribution to error in the sense voltage even at 10% analog dimming.

The two current sense loops of the LT3796-1 enable it to produce a high analog dimming ratio by dividing the job of current regulation between two loops. One loop features a low value sense resistor to limit power dissipation in the high current path, while the other loop uses a higher sense signal in the low current path to elevate accuracy, when power dissipation is less of a concern. Figure 4 shows the LT3796-1 configured to produce 100:1 analog dimming ratio in SEPIC mode by using the LTC1541 (a precision reference, op amp and a comparator in single package).

Assuming M2's  $R_{DS(ON)}$  is negligible, in the high current range between

The two current sense loops of the LT3796-1 enable it to produce a high analog dimming ratio by dividing the job of current regulation between two loops. One loop features a low value sense resistor to limit power dissipation in the high current path, while the other loop uses a higher sense signal in the low current path to elevate accuracy, when power dissipation is less of a concern.

**Figure 4. A SEPIC mode LED driver with 100:1 analog dimming ratio**



200mA and 1A, the ISP/ISN current loop regulates the output current at

$$I_{LED} = \frac{V_{CTRL\_IN} - 0.2V}{20 \cdot (R_{LED1} \parallel R_{LED2})}$$

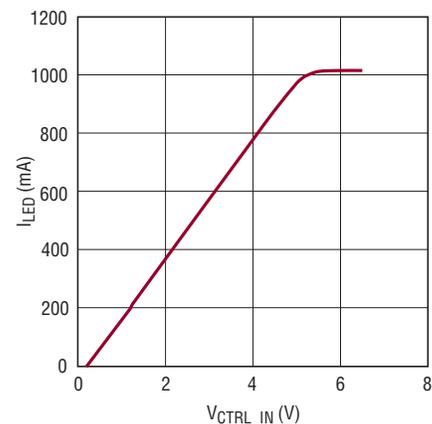
When  $v_{CTRL}$  drops below 1.2V, the LTC1541's comparator forces the TGEN pin low and disconnects M2. Thus the LED string is only sensed and regulated by CSP/CSN loop. When the CSP/CSN loop takes control, the FB2 pin voltage is regulated at 1.25V and the CTRL\_IN input sets the CSP/CSN threshold by pulling current out of CSOUT pin through R6. The CSP/CSN dimming range is from 668mV to 33.4mV, sensing low LED currents

between 200mA to 10mA while maintaining accuracy. Overall, the combined two current loops provide 100:1 analog dimming range as shown in Figure 5.

### CONCLUSION

The LT3796-1 is a LED controller that features two independent current sense amplifiers with reporting, two FB pins and a high side disconnect switch driver. It also features robust fault protection and a versatile toolset to address challenging LED applications such as high reliability and high performance analog dimming. ■

**Figure 5. I\_LED vs CTRL for the solution in Figure 4**



# High Efficiency PolyPhase Power Supply Delivers 30A per Phase with Ultralow Inductor DCR Sensing and Fast Transient Response

Jian Li and Gina Le

The LTC3875 is a feature-rich dual-output synchronous buck controller that meets the power density demands of modern high speed, high capacity data processing systems, telecom systems, industrial equipment and DC power distribution systems.

The LTC3875's features include:

- 4.5V to 38V input range and 0.6V to 3.5V output range
- Proprietary current mode architecture enhances the signal-to-noise ratio of the current sensing signal, allowing the use of ultralow DCR power inductors to maximize efficiency and reduce switching jitter.

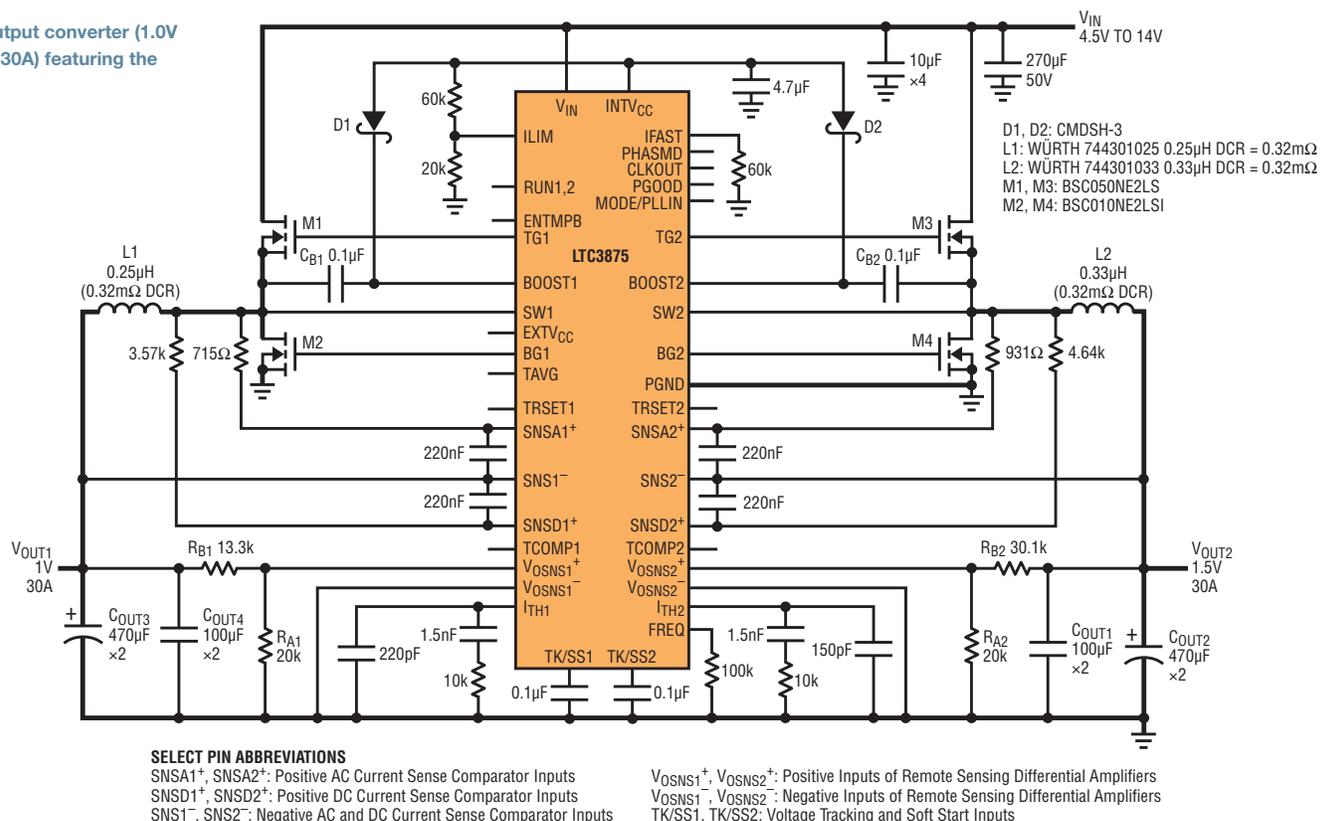
- Fast transient response, facilitating high density design with less output capacitance.
- Remote output voltage sensing and  $\pm 0.5\%$  reference (0.6V) window for accurate regulation.
- On-chip drivers in a 6mm x 6mm QFN package to satisfy demanding space requirements.
- Easy parallel multiphase operation for high current applications.

## DUAL-OUTPUT CONVERTER (1.0V AT 30A AND 1.5V AT 30A)

Figure 1 shows a typical 4.5V~14V input, dual-output solution. The LTC3875's two channels run relative to each other with a 180° phase shift, reducing the input RMS current ripple and capacitor size. Each phase has one top MOSFET and one bottom MOSFET to provide up to 30A of output current.

Similar to LTC3866, the LTC3875 employs a unique current sensing architecture to enhance its signal-to-noise ratio, enabling current mode control via a small sense signal of a very low inductor DCR, 1mΩ or less. As a result, the efficiency is greatly

Figure 1. A dual-output converter (1.0V at 30A and 1.5V at 30A) featuring the LTC3875



The LTC3875 can sense a DCR value as low as  $0.2\text{m}\Omega$  with careful PCB layout. Moreover, an additional temperature compensation circuit can be used to guarantee the accurate current limit over a wide temperature range.

improved and the jitter is reduced. The current mode control yields fast cycle-by-cycle current limit, current sharing and simplified feedback compensation.

The LTC3875 can sense a DCR value as low as  $0.2\text{m}\Omega$  with careful PCB layout. The LTC3875 uses two positive sense pins SNSD<sup>+</sup> and SNSA<sup>+</sup> to acquire signals. The filter time constant of the SNSD<sup>+</sup> should match the  $L/\text{DCR}$  of the output inductor, while the filter at SNSA<sup>+</sup> should have a bandwidth five times larger than that of SNSD<sup>+</sup>. Moreover, an additional temperature compensation circuit can be used to guarantee the accurate current limit over a wide temperature range.

Efficiency can be optimized with ultra-low DCR inductor. As shown in Figure 2, the total solution efficiency in forced continuous mode (CCM) is 87.3% at 1.0V/30A output, and 89.8% at 1.5V/30A. The hot spot (bottom MOSFET) temperature rise is  $57^\circ\text{C}$  without any airflow as shown in Figure 3, where the ambient temperature is about  $23^\circ\text{C}$ .

The LTC3875 features fast transient response and minimizes undershoot through a proprietary solution. Peak current mode control is widely adopted in switching converters due to its cycle-by-cycle peak current limit and easy compensation. However, the inherent switching cycle delay of peak current mode control results in large undershoot of the output voltage when there is a load step-up. The LTC3875 overcomes undershoot by using a dynamic switching frequency adjustment scheme. The internal

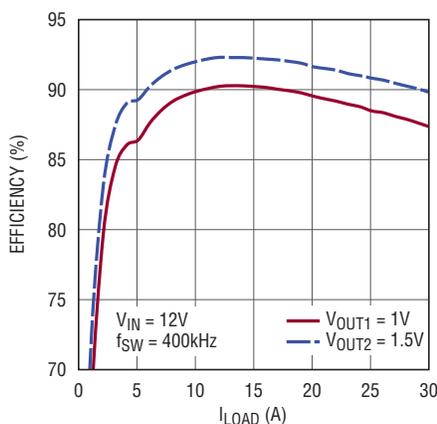
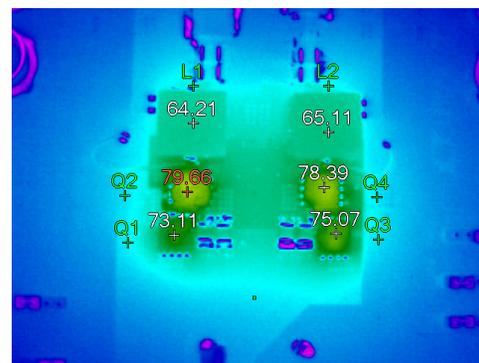


Figure 2. Efficiency comparison of the two channels



$V_{IN} = 12\text{V}$ ,  $V_{OUT1} = 1.0\text{V}/30\text{A}$ ,  $V_{OUT2} = 1.5\text{V}/30\text{A}$ , NO AIRFLOW

Figure 3. Thermal test results

transient detector can detect a large voltage undershoot, leading the LTC3875 to run the power stage at twice the preset switching frequency for about 20 cycles.

