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## Four New Amplifiers Serve Many Applications

This issue of *Linear Technology* features a quartet of exciting new amplifiers from LTC: The LT1792/LT1793 low noise JFET op amps, the LT1813 high slew rate op amp, the LT1787 high-side current sense amplifier and the LT1399/LT1399HV triple current feedback amplifiers.

The LT1792/LT1793 single JFET op amps offer both very low voltage noise and low current noise, providing the lowest total noise over a wide range of transducer impedances. These op amps are unconditionally stable for gains of one or more, even with capacitive loads of 1000pF. Their low offset voltage and high DC gain allow the LT1792/LT1793 to fit into precision applications, especially those involving high impedance, capacitive transducers.

The LT1813 is a 100MHz, 750V/μs dual operational amplifier. Requiring only 3mA of supply current, it uses LTC's advanced, low voltage complementary bipolar process and a few design tricks to exceed the performance of its older siblings. A key figure of merit for amplifiers is the ratio of gain bandwidth to supply current (expressed as MHz/mA). The new process employed by the LT1813 forsakes high supply voltage operation for a 3×-4× increase in MHz/mA. Blazing speed from such a modest amount of supply current (3mA) is extremely attractive for low power applications. The LT1813 extends the frequency response of applications such as filters, instrumentation amplifiers and buffers.

The LT1787 employs precision technology to build a superior high-side current sense amplifier. The LT1787 will find uses in cellular phones, portable instruments and wireless telecom devices for precisely monitoring the current into or out of a battery. The LT1787 monitors bidirectional currents via the voltage across an external sense resistor. It features a minute input offset voltage of 40μV with a full-scale input of up to 500mV. This translates to a dynamic range of over 12 bits. The LT1787HV features a 60V maximum input, which allows it to be used in telecom and industrial applications that require the sensing of higher voltages. The device is self-powered from the supply that it monitors and requires only 60μA supply current.

For video and computer display applications, LTC introduces the LT1399/LT1399HV triple current feedback amplifiers. These devices contain three independent 300MHz CFAs, each with a shutdown pin. Each CFA has 0.1dB gain flatness of 150MHz and a slew rate of 800V/μs. The LT1399 operates on supplies from 4V to ±6V. The LT1399HV operates on supplies ranging from 4V to ±7.5V. Each amplifier can be enabled in 30ns and disabled in 40ns, making them ideal in spread-spectrum and portable equipment applications. With the addition of a small series resistor, the parts can drive large capacitive loads. This feature, combined with the LT1399HV's high voltage operation, makes it ideal for driving LCD displays. □



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# Issue Highlights

In addition to the four new amplifiers described on page one, this issue of *Linear Technology* features a variety of power and data converter products.

In the power area, we introduce the LTC1702/LTC1703. The LTC1702 is the first in a new family of low voltage, high speed switching regulator controllers. It is designed to operate from the standard 5V logic rail and generate two lower voltage, high current regulated outputs. Running at a fixed 550kHz switching frequency, each side of the LTC1702 features a voltage feedback architecture using a 25MHz gain-bandwidth op amp as the feedback amplifier, allowing loop-crossover frequencies in excess of 50kHz. Large onboard MOSFET drivers allow the LTC1702 to drive high current external MOSFETs efficiently at 550kHz and beyond, allowing the use of small external inductors and capacitors while maintaining excellent output ripple and transient response. The LTC1703 is a modified LTC1702 with a 5-bit DAC controlling the output voltage at side 1. The DAC conforms to the Intel Mobile VID specification.


Two other new switchers debuted in this issue are the micropower LT1615 and LT1617. These devices can be used in a number of topologies, including boost, SEPIC and positive-to-negative. With an input voltage range of 1.2V to 15V, they are ideal for a wide range of applications and work with a variety of input sources. An internal 36V switch allows them to easily generate output voltages of up to  $\pm 34V$  without the use of costly transformers. The LT1615 is designed to regulate positive output voltages, whereas the LT1617 is designed to directly regulate negative output voltages without the need for level-shifting circuitry. Both parts use a current-limited, fixed

off-time control scheme, which helps achieve high efficiency operation over a wide range of load currents. Both devices use tiny, low profile inductors and capacitors to minimize the overall system footprint and cost.

A new  $\Delta\Sigma$  ADC is also featured in this issue: the LTC2408 is the offspring of the LTC2400 24-bit  $\Delta\Sigma$  converter, introduced in our November 1998 issue. The LTC2408 combines the high performance LTC2400 with an 8-channel analog input multiplexer. The resulting device offers many unique features. The single-cycle settling characteristics lead to simplified multiplexer hook-up and channel selection, without the added overhead seen with other converters. The exceptional noise performance of the device eliminates the need for a programmable gain amplifier (PGA). The unprecedented 10ppm absolute accuracy of the device allows measurement of microvolt signals superimposed upon large DC voltages. A wide range of sensor inputs and voltage levels can be applied simultaneously to the LTC2408.

Our final Design Feature spotlights some unusual applications for the LT1684 ring tone generator, introduced in the June issue. The LT1684 was specifically designed for OEM telephone equipment. Because of the versatility of the device, the LT1684 ring tone chip finds itself at home in motor drives, digital input amplified speakers, alarm systems and sine wave UPS systems.


Our Design Ideas section features an isolated 35 watt DC/DC converter designed to replace "half-brick" power modules at half the cost and a comparator circuit that provides automatic shutdown for the LT1795 high speed ADSL power amplifier.

We conclude with a trio of New Device Cameos. 

## LTC in the News...

On July 20, Linear Technology Corporation announced its financial results for fiscal year 1999. Robert H. Swanson, Chairman & CEO, stated, "We had record levels of sales and profits as we finished the year strongly. As we commence our first fiscal year in the new millennium, we are encouraged by the increasing opportunity for high performance analog circuits." The Company reported net sales for the fourth quarter of \$140,524,000 (a 6% increase over net sales for the fourth quarter of the previous year). Net income for the fourth quarter was \$54,179,000 compared with \$49,503,000 a year ago. For the year, the Company's net sales were \$506,669,000 (up 5% over the previous year), with net income of \$194,293,000 (up 7% versus the prior year).

The Company was also featured in "America's Favorite Stocks," an article in the July issue of *Money* magazine that detailed which stocks are most popular among investment clubs in the United States. Linear Technology ranked 94th in terms of number of clubs holding the stock, and 68th in terms of the value of Linear Technology stock held by clubs.

Linear Technology was included in *Business Week's* annual Global 1,000 rankings. The Company ranked 289th among U.S. companies and 570th among companies internationally. 

the  
**LINEAR**  
ZONE

<http://www.linear-tech.com/ezone/zone.html>  
Articles, Design Ideas, Tips from the Lab...

For more information on parts featured in this issue, see  
<http://www.linear-tech.com/go/ltmag>

# JFET Op Amps Equal Low Noise Bipolars and Have Picoamp Current Noise

by Alexander Strong

The LT1792 and LT1793 are single JFET op amps that offer both very low voltage noise ( $4\text{nV}/\sqrt{\text{Hz}}$  for the LT1792 and  $6\text{nV}/\sqrt{\text{Hz}}$  for the LT1793) and low current noise ( $10\text{fA}/\sqrt{\text{Hz}}$  for the LT1792 and  $0.8\text{fA}/\sqrt{\text{Hz}}$  for the LT1793), providing the lowest total noise over a wide range of transducer impedance. Traditionally, op amp users have been faced with a choice: which op amp will have the lowest noise for the transducer at hand. For high transducer impedance, the LT1792/LT1793 JFET op amps will win over the lowest voltage noise bipolar op amps due to lower current noise. The current noise ( $2qI_B$ ) of an amplifier is a function of the input bias current ( $I_B$ ). For lower transducer impedance, bipolar op amps usually win over typical JFET op amps due to lower voltage noise for the same tail current of the differential input pair. The LT1792/LT1793 op amps are designed to have voltage noise that approaches that of bipolar op amps. All of these op amps are unconditionally stable for gains of one or more, even with capacitive

loads of  $1000\text{pF}$ . The low offset voltage of  $250\mu\text{V}$  and high DC gain of four million allow the LT1792/LT1793 to fit into precision applications. Voltage noise, slew rate and gain-bandwidth product are 100% tested. All of the specifications are maintained in the SO-8 package.

The combination of low voltage and current noise offered by the LT1792/LT1793 makes them useful in a wide range of applications, especially with high impedance, capacitive transducers such as hydrophones, precision accelerometers and photo diodes. The total noise in such systems is the gain times the square root of the sum of the op amp input-referred voltage noise squared, the thermal noise of the transducer ( $4kTR$ ) and the op amp's bias current noise times the transducer resistance squared ( $2qI_B \times R^2$ ). Figure 1 shows total input voltage noise versus source resistance. In a low source resistance application ( $<5\text{k}$ ), the op amp's voltage noise will dominate the total noise. In this region of low source resistance, the LT1792/LT1793 JFET op amps are way ahead of other JFET op amps; only very low noise bipolar op amps such as the LT1007 and LT1028 have the edge. As source resistance increases from  $5\text{k}$  to  $50\text{k}$ , the LT1792/LT1793 will match the best bipolar or JFET op amp for noise performance, since the thermal noise of the transducer ( $4kTR$ ) will dominate the total noise. A further increase in source resistance, to above  $50\text{k}$ , brings us to the region where the op amp's current noise ( $2qI_B \times R_{\text{SOURCE}}$ ) will dominate the total noise. At these high source resistances, the LT1792/LT1793 will outperform the lowest noise bipolar op amp due to the inherently low current noise of FET input op amps. In some conditions it may be neces-

Table 1. LT1792/LT1793 specifications

Parameter	LT1792	LT1793	Units
$V_{OS}$ (Max)	0.56	0.73	mV
$I_B$ (Max)	450	10	pA
$e_N$ (1kHz)	4.2	6	$\text{nV}/\sqrt{\text{Hz}}$
$i_N$ (1kHz)	10	0.8	$\text{fA}/\sqrt{\text{Hz}}$
GBWP ( $f_0 = 100\text{kHz}$ )	6	5	MHz
$I_S$	4.2	4.2	mA

sary to add a capacitor in parallel with a source resistor to cancel the pole that is caused by the source impedance and the input capacitance ( $14\text{pF}$  for the LT1792 and  $1.5\text{pF}$  for the LT1793). Observe what happens to noise with source resistances over  $100\text{k}$ ; the overall noise for the LT1792 and LT1793 actually decreases.

The high input impedance JFET front end makes the LT1792 and LT1793 suitable for applications where very high charge sensitivity is required. Figure 2 illustrates the LT1792 and LT1793 in inverting and noninverting modes of operation. A charge amplifier is shown in the inverting mode example; here the gain depends on the principle of charge conservation at the input of the amplifier. The charge across the transducer capacitance,  $C_S$ , is transferred to the feedback capacitor,  $C_F$ , resulting in a change in voltage,  $dV$ , equal to  $dQ/C_F$ , resulting in a gain of  $C_F/C_S$ . For unity gain, the  $C_F$  should equal the transducer capacitance plus the input capacitance of the amplifier and  $R_F$  should equal  $R_S$ . In the non-inverting mode example, the transducer current is converted to a change in voltage by the transducer capacitance; this voltage is then buff-

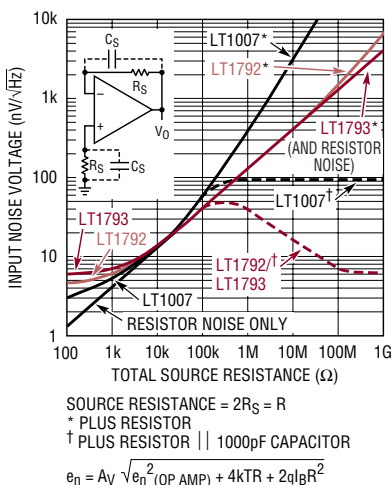
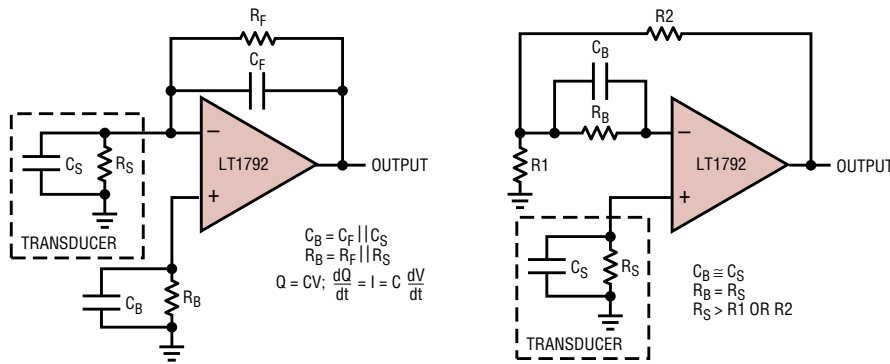


Figure 1. Comparison of LT1792/LT1793 and LT1007 input voltage noise vs source resistance



**Figure 2. LT1792/LT1793 inverting and noninverting gain configurations**

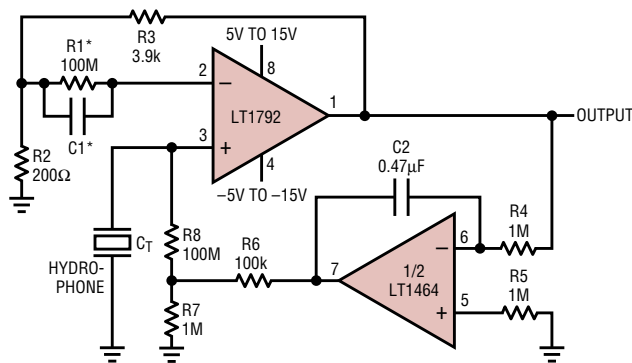
ered by the amplifier, with a gain of  $1 + R_2/R_1$ . A DC path is provided by  $R_S$ , which is either the transducer impedance or an external resistor. Since  $R_S$  is usually several orders of magnitude greater than the parallel combination of  $R_1$  and  $R_2$ ,  $R_B$  is added to balance the DC offset caused by the noninverting input bias current and  $R_S$ . The input bias currents, although small at room temperature, can create significant errors over increasing temperature, especially with transducer resistances of up to 1000M or more. The optimum value for  $R_B$  is determined by equating the thermal noise ( $4kTR_S$ ) to the current noise times  $R_S$ ,  $(2qI_B)R_S$ , resulting in  $R_B = 2V_T/I_B$  ( $V_T = kT/q = 26mV$  at  $25^\circ C$ ). A parallel capacitor,  $C_B$ , is used to cancel the phase shift caused by the op amp input capacitance and  $R_B$ .

The LT1792 has the lowest voltage noise ( $4nV/\sqrt{Hz}$ ) of the two, which makes it the best choice for transducer impedances of 5k or less. For transducer impedances over 100M, the LT1793, with a typical input bias current of only 3pA, will have lower output noise than the LT1792. The LT1793 has the additional advantage of very high input resistance ( $10^{13}$  ohms). Unlike most JFET op amps, the LT1792 and LT1793 have input bias currents that remains almost constant over the entire common mode range. The specifications for the LT1792 and LT1793 are summarized in Table 1.

The low noise of the LT1792 and LT1793 is achieved by maximizing the  $g_m$  of the input pair. The polygate JFETs have a higher  $g_m$ -to-area ratio than standard, single-gate JFETs. This is done by maximizing the tail current and the size of the input JFET geometries. Forty percent of the total supply current is used as the tail current for the LT1792 and LT1793. These op amps are best used with very high impedance transducers. The low noise hydrophone amplifier in Figure 3 is an application where the LT1792 excels. The AC current output of the hydrophone is converted to a voltage output by the 100M input resistor ( $R_8$ ). This signal is amplified

by the  $R_3/R_2$  ratio. DC leakage currents at the output of the hydrophone are subtracted by the servo action of the feedback amplifier. This amplifier need not have the low voltage noise of the LT1792; therefore, it can be chosen to minimize the overall system supply current. The LT1464 has less than an order of magnitude of supply current of the LT1792 and LT1793 and picoampere input bias current. This allows the time constant of this loop to be set using high value resistors and less expensive low value capacitors.

The LT1792 and LT1793 op amps are in a class by themselves when amplifying low level signals from high impedance sources. The design and process have been optimized to produce both low power consumption and low current and voltage noise. Most competing JFET op amps will have higher voltage noise or much higher supply current. Practically all bipolar op amps will have higher current noise. No other op amp will deliver the noise performance for a given supply current. For applications where low noise and power are issues, the LT1792 and LT1793 are the best choices.



DC OUTPUT  $\leq 2.5mV$  FOR  $T_A < 70^\circ C$   
 OUTPUT VOLTAGE NOISE =  $128nV/\sqrt{Hz}$  AT 1kHz (GAIN = 20)  
 $C_1 = C_T = 100pF$  TO  $5000pF$ ;  $R_4C_2 > R_8C_T$ ; \*OPTIONAL

**Figure 3. Low noise hydrophone amplifier with DC servo**

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# LT1813: 100MHz, 750V/ $\mu$ s Amplifier Draws Only 3mA

by George Feliz

## Introduction

The LT1813 is a 100MHz dual operational amplifier that has been optimized for supply voltages under 12V. It features an easy-to-use voltage feedback topology with high impedance inputs, yet it slews 750V/ $\mu$ s with only 3mA supply current. DC performance has not been neglected—the device has a 1.5mV maximum  $V_{OS}$  and a 400nA maximum  $I_{OS}$ .

## Performance

A summary of important specifications of the LT1813, compared to its higher voltage brethren, is shown in Table 1. A key figure of merit is the ratio of gain bandwidth to supply current (GBW/ $I_{SUPPLY}$ , expressed in units of MHz/mA). The new process employed by the LT1813 forsakes high supply voltage operation for a 3 $\times$ –4 $\times$  increase in MHz/mA compared to the LT1361 and LT1364. Blazing speed from such a modest amount of supply current is extremely attractive for low power applications. The LT1813 also propagates the family traits of matched, high input impedance inputs and low  $V_{OS}$ ,  $I_B$ ,  $I_{OS}$  and input noise. The improved common mode input range of the LT1813 adds to its utility in low supply voltage applica-

Table 1. Comparison of dual, high speed op amps ( $V_S = \pm 5V$ , 25°C)

	LT1813	LT1364	LT1361
Gain Bandwidth	100MHz	50MHz	37MHz
Supply Current per Amplifier	3.0mA	6.0mA	3.8mA
GBW/ $I_{SUPPLY}$	33.3MHz/mA	8.3MHz/mA	9.7MHz/mA
Slew Rate	750V/ $\mu$ s	450V/ $\mu$ s	350V/ $\mu$ s
Input Common Mode Range	$\pm 4.0V$	+3.4V, -3.2V	+3.4V, -3.2V
Output Swing	$\pm 4.0V$	$\pm 4.1V$	$\pm 4.0V$
Output Current ( $V_{OUT} = \pm 3V$ )	60mA	45mA	38mA
$V_{OS}$ (Max)	1.5mV	1.5mV	1.0mV
$I_B$ (Max)	4.0 $\mu$ A	2.0 $\mu$ A	1.0 $\mu$ A
$I_{OS}$ (Max)	400nA	350nA	250nA
$A_{VOL}$ (Min)	1.5V/mV	3.5V/mV	3V/mV
Input Noise Voltage	8nV/ $\sqrt{Hz}$	9nV/ $\sqrt{Hz}$	9nV/ $\sqrt{Hz}$
Input Noise Current	1pA/ $\sqrt{Hz}$	1pA/ $\sqrt{Hz}$	0.9pA/ $\sqrt{Hz}$
$C_{LOAD}$	1000pF	$\infty$	$\infty$
Max Supply Voltage ( $V^+$ to $V^-$ )	12.6V	36V	36V

tions. Stability with capacitive loading is another distinctive and desirable feature. Although the LT1813 is not stable with unlimited capacitive loads, it is stable with nearly two orders of magnitude more capacitance than competitors' high speed amplifiers. The small-signal transient response

in unity gain with  $C_{LOAD} = 100pF$ , 500pF and 1000pF is shown in Figure 1.

The LT1813 extends the frequency response of applications such as active filters, instrumentation amplifiers and buffers. Figure 2 shows the LT1813 converting a single-ended signal to a differential drive for the LTC1417 14-bit analog-to-digital converter (ADC). Note that the top amplifier provides unity voltage gain, but the amplifier is configured in a noise-gain of 2 to match the phase response of the bottom amplifier, which has a gain of -1. The filter in front of the ADC reduces broadband noise. The spurious free dynamic range (SFDR) of this circuit is -79dB for a 425kHz, 2V<sub>P-P</sub> input.

## Circuit Design

A simplified schematic of the circuit is shown in Figure 3. The circuit looks similar to a current feedback amplifier, but both inputs are high

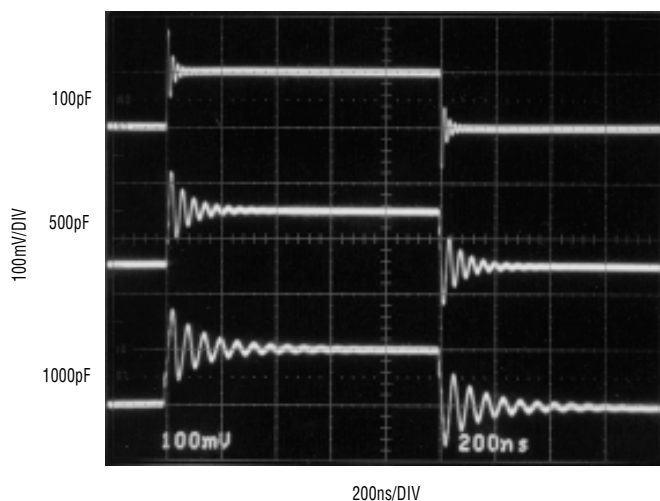
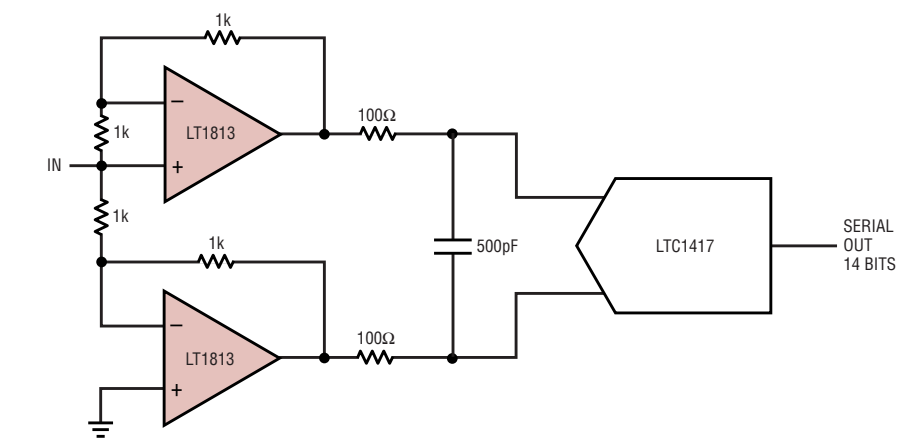


Figure 1. LT1813 in a gain-of-one configuration, no  $R_L$ ;  $C_L = 100pF$ , 500pF or 1000pF

impedance as in a traditional voltage feedback amplifier. A complementary cascade of emitter followers, Q1-Q4, buffers the noninverting input and drives one side of resistor R1. The other side of the resistor is driven by Q5-Q8, which form a buffer for the inverting input. The input voltage appears across the resistor, generating currents in Q3 and Q4 that are mirrored by Q9-Q11 and Q13-Q15 into the high impedance node. Transistors Q17-Q24 form the output stage. Bandwidth is set by R1, the  $g_m$ 's of Q3, Q4, Q7 and Q8 and the compensation capacitor,  $C_T$ .