Figure 4 shows that switching cycle delay is reduced from  $2.18\mu\text{s}$  to  $1.2\mu\text{s}$  and voltage undershoot is reduced from  $95\text{mV}$  to  $67.5\text{mV}$  (29% reduction) at 15A load up

after the fast transient is enabled. In other words, with fast transient enabled, the LTC3875 can achieve the same transient performance as without, but with 20% less output capacitance, increasing power density and reducing total cost. Compared to other nonlinear control methods, the response scheme used by the LTC3875 is linear, simplifying overall design.

Figure 4. Transient comparison

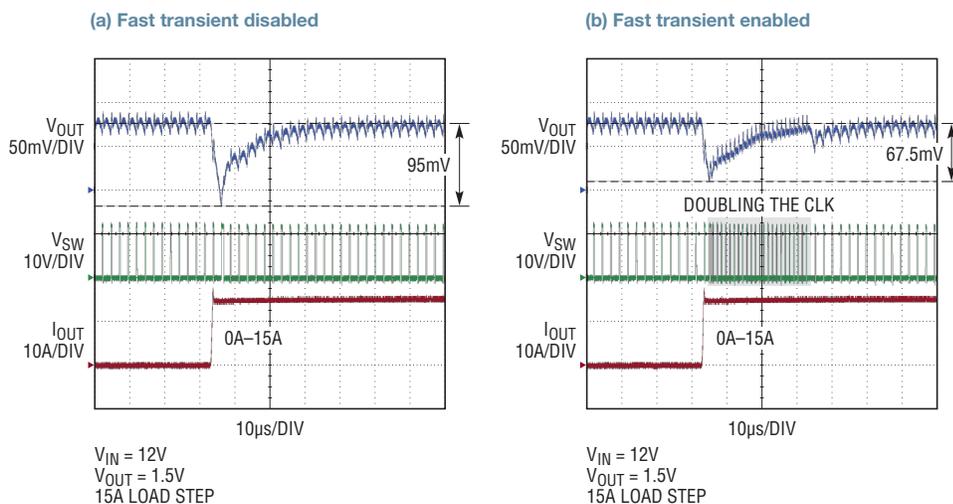
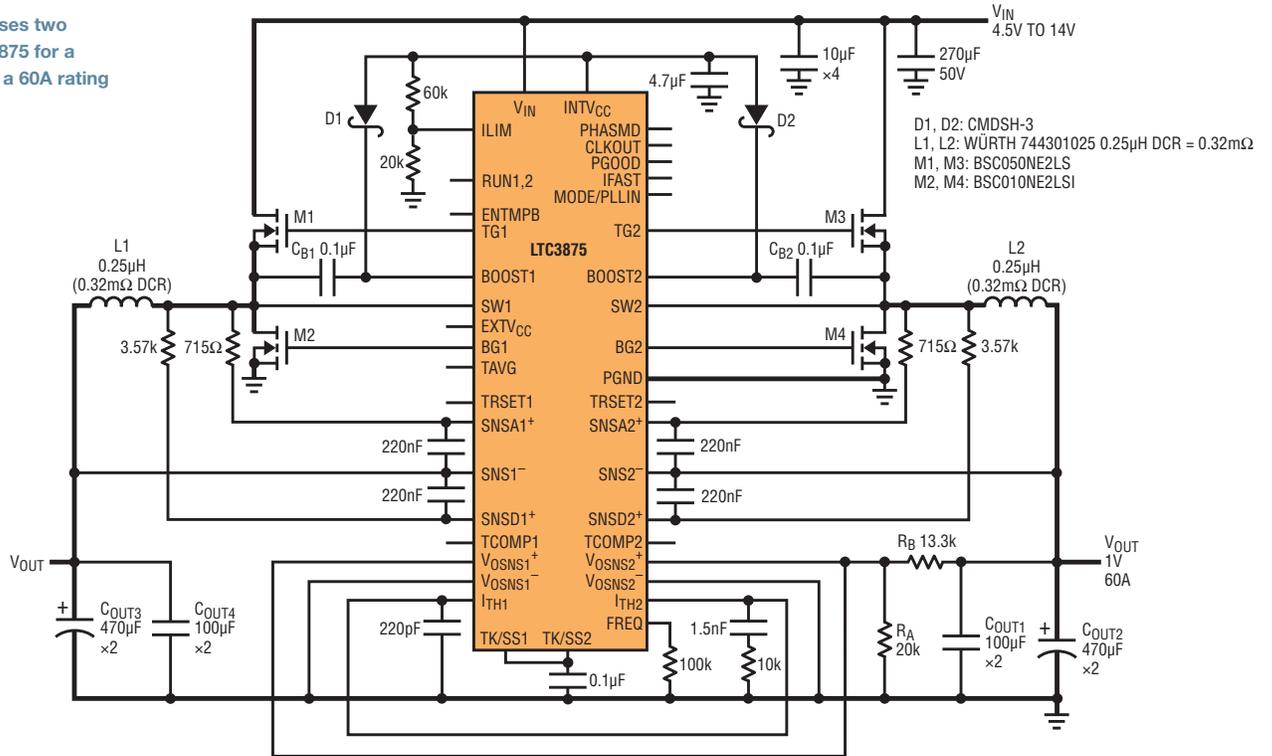


Figure 5. Converter uses two channels of the LTC3875 for a single 1V output with a 60A rating



**SELECT PIN ABBREVIATIONS**  
 SNSA1<sup>+</sup>, SNSA2<sup>+</sup>: Positive AC Current Sense Comparator Inputs  
 SNSD1<sup>+</sup>, SNSD2<sup>+</sup>: Positive DC Current Sense Comparator Inputs  
 SNS1<sup>-</sup>, SNS2<sup>-</sup>: Negative AC and DC Current Sense Comparator Inputs  
 VOSNS1<sup>+</sup>, VOSNS2<sup>+</sup>: Positive Inputs of Remote Sensing Differential Amplifiers  
 VOSNS1<sup>-</sup>, VOSNS2<sup>-</sup>: Negative Inputs of Remote Sensing Differential Amplifiers  
 TK/SS1, TK/SS2: Voltage Tracking and Soft Start Inputs

**SINGLE OUTPUT, DUAL PHASE, HIGH CURRENT CONVERTER (12V TO 1V AT 6A)**

The LTC3875 can be easily configured as a dual-phase single-output converter for higher current solutions. Figure 5 shows a buck converter that produces a 1V, 60A output from a 12V input. Multiple ICs

can be paralleled and phase-interleaved for even higher current if required.

The DC current sharing between the two channels is shown in Figure 6. The difference at full load is around 1.6A with 0.32mΩ DCR inductor. Thanks to the peak current mode control architecture,

the dynamic current sharing is also very good, as shown in Figure 7.

**CONCLUSION**

The LTC3875 delivers high efficiency with reliable current mode control, ultralow DCR sensing and strong integrated drivers in a 6mm × 6mm 40-pin QFN. It supports temperature-compensated DCR sensing for high reliability. Its fast transient response can help improve the transient response with minimum output capacitance. Tracking, multichip operation, and external sync capability fill out its menu of features. The LTC3875 is ideal for high current applications, such as telecom and datacom systems, industrial and computer systems applications. ■

Figure 6. DC current sharing of 60A solution shown in Figure 5

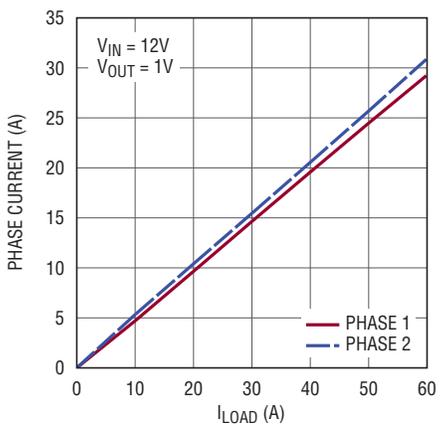
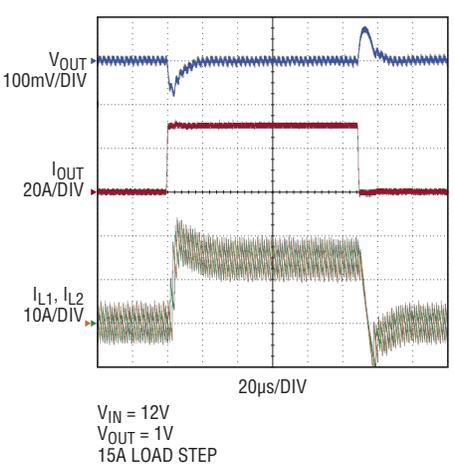


Figure 7. Dynamic current sharing of 60A solution shown in Figure 5



# Silent Switcher Meets CISPR Class 5 Radiated Emissions While Maintaining High Conversion Efficiency

Tony Armstrong & Christian Kück

When minimizing EMI is a design priority, a linear regulator can be a low noise solution, but heat dissipation and efficiency requirements may preclude that choice and point to a switching regulator. Even in EMI-sensitive applications, a switching regulator is typically the first active component on the input power bus line, and regardless of downstream converters, it significantly impacts overall converter EMI performance. Until now, there was no sure way to guarantee that EMI could be suppressed and efficiency requirements attained via power IC selection. The LT8614 Silent Switcher™ regulator now makes this possible.

The LT8614 reduces EMI by more than 20dB compared to current state-of-the-art switching regulators. In comparison, it lowers EMI by 10x in the frequency range above 30MHz without compromising minimum on- and off-times or efficiency in an equivalent board area. It accomplishes this with no additional components or shielding, representing a significant breakthrough in switching regulator design.

## A NEW SOLUTION TO EMI ISSUES

The tried and true solution to EMI issues is to use a shielding box for the complete circuit. Of course, this adds a significant

costs in required board space, components and assembly, while complicating thermal management and testing. Another method is to slow down the switching edges. This has the undesired effect of reducing the efficiency, increasing minimum on-, off-times, and their associated dead times and compromises the potential current control loop speed.

The LT8614 Silent Switcher regulator delivers the desired effects of a shielded box without using one (see Figure 1). The LT8614 features a low  $I_Q$  of 2.5μA total supply current consumed

by the device, in regulation with no load—important for always-on systems.

Its ultralow dropout is only limited by the internal top switch. Unlike alternative solutions, the LT8614's  $V_{IN}-V_{OUT}$  limit is not limited by maximum duty cycle and minimum off-times. The device skips its switch-off cycles in dropout and performs only the minimum required off cycles to keep the internal top switch boost stage voltage sustained, as shown in Figure 6.

At the same time, the minimum operating input voltage is only 2.9V typical (3.4V maximum), enabling it to supply

Figure 1. The LT8614 Silent Switcher minimizes EMI/EMC emissions while delivering high efficiency at frequencies up to 3MHz.

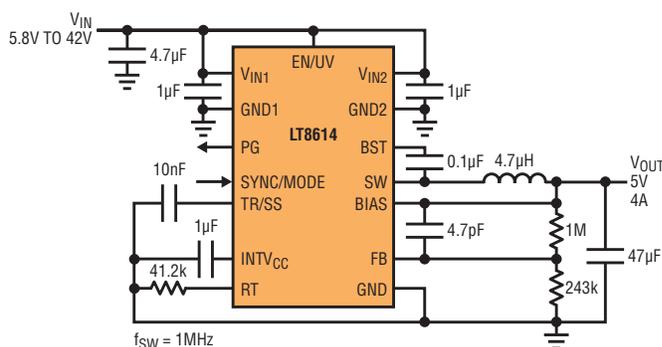
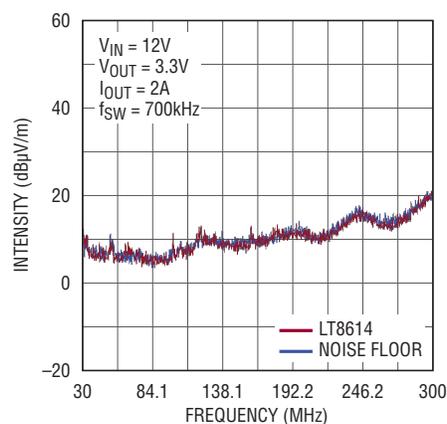


Figure 2. LT8614 board meets CISPR25 radiated standard in an anechoic chamber. Noise floor is equal to LT8614 radiated emissions.