The voltage drops of Q1-Q4 and the diodes Q10 and Q14 set the input common mode range of the amplifier. The emitters of Q3 and Q4 follow the noninverting input. As the input approaches either supply rail, the limiting voltage is determined by the saturation of Q3 or Q4, which occurs at approximately a  $V_{BE}$  plus a  $V_{SAT}$  from the supply rail. Typically, the input common mode range is 1V from either rail, and is guaranteed by the CMRR specification to be 1.5V from either rail. This excellent input range is achieved without compromising the output impedance of the mirrors Q9-Q11 and Q13-Q15, because Q25 and Q26 provide floating bias points for cascode devices Q9 and Q13. Lower bandwidth processes cannot successfully use this tech-



**Figure 2. Single-ended to differential ADC buffer: 2V<sub>P-P</sub> input at 425kHz yields -79dB SFDR**

nique and maintain high bandwidth, due to phase shift in the mirror.

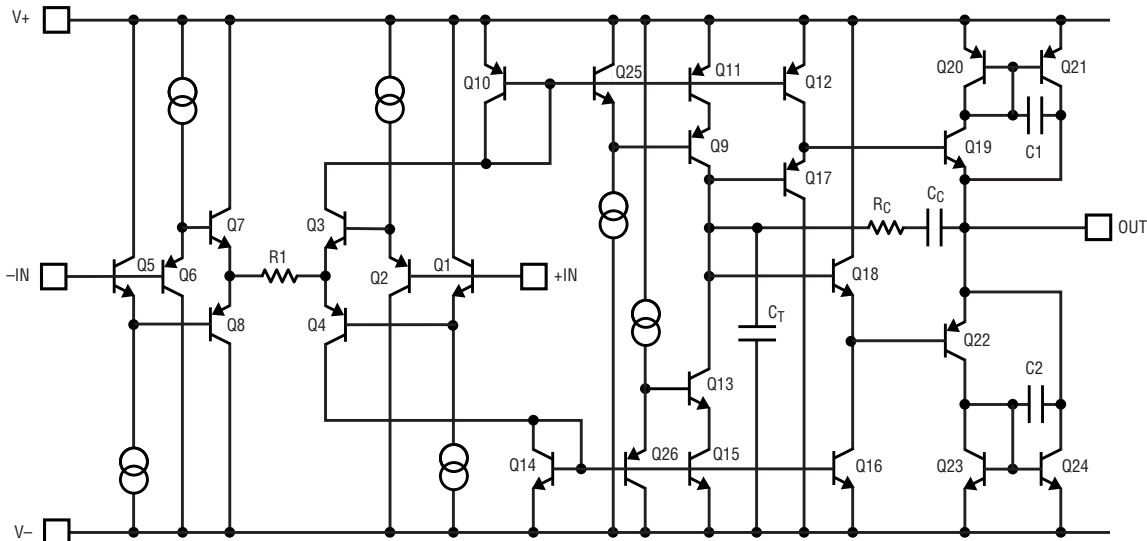
The current available to slew compensation capacitor  $C_T$  is proportional to the voltage that appears across R1. This method of "slew boost" achieves low distortion due to its inherent linearity with input step size. Large slew currents can be generated without increasing quiescent current. A low value for R1 reduces the input noise voltage to  $8nV/\sqrt{Hz}$  and helps reduce input offset voltage and drift. The LT1813 is built with small-geometry, multi-GHz transistors that produce abundant bandwidth with meager operating currents and allow for further reduction of idling supply current.

The output stage buffers the high impedance node from the load by

providing current gain. The simplest output stage would be two pairs of complementary emitter followers, which would provide a current gain of  $\beta_{NPN} \times \beta_{PNP}$ . Unfortunately, this gain is insufficient for driving even modest loads. Adding another emitter-follower or a Darlington configuration reduces output swing and creates instability with large capacitive loads.

The solution used on the LT1813 was to create a pair of composite transistors formed by transistors Q19-Q21 and Q22-Q24. The current mirrors attached to the collectors of emitter followers Q19 and Q22 provide additional current gain. The ratio of transistor geometries Q20 to Q21 and Q23 to Q24 increase the current gain by approximately fifteen. There

*continued on page 15*



**Figure 3. LT1813 simplified schematic**

# Micropower, Precision Current Sense Amplifier Operates from 2.5V to 60V

by Richard Markell, Glen Brisebois and Jim Mahoney

## Introduction

The LT1787 is a precision, high-side current sense amplifier designed for monitoring of the current either into or out of a battery or other element capable of sourcing or sinking current. The LT1787 features a miniscule 40 $\mu$ V (typical) input offset voltage with a 128mV full-scale input voltage. (The part is generally used at  $\pm$ 128mV full-scale, although it is specified for 500mV minimum full-scale.) This translates to a 12-bit dynamic range in resolving currents. A hefty 60V maximum input voltage specification allows the part to be used not only in low voltage battery applications but also in telecom and industrial applications where higher voltages may be present.

The device is self-powered from the supply that it is monitoring and requires only 60 $\mu$ A of supply current. The power supply rejection ratio of the LT1787 is in excess of 130dB.

The LT1787 allows the use of a user-selectable sense resistor, the value of which depends on the current to be monitored. This allows the input voltage to be optimized to the 128mV value, maximizing dynamic

range. The part has a fixed voltage gain of eight from input to output.

Additional LT1787 features include provisions for input noise filtering (both differential and common mode) and the ability to operate over a very wide supply range of 2.5V to 60V. The part is available in both 8-lead SO and MSOP packages.

## Operation of the LT1787

Figure 1 shows a function diagram of the LT1787 integrated circuit. When current is flowing from  $V_S^+$  to  $V_S^-$ , a sense voltage of  $V_{SENSE} = I_{SENSE} \cdot R_{SENSE}$  is generated. Because amplifier A1's positive and negative inputs are forced equal by feedback,  $V_{SENSE}$  also appears across the  $R_G$  side with the higher  $V_S$  potential. Hence, for the situation where  $V_S^+$  is greater than  $V_S^-$ ,  $V_{SENSE}$  appears across  $R_{G2A}$  and  $R_{G2B}$ . This current flows through Q2 and becomes the output current,  $I_{OUT}$ . Q1 is kept off by the amplifier and does not contribute to  $I_{OUT}$ .

For split-supply operation, where  $V_S^+$  and  $V_{EE}$  range from  $\pm$ 2.5V to  $\pm$ 30V; with  $V_{BIAS}$  at ground,  $V_{OUT}$  becomes  $(I_{OUT} \cdot R_{OUT})$ . In this case, and for the above described direction of sense current,  $V_{OUT}$  is positive or "above"  $V_{BIAS}$ ; thus Q2 is sourcing current.

When current flows from  $V_S^-$  to  $V_S^+$ , input  $V_S^-$  is at a higher potential than  $V_S^+$ , so  $V_{SENSE}$  will appear across  $R_{G1A}$  and  $R_{G1B}$ . The current,  $I_{OUT} = V_{SENSE} / (R_{G1A} + R_{G1B})$ , will be conducted through Q1, while Q2 remains off.  $I_{OUT}$  then duplicates itself through a one-to-one current mirror at  $V_{OUT}$ .  $V_{OUT}$  is negative or "below"  $V_{BIAS}$ ; Q1 sources current and the mirror sinks the current at the  $V_{OUT}$  node.

The output voltage across  $R_{OUT}$  is related to the input sense voltage by the following relationships:

$$V_{OUT} - V_{BIAS} = I_{OUT} \cdot R_{OUT}$$

$$V_{SENSE} = I_{SENSE} \cdot R_{SENSE}$$

$$I_{OUT} = V_{SENSE} / R_G, \\ R_G = R_{G1A} + R_{G1B} = R_{G2A} + R_{G2B}$$

$$R_{G(TYP)} = 2.5k$$

$$V_{OUT} - V_{BIAS} = (R_{OUT} / R_G) (V_{SENSE}) / 8 \\ R_{OUT(TYP)} = 20k$$

$$V_{OUT} - V_{BIAS} = 8 (V_{SENSE}), \\ V_{SENSE} = V_S^+ - V_S^-$$

$$V_{OUT} = 8 (V_{SENSE}) + V_{BIAS}$$

## Selection of $R_{SENSE}$

Maximum sense current can vary for each application. To sense the widest dynamic range, it is necessary to select an external sense resistor that fits each application. 12-bit dynamic range performance can be achieved regardless of the maximum current to be sensed, whether it is 10mA or 10A. The correct  $R_{SENSE}$  value is derived so that the product of the maximum sense current and the sense resistor value is equal to the desired maximum sense voltage (usually 128mV). For instance, the value of the sense resistor to sense a maximum current of 10mA is 128mV/10mA = 12.8 $\Omega$ . Since the LT1787 is capable of 12-bit resolution, the smallest measurable current is 10mA/4096 counts = 2.44 $\mu$ A/count. In terms of sense voltage, this translates to 128mV/4096 = 31.25 $\mu$ V/count. Other current ranges can be accommodated by a simple change in value in the sense resistor. Care should also be taken to ensure that the power dissipated in the sense resistor,  $I_{MAX}^2 \cdot R_{SENSE}$ , does not exceed the maximum power rating of the resistor.

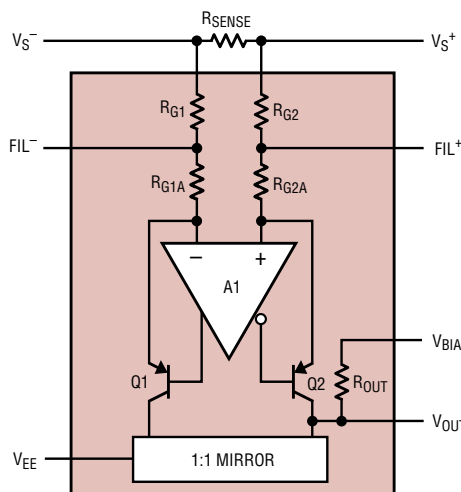


Figure 1. LT1787 function diagram

**Application Circuits**

**Dual-Supply, Bidirectional Current Sense**

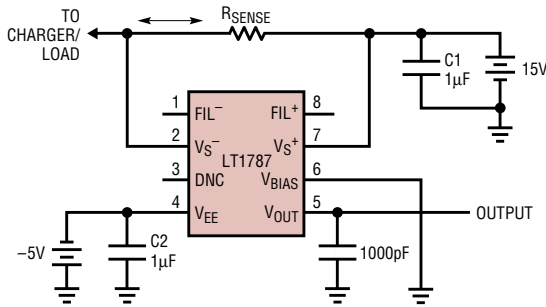
Figure 2 shows the schematic diagram of the LT1787 operated with dual supplies. This circuit can sense current in either direction, positive current flow being from  $V_S^+$  to  $V_S^-$ , where the output ranges from 0V to 1.024V for  $V_{SENSE} = 0mV$  to 128mV,

and negative current flow being from  $V_S^-$  to  $V_S^+$ , where the output varies from 0V to -1.024V for  $V_{SENSE} = 0mV$  to -128mV. Figure 3 shows the output voltage versus  $V_{SENSE}$  for this configuration.

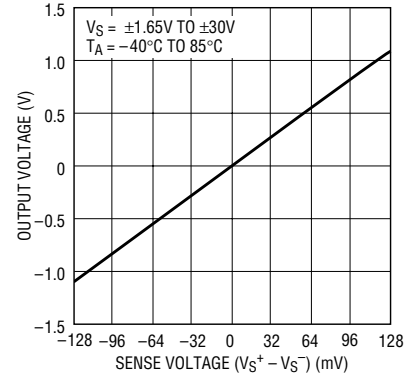
**Operation with Bias**

If a negative supply is not available, a voltage reference may be connected to the  $V_{BIAS}$  pin of the LT1787. The value of the reference is not critical, it

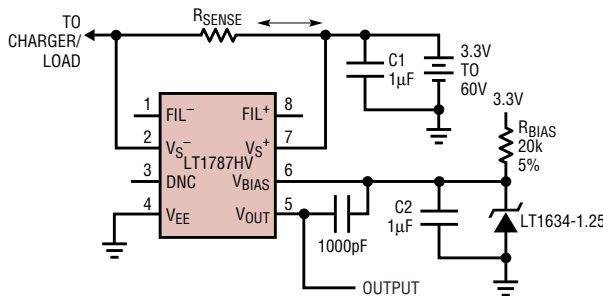
simply biases the output of the part to a new “zero” point (for  $V_{SENSE} = 0V$ ); zero is now at  $V_{BIAS} = V_{REF}$ , which, for the case of Figure 4, is equal to 1.25V. This configuration can be used for both unipolar and bipolar current sensing, with  $V_{OUT}$  ranging either “above” or “below”  $V_{BIAS}$ , depending on the direction of the current flow. This can be seen from the graph shown in Figure 5. (Note that the reference must be able to both source and sink



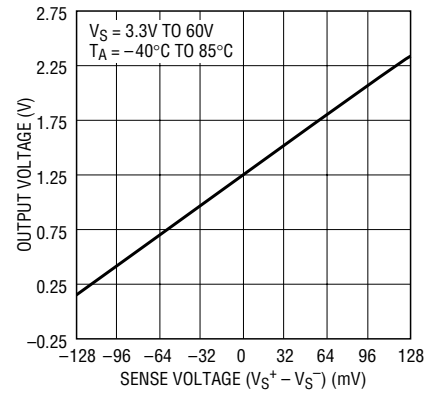
**Figure 2. Split-supply, bidirectional operation**



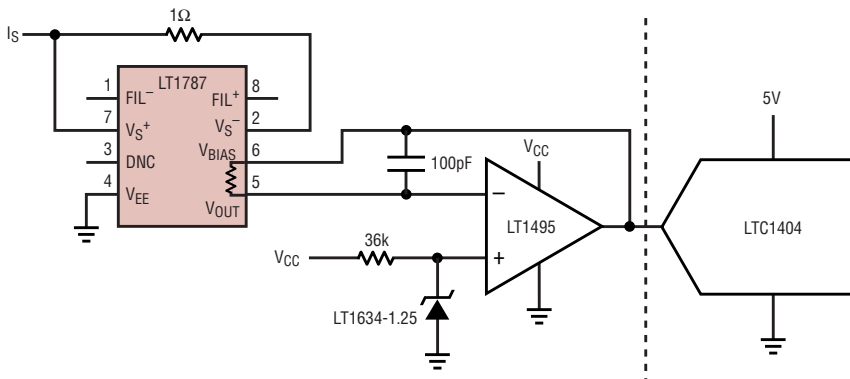
**Figure 3.  $V_{OUT}$  vs  $V_{SENSE}$  in bipolar mode**



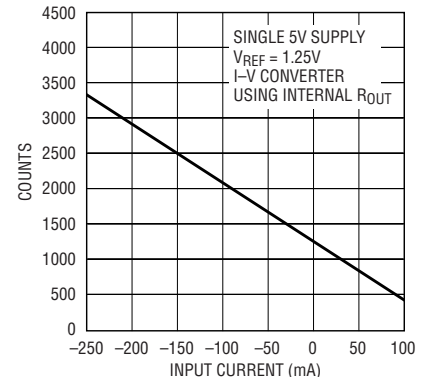
**Figure 4. Bidirectional operation with a reference**



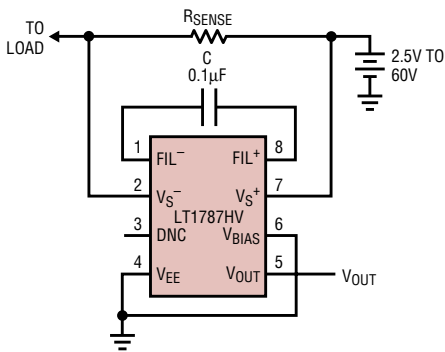
**Figure 5.  $V_{OUT}$  vs  $V_{SENSE}$  in bipolar mode**



**Figure 6. Operation with a buffer**



**Figure 7. Output counts vs input current for Figure 6's circuit**



**Figure 8. Output voltage referred to ground—unidirectional sensing mode**

current from the  $V_{BIAS}$  pin—refer to the block diagram in Figure 1.)

### Operation with a Buffer

Figure 6 uses a rail-to-rail op amp, the LT1495, as an I/V converter to buffer the LT1787's output. The LT1634-1.25 reference is used to bias the LT1787's output so that zero current is now represented by a 1.25V output. This allows the device to monitor current in either direction while the circuit operates on a single supply. This also allows lower voltage operation, since  $V_{OUT}$  of the LT1787 is held constant by the op amp. Figure 7 shows input current versus output counts (from the LTC1404 A/D converter), showing excellent linearity.

### Single-Supply Current Sense

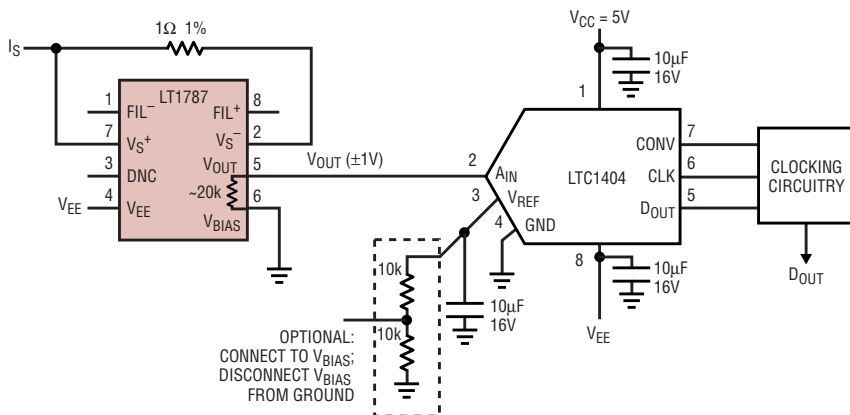
The circuit in Figure 8 provides good accuracy near full-scale, but has a limited dynamic range. In this circuit, the LT1787 is operated from a single supply of 2.5V minimum to 60V maximum. Current is allowed to flow through  $R_{SENSE}$  in both directions but

is measured in a single direction only, with current flow from  $V_{S^+}$  to  $V_{S^-}$ . In this connection,  $V_{BIAS}$  and  $V_{EE}$  are grounded. The output voltage ( $V_{OUT}$ ) of the LT1787 for this circuit is equal to  $8 \cdot V_{SENSE}$ , for  $V_{SENSE} = V_{S^+} - V_{S^-} = 0mV$  to 128mV. The dynamic range limits of this circuit can be seen in the graph shown in Figure 9.

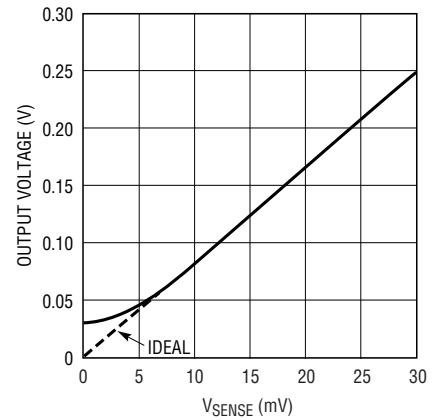
### Operation with an A/D Converter

Figure 10 shows the details of an LT1787 connected to a LTC1404 12-bit serial A/D converter. Details of the circuit are similar to those shown previously in Figures 2 and 8 and in the text detailing these circuits. The main difference in the applications is that the circuits in Figures 2 and 8 provide an analog output voltage proportional to the current, whereas the circuit shown in Figure 10 digitizes that analog voltage to provide a digital output. Figures 11 and 12 show the output of the LTC1404 A/D converter. The data in Figure 11 was collected with  $V_{EE}$  operated from -5V; in other words, both the LT1787 and the LTC1404 used a negative supply of -5V. Similarly, the data in Figure 12 was taken with  $V_{EE}$  connected to ground.

Connecting the optional section of the schematic (still operating the circuit from a single supply) allows the A/D's reference to "bias up" the LT1787 exactly as shown in Figure 5. Of course, the graph of the output would then be recast as similar to Figure 11 (counts versus  $V_{SENSE}$ ).



**Figure 10. Connection to an A/D converter (current-to-counts converter)**

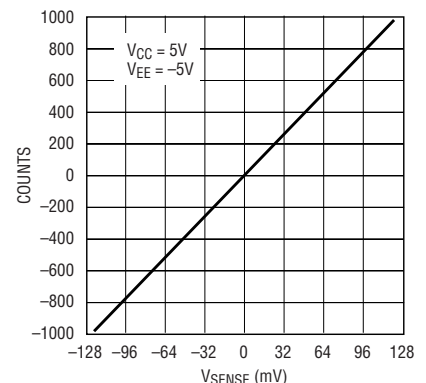


**Figure 9. Expanded scale of  $V_{OUT}$  vs  $V_{SENSE}$ , unidirectional current sensing mode**

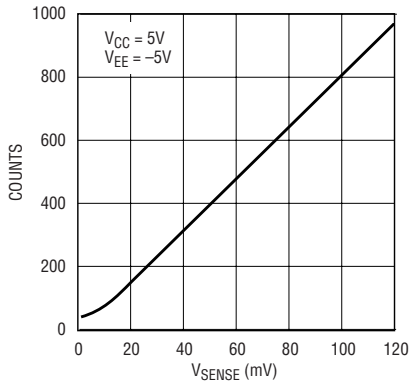
### Auto Shutdown Linear Regulator

Figure 13 shows the details of a linear regulator with high-side current sensing and latched shutdown capability. The circuit shuts down power to the load when the current reaches its overcurrent trip point. Power can then be restored only by cycling the main power off and on again. This circuit features the LT1787 and the LTC1440 precision comparator with on-chip reference. The LT1528 is a 3A low dropout linear regulator with shutdown.

The circuit uses the LT1787, U1, as a precision current sensor; the gain of the LT1787 allows the use of a 0.05Ω sense resistor, which dissipates a mere 0.312W of power. The LTC1440 ultralow power comparator, U2, with its internal reference, is used as a precision trigger, followed by a 74HC00 connected as an RS flip-flop (U3C and U3D) to latch the error condition until power is removed and reapplied. The other two NAND gates



**Figure 11. LT1787 input to LTC1404 ADC**



**Figure 12. LT1787 input to LTC1404 ADC—single supply**

are used to provide a power-on reset of the RS flip-flop. The 1M resistor, R7, and the 0.33μF capacitor, C3, at pin 2 of U3A provide a long enough time constant to properly initialize the flip-flop.

As shown, the circuit's trip point for shutdown is just under 2.5A. This may be changed by altering the value of current sense resistor R1. Consult the LT1787 data sheet for details on how to alter this resistor to sense different current ranges.

**Battery Fuel Gauge**

The industry standard method for gauging battery charge is to keep track of the endpoints, the full charge and discharged states, and, in between, to measure how much discharge has occurred since the last full charge (and vice versa). In an automobile, this would be analogous to having only a “full” reading and an “empty” reading on the gas tank, and, in between, to keep track of mileage. Applying this strategy to batteries is called “Coulomb-counting”; it is achieved by measuring (and numerically accumulating in a microcontroller) the current flow from the battery over time. Keeping track of the history of battery currents and voltages allows the present state of the battery charge to be determined. Hence the need for an accurate current sense amplifier like the LT1787.

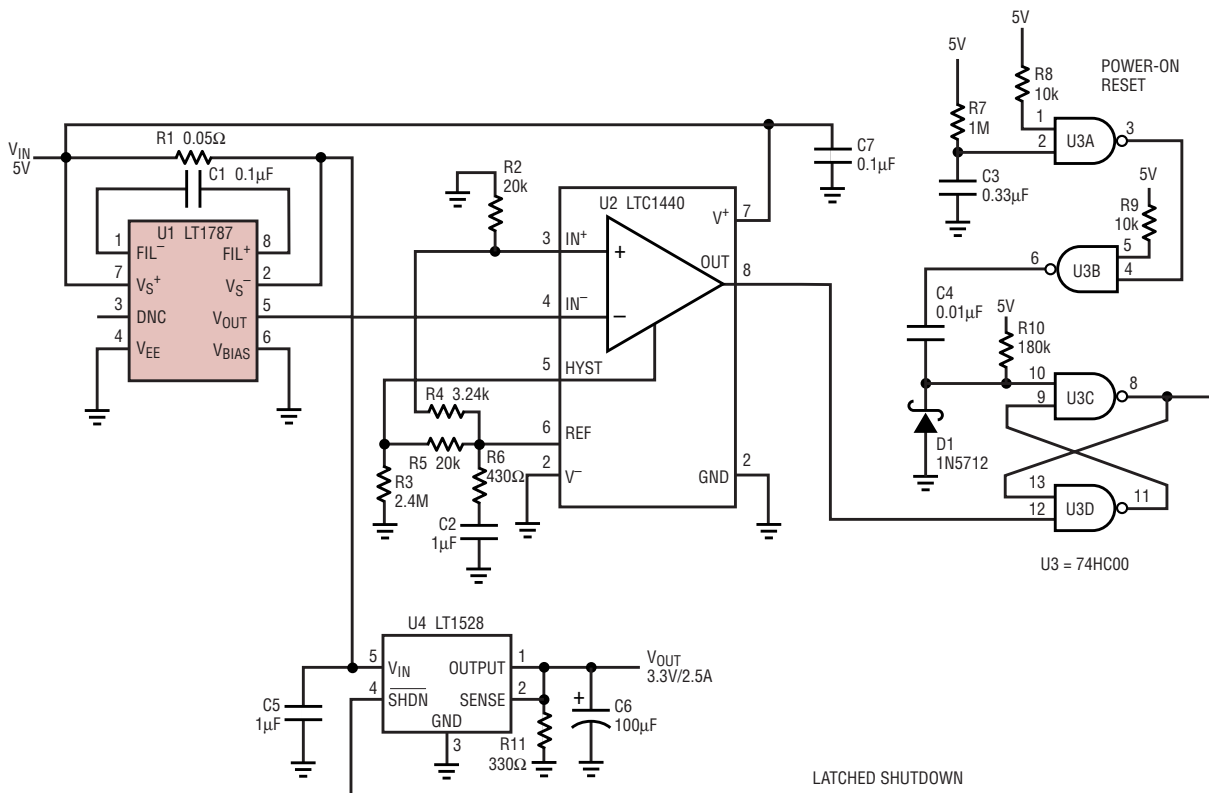
Figure 14 shows a schematic for measuring battery current using the LT1787 and measuring battery voltage using the micropower LT1635 with 200mV internal reference. The LT1787 is configured for single-supply, bidirectional operation, with the

2.0V reference coming from U2B, the LT1635 (created by amplifying its internal 200mV reference by a gain of ten.) Note that the reference voltage is fed to the ADC, so its absolute value is not critical, except in that it will form the center point for the battery voltage measurement and will thus determine the valid battery input voltage range. Resistors R1 and R2 form a divide-by-five, bringing the battery voltage down from ~10.8V to ~2.1V to put it within the input range of a downstream ADC. Op amp U2A with resistors R3 and R4 level shift this by the reference voltage and apply a gain of five. If a 12-bit ADC with a 5V reference is used, the following equations apply:

$$V_{BATT} = 4 (V_{REF}) + V_B = (5/4095) (4(Ch1) + Ch2)$$

$$I_{BATT} = V_S / R_S = (1/8) (V_I - V_{REF}) / 0.05 = 2.5 (V_I - V_{REF}) = 2.5 (5/4095) (Ch0 - Ch1)$$

where Ch0, Ch1 and Ch2 are in counts from 0 to 4095.




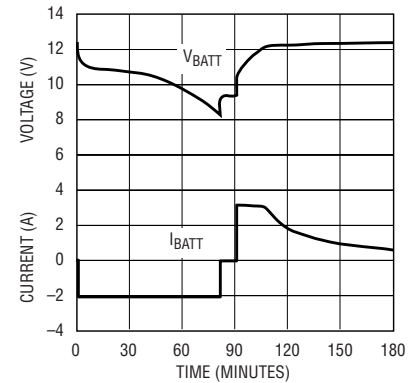
**Figure 13. LT1787 auto shutdown with latch**

Figure 15 shows a typical discharge and charge cycle for a 10.8V, 4A-hour Li-Ion battery.

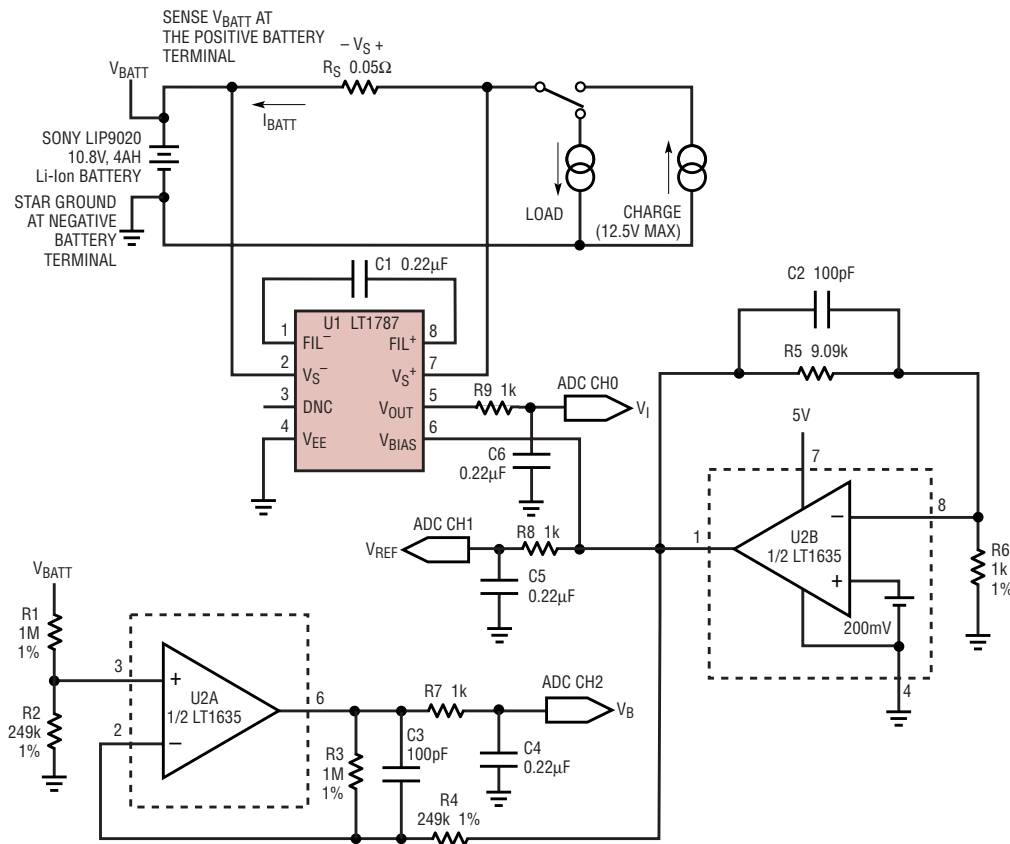
**Conclusion**

The LT1787 high-side current sense amplifier provides an easy-to-use method of sensing current with 12-bit resolution for a multiplicity of application areas. The part can oper-

ate to 60V, making it ideal for higher voltage topologies such as might be used in telecom or industrial applications. Additionally, the part can find homes in battery-powered, handheld equipment and computers, where the need for gauging the amount of current consumed and/or the amount of charge remaining in the battery is critical. 



**Figure 15. Discharge and charge cycle of a 10.8V Li-Ion battery**



**Figure 14. Battery "fuel gauging" system**

For more information on parts featured in this issue, see <http://www.linear-tech.com/go/ltmag>



<http://www.linear-tech.com/ezone/zone.html>  
Articles, Design Ideas, Tips from the Lab...

# Triple 300MHz Current Feedback Amplifiers Drive RGB/Component Video and LCD Displays

by Brian Hamilton

## Introduction

With the advent of HDTV and DVD video, there is a renewed focus on RGB/component video to maximize picture quality. LCD displays have also entered the mainstream for high end and portable computer displays. Both of these applications require high speed triple amplifiers for routing and conditioning video signals. LCD displays also require voltage swings of over 10V, with fast settling, into large capacitive loads. With these applications in mind, Linear Technology has introduced the LT1399 and LT1399HV triple current feedback amplifiers. The LT1399 and LT1399HV contain three independent 300MHz current feedback amplifiers, each with a shutdown pin. Each amplifier has exceptional 0.1dB gain flatness of 150MHz and a slew rate of 800V/ $\mu$ s. Output drive current is a minimum of 80mA over temperature.

The LT1399 operates on all supplies from a single 4V to  $\pm$ 6V. The LT1399HV supports higher supply voltages and will operate on supplies ranging from a single 4V to  $\pm$ 7.5V. Each of the three amplifiers draws 4.6mA when active. When disabled,

each amplifier draws zero supply current and its output becomes high impedance. Each amplifier can be enabled in 30ns and disabled in 40ns, making the LT1399 and LT1399HV ideal in spread-spectrum and portable equipment applications.

With the addition of a small series resistor at the output, the LT1399 and LT1399HV are capable of driving large capacitive loads. The LT1399HV's high voltage operation, when combined with its ability to drive capacitive loads, makes it ideal for driving LCD displays.

## What's Inside

Figure 1 is a simplified schematic for one of the three amplifiers found in the LT1399/LT1399HV. Tying EN low allows current to flow through J1, Q1-Q4 and R1. Q3 and Q6 mirror this current on top, while Q5 and Q7 mirror the current on the bottom. Q6 and Q7 thus act as current sources for input-stage transistors Q8-Q11. +IN is a high-impedance input, driving the bases of Q8 and Q9. The emitters of these transistors then drive the bases of Q10 and Q11, which

have their emitters tied together and form a buffered representation of +IN. This node is the inverting input -IN. Any current flowing into or out of -IN modulates the collector currents of Q12 and Q14. This, in turn, modulates the collector currents of Q13 and Q15, which drive the high impedance node. Transistors Q16-Q21 and resistors R2 and R3 form the output stage that buffers the signal at the high impedance node from the output.

## Using the LT1399HV to Drive LCD Displays

Driving present-day XGA and UXGA LCD displays can be a difficult problem because they require drive voltages of up to 12V, usually present a capacitive load of over 300pF and require fast settling. The LT1399HV is particularly well suited for driving these LCD displays because it is capable of swinging more than  $\pm$ 6V on  $\pm$ 7.5V supplies and, with a small series resistor  $R_S$  at the output, can drive large capacitive loads with good settling time. This circuit topology can be seen in Figure 2. As seen in Figure 3, at a gain of three, with a 16.9 $\Omega$  output series resistor and a 330pF load, the LT1399HV is capable of settling to 0.1% in 30ns for a 6V step. Similarly, as seen in Figure 4, a 12V output step will settle in 70ns.

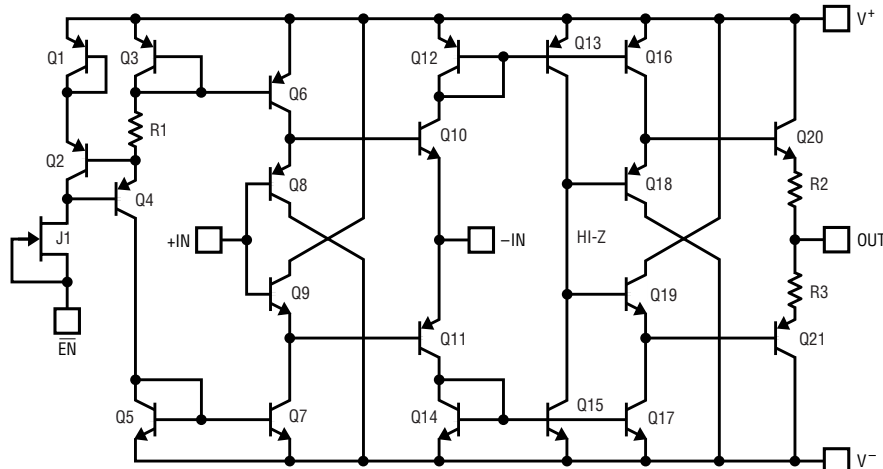


Figure 1. LT1399 simplified schematic (one amplifier)

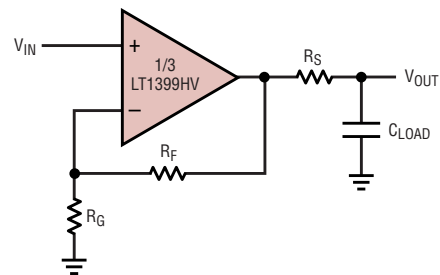
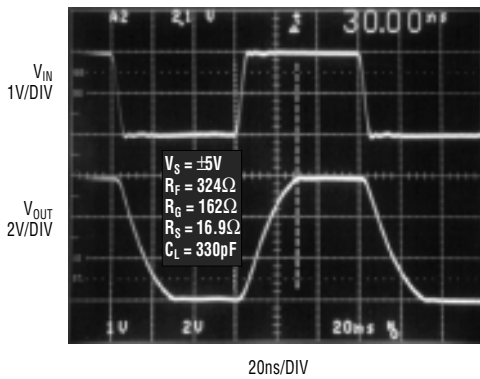
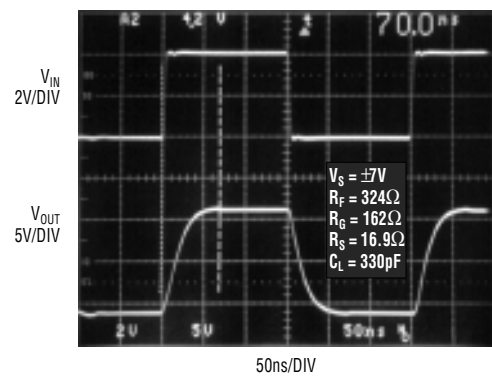


Figure 2. Adding an output series resistor for driving capacitive loads



**Figure 3. LT1399/LT1399HV large-signal pulse response driving 330pF (typical LCD loading)**



**Figure 4. LT1399HV output-voltage swing is >12V with ±7V supplies**

### A 3-Input Video Multiplexer and Cable Driver

Figure 5 shows a low cost, 3-input video MUX cable driver. The scope photo in Figure 6 displays the cable output of a 30MHz square wave driving 150Ω. In this circuit, the active amplifier is loaded by the sum of  $R_F$  and  $R_G$  of each disabled amplifier. Resistor values have been chosen to keep the total back termination at 75Ω while maintaining a gain of one at the 75Ω load. Figure 7 shows the envelope of the output signal as the multiplexer is switched from channel A to channel B. Channel A is being driven by a 2V<sub>P-P</sub>, 3.58MHz sine wave. The switching envelope of the output is well behaved and the switching time is approximately 32ns.

### Buffered RGB to Color-Difference Matrix

Two LT1399s can be used to create buffered color-difference signals from RGB inputs. In the application shown in Figure 8, a total of four amplifiers is used to create color-difference signals. The luminance signal Y is created using amplifiers A2 and A3. The remaining color-difference signals each use a single amplifier and the newly created Y output to perform the appropriate difference function.

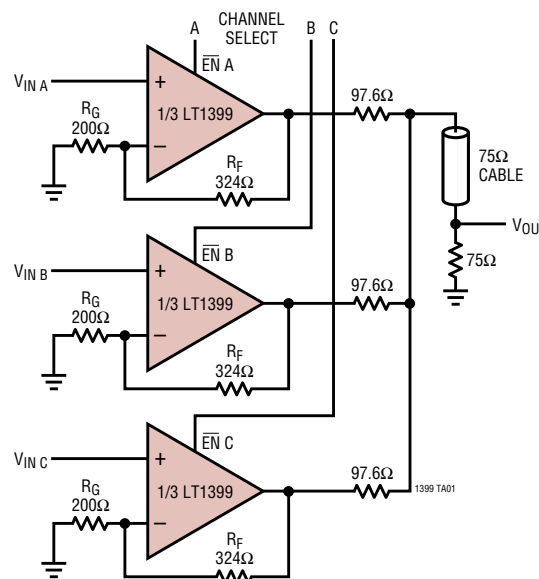
The R input arrives via 75Ω coax and is routed to 1082Ω resistor R8 and the noninverting input of LT1399 amplifier A1. There is also an 80.6Ω termination resistor, R11, which yields a 75Ω input impedance at the R input when considered in parallel with R8. R8 connects to the inverting

input of a second amplifier, A2, which also sums the weighted G and B inputs to create a  $-0.5 \cdot Y$  output. Amplifier A3 then takes the  $-0.5 \cdot Y$  output and amplifies it by a gain of minus two, resulting in the Y output. Amplifier A1 is configured in a noninverting gain-of-two configuration, with the bottom of the gain resistor R2 tied to the Y output. The output of amplifier A1 thus results in the color-difference output  $R - Y$ .

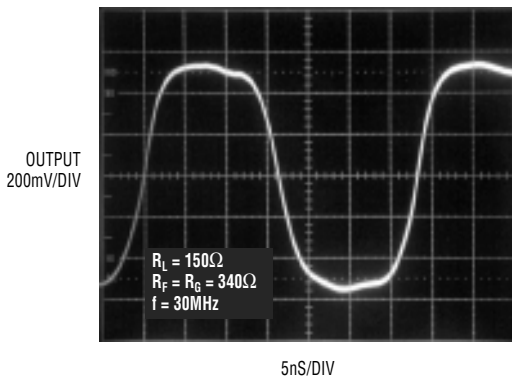
The B input is similar to the R input. It arrives via 75Ω coax and is routed to 2940Ω resistor R10 and the noninverting input of amplifier A4. There is also a 76.8Ω termination resistor R13, which yields a 75Ω input

impedance when considered in parallel with R10. R10 also connects to the inverting input of amplifier A2, adding the B contribution to the Y signal as discussed above. Amplifier A4 is configured in a noninverting gain-of-two configuration with the bottom of the gain resistor R4 tied to the Y output. The output of amplifier A4 thus results in the color-difference output  $B - Y$ .

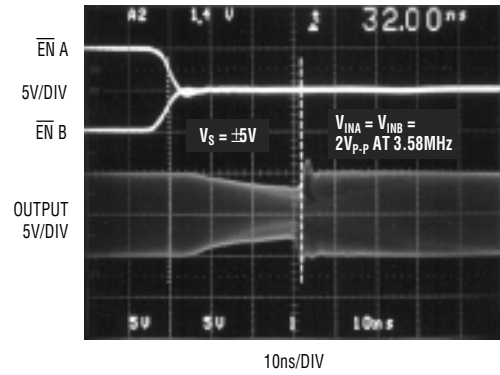
The G input also arrives via 75Ω coax and adds its contribution to the Y signal via a 549Ω resistor R9 that is tied to the inverting input of amplifier A2. There is also an 86.6Ω termination resistor, R12, which yields a 75Ω termination when considered in par-



**Figure 5. 3-input video mux/cable driver**



**Figure 6. 30MHz square wave response**



**Figure 7. 3-input video mux switching response**

allel with R9. Using superposition, it is straightforward to determine the output of amplifier A2. Although inverted, it sums the R, G and B signals in the standard proportions of 0.3R, 0.59G and 0.11B, which are used to create the Y signal. Amplifier A3 then inverts and amplifies the signal by two, resulting in the Y output. Two additional LT1399 amplifiers remain unused, available for additional signal conditioning as needed.

**Buffered Color-Difference to RGB Matrix**

The LT1399 can also be used to create buffered RGB outputs from color-difference signals. As seen in Figure 9, the R output is a back-terminated 75Ω signal created using resistor R5 and LT1399 amplifier A1 configured for a gain of two via 324Ω resistors R3 and R4. The noninverting input of amplifier A1 is connected via 1k resistors R1 and R2 to the Y and R - Y inputs, respectively, resulting in cancellation of the Y signal at the amplifier input. The remaining R signal is then amplified by A1.

The B output is also a back-terminated 75Ω signal created using resistor R16 and amplifier A3 configured for a gain of two via 324Ω resistors R14 and R15. The non-inverting input of amplifier A3 is connected via 1kΩ resistors R12 and R13 to the Y and B - Y inputs respectively, resulting in cancellation of the Y signal at the amplifier input. The remaining B signal is then amplified by A3.

The G output is the most complicated of the three. It is a weighted sum of the Y, R - Y and B - Y inputs. The Y input is attenuated via resistors R6 and R7 such that amplifier A2's noninverting input sees 0.83Y. Using superposition, we can calculate the positive gain of A2 by assuming that R8 and R9 are grounded. This results in a gain of 2.41 and a contribution at the output of A2 of 2Y. The R - Y input is amplified by A2 and resistors R8 and R10, giving a gain of -1.02. This results in a contribution at the output of A2 of 1.02Y - 1.02R. The B - Y input is amplified by A2 and resistors R9 and R10, giving a gain of -0.37. This results in a contribution at the output of A2 of 0.37Y - 0.37B.

If we now sum the three contributions at the output of A2, we get:

$$A2_{OUT} = 3.40Y - 1.02R - 0.37B$$

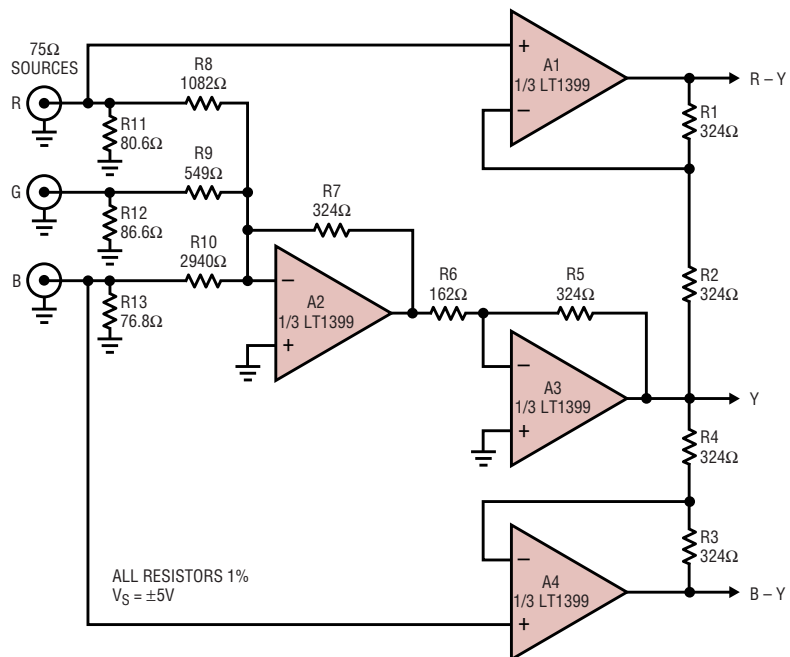
It is important to remember though, that Y is a weighted sum of R, G, and B such that:

$$Y = 0.3R + 0.59G + 0.11B.$$

If we substitute for Y at the output of A2, we then get:

$$A2_{OUT} = (1.02R - 1.02R) + 2G + (0.37B - 0.37B) = 2G$$

The back-termination resistor R16 then halves the output of A2, resulting in the G output.