The LT8614 reduces EMI by more than 20dB when compared current state-of-the-art switching regulators. In comparison, it lowers EMI by 10x in the frequency range above 30MHz without compromising minimum on- and off-times or efficiency in an equivalent board area. It accomplishes this feat with no additional components or shielding, representing a significant breakthrough in switching regulator design.

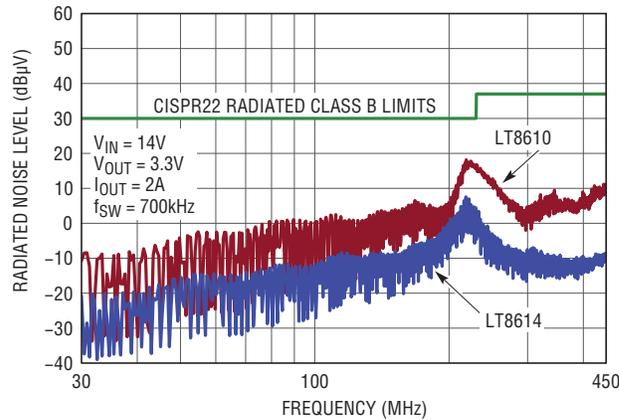


Figure 3. Comparison of radiated emissions for LT8614 and LT8610

The Figure 2 shows measurements taken in an anechoic chamber at 12V input, 3.3V output at 2A with a fixed switching frequency of 700kHz. To compare the LT8614 Silent Switcher technology against another current state-of-the-art switching regulator, the part was measured against the LT8610 (see Figure 3). The test was performed in a GTEM cell using the same load, input voltage and the same inductor on the standard demo boards for both parts.

a 3.3V rail with the part in dropout. At high currents, the LT8614 has higher efficiency than comparable parts since its total switch resistance is lower.

The LT8614 can be synchronised to an external frequency operating from 200kHz to 3MHz. AC switching losses are low, so it can be operated at high switching frequencies with minimal efficiency loss. In EMI-sensitive applications, such

as those commonly found in many automotive environments, a good balance can be attained and the LT8614 can run either below the AM band for even lower EMI, or above the AM band. In a setup with 700kHz operating switching frequency, the standard LT8614 demo board does not exceed the noise floor in a CISPR25, Class 5 measurement.

One can see that up to a 20dB improvement is attained using the LT8614 Silent Switcher technology compared to the already very good EMI performance of the LT8610, especially in the more difficult to manage high frequency area.

In the time domain, the LT8614 shows benign behavior on the switch node edges, as shown in Figures 4 and 5. Even at 4ns/div, the LT8614 Silent Switcher regulator shows minimal ringing. In contrast, the LT8610 successfully damps

Figure 4. Comparison of switch node rising edges for LT8614 Silent Switcher and the LT8610

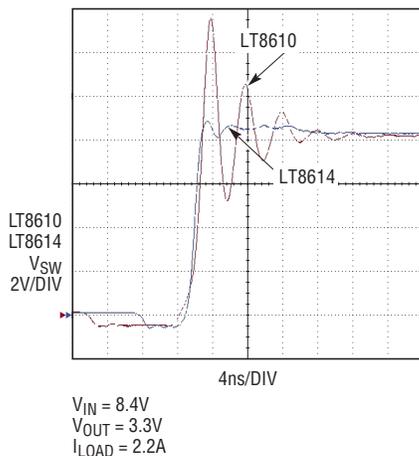


Figure 5. Near ideal square wave switch waveform of LT8614 enables low noise operation

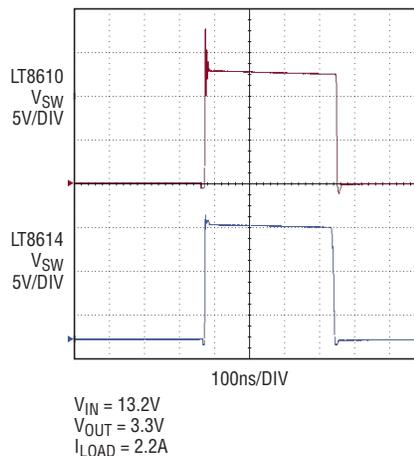
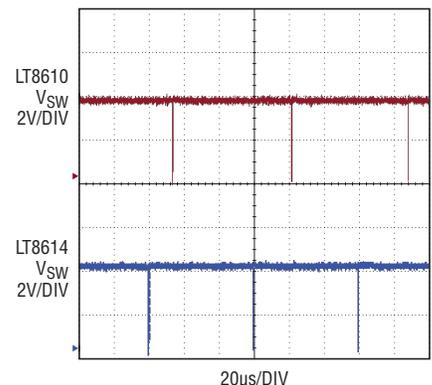


Figure 6. Dropout behavior of switch node for LT8614 and LT8610

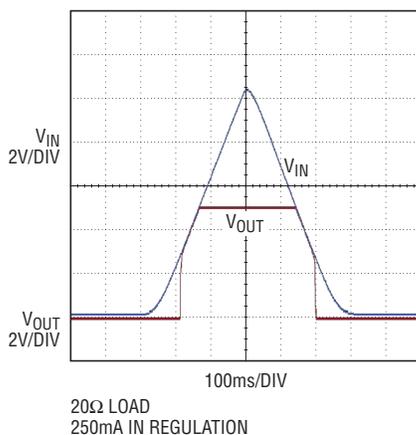


ringing as shown in Figure 4, but one can see the higher energy stored in the hot loop compared to the LT8614 (Figure 4).

Figure 5 shows the switch node at 13.2V input, and how the LT8614 achieves a near ideal square wave at the switch node. All time domain measurements in Figures 4, 5 and 6 are performed with 500MHz Tektronix P6139A probes with close probe tip shield connection to the PCB GND plane. Off-the-shelf demo boards are used for both parts.

The 42V absolute maximum input voltage ratings of the LT861x family is important for automotive and industrial environments. Just as important, especially in automotive situations, is dropout behavior. Often critical 3.3V logic supplies must be supported through cold crank situations. The LT8614 Silent Switcher regulator maintains the close to ideal behavior of the LT861x family in this case. Instead of higher undervoltage lockout voltages and maximum duty cycle clamps of alternative parts, the LT8610/11/14 devices operate down to 3.4V and start skipping off cycles as soon as necessary, as shown in Figure 6. This results in the ideal dropout behavior as shown in Figure 7.

**Figure 7. LT8614 dropout performance. Like other LT861x devices, it operates down to 3.4V and starts skipping off cycles as soon as necessary.**



The LT8614's low minimum on-time of 30ns enables large step-down ratios even at high switching frequencies. As a result, it can supply logic core voltages with a single step-down from inputs up to 42V.

## CONCLUSION

It is well known that EMI considerations require careful attention during the initial converter design in order to

pass EMI testing at system completion. The LT8614 Silent Switcher regulator makes it possible to ensure success with a simple power IC selection. The LT8614 reduces EMI from current state-of-the-art switching regulators by more than 20dB, even as it increases conversion efficiencies—no additional components or extra shielding are required. ■

## Switching Regulators and EMI

Printed circuit board layout determines the success or failure of every power supply. It sets functional, electromagnetic interference (EMI) and thermal behavior. While switching power supply layout is not a black art, it can often be overlooked in the initial design process. Since functional and EMI requirements must be met, what is good for functional stability of the power supply is also usually good for its EMI emissions. It should be noted that good layout from the beginning does not add cost, but can actually produce cost savings, eliminating the need for EMI filters, mechanical shielding, EMI test time and PC board revisions.

There are two types of EMI emissions: conducted and radiated. Conducted emissions ride on the wires and traces that connect to a product. Since the noise is localized to a specific terminal or connector in the design, compliance with conducted emissions requirements can often be assured relatively early in the development process with a good layout and filter design.

Radiated emissions, however, are another story. Everything on the board that carries current radiates an electromagnetic field. Every trace on the board is an antenna, and every copper plane is a mirror. Anything, other than a pure sine wave or DC voltage, generates a wide signal spectrum. Even with careful design, a designer never really knows how bad the radiated emissions are going to be until the system gets tested. And radiated emissions testing cannot be formally performed until the design is essentially complete.

Filters are often used to reduce EMI by attenuating the strength at a certain frequency or over a range of frequencies. A portion of this energy that travels through space (radiated) is attenuated by adding sheet metal as magnetic shields. The lower frequency part that rides on PCB traces (conducted) is tamed by adding ferrite beads and other filters. EMI cannot be eliminated, but can be attenuated to a level that is acceptable by other communication and digital

components. Moreover, several regulatory bodies enforce standards to ensure compliance.

Modern input filter components in surface mount technology have better performance than through-hole parts. Nevertheless, this improvement is outpaced by the increase in operating switching frequencies of switching regulators. Higher efficiency, low minimum on- and off-times result in higher harmonic content due to the faster switch transitions. For every doubling in switching frequency, the EMI becomes 6dB worse while all other parameters, such as switch capacity and transition times, remain constant. The wideband EMI behaves like a first order high pass with 20dB higher emissions if the switching frequency increases by 10 times.

Savvy PCB designers will make the hot loops small and use shielding ground layers as close to the active layer as possible. Nevertheless, device pinouts, package construction, thermal design requirements and package sizes needed for adequate energy storage in decoupling components dictate a minimum hot loop size. To further complicate matters, in typical planar printed circuit boards, the magnetic or transformer style coupling between traces above 30MHz will diminish all filter efforts since the higher the frequencies become the more effective the unwanted magnetic or antenna coupling becomes.

The potential problem for interference and noise can be exacerbated when multiple DC/DC switch mode regulators are paralleled for current sharing and higher output power. If all are operating (switching) at a similar frequency, the combined energy generated by multiple regulators in a circuit is concentrated at that frequency and its harmonics. Presence of this energy can become a concern especially to the rest of ICs on the PC board and other system boards close to each other and susceptible to this radiated energy. This can be particularly troubling in automotive systems which are densely populated and are often in close proximity to audio, RF, CAN bus and various receiving systems.

# What's New with LTspice IV?

Gabino Alonso



New Video: "Stability of Op Amp Circuits"  
[www.linear.com/solutions/4449](http://www.linear.com/solutions/4449)

## NOW NATIVE ON MAC OS X

LTspice for Mac OS X 10.7+ platforms is now available at [www.linear.com/LTspice](http://www.linear.com/LTspice).



This new release has similar capabilities and features as its Windows counterpart. To access the menu you can right-click or use shortcuts. A guide to Mac OS X shortcuts is also available online.

## BLOG UPDATE

Check out the LTspice blog ([www.linear.com/solutions/LTspice](http://www.linear.com/solutions/LTspice)) for tech news, insider tips and interesting points of view regarding LTspice.