**Figure 8. Buffered RGB to color-difference matrix**

## Single-Supply RGB Video Amplifier


The LT1399 can be used with a single supply voltage of 6V or more to drive single ground-referenced RGB video. As seen in Figure 10, two 1N4148 diodes, D1 and D2, have been placed in series with the output of the amplifier A1, but within the feedback loop formed by resistor R8. These diodes effectively level-shift A1's output downward by 2 diodes, allowing the circuit output to swing to ground.

Amplifier A1 is used in a positive gain configuration. The feedback resistor R8 is 324Ω. The gain resistor is created from the parallel combination of R6 and R7, giving a

Thevenin-equivalent 80.4Ω connected to 3.75V. This gives an AC gain of five from the noninverting input of amplifier A1 to the cathode of D2. However, the video input is also attenuated before arriving at A1's positive input. Assuming a 75Ω source impedance for the signal driving  $V_{IN}$ , the Thevenin-equivalent signal arriving at A1's positive input is  $3V + (0.4 \cdot V_{IN})$ , with a source impedance of 714Ω. The combination of these two inputs gives an output at the cathode of D2 of  $2 \cdot V_{IN}$  with no additional DC offset. The 75Ω back termination resistor R9 halves the signal again such that  $V_{OUT}$  equals a buffered version of  $V_{IN}$ .

It is important to note that the 4.7μF capacitor C1 is required to maintain the voltage drop across diodes D1 and D2 when the circuit output drops low enough that the diodes might otherwise be reverse biased. This means that this circuit works fine for continuous video input, but will require that C1 be charged after a period of inactivity at the input.

## Conclusion

Linear Technology has introduced the LT1399 and LT1399HV triple 300MHz current feedback amplifiers. Both of these products are well suited for use in component video applications. The higher supply voltage rating of the LT1399HV makes it an excellent choice for LCD driver applications. Both products feature 4.6mA of supply current per amplifier, 300MHz -3dB bandwidth, an exceptional 0.1dB gain flatness of 150MHz, 800V/μs slew rate and a shutdown pin for each channel. 

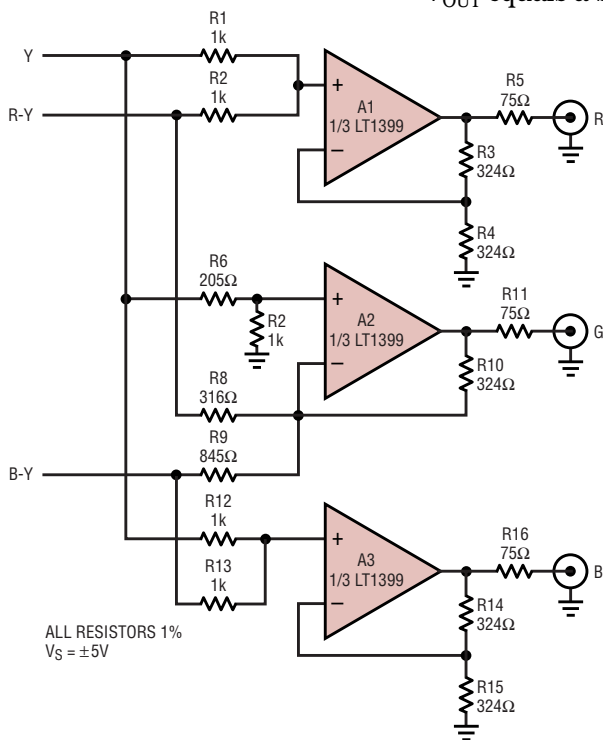


Figure 9. Buffered color-difference to RGB matrix

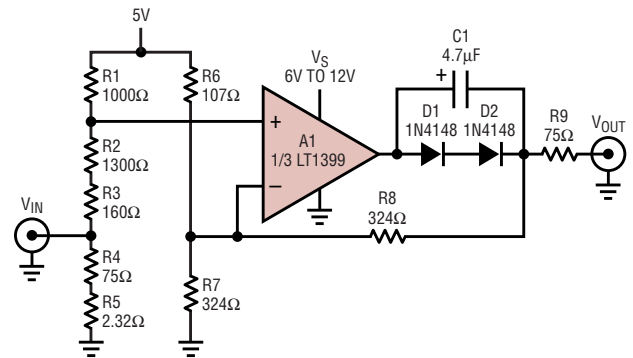


Figure 10. Single-supply RGB video amplifier (one of three channels)

LT1813, continued from page 6


is no output swing penalty as the swing is limited at the collectors of Q9 and Q13. The dynamics of the composites are not as benign as those of emitter followers, so compensation is required and is provided by C1 and C2.

The stability with capacitive loads is provided by the  $R_C$ ,  $C_C$  network between the output stage and the

gain node. When the amplifier is driving a light or moderate load, the output can follow the high impedance node and the network is bootstrapped and has no effect. When driving a heavy load such as a capacitor or small-value resistor, the network is incompletely bootstrapped and adds to the compensation provided by  $C_T$ . The added capacitance provided by  $C_C$  slows down the amplifier and the

zero created by  $R_C$  adds phase margin to increase stability.

## Conclusion

The combination of a high slew rate, DC accuracy and a frugal 3mA-per-amplifier supply current make the LT1813 a compelling choice for low voltage and low power, high speed applications. 

# LTC1702/LTC1703 Switching Regulator Controllers Set a New Standard for Transient Response

by Dave Dwelley

## Introduction

The LTC1702 dual switching regulator controller uses a high switching frequency and precision feedback circuitry to provide exceptional output regulation and transient response performance. Running at a fixed 550kHz switching frequency, each side of the LTC1702 features a voltage feedback architecture using a 25MHz gain-bandwidth op amp as the feedback amplifier, allowing loop-crossover frequencies in excess of 50kHz. Large onboard MOSFET drivers allow the LTC1702 to drive high current external MOSFETs efficiently at 550kHz and beyond. The high switching frequency allows the use of small external inductors and capacitors while maintaining excellent output ripple and transient response, even as load currents exceed the 10A level. The dual-output LTC1702 is packaged in a space-saving 24-pin narrow SSOP, minimizing board space consumed.

Mobile PCs using the most recent Intel Pentium® III processors require LTC1702-level performance coupled with a DAC-controlled voltage at the core supply output. The LTC1703 is designed specifically for this application and consists of a modified LTC1702 with a 5-bit DAC controlling the output voltage at side 1. The DAC conforms to the Intel Mobile VID

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specification. Figure 6 shows an example of a complete mobile Pentium III power supply solution using the LTC1703. The LTC1703 is packaged in the 28-pin SSOP package, conserving valuable PC board real estate in cramped mobile PC designs.

## LTC1702/LTC1703 Architecture

The LTC1702/LTC1703 each consist of two independent switching regulator controllers in one package. Each controller is designed to be wired as a voltage feedback, synchronous step-down regulator, using two external N-channel MOSFETs per side as power switches (Figure 1). A small external charge pump ( $D_{CP}$  and  $C_{CP}$  in Figure 1) provides a boosted supply voltage to keep M1 turned fully on. The switching frequency is set internally at 550kHz. A user-programmable current limit circuit uses the synchronous MOSFET switch, M2, as a current sensing element, eliminating the need for an external low value current sensing resistor. The LTC1702/LTC1703 are designed to operate from a 5V or 3.3V input supply, provided either by the main off-line supply in an AC powered system or a primary switching regulator in battery powered systems. Maximum input voltage is 7V.

Synchronous operation maximizes efficiency at full load, where resistive drops in the switching MOSFET and the synchronous rectifier dominate the power losses. As the load drops and switching losses become a larger factor, the LTC1702/LTC1703 automatically shifts into discontinuous mode, where the synchronous rectifier MOSFET turns off before the end of a switching cycle to prevent reverse current flow in the inductor. As the load current continues to decrease, the LTC1702/LTC1703 switches modes again and enters Burst Mode™, where it will only switch as required to keep the output in regulation, skipping cycles whenever possible to reduce switching losses to a bare minimum. With no output load in Burst Mode, the supply current for the entire system drops to the 3mA quiescent current drawn by each side of the LTC1702/LTC1703. Each side can be shutdown independently; with both sides shut down, the LTC1702/LTC1703 enters a sleep mode where it draws less than 50µA.

## Inside the LTC1702/LTC1703

The LTC1702/LTC1703 features peerless regulation and transient response, due to both to its high switching frequency and a carefully designed internal architecture (Figure 2). Much of the transient response improvement comes from a new feedback amplifier design. Unlike conventional switching regulator designs, the LTC1702/LTC1703 use a true 25MHz gain-bandwidth op amp as the feedback amplifier (FB in Figure 2). This allows the use of an optimized compensation scheme that can tailor the loop response more precisely than the traditional RC from COMP to ground. A "type 3" feedback circuit (Figure 3) typically allows the

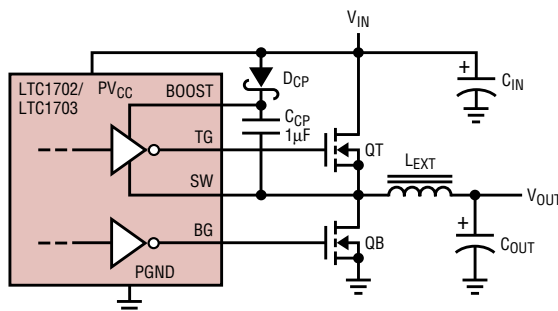


Figure 1. LTC1702/LTC1703 switching architecture

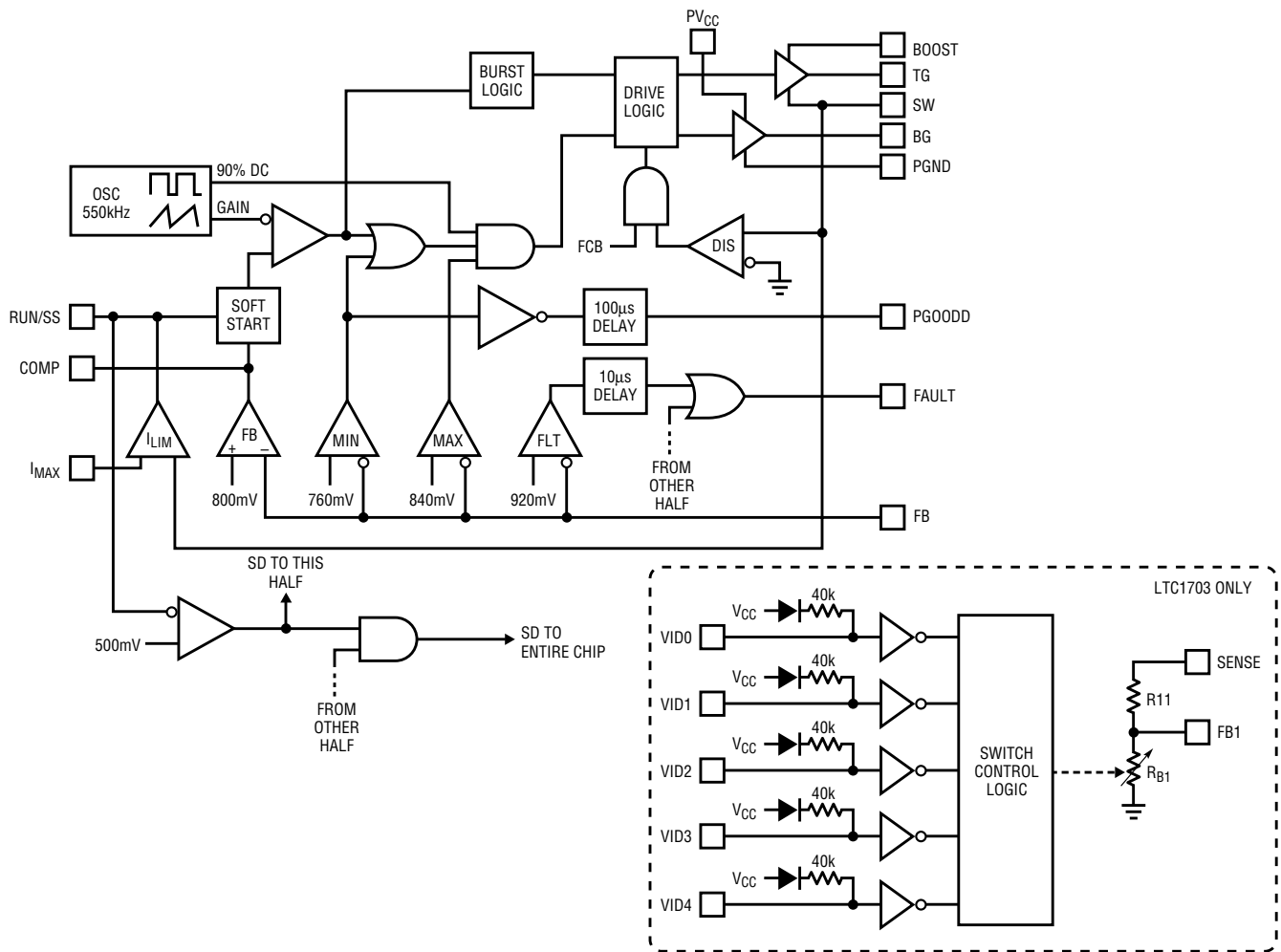


Figure 2. LTC1702/LTC1703 block diagram

loop to be crossed over beyond 50kHz while maintaining good stability, significantly enhancing load transient response. Two additional high speed comparators (MIN and MAX in Figure 2) run in parallel with the main feedback amplifier, providing virtually instantaneous correction to sudden changes in output voltage. In a typical application, the LTC1702/LTC1703 will correct the duty cycle and have the output voltage headed back in the right direction the very next switching cycle after a transient load is applied.

The positive input of the feedback op amp is connected to an onboard reference trimmed to 800mV  $\pm$ 3mV. DC output error due to the reference and the feedback amplifier are inside 0.5% and DC load and line regulation are typically better than 0.1%, providing excellent DC accuracy. The 800mV reference level allows the

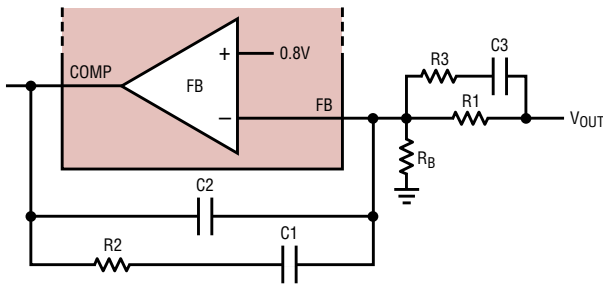
LTC1702/LTC1703 to provide regulated output voltages as low as 0.8V without additional external components. This reference performance, combined with the high speed internal feedback amplifier and properly chosen external components, allows the LTC1702 to provide output regulation tight enough for virtually any microprocessor, today or in the future. For those Intel processors that don't know what voltage they want until they actually get powered up, the LTC1703 with its onboard 5-bit VID output voltage control is the best solution.

Another architecture trick inside the LTC1702/LTC1703 reduces the required input capacitance with virtually no performance penalty. The LTC1702/LTC1703 includes a single master clock, which drives the two sides such that side 1 is 180° out of phase from side 2. This technique,

known as 2-phase switching, has the effect of doubling the frequency of the switching pulses seen by the input capacitor and significantly reducing their RMS value. With 2-phase switching, the input capacitor is sized as required to support a single side at maximum load. As the load increases at the other side, it tends to cancel, rather than add to, the RMS current seen by the input capacitor; hence, no additional capacitance needs to be added.

### External Components

The other half of the performance equation is made up by the external components used with the LTC1702/LTC1703. The 550kHz clock frequency and the low 5V input voltage allow the use of external inductors in the 1µH range or lower ( $L_{EXT}$  in Figure 1) while still keeping inductor ripple current under control. This low inductance



**Figure 3. Type-3 feedback loop**

value helps in two ways: it reduces the energy stored in the inductor during each switching cycle, reducing the physical core size required; and it raises the attainable di/dt at the output of the circuit, decreasing the time that it takes for the circuit to correct for sudden changes in load current. This, in turn, reduces the amount of output capacitance ( $C_{OUT}$  in Figure 1) required to support the output voltage during a load transient. Together with the reduced capacitance at the input due to the LTC1702/LTC1703's 2-phase internal switching, this significantly reduces the amount of total capacitance needed, compared to a conventional design running at 300kHz or less.

Each side of an LTC1702/LTC1703 circuit requires a pair of N-channel power MOSFETs to complete the power switching path. These are chosen for low  $R_{DS(ON)}$  and minimum gate charge, to minimize conductive losses with heavy loads and switching losses at lighter loads. MOSFET types that work well with the LTC1702/LTC1703 include the IRF7805 from International Rectifier, the Si9802 and Si9804 from Siliconix and the FDS6670A from Fairchild.

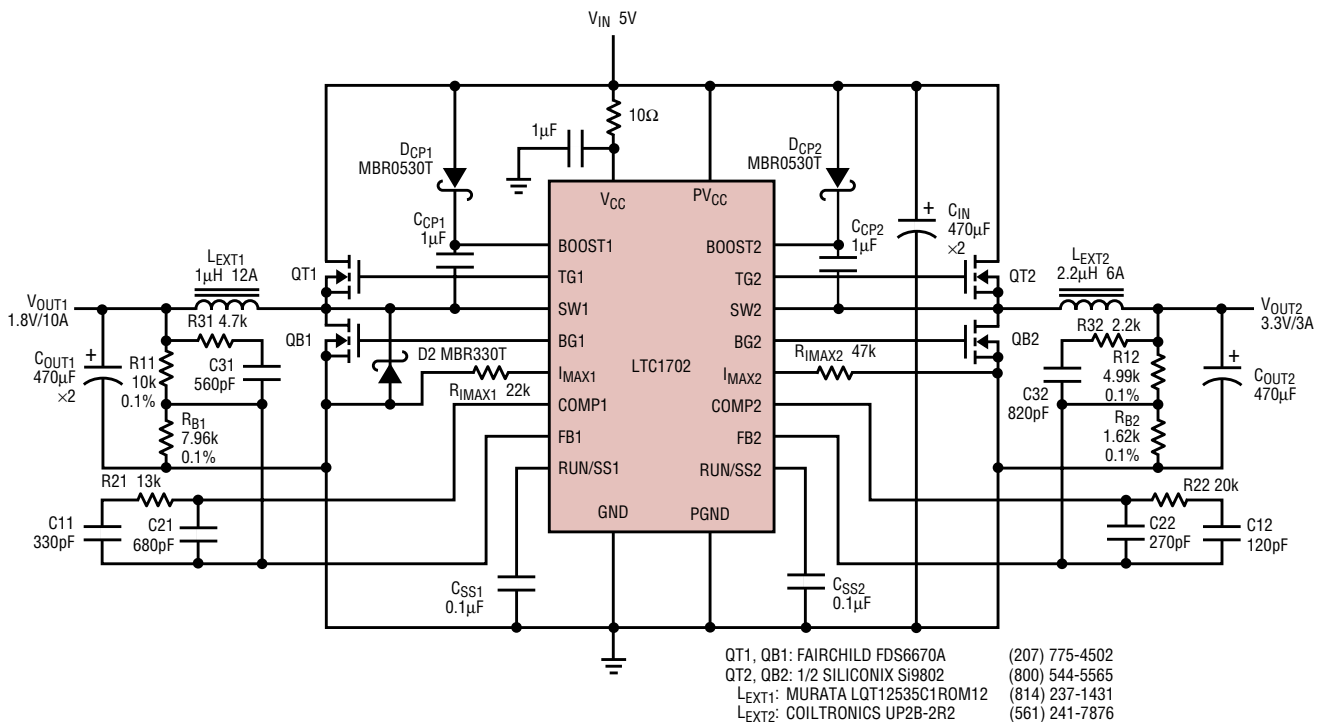
The compensation components round out the list of external parts required to complete an LTC1702/LTC1703 circuit. Because the LTC1702/LTC1703 uses an op amp as the feedback amplifier, the compensation network is connected between the COMP pin (at the output

of the op amp) and the FB pin (the inverting input) as a traditional op amp integrator (Figure 3). A bias resistor is added to set the DC output voltage and two pole/zero pairs are added to the circuit to compensate for phase shift caused by the inductor/output capacitor combination. Current limit and soft start time for each side are programmed with a single resistor ( $R_{IMAX}$ ) at each  $I_{MAX}$  pin and a single capacitor ( $C_{SS}$ ) at each RUN/SS pin. Optional FAULT (LTC1702/LTC1703) and PWRGD (LTC1702 only) flags are available to provide status information to the host system.

## Applications

### Dual Outputs from a 5V Supply

A typical LTC1702 application is shown in Figure 4. The input is taken from the 5V logic supply. Side 1 is set up to provide 1.8V at 10A and side 2 is set to supply 3.3V at a lower 3A load level. System efficiency peaks at greater than 90% at each side. This circuit shows examples of both high power and lower power output designs possible with the LTC1702 controller. Side 1 uses a pair of ultralow  $R_{DS(ON)}$  Fairchild FDS6670A SO-8 MOSFETs



**Figure 4. Dual outputs from a 5V supply**

QT1, QB1: FAIRCHILD FDS6670A (207) 775-4502  
 QT2, QB2: 1/2 SILICONIX SI9802 (800) 544-5565  
 L\_EXT1: MURATA LQT12535C1R0M12 (814) 237-1431  
 L\_EXT2: COILTRONICS UP2B-2R2 (561) 241-7876

## 2-Step Conversion


As microprocessor operating voltages continue to decrease, power conversion for CPU core power is becoming a daunting challenge. A core power supply must have fast transient response, good efficiency and low heat generation in the vicinity of the processor. These factors will soon force a move away from 1-step power conversion directly from battery or wall adapter to processor, to 2-step conversion, where the CPU core power is obtained from the 5V or 3.3V supply.

Several benefits result from 2-step conversion: more symmetrical transient response, lower heat generation in the vicinity of the processor and easy modification for lower processor voltages in the future. Peak currents taken from the battery are also reduced, which leads to improved battery chemical efficiency that can often compensate for the slight difference in electrical efficiency measured using laboratory power supplies. Battery life in a real notebook computer is virtually identical for 1-step and 2-step architectures.

The duty cycle for a step-down switching regulator is given by the ratio of  $V_{OUT}$  to  $V_{IN}$ . In 1-step power conversion, the duty cycle must be very low because the step-down ratio is large. This gives a very fast inductor current rise time and a much slower current decay time. The inductor size must be large enough to keep the current under control during the ramp-up. Fast current rise and slow current decay mean that the transient response of the regulator is good for load increases but poor for load decreases. The lower, constant input voltage for a 2-step conversion process yields a more symmetrical transient response and allows smaller, lower cost external components to be used. Because there is less switching loss due to the lower voltage swings, the switching frequency may also be increased.

Thermal concerns are also eased with the 2-step approach. To minimize high current PCB trace lengths, the core supply must be located near the processor. Core-voltage-level 1-step converters usually run at mid-

80% efficiencies, while the second step of a 2-step solution (like the LTC1703) runs near 90% efficiency, minimizing heat generation near the processor.

The biggest argument against 2-step conversion is the perceived drop in efficiency. "Off the cuff" calculations give a false impression that efficiency decreases. In fact, accurate calculations of efficiency for 2-step power conversion based on actual circuit measurements show efficiency numbers within 1% of 1-step, high efficiency converters. As time goes forward, microprocessor fabrication lithography will continue to shrink and force still lower CPU core operating voltages and higher operating currents; 1.1V supplies and 15A operating currents are already on the horizon for portable systems. These demands will render the traditional 1-step conversion approaches unworkable as a result of infinitesimal duty cycles and severely skewed transient behavior. 

For more information on 2-step conversion, see [www.linear-tech.com/ezone/2-step.html](http://www.linear-tech.com/ezone/2-step.html)

and a large 1 $\mu$ H/12A Murata surface mount inductor.  $C_{IN}$  consists of two 470 $\mu$ F, low ESR tantalum capacitors to support side 1 at full load, and  $C_{OUT1}$  uses two more of the same to provide better than 5% regulation with 0A-10A transients.

Side 2 uses a single SO-8 package with two smaller MOSFETs inside (the Siliconix Si9402) and a smaller 2.2 $\mu$ H/6A inductor.  $C_{OUT2}$  is a single

470 $\mu$ F tantalum to support 0A-3A transients while maintaining better than 5% regulation. As the load current at side 2 increases, the LTC1702 2-phase switching actually reduces the RMS current in  $C_{IN}$ , removing the need for additional capacitance at the input beyond what side 1 requires. Both sides exhibit exceptional transient response (Figure 5). The entire circuit can be laid out in less than 2 square inches when a double-sided PC board is used.

and the CPU I/O ring supply voltage. Both the LTC1628 and the LTC1703 use 2-phase switching to minimize capacitance required by the circuit; the entire 4-output circuit requires barely 2000 $\mu$ F while generating 60W of output power.

The 2-step conversion used in this circuit provides improved transient response compared to the traditional single-step approach where each

## 2-Step Converter for Notebook Computers

Figure 6 is a complete power supply for a typical notebook computer using the next generation of Intel mobile Pentium III processor. The circuit uses the LTC1628 to generate 5V and 3.3V from the input battery and uses the LTC1703 to generate the processor core voltage (with 5-bit VID control)

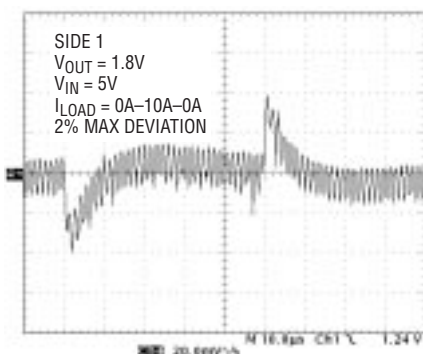


Figure 5a. Transient response, side 1

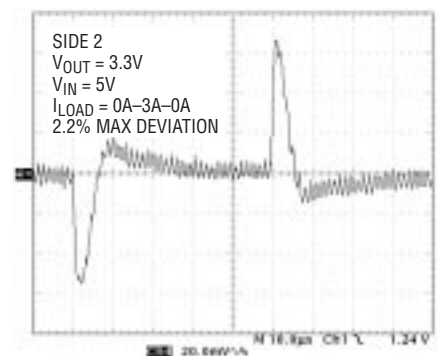
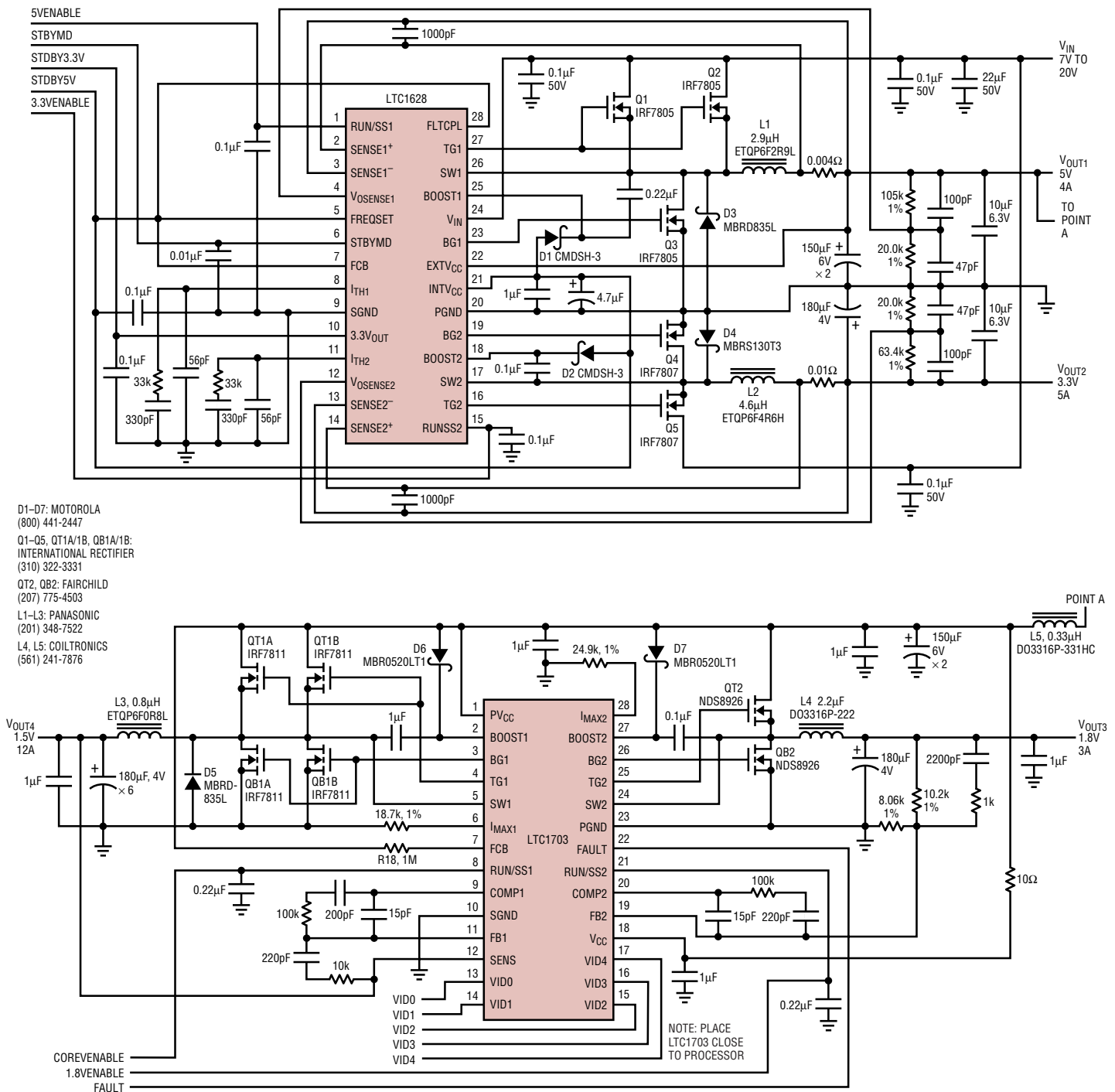


Figure 5b. Transient response, side 2



**Figure 6. 4-output notebook computer power supply**

**Conclusion**

voltage is derived directly from the battery voltage. 2-step also allows the use of smaller external components without paying an efficiency or performance penalty and it eases layout and thermal management concerns. See the “2-Step Conversion” sidebar for more information.

The LTC1702 and LTC1703 achieve DC and AC regulation performance that tops the best switching regulator controllers available today. As logic densities continue to climb, more applications are appearing where the input voltage is limited to below 7V and the output voltage is low, the output current is high and multiple

outputs are required. The LTC1702 and LTC1703 provide the best combination of regulation performance, high efficiency, small size and low system cost for such applications, whether they appear in advanced notebook computers or complex logic systems. **LT**

# An 8-Channel, High-Accuracy, No Latency $\Delta\Sigma$ 24-Bit ADC

by Michael K. Mayes

## Introduction

Recently, Linear Technology introduced the world's most accurate, simplest to use, 24-bit analog-to-digital converter, the LTC2400. With its on-chip oscillator, 120dB line-frequency rejection, user-transparent offset/full-scale calibration, 10 parts-per-million (ppm) total unadjusted error and  $1.5\mu\text{V}_{\text{RMS}}$  noise, the LTC2400 has become a key building block in many system designs. The LTC2400's ease of use and high performance enable faster design cycles and better performance than other  $\Delta\Sigma$  converters.

This article introduces the LTC2408, a device combining the high performance LTC2400 ADC core with an 8-channel analog input multiplexer (see Figure 1). This device offers many unique features. The single-cycle settling characteristics lead to simplified multiplexer hook up and channel selection, without the added overhead required other converters. The exceptional noise performance of the device eliminates the need for a programmable gain amplifier (PGA). This allows direct digitization of a variety of voltage levels. Its 10ppm absolute accuracy ensures a minimum perfor-

mance in excess of 16 bits. A unique analog modulator implementation allows measurement of microvolt signals superimposed upon large DC voltages. A wide range of sensor inputs and voltage levels can be applied simultaneously to the LTC2408. These signals can extend below ground, above  $V_{\text{CC}}$  or anywhere in between, with the same 10ppm absolute accuracy.

## Single-Cycle Settling Ensures No Latency

Many applications requiring 16-bit to 24-bit resolution use delta-sigma ( $\Delta\Sigma$ ) ADCs. These applications typically measure slow-moving signals, such as those found in temperature measurement, weight scales, strain-gage transducers, gas analyzers, battery monitoring circuits and DVMs. One advantage delta-sigma converter architectures offer over conventional ADCs is on-chip digital filtering. For the low frequency applications described above, this filter is designed to provide rejection of line frequencies at 50Hz or 60Hz and their harmonics.

A disadvantage of conventional digital filters, prior to the release of

the LTC2400, was digital filter settling time. If the input signal changes abruptly, the conversion result is invalid for the following 3–4 conversion cycles (see Figure 2a). This makes multiplexing the input difficult. The LTC2400 does not exhibit a filter settling time; hence, it is easy to multiplex (see Figure 2b); There is a one-to-one correspondence between the conversion result and the applied input signal. Each conversion result is independent from the previous conversion result. The 10ppm total error is maintained for each conversion cycle, even in the extreme case of sequentially measuring 0V and 5V on adjacent channels.

## The Advantages of Not Using a Programmable Gain Amplifier (PGA)

The exceptional noise performance of the LTC2408 ( $1.5\mu\text{V}_{\text{RMS}}$ ) corresponds to an effective resolution of 21.6 bits for a 5V input range. Low level input signals within a 100mV range achieve better than 16-bit effective resolution without the use of a PGA. On the other hand, conventional  $\Delta\Sigma$  ADCs are significantly noisier than the LTC2408 for a 5V input range. These converters require internal PGAs in order to improve the noise performance for low level input voltages.

The LTC2408 offers several significant advantages over those converters requiring a PGA. One advantage the LTC2408 offers is the ability to measure small signals (microvolts) superimposed upon large DC voltages (volts). For example (see Figure 3), a 100mV signal sitting on 2V (2V to 2.1V) can be measured with the same accuracy and noise performance as a 100mV signal sitting on ground (0V to 0.1V). Conversely, an ADC operating with a programmable gain of 50 is limited to an input range of 0V to 0.1V with a 5V reference (see

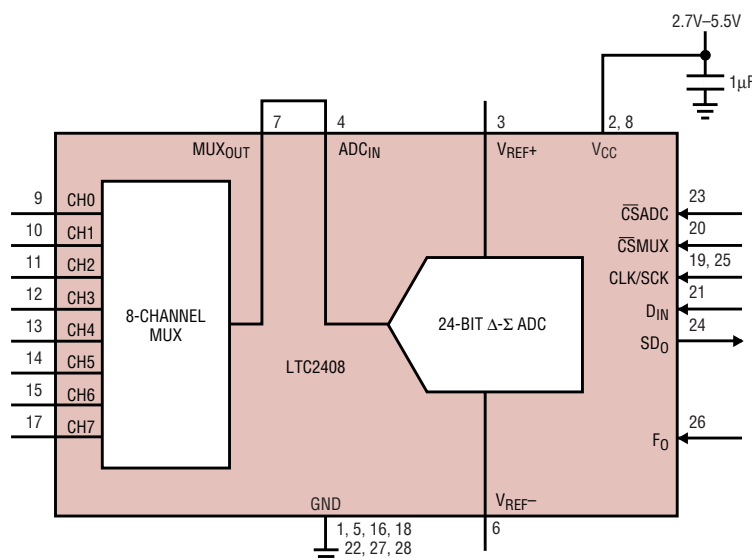
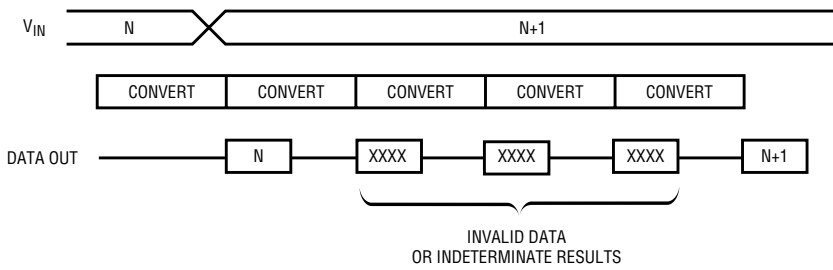
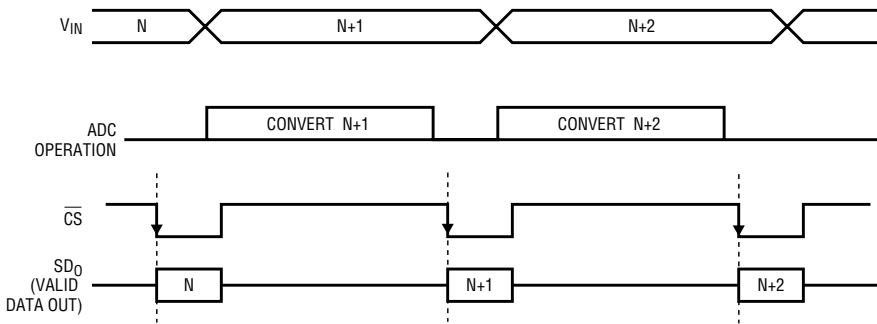


Figure 1. LTC2408 block diagram



**Figure 2a. Effect of conventional digital filter settling time**



**Figure 2b. The LTC2408 has no digital filter settling time.**

Figure 3). It cannot digitize any signal larger than 100mV full-scale.

A second advantage the LTC2408 offers is full-scale accuracy. Since the total unadjusted error is less than 10ppm, the absolute accuracy of any input voltage within the 0V to 5V range is within 10ppm or 16 bits. Alternatively, devices using PGAs exhibit full-scale errors limited by the matching of internal components. The user is burdened with removal of these errors. The user must first apply the system's full-scale voltage to the device and then perform a system calibration.

The use of a PGA in conventional  $\Delta\Sigma$  adds complexity. Each channel requires a system full-scale and offset calibration. Each channel may have a different PGA gain and input-signal range settings, corresponding to different offset and full-scale calibration coefficients. This requires

programming and maintaining configuration/status registers, gain/offset registers and channel/PGA-gain registers. The LTC2408 does not require any registers. The offset and full-scale error corrections are performed during each conversion cycle and are transparent to the user.

**“Microvolts on Volts”**

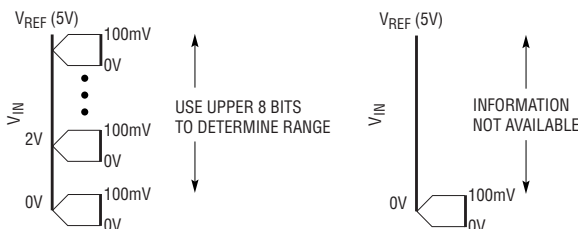
In order to measure a small level signal (microvolts) superimposed upon a large signal (volts), the converter must exhibit extremely good DC performance. The device must have very low offset and full-scale errors and excellent linearity performance in order to accurately digitize small signals with large fixed DC levels. Additionally, the temperature coefficients of offset, full-scale and linearity errors must be low. The LTC2408's offset error is less than 1ppm and its offset drift is less than

0.01ppm/°C (see Figure 4a). The full-scale error is less than 4ppm while its drift is less than 0.02ppm/°C (see Figure 4b). Combined with an integral nonlinearity error of 4ppm, the LTC2408 can consistently resolve low level signals in the microvolt range, regardless of the fixed DC level (within the 0V to  $V_{REF}$  range).

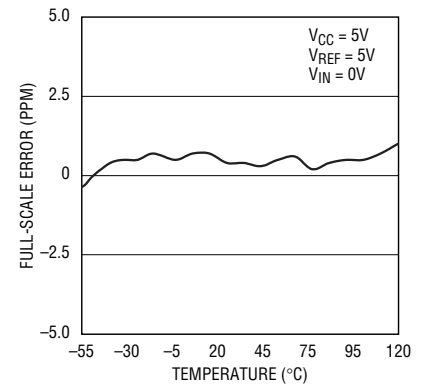
The accuracy, noise performance, and temperature stability of the LTC2408 enable the converter to measure input signals from a multitude of sensors (see Figure 5). In addition to the LTC2408's ability to measure signals from 0 to  $V_{REF}$ , the device also has overrange/underrange capabilities. The device can measure an input signal 100mV below ground and 100mV above  $V_{REF}$ , even if  $V_{REF}$  is equal to  $V_{CC}$ .

**A Simple 4-Wire SPI Interface**

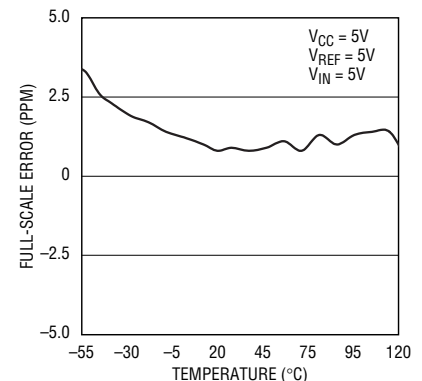
Interfacing to the LTC2408 is simple. The individual  $\overline{CS}$  and CLK signals (see Figure 6) can be common to both the ADC and multiplexer or driven independently to allow separate con-



**Figure 3. Full range without PGA (left); limited range with PGA (right)**



**Figure 4a. Offset error drift**



**Figure 4b. Full-scale error drift**

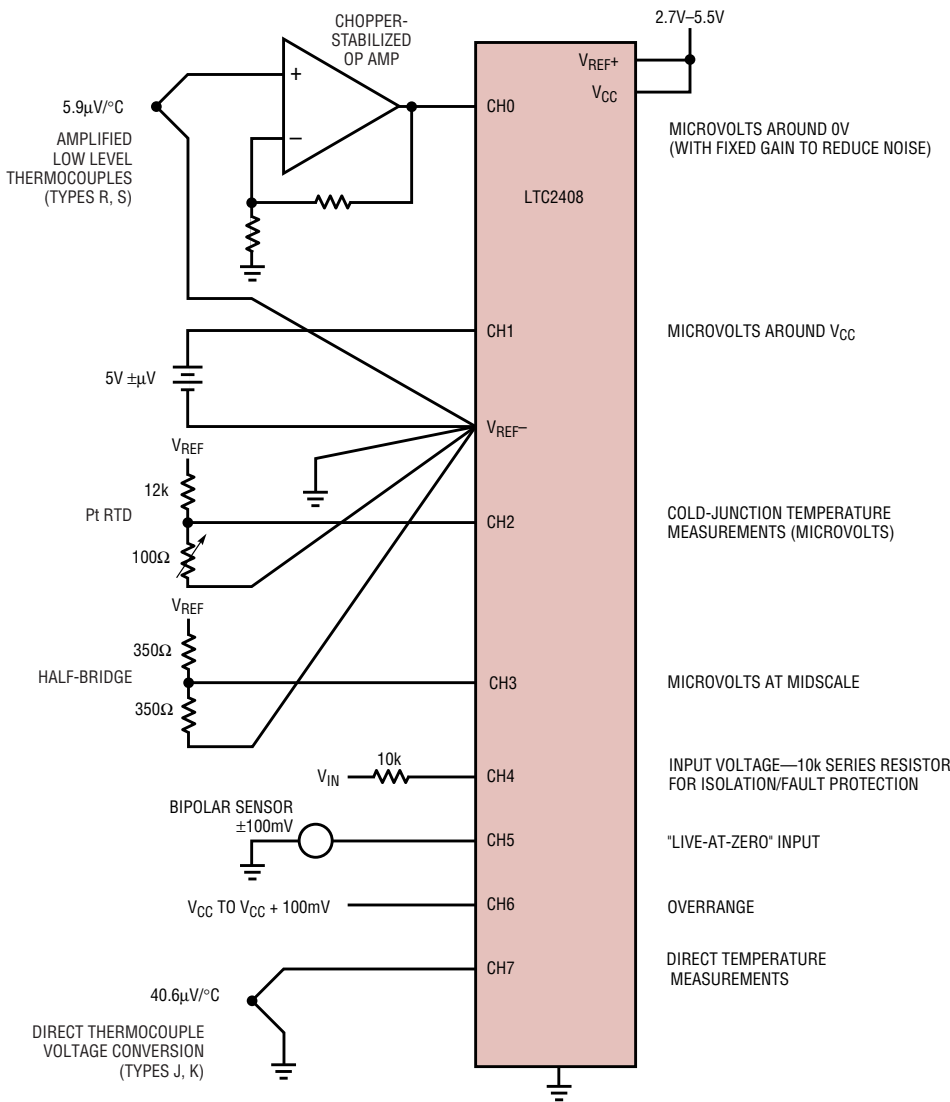



Figure 5. The LTC2408 simultaneously measures many input devices.

control of the ADC and the mux.  $D_{IN}$  is serially programmed to select the desired input channel;  $SDO$  is the serial output data of the converter.  $D_{IN}$  and  $SDO$  may be shared by using an external driver with a high impedance output state. Since the LTC2408 exhibits single-cycle settling, there is no overhead associated with digital filter settling time. At the conclusion of each conversion, a new channel may be selected by a 4-bit serial input word, or the same channel can be retained by not shifting in a new word. A new input channel may be selected up to 66ms after the data-output read has been completed. This 66ms period may be used to allow the input signal to settle or offer the user flexibility in the timing of the mux channel selection.

### Conclusion

The LTC2408 is a highly accurate No Latency  $\Delta\Sigma$  converter capable of digitizing a variety of input signals. Its exceptional noise performance allows direct digitization of sensors. The device can measure microvolts on one channel and volts on another, all with 10ppm accuracy. The LTC2408 requires no user calibration or PGA, and there is no overhead associated with the input multiplexer. The LTC2408's exceptional accuracy, ease-of-use and eight input channels make it an ideal multichannel ADC for complete system monitoring. 