**New Video on the Blog, "Stability of Op Amp Circuits"**—The latest video topic is available at [www.linear.com/solutions/4449](http://www.linear.com/solutions/4449).

We all know that feedback circuits can oscillate. We may even know a few tricks to fix the problem, but wouldn't it be

## What is LTspice IV?

LTspice® IV is a high performance SPICE simulator, schematic capture and waveform viewer designed to speed the process of power supply design. LTspice IV adds enhancements and models to SPICE, significantly reducing simulation time compared to typical SPICE simulators, allowing one to view waveforms for most switching regulators in minutes compared to hours for other SPICE simulators.

LTspice IV is available free from Linear Technology at [www.linear.com/LTspice](http://www.linear.com/LTspice). Included in the download is a complete working version of LTspice IV, macro models for Linear Technology's power products, over 200 op amp models, as well as models for resistors, transistors and MOSFETs.

nice if our simulation tool could show us exactly what is happening, and why? This video illustrates how to use the .AC analysis to look at open loop gain and phase of operational amplifier feedback circuits in LTspice IV. It explains how to break the feedback loop in an op amp circuit while maintaining the correct operating point so that the open loop transfer function of the circuit can be obtained and the phase margin measured. This video also covers some common techniques on how to improve the phase margin of a design and improve your circuit intuition.

## SELECTED DEMO CIRCUITS

For a complete list of example simulations utilizing Linear Technology devices, please visit [www.linear.com/democircuits](http://www.linear.com/democircuits).

### Linear Regulators

- **LT3007:** 3.3V, 20mA linear regulator with shutdown (3.8V–45V to 3.3V at 20mA) [www.linear.com/LT3007](http://www.linear.com/LT3007)
- **LT3090:** Negative linear regulator with current monitor (–5V to –1.25V at 600mA) [www.linear.com/LT3090](http://www.linear.com/LT3090)

### µModule Regulators

- **LTM4620A/LTM4676:** High current, parallel µModule buck regulators with power system management (4.5V–16V to 1V at 100A) [www.linear.com/LTM4620A](http://www.linear.com/LTM4620A)
- **LTM4676:** Dual 13A µModule buck regulator with digital interface for control & monitoring (5.75V–26.5V to 1V at 13A & 1.8V at 13A) [www.linear.com/LTM4676](http://www.linear.com/LTM4676)

- **LTM4630:** High efficiency 8-phase 140A step-down regulator (4.5V–15V to 1V at 140A) [www.linear.com/LTM4630](http://www.linear.com/LTM4630)

### Flyback Controller

- **LT8302:** µPower no-opto isolated flyback converter (10V–30V to 5V at 2.2A) [www.linear.com/LT8302](http://www.linear.com/LT8302)

### LED Drivers

- **LT3955:** 20W boost LED driver with internal PWM dimming (5V–60V to 67V LED string at 300mA) [www.linear.com/LT3955](http://www.linear.com/LT3955)

### Boost Regulators

- **LTC3788-1 Demo Circuit:** High efficiency dual 12V/24V boost converter with R<sub>SENSE</sub> current sensing (4.5V–24V to 24V at 5A & 12V at 10A) [www.linear.com/LTC3788-1](http://www.linear.com/LTC3788-1)

### Current Sense Amplifiers

- **LT1999-20:** High voltage bidirectional current sense –5V to 80V Input) [www.linear.com/LT1999](http://www.linear.com/LT1999)
- **LT6105:** Unidirectional current sense amplifier for a 1V supplies (0A to 10A) [www.linear.com/LT6105](http://www.linear.com/LT6105)

### ADC Driver

- **LTC6360:** ±10V input signal to a 5V single-ended ADC driver [www.linear.com/LTC6360](http://www.linear.com/LTC6360)

### Oscillator

- **LTC6991:** Low frequency voltage controlled oscillator (250Hz to 1kHz) [www.linear.com/LTC6991](http://www.linear.com/LTC6991)

 Follow @LTspice on Twitter for up-to-date information on models, demo circuits, events and user tips: [www.twitter.com/LTspice](http://www.twitter.com/LTspice)

 Like us on Facebook at [facebook.com/LTspice](http://facebook.com/LTspice)

## SELECTED MODELS

### Buck Regulators

- **LT8610A/AB:** 42V, 2.5A synchronous step-down regulator with 2.5 $\mu$ A quiescent current [www.linear.com/LT8610](http://www.linear.com/LT8610)
- **LT8614:** 42V, 4A synchronous step-down Silent Switcher™ regulator with 2.5 $\mu$ A quiescent current [www.linear.com/LT8614](http://www.linear.com/LT8614)
- **LTC3607:** Dual 600mA 15V monolithic synchronous step-down DC/DC regulator [www.linear.com/LTC3607](http://www.linear.com/LTC3607)

### Charge Pump

- **LTC3255:** Wide  $V_{IN}$  range fault protected 50mA step-down charge pump [www.linear.com/LTC3255](http://www.linear.com/LTC3255)

### MOSFET Driver

- **LTC4440A-5:** High speed, high voltage, high side gate driver [www.linear.com/LTC4440A-5](http://www.linear.com/LTC4440A-5)

### Ideal Diodes & Hot Swap Controllers

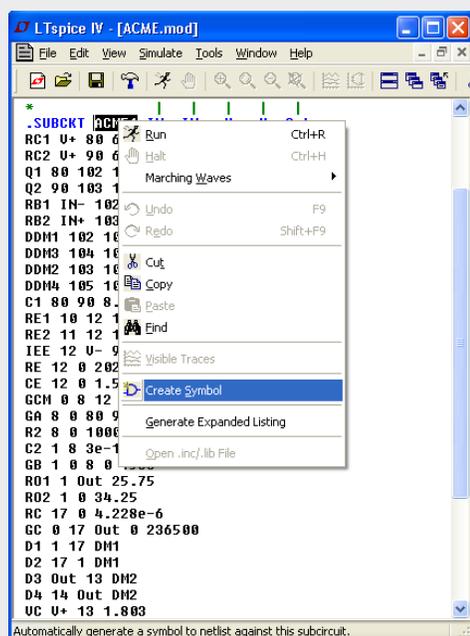
- **LTC4221:** Dual Hot Swap™ controller/power sequencer with dual speed, dual level fault protection [www.linear.com/LTC4221](http://www.linear.com/LTC4221)
- **LTC4229:** Ideal diode and Hot Swap controller [www.linear.com/LTC4229](http://www.linear.com/LTC4229)
- **LTC4280:** Hot Swap controller with I<sup>2</sup>C-compatible monitoring [www.linear.com/LTC4280](http://www.linear.com/LTC4280)

## Power User Tip

### IMPORTING THIRD-PARTY MODELS

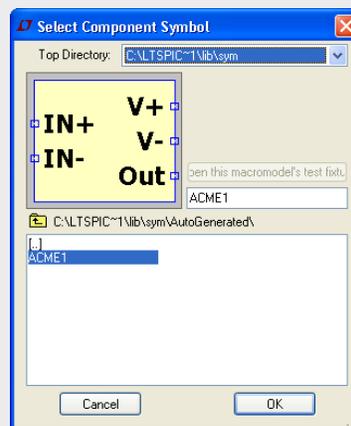
It is possible in LTspice to create a new symbol from scratch for a third-party model but who has the time. Follow these easy steps to generate a new symbol for a third-party model defined in a subcircuit (.SUBCKT statement).

1. Open the netlist file that contains the subcircuit definitions in LTspice.  
(File > Open or drag file into LTspice)
2. Right-click the line containing the name of the subcircuit, and select Create Symbol:



3. Edit the symbol if needed and save.

To use the new symbol (and associated third party model) in a schematic, select the symbol from the AutoGenerated directory in the component library (F2) and place it in your schematic:



By using the automatic symbol generation you can focus on your simulations, not creating new symbols. For a more information on how to import third party models that use intrinsic SPICE device (.MODEL statement) see the video at [www.linear.com/solutions/1083](http://www.linear.com/solutions/1083).



Happy simulations!

# Complete 20V, 20A High Performance $\mu$ Module DC/DC Regulator in 15mm $\times$ 15mm $\times$ 4.32mm Package

Sam Young

The LTM4637 is a complete 20A high efficiency switch mode power supply in a compact 15mm  $\times$  15mm  $\times$  4.32mm LGA package. The controller, power MOSFETs, inductor and compensation circuitry are all integrated within its low profile package, allowing use on the bottom side of the PCB. The LTM4637 has an input voltage range of 4.5V to 20V and supports an output voltage range of 0.6V to 1.8V. It has an operating frequency range of 440kHz–770kHz where it can provide up to 20A with just a few additional external components.

## VERSATILITY

The LTM4637 features a programmable soft-start to limit inrush currents as well as output voltage tracking for sequencing requirements. Also included are an output voltage power good indicator, an external  $V_{CC}$  pin allowing for bypassing of the internal LDO for an extra efficiency benefit, selectable pulse-skipping mode and Burst Mode operation for increased light load efficiency, as well as internal temperature monitoring.

## EASY COMPENSATION; EASY PARALLEL OPERATION

The LTM4637's constant-frequency current mode control and internal feedback loop compensation provide excellent stability and transient performance for a wide range of output capacitors. The LTM4637 also includes an internal differential remote sense amplifier, as well as a phase locked loop and voltage controlled oscillator for synchronization to an external clock. These features make the LTM4637 ideal for paralleling modules to satisfy even higher load current demands. The current mode control architecture of the LTM4637 ensures balanced load current sharing to maximize efficiency and thermal benefits. The differential remote sense

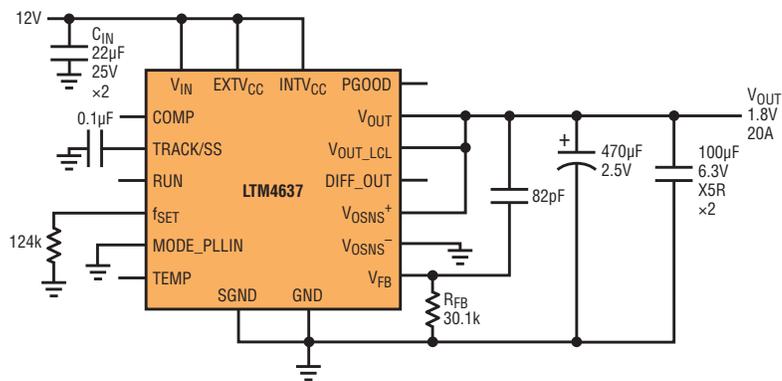
amplifier allows for paralleled LTM4637s to regulate a precise output voltage  $\pm 1.5\%$  at the point-of-load, compensating for error caused by PCB IR voltage drops. The internal phase locked loop allows synchronization and interleaving of paralleled phases, minimizing voltage and current ripple in the system.

## PROTECTION

The LTM4637 features overvoltage protection, foldback current protection and overtemperature protection. In the event of an overvoltage condition where the output voltage exceeds 10% of the programmed value, the top MOSFET is forced off and bottom MOSFET is turned on until

the overvoltage condition is cleared. In the event of an output short-circuit, if the output voltage falls 50% below its nominal output level, the maximum output current is progressively lowered to one-third of the original value, limiting the current dumped into the short. The overtemperature protection shuts down the internal controller at  $\sim 130^{\circ}\text{C}$ – $137^{\circ}\text{C}$ , restarting the controller again when the temperature cools. These protection features help ensure a low risk design. Furthermore, the LTM4637's dependability can be assured by rigorous electrical and reliability testing performed on each module.