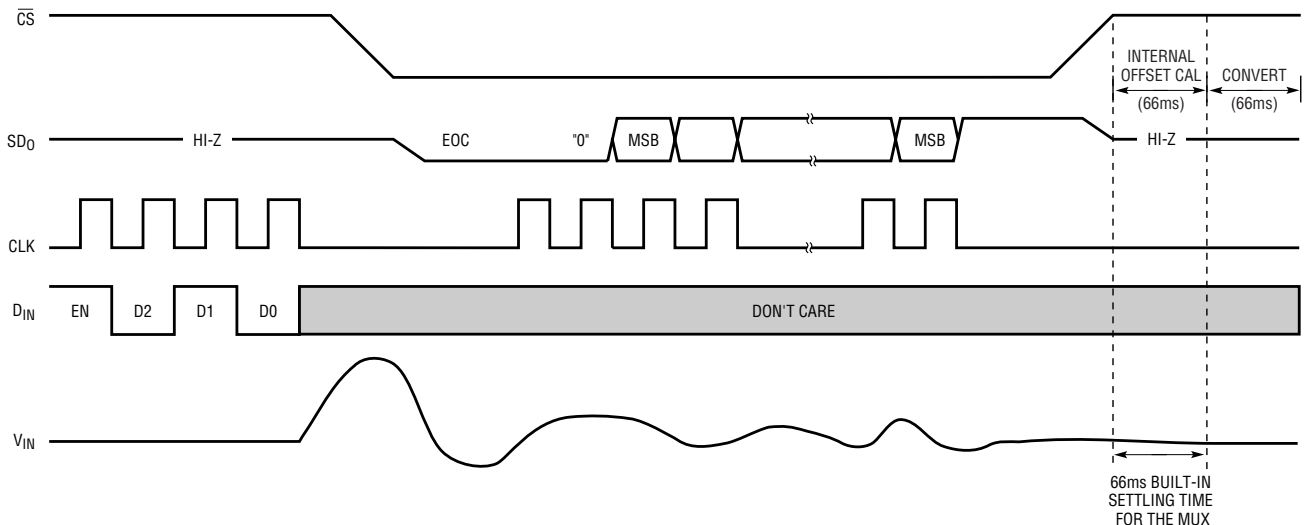


Figure 6. Mux/ADC timing and look ahead

# Micropower 5-Lead SOT-23 Switching Regulators Extend Battery Life in Space-Sensitive Applications

by Bryan Legates

## Introduction

The LT1615 and LT1617 are designed for portable electronics that need a power solution with a minimum footprint and long battery life. These devices can be used as step-up or boost converters, single-ended primary inductance converters (SEPIC) or positive-to-negative converters. With an input voltage range of 1.2V to 15V, these devices are ideal for a wide range of applications and work with a variety of input sources. An internal 36V switch allows the two devices to easily generate output voltages of up to  $\pm 34V$  without the use of costly transformers. The LT1615 is designed to regulate positive output voltages, whereas the LT1617 is designed to directly regulate negative output voltages without the need for level-shifting circuitry. Both parts use a current-limited, fixed off-time control scheme, which helps achieve high efficiency operation over a wide range of load currents. With a no-load quiescent current of only  $20\mu A$  (with the output in regulation) and a shutdown quiescent current of  $0.5\mu A$ , these devices squeeze the most life out of any battery application. Both devices use tiny, low profile inductors and capacitors to minimize the overall system footprint and cost.

The LT1615 and LT1617 are pin compatible with two other members of the PowerSOT™ family, the LT1613 and LT1611, respectively. This allows the same board layout to be used to evaluate the performance of multiple devices. The LT1613 and LT1611 are both current mode, constant frequency devices, capable of producing larger output currents than the LT1615 and LT1617.<sup>1</sup>

## LT1615 2-Cell to 3.3V Boost Converter

A popular supply for many portable electronic devices, a 2-cell alkaline to 3.3V converter with the LT1615 can deliver 60mA of load current. The circuit is shown in Figure 1 and the system efficiency appears in Figure 2. The efficiency peaks at 84% with a fresh 2-cell battery and averages 78% over the entire 1.5V to 3V input voltage range. Switching waveforms with an input voltage of 2.8V and a 60mA load appear in Figure 3. This photo illustrates the Burst Mode operation of the LT1615, as the system delivers energy to the output in short bursts about every  $15\mu s$ . The device is in standby (drawing only  $20\mu A$  of quies-

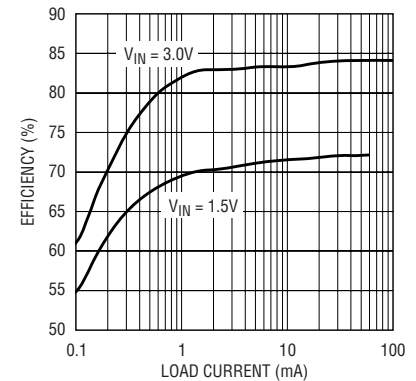


Figure 2. 3.3V boost converter efficiency reaches 84%

cent current) for about  $12\mu s$  of each  $15\mu s$  burst cycle, which greatly increases the overall converter efficiency.

## LT1615 1-Cell Li-Ion to 15V Boost Converter

The internal 36V switch of the LT1615 makes it an attractive choice for applications that need a high output voltage at a relatively low current. Figure 4's circuit is a typical system that provides 15V at 15mA from a single-cell Li-Ion battery. Efficiency, shown in Figure 5, reaches 82% from a fully charged battery.

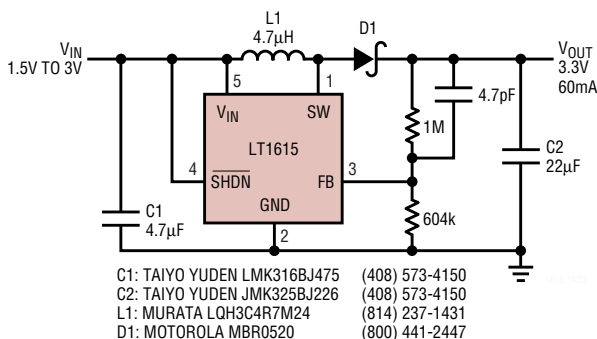


Figure 1. 2-cell to 3.3V boost converter

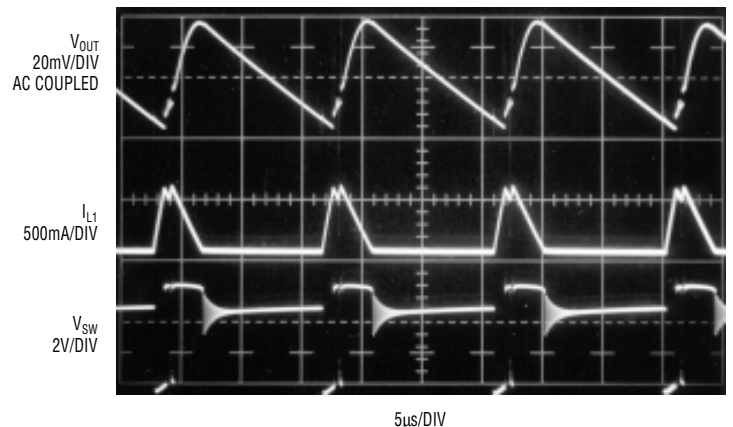


Figure 3. Switching waveforms of 3.3V boost converter with 60mA load

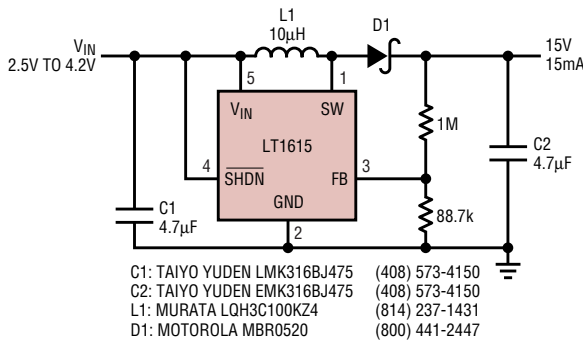


Figure 4. Li-Ion to 15V boost converter

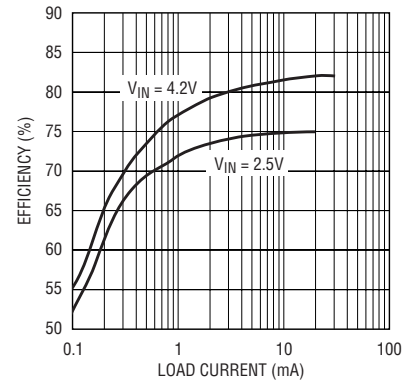


Figure 5. Efficiency of Figure 4's boost converter

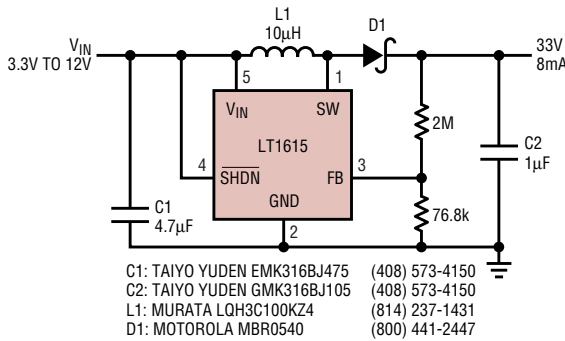


Figure 6. 33V boost converter

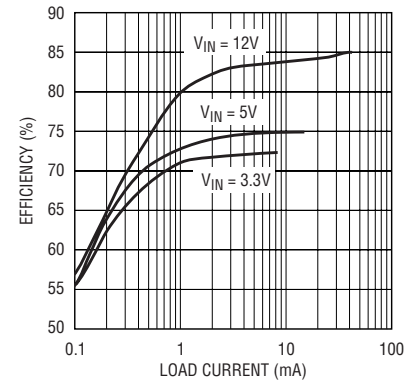


Figure 7. 33V boost converter efficiency

### LT1615 33V Boost Converter

The converter in Figure 6 displays the exceptional input and output voltage range of the LT1615. A 33V output is easily generated from a wide ranging input voltage using a simple boost topology. A small, 1µF ceramic capacitor is all that is needed at the output, making the total footprint much smaller than other systems that need much larger tantalum capacitors. The efficiencies for inputs of 3.3V, 5V and 12V are shown in Figure

7. With a 12V input, this converter can deliver up to 1.32W (40mA at 33V) of power at an efficiency of 85%, all from a tiny SOT-23 package.

### LT1615 1-Cell Li-Ion to 3.3V SEPIC

Lithium-ion (Li-Ion) is the battery of choice for systems needing the most energy with the lightest weight, but with a cell voltage ranging from 4.2V down to 2.5V, a simple boost topology cannot be used to provide a 3.3V

output. Figure 8's circuit is a SEPIC converter that can easily do the job, providing 100mA of load current. The circuit shown uses two separate inductors, but a single, dual-winding inductor (a 1:1 transformer) can be substituted. Figure 9 shows the switching waveforms for this SEPIC converter with an input voltage of 2.7V and a 50mA load. Notice that this circuit uses the same basic Burst Mode operation as the boost converter, but the inductor current is

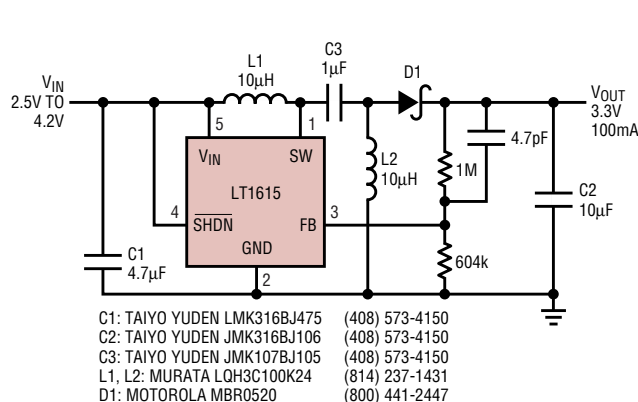


Figure 8. Li-Ion to 3.3V SEPIC converter

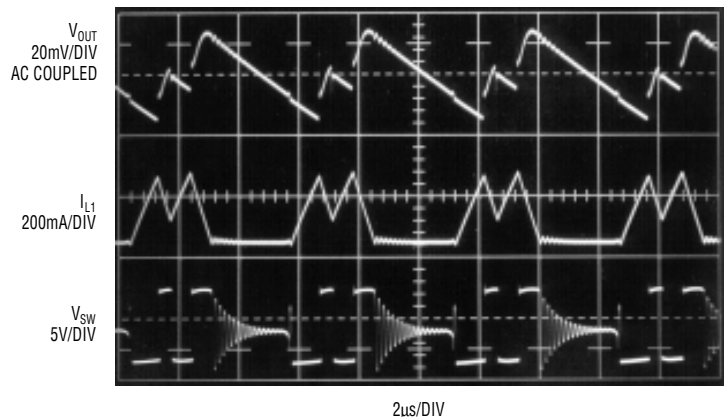
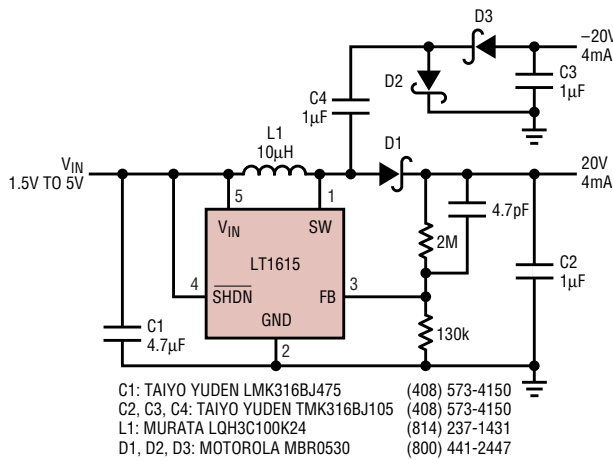
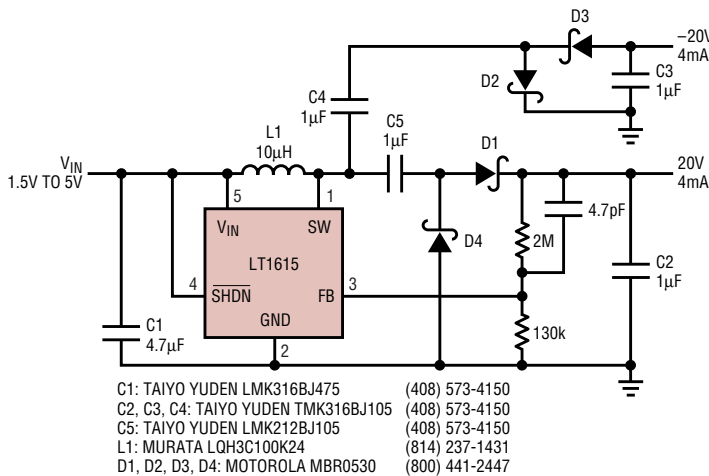


Figure 9. Switching waveforms of the 3.3V SEPIC converter with 50mA load



**Figure 10. ±20V dual-output converter**



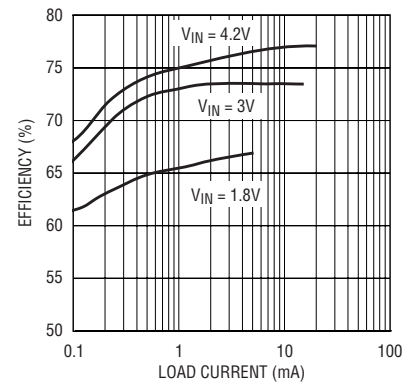
**Figure 12. ±20V dual-output converter with load disconnect**

about one half that of the 3.3V boost circuit whose waveforms are shown in Figure 3. For the SEPIC, the switch current is split equally between the two inductors, with both inductors providing current to the load when the switch is turned off. Typical efficiency for this converter is 70%.

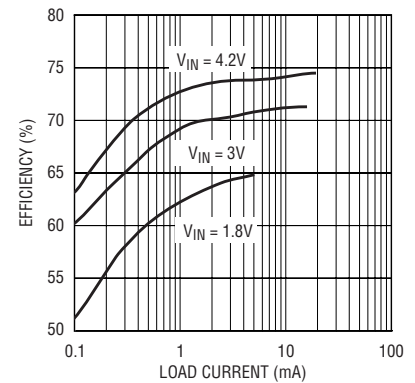
### LT1615 ±20V Dual-Output Converter

Figure 10 shows a single-inductor, dual-output converter ideal for use in applications needing both a positive and a negative voltage. The positive output is generated using a traditional boost converter, whereas the negative output is generated using an inverting charge pump. Regulation is achieved by sensing the positive output, but by using identical output capacitors and rectifying diodes, the negative output is also very well regulated. For a 2× difference in output

currents, the positive and negative output voltages differ less than 3%; for a 10× difference, they differ less than 5%. This converter provides 8mA total output current from a 1.5V input (two fully discharged alkaline batteries) and 12mA total output current from a 2.5V input (a fully discharged single-cell Li-Ion battery). Increasing the value of L1 to 22µH increases the available output current by about 15%. If even larger load currents are needed, the same converter can be implemented using the LT1613 in place of the LT1615. If the accuracy of the negative output is more critical than the accuracy of the positive output, try the same topology using the LT1617 to regulate the negative output. Efficiency for this dual-output converter is quite good, reaching 77% with a single Li-Ion battery. See Figure 11 for efficiency curves at several different input voltages.



**Figure 11. Efficiency of Figure 10's circuit**



**Figure 13. Efficiency of Figure 12's circuit**

### LT1615 ±20V Dual-Output Converter with Load Disconnect

One drawback to the circuit shown in Figure 10 is that during shutdown, the positive output is one diode drop below the input voltage. This is an undesirable condition for many systems, and can be easily corrected with the circuit in Figure 12. This is a dual-output converter where both outputs are developed using charge pumps, so that both are disconnected from the input when the LT1615 is turned off. An additional benefit is that cross regulation is improved because both outputs are generated in the same manner. For a 5× difference in output currents, the positive and negative output voltages differ less than 1%; for a 10× difference, they differ less than 2%. The improvements this circuit provides do come at a cost: slightly lower efficiency. Figure 13 shows that the efficiency curves are about 3% lower for load currents greater than 1mA, but the efficiency still reaches a respectable 74%.

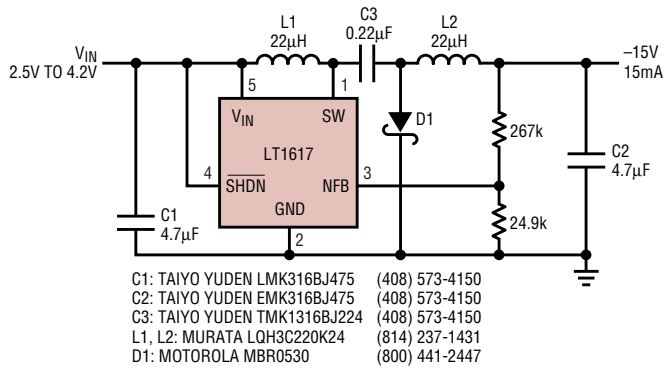


Figure 14. Li-Ion to -15V inverting converter

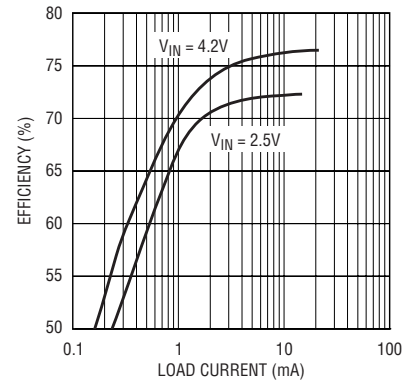


Figure 15. Li-Ion to -15V inverting converter efficiency

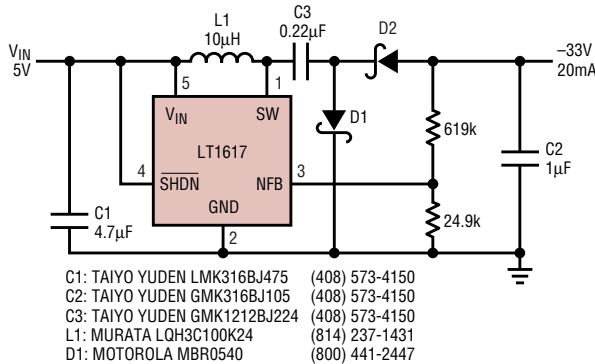


Figure 16. -33V inverting charge pump

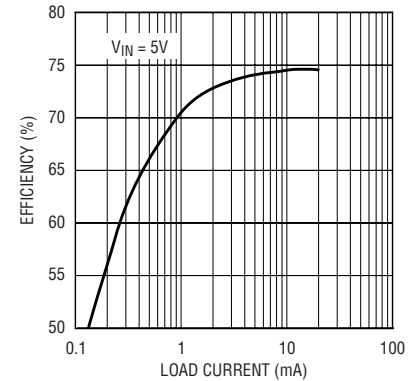


Figure 17. 5V to -33V inverting charge pump efficiency

### LT1617 1-Cell Li-Ion to -15V Inverting Converter


Many electronic systems need a negative supply but have only a positive input voltage to work with. A well regulated, positive-to-negative converter can be easily designed using the LT1617. A Li-Ion to -15V inverting converter capable of providing 15mA of load current is detailed in Figure 14. Efficiency for this inverter, shown in Figure 15, peaks at 76%.

### LT1617 5V to -33V Inverting Charge Pump

For the previous inverting converter, the maximum voltage seen by the

power switch is equal to the sum of the input and output voltages; this, along with the 36V switch rating, limits the output voltage that can be provided using the inverting topology. If higher negative voltages are needed, use an inverting charge pump, in which the maximum voltage seen by the switch is equal to the output voltage. Figure 16 shows a -33V, 20mA inverting charge pump that can provide 20mA of load current. Efficiency reaches 74%, as seen in Figure 17.

### Conclusion

The applications presented show the versatility of the LT1615 and LT1617. These devices are capable of producing a wide range of positive and negative outputs from a variety of input sources. Their tiny SOT-23 packages, along with small external components, combine to minimize footprint and cost in space-conscious applications. 

### References

1. Pietkiewicz, Steve. "SOT-23 Switching Regulators Deliver Low Noise Outputs in a Small Footprint." *Linear Technology* IX:1 (February 1999), pp.11-13, 23.



# Versatile Ring Tone Generator Finds Uses in Motor Drivers and Amplifiers

by Dale Eagar

## Overview

The LT1684 was specifically designed for OEM telephone equipment. Its function is to interface between the digital control logic and the high voltage analog phone line. When used in the design of a phone system, the LT1684 allows software to control the frequency, voltage and cadence of the ring signal. Because of similarities of application to telephone systems, the LT1684 ring tone chip finds itself at home in motor drives, digital input amplified speakers, alarm systems and sine wave UPS systems.

## Introducing the LT1684

The LT1684 provides the tool set to easily implement a digital, pulse width modulated (PWM) signal to DC-coupled voltage converter (at high currents), while providing isolation, switching frequency filtering and output protection. Figure 1 is a system-level diagram of the LT1684 in action. By controlling a pair of external MOSFETs, the LT1684 utilizes their inherent robustness while pro-

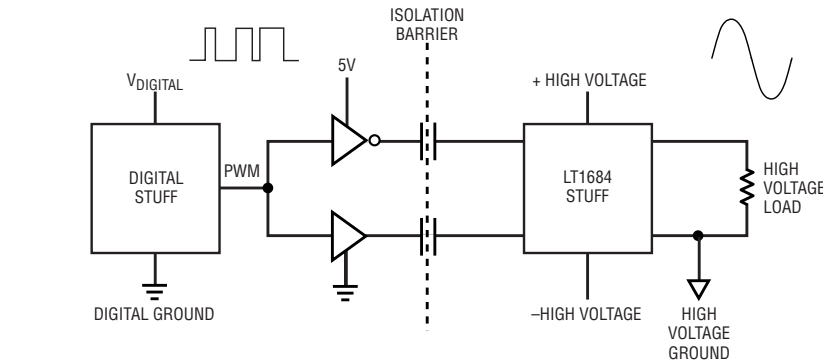


Figure 1. The LT1684 uses differential pulse width modulation to provide isolation for digitally controlled analog power solutions.

viding control of the output voltage and current. In its telecom application circuit, the LT1684 provides up to  $\pm 240V$  of smooth, clean output at up to 200mA of output current. Higher output voltages are obtained by cascoding MOSFETs, while higher currents are readily achieved by using the LT1166 MOSFET automatic bias generator chip as a companion.

## The Circuit

Figure 2 is the schematic of the LT1684 implementing a digital-PWM-to-ringing-telephone converter. This is something like a high power, high voltage, isolated, output filtered D/A converter. Like its DAC counterpart, the LT1684 has a precision reference, switches and an output amplifier. Unlike its DAC counterpart, it includes post-conversion ripple filtering, isolation and a robust high voltage output.

In addition to the isolation, filtering and amplification, the LT1684 provides the gate-bias control and gate voltage protection for the two external MOSFETs. Providing such a plethora of functions from a single, monolithic IC requires the use of a somewhat tricky circuit. This circuit

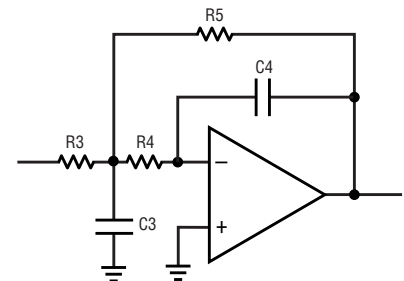


Figure 3. The basic 2nd order lowpass MFB filter, as implemented by the LT1684

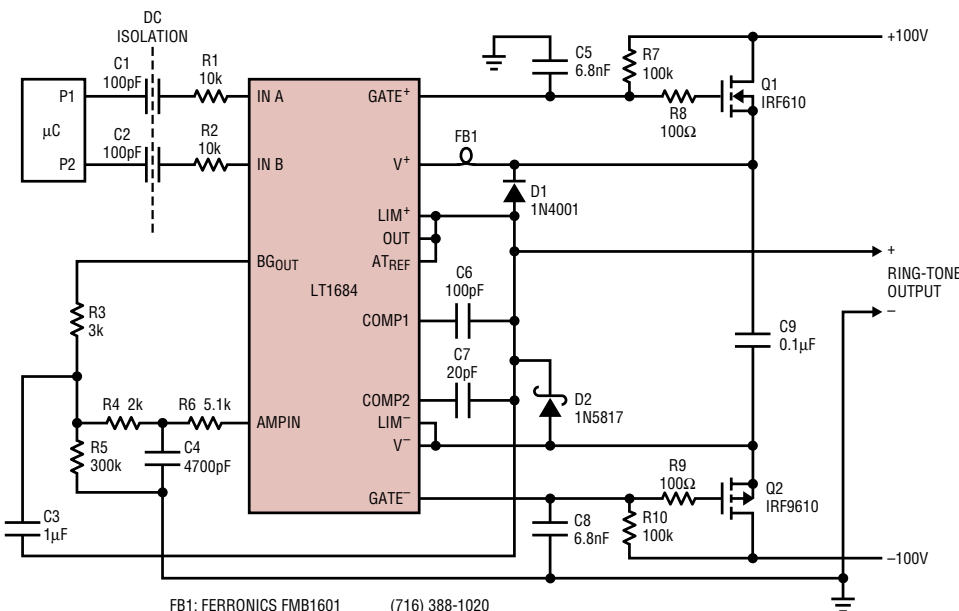


Figure 2. Typical LT1684 digital-PWM-to-ringing-telephone converter application

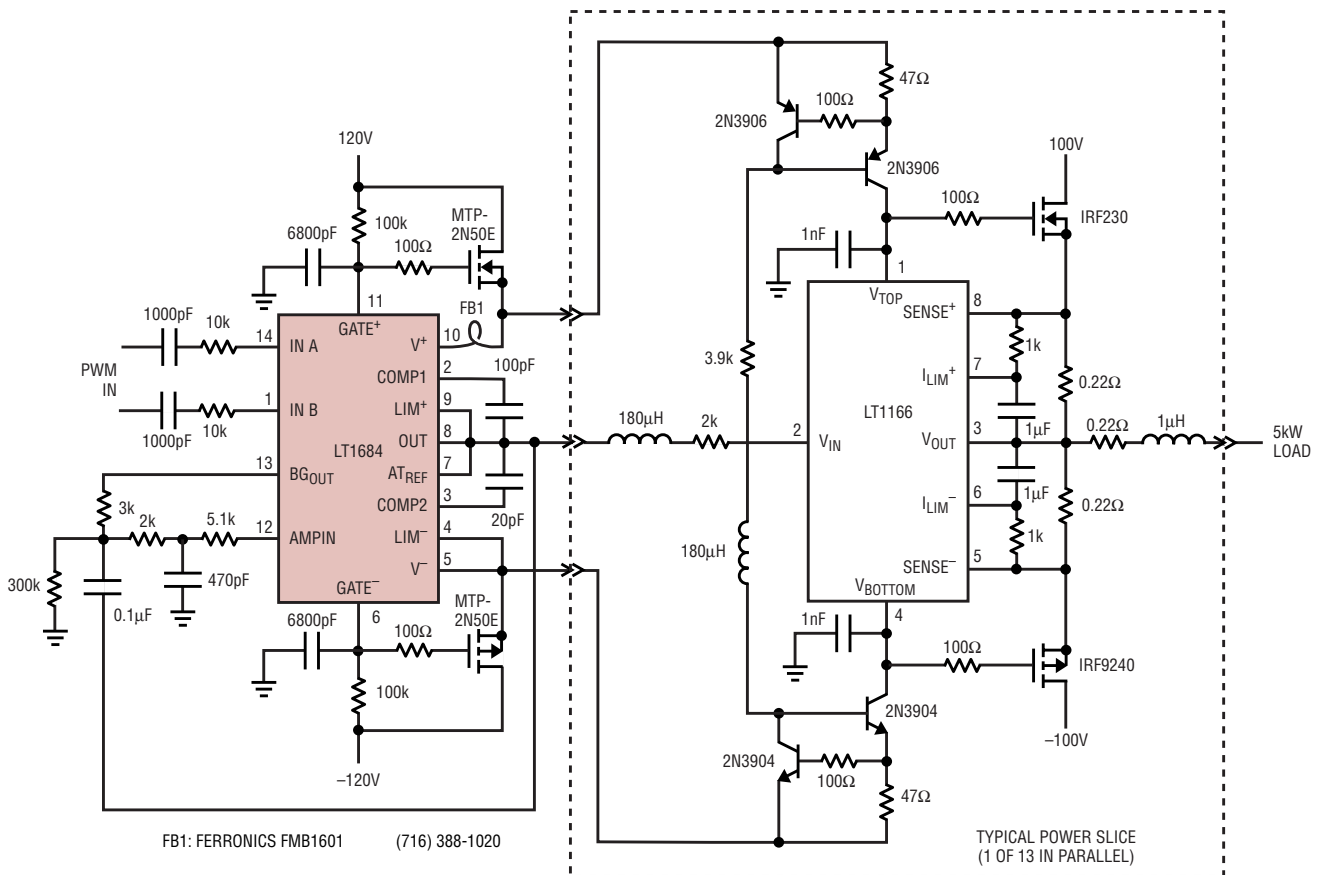


Figure 4. 5kW PWM-to-analog converter

is arrived at by applying a circuit transformation to a simple filter circuit. This transformation is performed on the basic 2nd order lowpass multiple feedback filter (MFB) circuit and ends up looking somewhat like the filter/amplifier shown in Figure 3. The filter/amplifier components in Figure 2 are R3–R5 and C3 and C4. Looking backwards through the circuit transformation, these compo-

nents, in fact, form the 2nd order MFB filter shown in Figure 3. The values chosen for these components in Figure 2 implement a 2nd order Butterworth MFB lowpass filter with a cutoff frequency of 100Hz and a DC gain of 100. These were chosen to provide  $\pm 80V$  of output swing with PWM duty factors of 10% to 90%, while filtering the 10kHz PWM ripple to meet telephone specifications.

### Stealing the LT1684 for Use In Other Applications

The LT1684, used as shown in Figure 1, outputs a ring signal that meets Belcore specification. This means we can ring a phone 22,000 feet away. The LT1684 fits another role, where 22,000 feet of separation would be a nice minimum. This is the application of the LT1684 in the scaleable power amplifier, as detailed in Figure 4. This amplifier can be used to drive motors, simulate the power company in sine wave UPS systems and operate large audio drivers. Because of its scaleable nature, this design can be used at any power level. The circuit in Figure 4 is shown implementing a 5000W bits-to-decibels converter. When this converter is implemented with the appropriate audio drivers and enclosures, the output sound pressure level can be significant—so significant, in fact, that the author suggests giving it a wide berth of at least 22,000 feet.

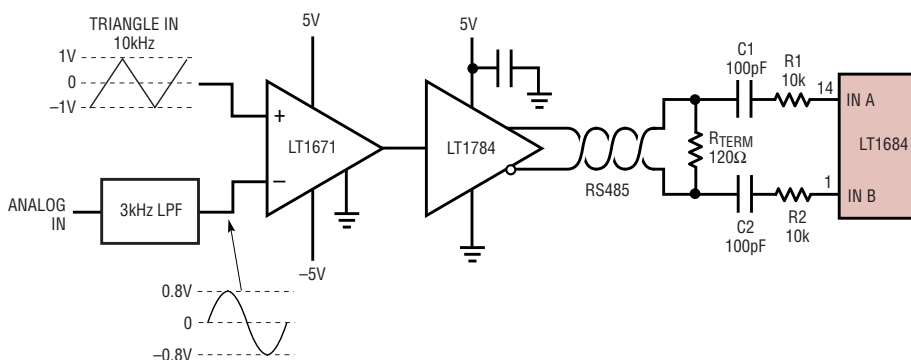


Figure 5. A remote, isolated, analog input amplifier using a robust RS485 driver and a terminated, twisted-pair line

continued on page 35

# 35 Watt Isolated DC/DC Converter Replaces Modules at Half the Cost

by Robert Sheehan

## Introduction

The choice between building or buying an isolated DC/DC converter can be a complex decision. If you use an off-the-shelf, module you are constrained by what the module makers offer in their catalogs. In many cases, this may not precisely meet the requirements for a particular project. Also, while simple to use, the cost of these modules can be significantly higher than the cost of "rolling your own." The complexity of the DC/DC

design can be daunting and leads many to the decision to buy. Demonstration circuit DC227 provides a DC/DC solution that can serve the needs of many "standard" module applications and offers the designer the option of customizing the design to suit any slightly unusual system requirements. The power supply now becomes merely another collection of parts in the system.

## Features

Demonstration circuit DC227 is a board level replacement for "half-brick" DC/DC converters. It can provide 5V or 3.3V at up to 7A from an isolated 48V (36V to 72V) input. The isolation voltage is 500VDC with an option for 1500VDC. The circuit has low input capacitance, fast turn-on time, low shutdown power consumption and overtemperature protection. Continuous short-circuit protection eliminates any restriction on maximum capacitive load. The output overvoltage circuit provides protection for open or short circuits on the output power or sense lines. The standard footprint allows the circuit to fit

directly into the module's socket. Figure 1 shows a typical layout for a 2.28" by 2.40" circuit board.

DC227A-A is designed for 500VDC isolation and lowest cost; it uses a standard Coiltronics VERSA-PAC™ transformer and a Pulse Engineering inductor for the output filter. DC227A-B has 1500V isolation and uses a semicustom transformer, also from Coiltronics. DC227A-C has 500VDC isolation and achieves the highest efficiency using a Panasonic type PCC-S1 inductor for the output filter. The efficiency curves in Figures 2-5 are quite competitive, reaching 85% for the DC227A-C with a 5V output. The efficiency at 3.3V out is somewhat lower, due to the fixed losses of the output rectifier.

## Circuit Description

This single-ended forward converter operates at a nominal switching frequency of 200kHz. Referring to the schematic in Figure 6, pulse width modulation is controlled by U1, an LT1247 current mode PWM controller. Transformer T2 and optocoupler Q7 provide galvanic isolation. C2 is a

VERSA-PAC is a trademark of Coiltronics, Inc.

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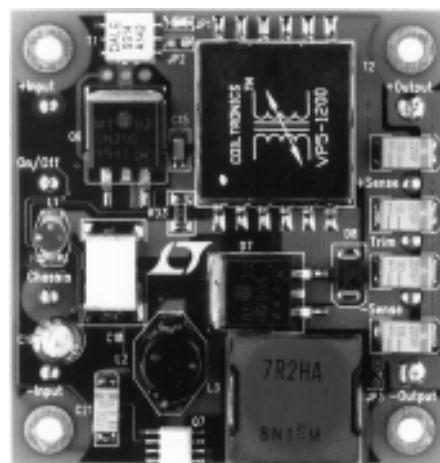
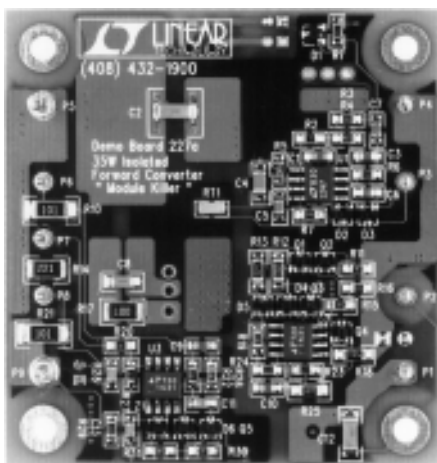
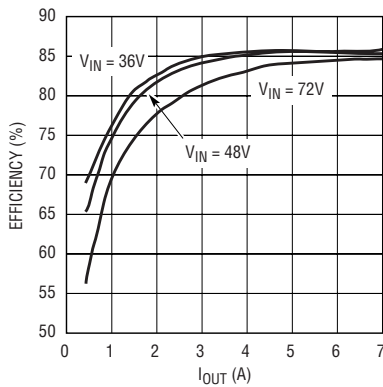


Figure 1. Control (left) and power component (right) views of demonstration circuit DC227, a complete 35W DC/DC converter in a 2.28" by 2.40" footprint

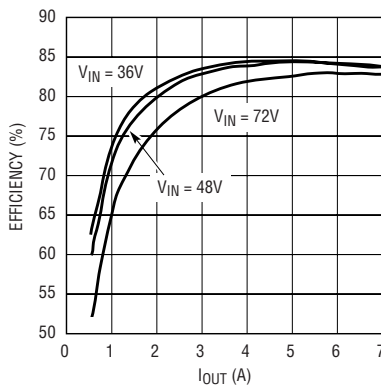


**Figure 2. DC227A-C 5V output efficiency (typical)**

local bypass cap to reduce common mode-induced current.

To achieve fast start-up time, a hysteretic buck regulator is used for the bias supply power. U2, an LT1431 shunt voltage regulator, provides control for this function, with Q1 acting as the switch element; L2 and C21 provide output filtering. Q2 and Q4 protect the circuit during a hot plug, making this a very robust design; it is also impervious to output short circuits. The input surge voltage is limited to 80V by the rating of Q1-Q4.

The main switching power path through T2 comprises L1 and C18 as the input filter, Q6 as the primary switch, D7 as the secondary rectifier and L3 and C14, C16, C17 and C20 as the secondary filter. Transient volt-



**Figure 3. DC227A-A/B 5V output efficiency (typical)**

age suppressor D8 is used to protect Schottky diode D7 during large-signal transient conditions. Power is transferred during the on cycle of Q6 and integrated by the output filter, just as in a buck regulator. The input filter component values for L1 and C18 are optimal and should not be changed without careful evaluation. C19 damps the input filter and will provide adequate stability for large values of input inductance. See LTC Application Note 19 for a discussion of input filter stability analysis.

Output voltage feedback is controlled using U3, another LT1431 shunt voltage regulator, as an error amplifier. In the event of a fault on the output power or sense lines, Z1/Q5 will override U3 and provide overvoltage protection. R10 and R21 are sized to handle any overvoltage condition.

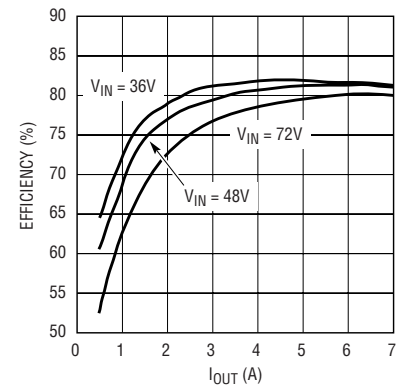
During an output short-circuit condition, the LT1247 is able to decrease the on time of Q6 to less than 200ns. This results in good control of the output short-circuit current, keeping power dissipation to a manageable level.

The demonstration circuit uses surface mount devices for Q6 and D7. For elevated temperature operation at the full rated load, TO-220 devices can be mounted on a standard half-brick heat sink.

For -48V inputs that require hot swap capability, the LT1640H negative voltage HotSwap™ controller provides a seamless interface. Demonstration circuit DC223A-B using the LT1640HCS8 is the recommended solution for use with the DC227A.



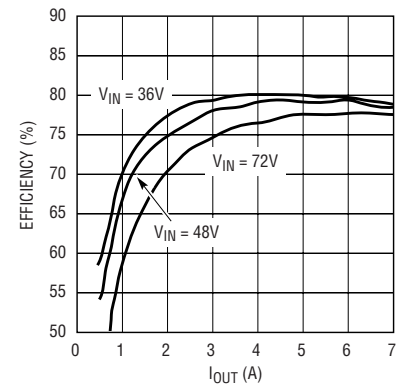
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Articles, Design Ideas, Tips from the Lab...



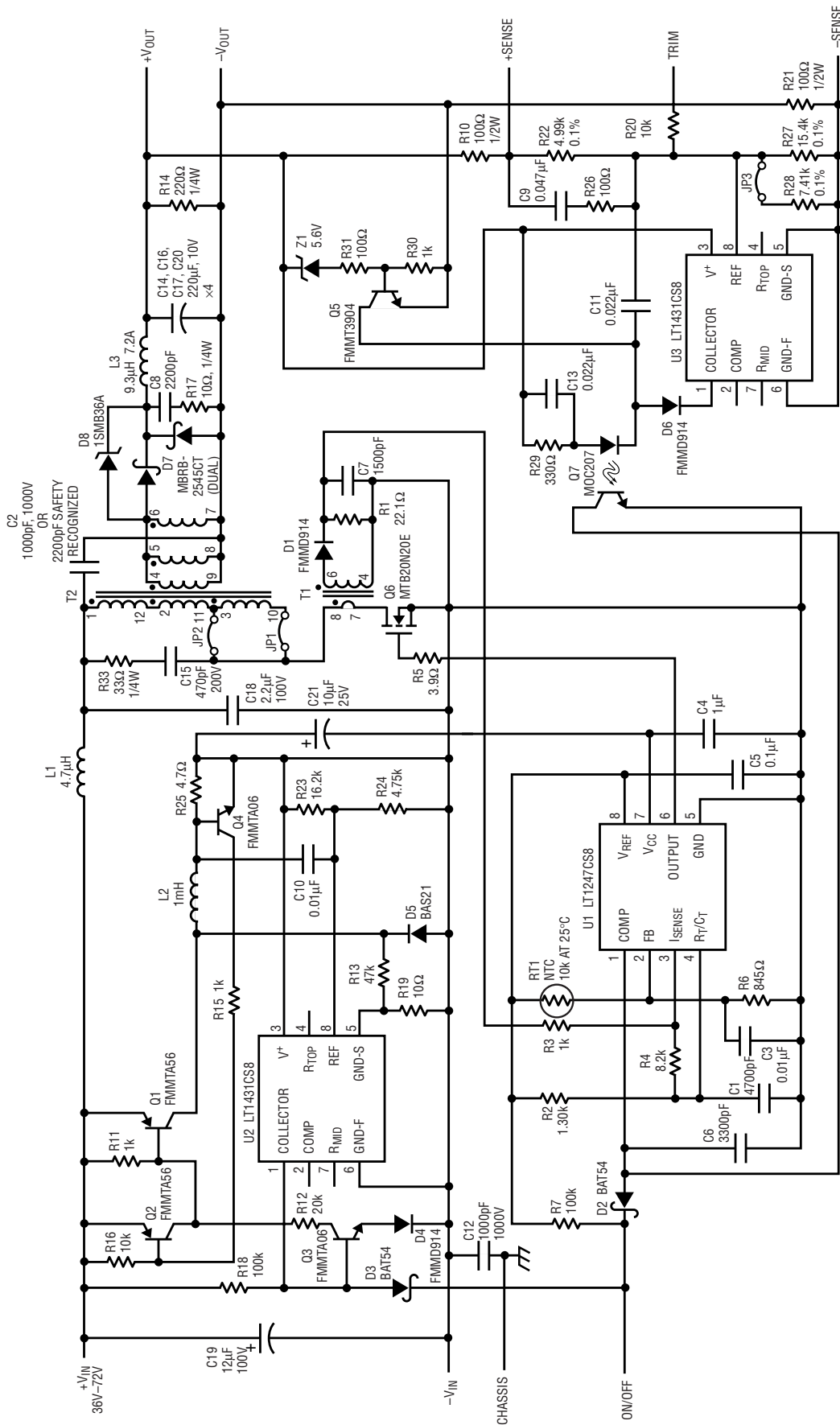
**Figure 4. DC227A-C 3.3V output efficiency (typical)**

## Conclusion

At 35 watts, the topology presented here is one of the most common used by the module manufacturers. This is only one solution for isolated power, and opens up many possibilities for other input and output voltage combinations. For lower power, demonstration circuit DC211 using the LT1425 isolated flyback switching regulator is designed for 10 watts. Demonstration circuit DC259 using the LT1339 adds synchronous rectification, providing a high efficiency solution for 50 watts. See the DC/DC Converter Module section of LTC's Volume 1 1999 New Products Catalog for additional information.



**Figure 5. DC227A-A/B 3.3V output efficiency (typical)**



NOTES:  
 1. CHANGE Z1 TO A 3.3V ZENER FOR 3.3V OUT  
 2. JP3: OPEN = 3.3V; SHORTED = 5V

	T2 AND JUMPER	L3	C2
DC227A-A	500VDC ISOLATION	COLLITRONICS VP5-1200; JP1	VITRIMON VJ1808Y102KXGAT
DC227A-B	1500VDC ISOLATION	COLLITRONICS CTX02-14281-X2; JP2	MURATA GHM3045X7R22K-GC
DC227A-C	500VDC ISOLATION HIGH EFFICIENCY	COLLITRONICS VP5-1200; JP1	VITRIMON VJ1808Y102KXGAT

(847) 639-6400  
 (605) 665-1627  
 (803) 946-0362  
 (804) 239-6841  
 (619) 661-6835

L1: COILCRAFT D01608C-472  
 L2: COILCRAFT D03316P-105  
 T1: DALE LPE-3325-A142  
 C14, C16, C17, C20: AVXTPSE227M010R0100  
 C18: ITW PAKTRON 225100ST3827T  
 C19: SANYO 100MM72GX

Figure 6. 35W isolated DC/DC converter schematic diagram

# Comparator Circuit Provides Automatic Shutdown of the LT1795 High Speed ADSL Power Amplifier

by Tim Regan

## Introduction

Data transmission standards such as HDSL2, ADSL (both Full Rate and G.Lite) and VDSL (collectively known as xDSL) require the combined speed, output power and dynamic range capabilities of the LT1795 to drive the telephone line. In a typical data communications installation, in both the telephone central office and office building sites, hundreds of telephone-wire pairs are brought together into a line multiplexer. These multiplexers compact the individual line driver and receiver circuits to save space. Eight lines per PC card are often implemented. This tight partitioning raises the challenges of both power and heat management in each installation.