Figure 1. Complete 1.8V, 20A regulator requires only a few components



The LTM4637 is a complete POL power supply solution for applications that require top-notch performance in a small footprint with minimal design effort and user-adjustable feature versatility. Its low power loss and thermal resistance allow for up to 20A of load current per phase, while its current mode structure allows the load current to be scaled up via parallel configurations.

Figure 2. Efficiency of the 1.8V regulator in Figure 1

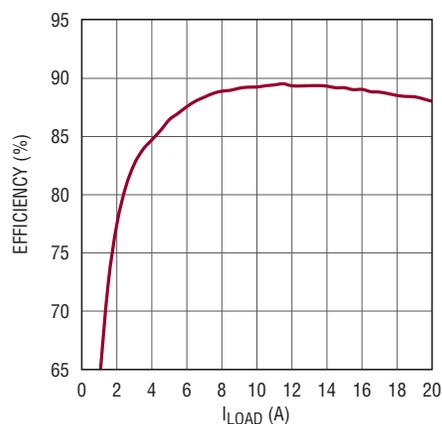
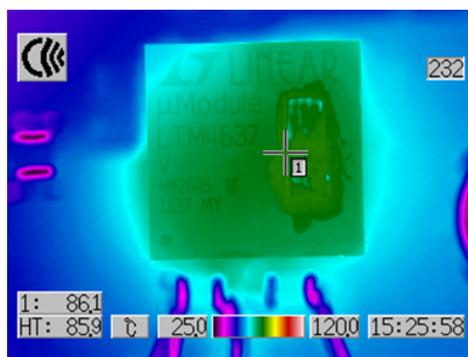


Figure 3. Thermal scan of the LTM4637 under full load for the solution in Figure 1.



$T_A = 25^\circ\text{C}$ , NO AIR, NO HEAT SINK  
 $I_{LOAD} = 20\text{A}$ ,  $V_{IN} = 12\text{V}$ ,  $V_{OUT} = 1.8\text{V}$

### MINIMAL DESIGN EFFORT

Little effort is required to create a high performance design because a complete power supply solution is integrated into the LTM4637. Just a few external components are required to conclude an application's design, simplifying the schematic and PCB layout.

For example, set the output voltage and switching frequency each with a single resistor from their respective programming pins to ground, then select input and output capacitors to satisfy voltage and current ripple requirements and the base design is complete. The LTM4637 minimizes total design effort, an important factor when time to market is important or limited power design resources are available. Figures 1 and 2 show a typical 12V input to 1.8V, 20A output application circuit and its efficiency, exemplifying the simplicity and high performance benefits of designing with the LTM4637.

### EXCEPTIONAL THERMAL PERFORMANCE

Exceptional thermal performance allows for increased current capability, which translates to a reduced total solution cost and increased reliability in high current density applications. The LTM4637 can provide up to 20A while remaining cool due to low power dissipation and thermal resistances. Figure 3 shows the  $\mu$ Module regulator's outstanding thermal performance for the setup in Figure 1. Even while providing a hefty 20A of load current, with no forced air cooling, the LTM4637 easily provides 36W of high current output power all from a footprint of less than 0.35in<sup>2</sup> without risking thermal overload. Attach a heat sink to the exposed metal on the top of the package and add forced airflow conditions for further reduction of internal temperatures.

### CONCLUSION

The LTM4637 is a complete POL power supply solution for applications that require top-notch performance in a small footprint with minimal design effort and user-adjustable feature versatility. Its low power loss and thermal resistance allow for up to 20A of load current per phase, while its current mode structure allows the load current to be scaled up via parallel configurations. Its built-in protection features and rigorous factory electrical testing ensure a reliable, low risk solution. ■

# Increase Current Capability and Simplify Thermal Design of Flyback Converters with Secondary-Side Synchronous Rectifier Driver in a 5-Pin SOT-23

Wei Gu

In a flyback converter, current levels in high current applications are constrained by heat produced in the output rectifier diode. The clear way to lift this restriction is to replace the diode with a much lower voltage drop MOSFET, significantly reducing heat produced in the rectifier—reduced heat dissipation increases output current capability and efficiency, and simplifies thermal design. The LT8309 is a secondary-side synchronous MOSFET driver that replicates the behavior of the output diode by sensing the drain-to-source voltage of the MOSFET to determine if it should be on or off, allowing it to replace a less efficient rectifier diode.

The LT8309 can be combined with any of Linear's line of no-opto boundary-mode flyback ICs (such as the LT3748 primary side controller) to produce high performance isolated power supplies with a minimal number of components.

## 5V, 8A ISOLATED SUPPLY

Figure 1 shows a low voltage, high current, low parts count flyback supply. The traditional output diode is replaced by an ideal diode consisting of the LT8309, a MOSFET and a few small external components.

For a MOSFET to act as a diode it must turn on as soon as the body diode starts

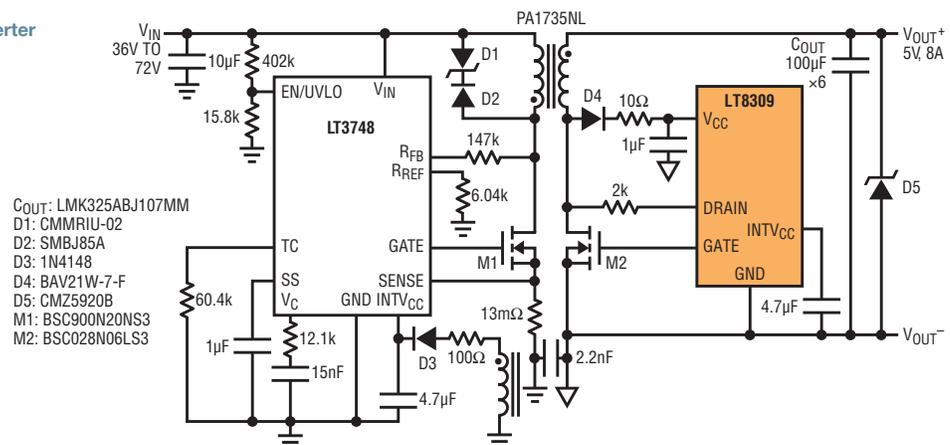
conducting current, and turn off as soon as its current decreases to zero. The LT8309's fast comparator produces the required near instantaneous action. The current sensing comparator monitors the drain voltage of the MOSFET. When the body diode begins to conduct, the drain voltage goes well below ground, the comparator trips and turns on the MOSFET. After a minimum on-time, the LT8309 waits for the MOSFET turn-off trip point to be reached to turn off the MOSFET. The turn-off trip point can be adjusted through an external resistor connecting the part's DRAIN pin to the drain of the MOSFET. The DRAIN pin has

a 150V voltage rating, making it suitable for wide input voltage design as well.

LT8309's internal LDO generates a 7V output at the INTV<sub>CC</sub> pin for the MOSFET gate drive. The strong gate drive with 1Ω pull-down resistance speeds up turn-on and turn-off of the MOSFET, resulting in better efficiency.

Efficiency is shown in Figure 2 with a comparison to a diode-only design. Thanks to higher efficiency, the operating temperature of a board built around an LT8309-based design remains significantly lower than a diode-only design, as shown in Figures 3 and 4.

Figure 1. A low voltage high current flyback converter



The LT8309 can be combined with any of Linear's line of no-opto boundary-mode flyback ICs, such as the LT3748 primary side controller, to produce high performance isolated power supplies with minimal components.

The 40V  $V_{CC}$  pin rating allows the LT8309 to be driven from the output voltage or the rectified drain voltage of the MOSFET. If The  $V_{CC}$  pin is connected to the output of the flyback converter, during an output short-circuit condition, the LT8309 is off and the body diode of the MOSFET must handle the short-circuit condition. This puts additional thermal requirements on the MOSFET. Instead, if  $V_{CC}$  is connected to the drain voltage of the MOSFET as shown in Figure 1,  $V_{CC}$  is equal to  $V_{IN}/N$  in short-circuit, allowing the LT8309 to operate during a short. The short-circuit current flows through the MOSFET instead of the body diode.

## CONCLUSION

LT8309 is an easy-to-use, fast, secondary-side synchronous flyback rectifier driver in SOT-23 package. High efficiency, high current isolated power supplies require a minimal number of components, and thermal

design is simplified when the LT8309 is combined with one of Linear's line of no-opto boundary-mode flyback ICs. ■

Figure 2. Efficiency of LT8309-based flyback converter compared same converter with a traditional secondary-side diode rectifier

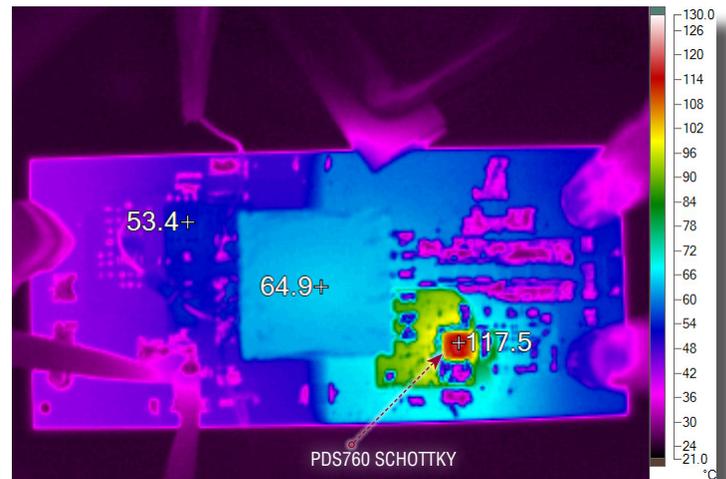
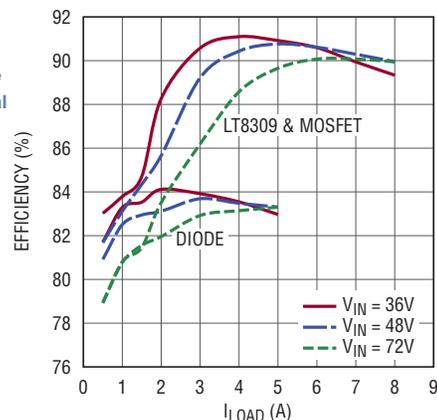


Figure 3. Thermal image at 5V/5A output with diode PDS760

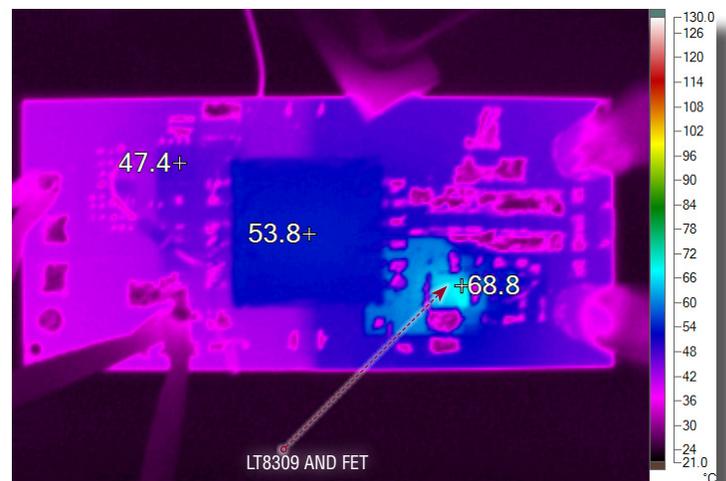


Figure 4. Thermal image shows 5V/5A output runs much cooler with the LT8309

# High Speed ADC Sampling Transients

Doug Stuetzle

High speed analog to digital converters (ADCs) are, at the analog signal interface, track and hold devices. As such, they include sampling capacitors and sampling switches. The action of these elements as they alternately track the input signal and hold its voltage produces small transient voltages and currents. These transients can introduce distortion into the circuitry that drives the ADC analog input.

High resolution ADCs are particularly susceptible to these effects. Consider a typical 16-bit ADC with a 2V peak to peak input range. The analog input voltage change required to affect the digital output by one bit is on the order of  $30\mu\text{V}$ .

These high performance converters are almost always differential input devices, but the analysis here starts with a single-ended circuit model to show the transient effects. The degree to which these effects appear in a true differential setting depend heavily on the common mode rejection ratio of the signal source driving the ADC and the ADC itself.

## INPUT TRACK AND HOLD MODEL

The input circuit to implement the track and hold function is shown in Figure 1. This is a simplified model, but it serves to illustrate the charge transfer effects that can distort the input waveform.

There are three CMOS switches, a sample capacitor and an op amp. The circuit is driven by a sample clock, which can operate at 100MHz or more. It alternates between two phases: the track phase, in which the input analog voltage is placed onto a sample capacitor, and the hold phase, which disconnects the sample capacitor from the analog input and makes the capacitor voltage available to the digital circuitry that quantizes it. One edge of the sample

clock—say the rising edge—initiates the track phase, and the next edge—say the falling edge—initiates the hold phase.

Note that the transition from the track phase to the hold phase is comprised of two transition states. The transition states exist because the three switches change state in sequence—not all at once. The transition from track phase to hold phase, along with the transition states, is shown in Figure 2. Note the directions of the transient currents on the sample and switch capacitances. In some cases the current changes direction as the switch control voltage ramps up/down.

During the transition from track to hold phase, transient charges can be imposed on the sample capacitor, altering the final voltage on this capacitor at the beginning of the hold phase. The reason for these errant charges is the inherent channel capacitance of the CMOS switch. Each of the three CMOS switches includes a parasitic capacitance from gate to channel. When the control voltage on the switch changes, a small amount of charge is transferred to the channel.

In the track phase, the conditions are:

- Switches SW1 and 2 are closed; the control voltage on these switches is 1.8VDC. The channel capacitances of the switches are charged to 1.8V.
- Switch SW3 is open; the control voltage on this switch is zero volts DC. The channel capacitance of this switch is charged to zero volts.

Figure 1. Track and hold model for single-ended signal

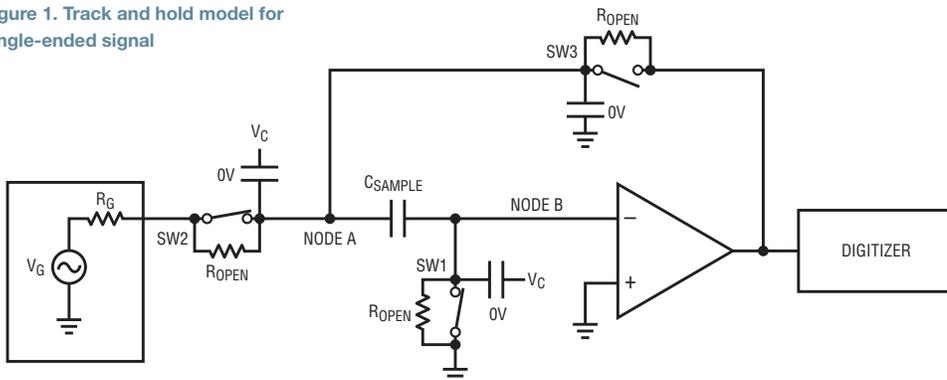
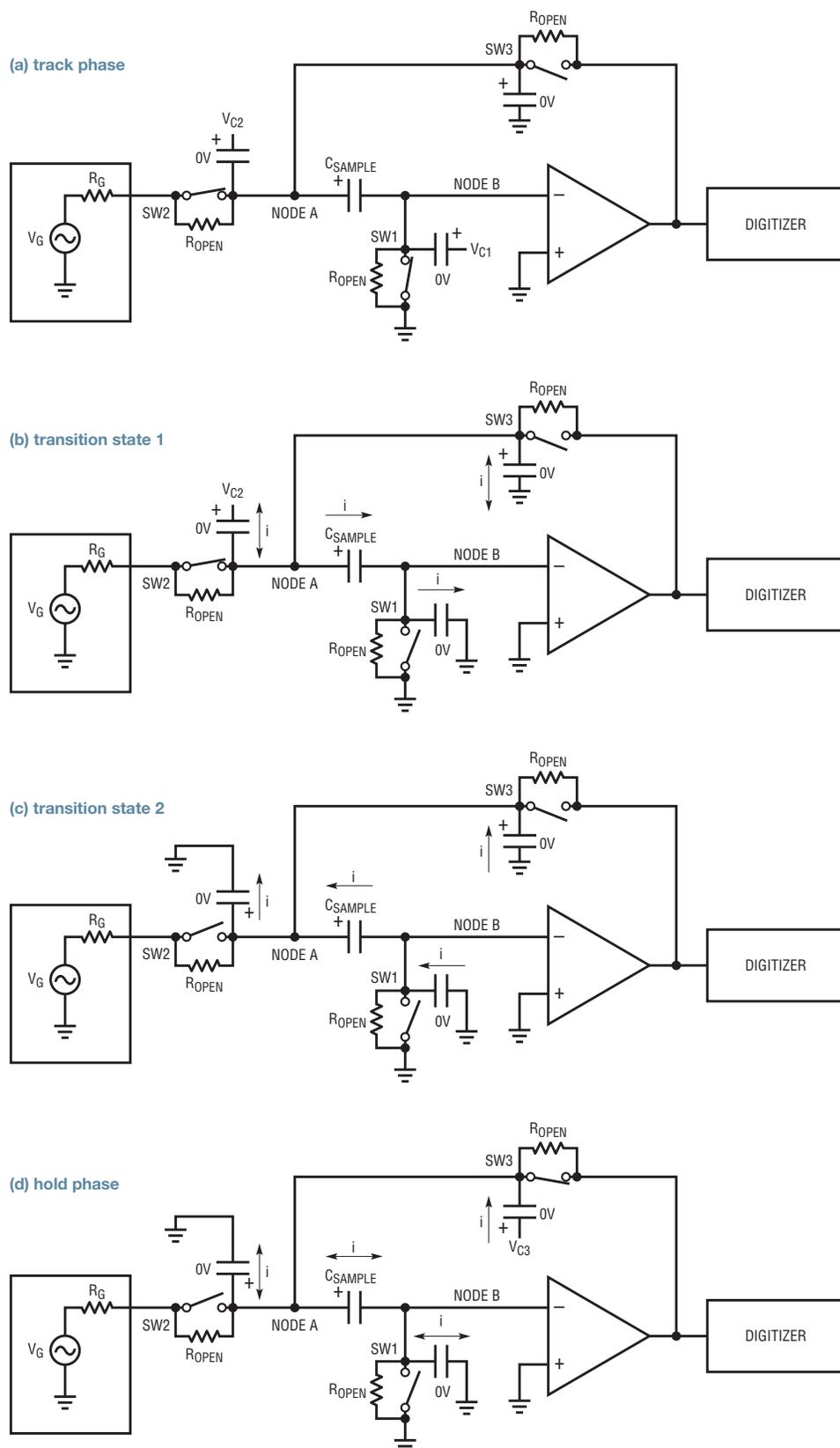


Figure 2. The transition from track phase to hold phase via transition states



Here is the sequence of events that make up the transition from track to hold phase:

1. The control voltage  $V_{C1}$  for  $sw1$  ramps to zero volts DC.

- The voltage at Node B is pulled lower by the charge on  $sw1$ .
- Some of the charge on the channel of  $sw1$  flows to the sample capacitor; the effect is to increase the charge on the sample capacitor.

2. The control voltage  $V_{C2}$  for  $sw2$  ramps to zero volts DC.

- The voltage at Node A is pulled lower by the charge on  $sw2$ .
- Some of the charge on the channel of  $sw2$  flows to the sample capacitor; the effect is to decrease the charge on the sample capacitor.

3. The control voltage  $V_{C3}$  for  $sw3$  ramps to 1.8VDC.

- The voltage at Node A is pulled higher by the channel capacitance of  $sw3$ .
- The voltage across the sample capacitor increases.
- At the end of this sequence, the charge, and hence the voltage, on the sample capacitor has changed.

The input network that links the analog signal source to the ADC input contributes to the transient charge effects outlined here. In particular, elements that store charge, or delay the transit of charge, can produce additional artifacts as the switches change state. Generally the input network consists of a driver amplifier, a simple RC network—usually lowpass—and some length of transmission line. Some length of transmission line is unavoidable, particularly as ADC packages become smaller.

Figure 3. End of tracking phase, with switch parasitics

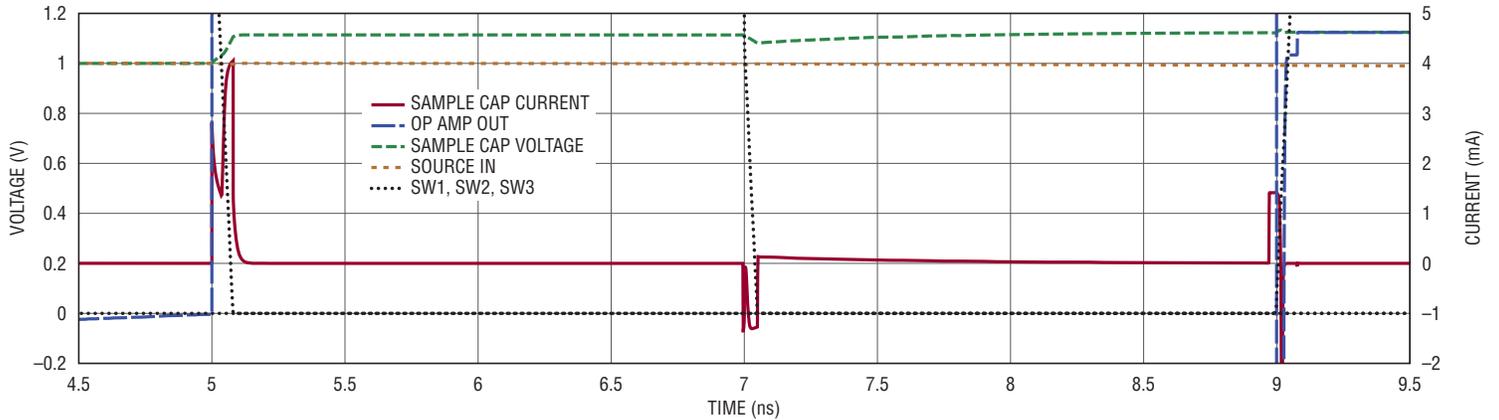


Figure 4. End of tracking phase, with switch parasitics, RC network and 200ps transmission line

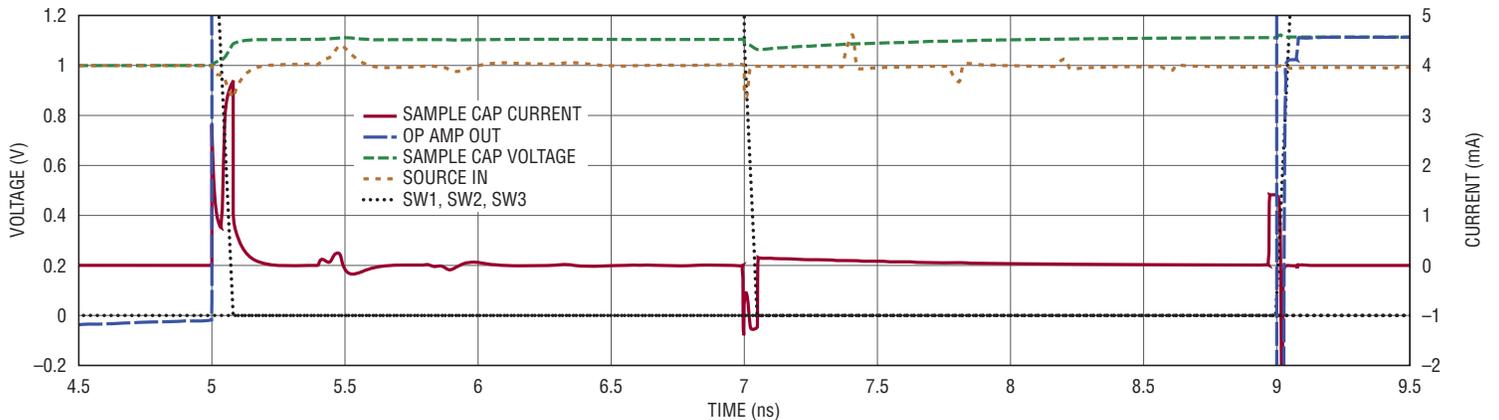
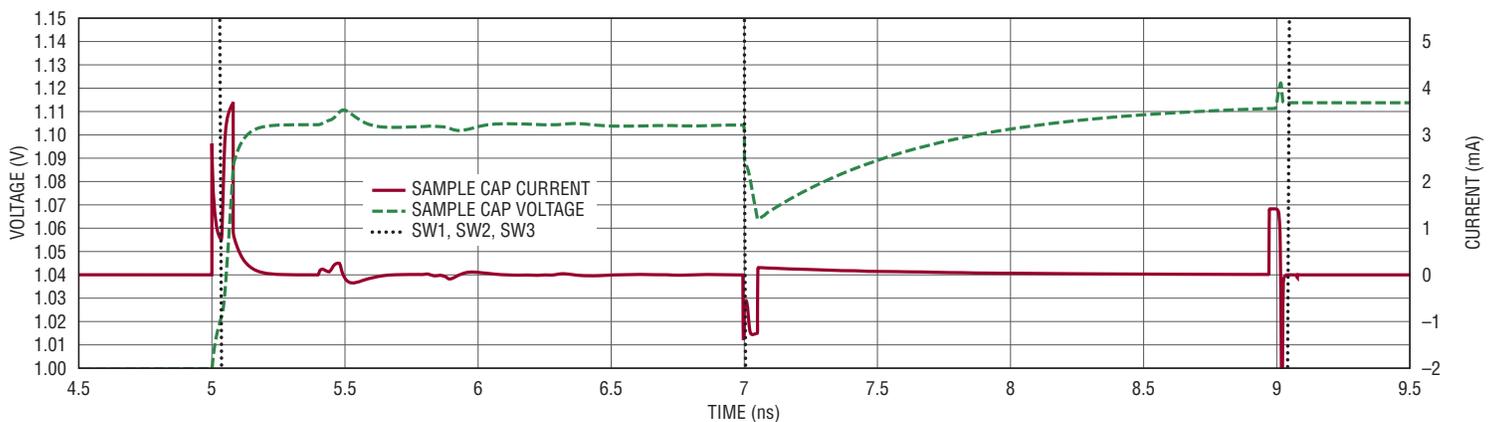


Figure 5. Detail of transients on the sample capacitor voltage



## TIME DOMAIN MODELING

The time domain behavior of a typical input track and hold circuit can be modeled in PSpice. Here are the assumed parameters:

- Input signal = 5MHz sinewave, 1V peak
- Sample capacitance = 2pF
- Switch channel capacitance = 0.2pF
- Switch control voltages = 0V/1.8V
- Switch ramp time = 100ps

The effects of the switching events are more evident if the three events are separated by 2ns each. Figure 3 shows each of the switching events, starting with switch SW<sub>1</sub> opening at 5ns. At 7ns, SW<sub>2</sub> opens; SW<sub>3</sub> closes at 9ns. Note that once the tracking phase is complete, the op amp output voltage matches that of the sample capacitor. This voltage, however, differs from the input voltage because of the charge injected by the switches.

## INPUT NETWORK EFFECTS

The input network that links the analog signal source to the ADC input contributes to the transient charge effects outlined above. In particular, elements that store charge, or delay the transit of charge, can produce additional artifacts as the switches change state. Generally the input network consists of a driver amplifier, a simple RC network—usually low-pass—and some length of transmission line. Some length of transmission line is unavoidable, particularly as ADC packages become smaller. Simply conveying the input signal, which is always differential, from the discrete network to the ADC input pins involves distance. So a differential transmission line structure is used.

Examination of these artifacts in Figure 4 shows the transients imposed upon the input signal. In this case the input network consists of a pi section RC lowpass and a 200ps transmission line. The input signal shows the initial transient once SW<sub>1</sub> opens, and echoes of this transient appear

at 400ps intervals. Less evident but still present are the transients and their echoes on the sample capacitor. Note that if the switch events are separated enough in time for the transients to settle out before entering the hold phase, their effect on the final sample capacitor voltage decays to zero. Figure 5 shows these transients on the sample capacitor voltage in greater detail.

In practice the three switch events are separated not by 2ns, but by 100ps or so. In this short window, transients have little time to decay before the last switch is closed and the voltage on the sample capacitor is captured. This gains importance with regard to the echoes from the transmission line structure. Figure 6 shows this behavior.

The transients on the sample capacitor are shown in detail in Figure 7. It has been

shown that charge injection from the sample switches affects the final charge stored on the sample capacitor. This is a classic issue with track and hold circuits. We have also seen that settling transients can further affect this final charge, and hence the voltage passed on to the digital quantization portion of the ADC for that sample.

## MULTIPLE SAMPLE EFFECTS

Thus far we have examined the effect of these artifacts upon a single sample. The next step is to expand the field of view to see how the sample capacitor voltages differ from the input voltages at several sample instants. Figure 8 illustrates this. The input signal is a 1MHz sine wave, and the sample rate is 20MSPS. The end of each track phase is denoted by a blue dashed vertical line. The error voltage on the sample capacitor is defined as the

Figure 6. Fast switch behavior

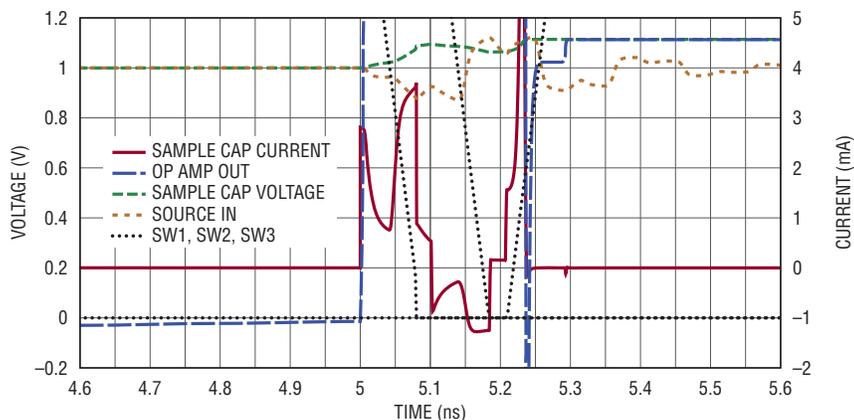
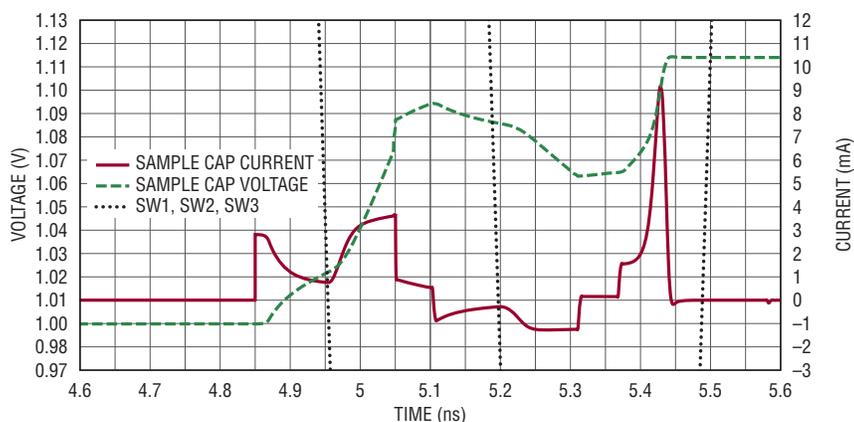


Figure 7. Detail of transients on the sample capacitor with fast switch events



The track and hold circuit at the input of an ADC necessarily injects transient charges into the circuit that drives it. These charges can affect the accuracy of the voltage stored on the sample capacitor at each sample instant. The key interval is the time it takes to transition from the track phase to the hold phase. During this time switches are toggled and the voltage stored on the sample capacitor can be perturbed.

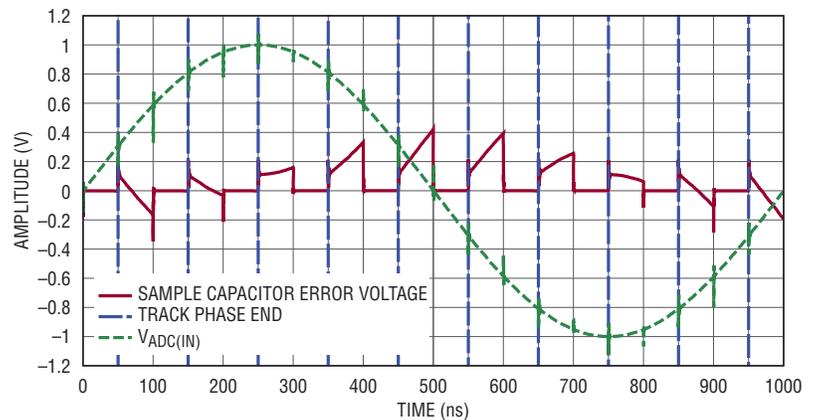
difference between the sample capacitor voltage and the input signal voltage.

The real error of interest consists of the difference between the sample capacitor voltage at the end of the hold phase and the input voltage at the end of the track phase. A set of these error samples has been plotted for one signal cycle and plotted in Figure 9. Note that the deviation from true input signal values is a time-varying quantity; it does not represent a DC offset that can be removed or ignored. It appears, most likely, as multiple harmonic and spurious signals in the digitized ADC output.

### COMMON MODE VERSUS DIFFERENTIAL MODE

Having explored all of this, there is an important factor that greatly reduces the effect of these transients on the output signal. This factor is the common mode rejection of the ADC. The ADC input is differential, and accepts a differential input signal. The switch effects, however, are not differential; they are the same on both sides of the input circuit. Since all of the switch charge transfers are common mode, they are rejected by the ADC. So the net effect of these distortions on the digitally represented output signal depends not only upon the magnitude of the errant charges, but also on the common mode rejection of the ADC.

Figure 8. Sample capacitor values vs source voltages



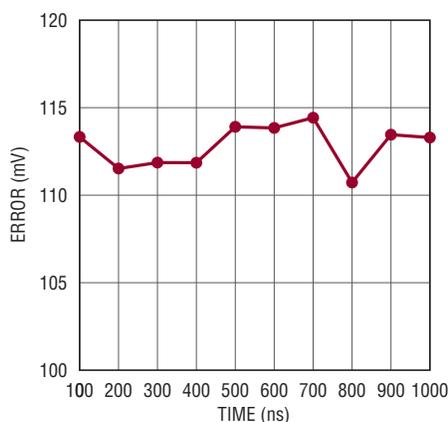
### CONCLUSION

The track and hold circuit at the input of an ADC necessarily injects transient charges into the circuit that drives it. These charges can affect the accuracy of the voltage stored on the sample capacitor at each sample instant. The key interval is the time it takes to transition from the track phase to the hold phase. During this time switches are toggled and the voltage stored on the sample capacitor can be perturbed.

Fortunately the ADC input is typically differential and the undesired transients are common mode. Nevertheless, any input circuit elements that store large amounts of charge, or cause charge to bounce back and forth will exacerbate the charge disturbances. The latter effect is usually the result of transmission line structures whose round trip delay falls within the transition time between the track and hold phases. In particular, if there are circuit elements at the input end of the line that are reflective relative to the transmission line, these contribute to the magnitude of the reflections.

All of the preceding modeling was done using LTspice and a simple input circuit model. There is no practical way to estimate the magnitude or spectral distribution of the distortions outlined, but the analysis suggests that minimizing the storage and reflection of transient charges will reduce them. ■

Figure 9. Error voltage at sample instants



# New Product Briefs

## 60V FAULT-PROTECTED, 50mA STEP-DOWN CHARGE PUMP PRODUCES REGULATED OUTPUT WITH INPUT CURRENT DOUBLING

The LTC3255 is a versatile high voltage step-down switched capacitor converter that delivers up to 50mA of output current. In applications where the input voltage exceeds twice the output voltage, a charge pump provides nearly twice the efficiency of an equivalent linear regulator, while offering a space-saving inductor-free alternative to switching DC/DC regulators.

The LTC3255 produces a regulated output (2.4V to 12.5V adjustable) from a wide 4V to 48V input that is 60V/-52V tolerant. At no load, Burst Mode® operation reduces  $V_{IN}$  quiescent current to 16 $\mu$ A, and 2:1 capacitive charge pumping extends output current capability to approximately twice the input current. The LTC3255 is ideal for industrial control, factory automation, sensors and supervisory control and data acquisition (SCADA) systems, housekeeping power supplies, and current-boosting voltage regulators for 4mA to 20mA current loops.

The LTC3255 operates either as a general purpose step-down charge pump with 2:1 or 1:1 conversion ratios, or as a current doubling shunt regulator. In normal mode, the conversion ratio is chosen based on  $V_{IN}$ ,  $V_{OUT}$  and load conditions, and switching between conversion modes is automatic. In shunt mode, the device is forced into 2:1 mode, enabling the LTC3255 to provide a regulated output voltage from a current source input with nearly two times the input current available at the load. For

example, this feature enables a 4mA current loop to power a 7.4mA load continuously with a 3.3V regulated output voltage.

The LTC3255 withstands reverse-polarity input supplies down to -52V and output short circuit without damage. Safety features, including an output current limit and overtemperature protection further enhance robustness.

The LTC3255 is available in low profile (0.75mm) 3mm  $\times$  3mm 10-lead DFN and 10-lead MSOP packages, both with backside thermal pads.

## 600mA NEGATIVE LINEAR REGULATOR FEATURES RAIL-TO-RAIL OPERATION, PROGRAMMABLE CURRENT LIMIT AND OUTPUT CURRENT MONITOR

The LT3090 600mA low dropout negative linear regulator features low noise, rail-to-rail operation, precision programmable current limit and a bidirectional output current monitor. The device is cable-drop compensation capable, easily parallelable for higher current or PCB heat spreading and configurable as a floating 3-terminal regulator. The LT3090 is ideal for negative logic supplies, low noise instrumentation and RF supplies, rugged industrial supplies and for post-regulating switching supplies.

The LT3090's input voltage range is -1.5V to -36V. A single resistor programs the adjustable rail-to-rail output voltage from 0V to -32V and dropout voltage is only 300mV (typical at full load). The device features a trimmed,  $\pm 1\%$  accurate 50 $\mu$ A current source reference and provides  $\pm 2\%$  output voltage tolerance over

line, load and temperature. Output voltage regulation, bandwidth, transient response and output noise (18 $\mu$ V<sub>RMS</sub> over a 10Hz to 100kHz bandwidth) remain independent of output voltage due to the device's unity-gain voltage follower architecture.

The LT3090 exhibits excellent stability with a wide range of output capacitors including small, low cost ceramic capacitors. It is stable with a minimum 4.7 $\mu$ F output capacitor.

A single resistor adjusts the precision programmable current limit. The device's bipolar current monitor either sources or sinks a current proportional to output current, useful for system monitoring.

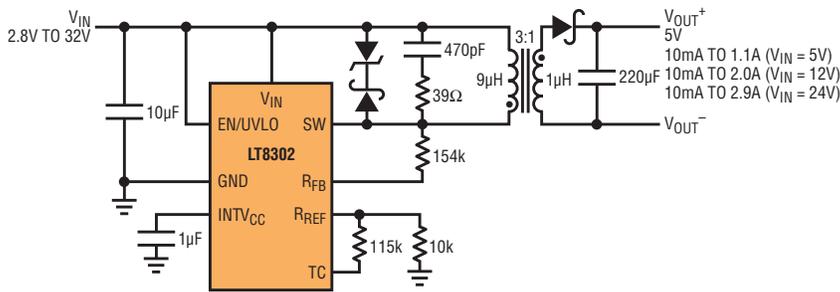
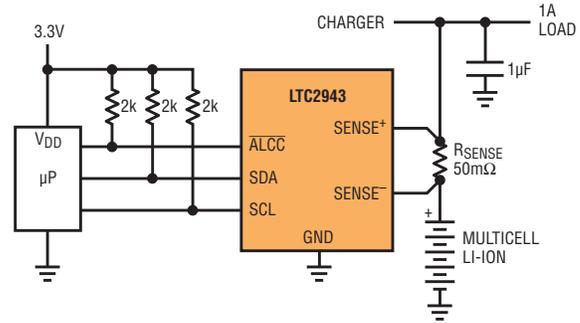
The LT3090's bidirectional shutdown capability allows the device to operate with either positive or negative logic levels. In addition, the LT3090's accurate shutdown thresholds enable a programmable  $UV_{LO}$  threshold for either the regulator's input supply or a positive system supply voltage.

The LT3090's internal protection circuitry includes a precision current limit with foldback and thermal shutdown with hysteresis. In bipolar supply applications where the regulator's load returns to a positive supply, the OUT pin can be pulled above GND, up to 40V, and the LT3090 will safely start up.

The LT3090 is available in low profile (0.75mm) 3mm  $\times$  3mm 10-lead DFN and 12-lead MSOP packages, both with backside thermal pads. ■

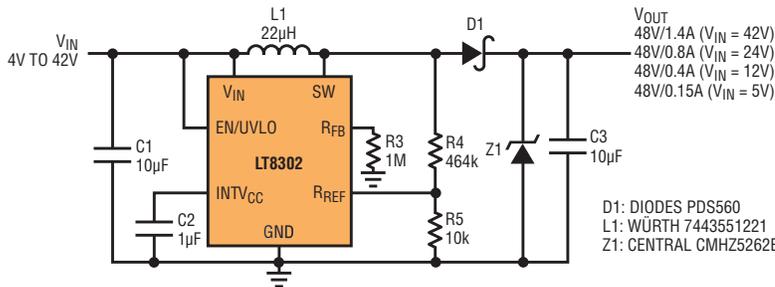
**MULTICELL BATTERY GAS GAUGE WITH TEMPERATURE, VOLTAGE AND CURRENT MEASUREMENT**

The LTC2943 measures battery charge state, battery voltage, battery current and its own temperature in portable product applications. The wide input voltage range allows use with multicell batteries up to 20V. A precision coulomb counter integrates current through a sense resistor between the battery's positive terminal and the load or charger. Voltage, current and temperature are measured with an internal 14-bit No Latency  $\Delta\Sigma$ ™ ADC. The measurements are stored in internal registers accessible via the onboard I<sup>2</sup>C/SMBus Interface. [www.linear.com/solutions/4467](http://www.linear.com/solutions/4467)



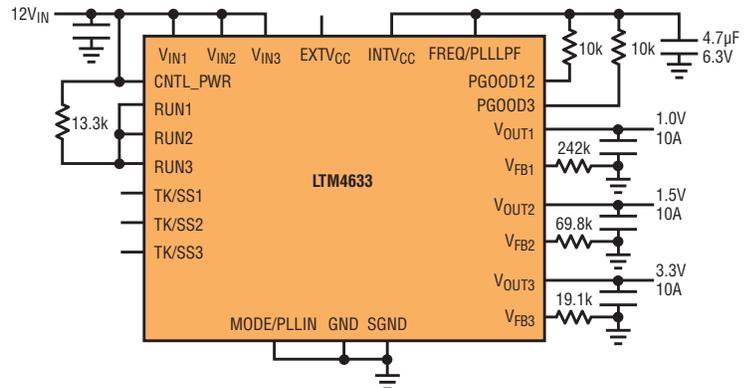
**2.8V–32V INPUT TO 5V AT 2.9A OUTPUT ISOLATED FLYBACK CONVERTER AND 4V–42V INPUT TO 48V AT 1.4A OUTPUT BOOST CONVERTER**

The LT8302 is a monolithic micropower isolated flyback converter. By sampling the isolated output voltage directly from the primary-side flyback waveform, the part requires no third winding or opto-isolator for regulation. The output voltage is programmed with two external resistors and a third optional temperature compensation resistor. Boundary mode operation provides a small magnetic solution with excellent load regulation. Low ripple Burst Mode operation maintains high efficiency at light load while minimizing the output voltage ripple. A 3.6A, 65V DMOS power switch is integrated along with all the high voltage circuitry and control logic into a thermally enhanced 8-lead SO package. [www.linear.com/solutions/4465](http://www.linear.com/solutions/4465) [www.linear.com/solutions/4466](http://www.linear.com/solutions/4466)



**TRIPLE OUTPUT µModule BUCK REGULATOR: 12V INPUT TO 1.0V, 1.5V AND 3.3V AT 10A EACH**

The LTM4633 µModule regulator combines three complete 10A switching mode DC/DC converters into one small package. Included in the package are the switching controllers, power FETs, inductors, and most support components. The LTM4633's three regulators operate from 4.7V to 16V input rail(s) or 2.375V to 16V with an external 5V bias. The V<sub>OUT1</sub> and V<sub>OUT2</sub> output range is 0.8V to 1.8V, while the V<sub>OUT3</sub> output range is 0.8V to 5.5V. Each output is set by one external resistor. [www.linear.com/solutions/4468](http://www.linear.com/solutions/4468)



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