A simple comparator circuit can be used to monitor the activity of an individual phone line and completely shut down the line driver when not in use. This provides a means of implementing activity-based power consumption with reduced overall heat generation.

## Controlling Power Dissipation

The LT1795 is a dual, high speed, current feedback amplifier with high output current capability. With a

50MHz Gain Bandwidth product, 900V/ms slew rate and an output stage that can source and sink 500mA, the LT 1795 is ideal for use as a low distortion differential line driver in very high data rate modem applications.

For the LT1795 to obtain full performance, a fair amount of quiescent operating current is required—typically 60mA. When supplied by  $\pm 15V$  power supplies to obtain maximum dynamic output range, the quiescent power dissipation with no load is typically 1.8 watts ( $30V \times 60mA$ ). In these applications, however, full performance is not required at all times or for all applications. To address this issue, the LT1795 provides the ability to completely shut down the driver or to tailor the quiescent operating current to match the actual requirements of a particular application. Figure 1 illustrates how this control is implemented.

Pins 10 and 11 of the LT1795 combine to control the operating current of the amplifiers. If the Shutdown Reference (pin 11) is grounded and the Shutdown input (pin 10) is driven to a voltage greater than two diode drops above ground, both amplifiers are biased to “full speed ahead” with maximum AC performance and also maximum quiescent power dissipation. The current through the two diodes shown in Figure 1 is internally limited to  $300\mu A$  and results in 60mA of quiescent current for the amplifiers. Resistance can be added between the Shutdown Reference pin and ground to limit the maximum operating current of the amplifiers. This programmability can optimize the trade-off between AC performance (primarily slew rate and bandwidth) and quiescent package power dissipation. Many applications require the

peak current drive capability of the LT1795 but do not need the full bandwidth and slew performance. Details of this control can be found on the LT1795 data sheet.

If the Shutdown input is grounded, both amplifiers are disabled and the total quiescent current drain for the package is reduced to only  $200\mu A$ . This feature can be used to save a substantial amount of power when the line drivers are not in use at all times. A simple comparator circuit, as shown in Figure 2, provides a timed, automatic shutdown when no input signals are applied to the amplifiers.

## Timed Automatic Shutdown

In this circuit, an LT1795 is configured as a unity gain differential driver for a  $100\Omega$  transformer-coupled wire pair. The two comparators in an LT1720 package monitor the signals on each of the input lines to the driver amplifier. If the signal on either input exceeds the threshold set by the sensitivity adjustment (shown to provide a range from 65mV to 500mV peak), the output of one or the other of the comparators goes high immediately and puts the LT1795 into action. The LT1720 comparators feature a propagation delay of only 4.5ns, allowing them to respond to input signals well beyond 10MHz in frequency.

The LT1720 outputs are wire-ORed through a simple timing network. When either output is high, timing capacitor  $C_T$  charges quickly to 5V through the diodes shown. While a signal is present, one or the other comparator output keeps  $C_T$  charged. The voltage on the timing capacitor is buffered by a third comparator, an LTC1440, to provide a sharp 0V to 5V shutdown/enable signal to the LT1795.

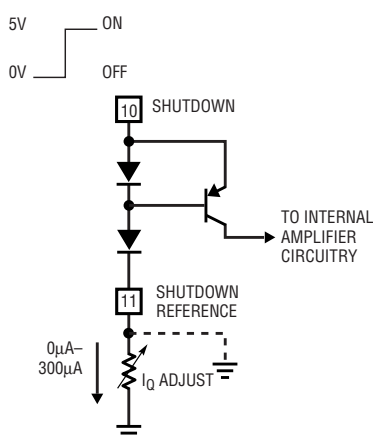
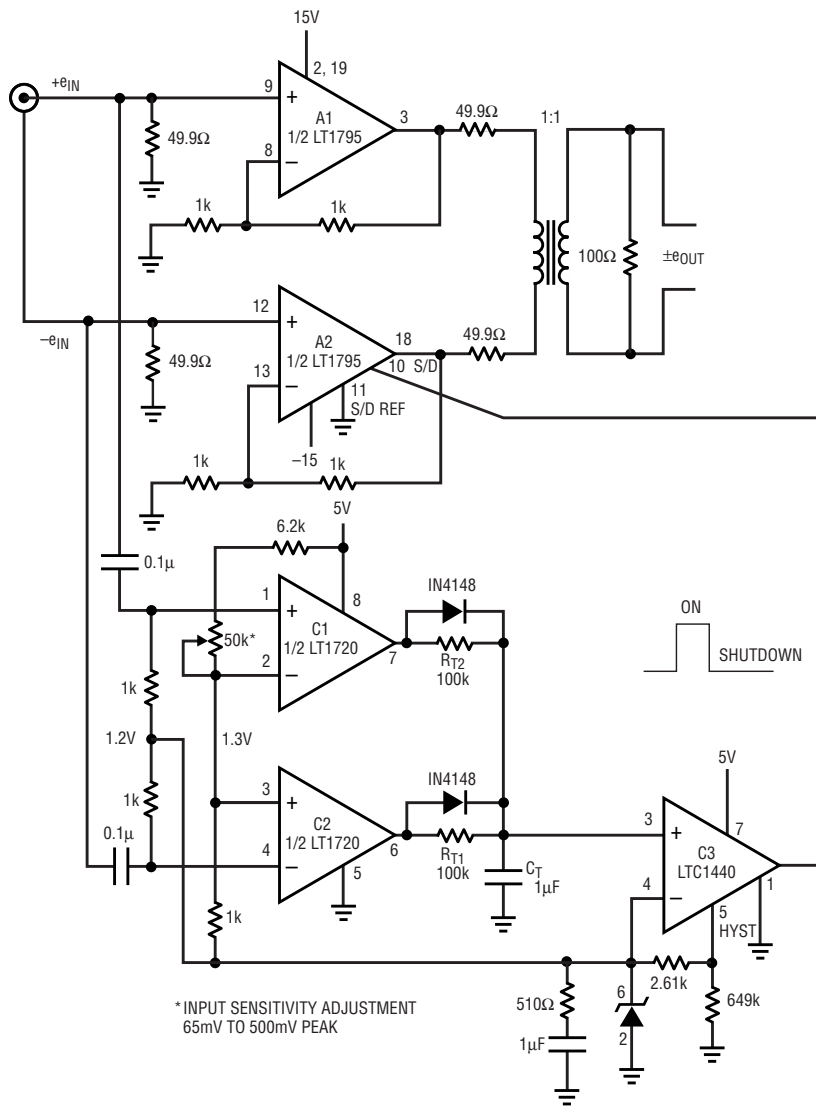


Figure 1. Shutdown/ $I_q$  adjust for the LT1795



**Figure 2. Automatic shutdown of the LT1795 power amplifier**

The LTC1440 is a CMOS comparator with a built-in voltage reference and programmable hysteresis. The 1.18V reference is used for DC biasing to set a stable input threshold for the circuit. The input current of this comparator, 10pA, does not present a significant load on the timing capacitor.


When the signals of both inputs to the driver amplifier drop below the threshold, both outputs of the LT1720 go low. While low, the timing capacitor discharges exponentially through the two 100k resistors,  $R_{T1}$  and  $R_{T2}$ . If the voltage on  $C_T$  is given time to discharge below the 1.18V threshold

of the LTC1440—that is, when no further signals are applied to the input of the circuit—the output of the LTC1440 snaps low and immediately shuts down the LT1795 power amplifier. The discharge voltage of  $C_T$  is fairly slow moving but the input hysteresis set between pins 4 and 5 of the LTC1440 allows a clean ON-to-OFF transition.

With the component values provided in Figure 2, the circuit “wakes up” and properly amplifies the input signal in approximately 50μs. Most of this delay is in the charging of  $C_T$  and the propagation delay of 8μs through the LTC1440. With the Shutdown pin driven high, the LT1795 is fully up to speed in only 1μs. For a DSL application, this wake-up time occurs during the initial “training-up” interval of the data transmission sequence.

When the input signal is removed,  $C_T$  discharges in approximately 65ms to put the driver into a low power dissipation state until the next data transmission. This time-out interval and wake-up time can be easily tailored through the selection of  $C_T$  and  $R_{T1}$  and  $R_{T2}$ . If the two timing resistors are of equal value, the OFF time interval is set by the relationship:  $t = 0.713 \cdot R_{TX} \cdot C_T$ .

### Conclusion

This simple comparator sensing and timing circuit provides automatic control over the power dissipation of a high speed power amplifier line driver. With no signal present there is no wasted quiescent power. This same system enhancement can also be achieved through direct logic control of the LT1795 shutdown feature. This would require a 0V to 5V control signal for each line, which is set or cleared in synchronization with each data transmission interval. 

For more information on parts featured in this issue, see <http://www.linear-tech.com/go/ltmag>


# SMBus Controlled CCFL Power Supply

by Jim Williams

Figure 1 shows a cold cathode fluorescent lamp (CCFL) power supply that is controlled via the popular SMBus interface. The LT1786 CCFL switching regulator receives the SMBus instruction. The IC converts this instruction to a current, which appears at the  $I_{OUT}$  pin. This current, routed to the  $I_{CCFL}$  pin, provides a set point for switching regulator operation. The resultant duty cycle at the  $V_{SW}$  pin pulls current through L2. L2, acting as a switched current sink, drives a resonant Royer converter composed of Q1-Q2, C1 and L1. The high voltage sine wave produced at

L2's secondary drives the floating lamp.

Current flow into the Royer converter is monitored by the IC at pin 13 ("Royer" in Figure 1).<sup>1</sup> Royer current correlates tightly with lamp current, which, in turn, is proportional to intensity. The IC compares the Royer current to the SMBus-derived current, closing a lamp-intensity control loop. The SMBus permits wide-range regulated lamp-intensity control and allows complete IC shutdown. Optimal display and lamp characteristics permit 90% efficiency. The circuit is calibrated by correlating SMBus

instruction codes with attendant RMS lamp current. Detailed information on circuit operation and measurement techniques appears in the references below. 

## References:

1. Williams, Jim. Linear Technology Application Note 65: A Fourth Generation of LCD Backlight Technology. November 1995.
2. LT1786F Data Sheet. Linear Technology Corp. 1998.

<sup>1</sup> Local historians can't be certain, but this may be the only IC pin ever named after a person.

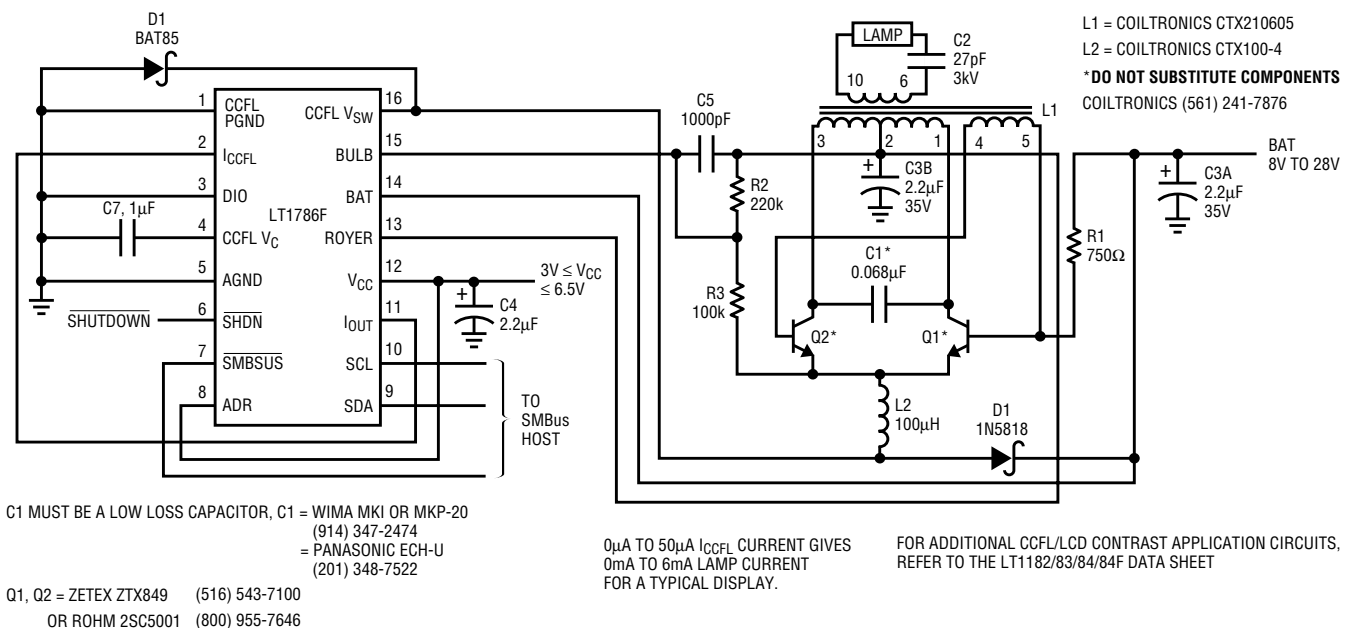


Figure 1. 90% efficient floating CCFL with 2-wire SMBus lamp-current control


Ring Tone, continued from page 29

## Analog Inputs Welcome

The scalable amplification system detailed in Figure 4 can be driven with analog inputs while still maintaining full isolation. Such a system is detailed in Figure 5, where the analog input is filtered (to prevent aliasing) and converted to PWM. Figure 5 goes on to show the use of an

RS485 differential driver to drive a twisted pair line. The receiver end of the twisted pair line is terminated with a resistor and put across the isolation barrier. This provides very good ESD protection on both ends of the line.

## Conclusion

The LT1684 is useful in a wide variety of applications. The LT1684 is a highly integrated solution for use in any system that requires digital control of high output voltage or high output power. 

# Triple Output TFT-LCD Bias Supply Uses All Ceramic Capacitors

by Gary Shockey

Current power supply requirements for TFT-LCD panels call for an 8V or 10V main supply plus two or more auxiliary outputs. The overall layout must be small and meet tight height requirements of under 2mm. Bulky inductors and capacitors must be eliminated if the design is to meet space requirements. The circuit described in this design idea features the new LT1949 and delivers 8V at 200mA from 3.3V while generating auxiliary 24V and -8V outputs capable of 10mA of output current.

The LT1949 is a boost switching regulator that comes in the MSOP-8 package and has an integrated 1.1A NPN power transistor. For this appli-

cation to be small and have a low profile, the boost converter must switch at a high frequency, which allows compact inductors to be used, and must be able to work with ceramic

output capacitors. The LT1949 does both of these things. The switching frequency is fixed at 600kHz and the external compensation pin allows for loop characteristics to be tuned so that tiny ceramic output capacitors can be used.

To understand circuit operation, refer to Figure 1. The LT1949 generates the 8V output in the normal boost mode configuration, while using charge pumps for the 24V and -8V outputs. During boost operation, the SW pin is switching between  $V_{OUT}$  and ground. When at  $V_{OUT}$ , capacitor C6 is charged to  $V_{OUT}$  through D5. When the SW pin flies to ground, C6 holds its charge, causing D6 to be forward biased, charging C5 to -8V. The positive 24V output is developed in a similar fashion except that  $V_{OUT}$  is tripled. Figure 2 details the transient response of  $V_{OUT}$  to an 80mA to 200mA load step.

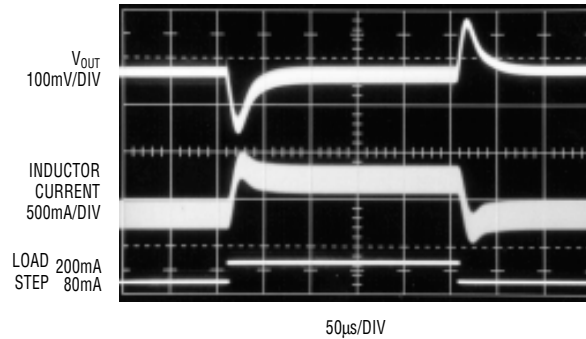


Figure 2. Transient response for an 80mA to 200mA load step

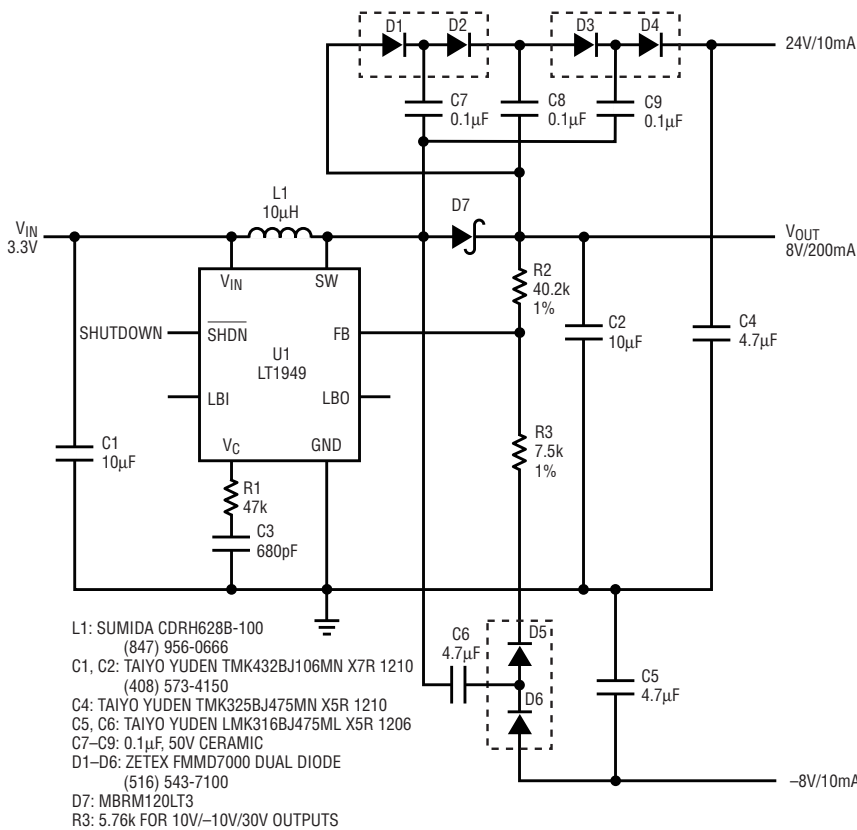


Figure 1. 3.3V to 8V/200mA DC/DC converter with auxiliary 24V and -8V outputs



<http://www.linear-tech.com/ezone/zone.html>  
 Articles, Design Ideas, Tips from the Lab...

# Low Noise Boost VCO Power Supply for Portable Applications

by Ted Henderson

## Introduction

Many portable RF products use voltage controlled oscillators (VCOs) to generate the RF carrier frequency. These applications often require low noise VCO power supply voltages that are greater than the primary battery supply. A DC/DC converter powering a low noise linear regulator is often used. Unfortunately, there are several disadvantages to this solution. The DC/DC converter tends to produce noise that may not be rejected by the regulator, resulting in regulator output noise far greater than the thermal noise levels. The linear regulator may require a large output


compensation capacitor with specific ESR requirements. The board area for both devices and support components can be large. The LTC1682 charge pump DC/DC voltage converter has been designed to minimize these issues. The charge pump and linear regulator have been mutually optimized for minimum regulator output noise. The linear regulator was designed to operate with several different types of output capacitors including small, low value, low ESR ceramic capacitors. The charge pump voltage converter and low dropout linear regulator are combined on one

die and assembled in an extremely small MS8 package. Both fixed and adjustable output voltage versions are available to cover the widest possible output voltage range.

## VCO Power Supply

Figure 1 shows the LTC1682 generating a 4.2V low noise power supply for a 900MHz VCO with an input voltage range of 2.5V to 4.4V. Figure 2 shows the close-in phase noise of the VCO operating open loop and Figure 3 shows the typical peak-to-peak noise voltage at  $V_{OUT}$ .

## Conclusion

The new LTC1682 family of charge pump DC/DC voltage converters represents a complete low noise solution for VCO power supplies. A wide input voltage range of 1.8V to 4.4V, low dropout voltage, low quiescent current, low external parts count and small board area make these devices ideal for portable applications. 

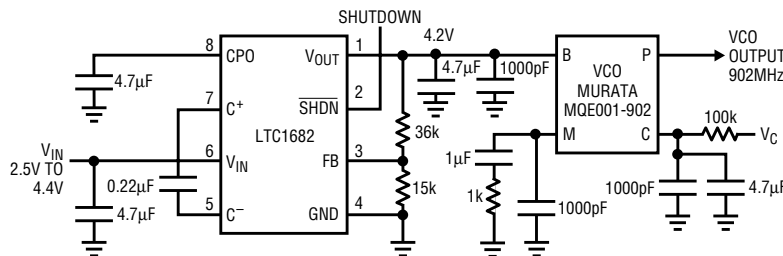


Figure 1. 4.2V VCO power supply

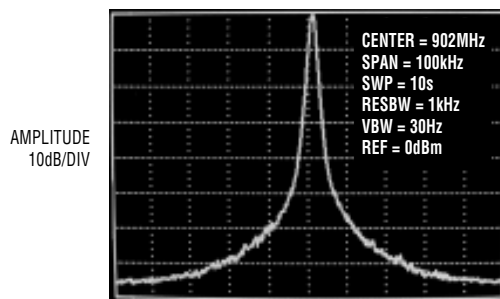


Figure 2. Close-in phase noise

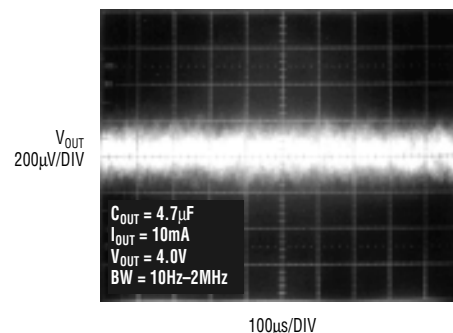


Figure 3. Output noise voltage

For more information on parts featured in this issue, see <http://www.linear-tech.com/go/ltmag>

# New Device Cameos

## **LT1762/LT1763 Low Noise, Micropower, Low Dropout Regulators Save Current in Battery-Powered Applications**

The LT1762 and LT1763 are low noise, low dropout linear regulators. The LT1762 is rated for 150mA of output current, whereas the LT1763 is rated for 500mA. The typical dropout voltage for either regulator at the rated output current is 300mV. The regulators are designed for use in battery-powered systems, with 25 $\mu$ A operating current for the LT1762 and 30 $\mu$ A for the LT1763; both regulators feature a 0.1 $\mu$ A shutdown state. Quiescent current is well controlled for these devices; it does not rise in dropout as is the case with many other regulators.

The LT1762 and LT1763 regulators feature low noise operation. With the addition of an external 0.01 $\mu$ F bypass capacitor, output voltage noise over the 10Hz to 100kHz bandwidth is reduced to 30 $\mu$ V<sub>RMS</sub> for both regulators. Both regulators can operate with small capacitors, as low as 2.2 $\mu$ F for the LT1762 and 4.7 $\mu$ F for the LT1763. Small ceramic capacitors can be used with either device without the need for additional series resistance as is common with other regulators. Internal protection circuitry on both regulators includes reverse-battery protection, current

limiting, thermal limiting and reverse-current protection.

Both regulators are available in fixed output voltages of 2.5V, 3V, 3.3V and 5V, or as an adjustable device with an output voltage range of 1.22V to 20V. The LT1762 regulators are packaged in the 8-lead MSOP package and the LT1763 regulators are available in the 8-lead SO package.

## **LTC2050: Zero-Drift Operational Amplifier in SOT-23**


The LTC2050 is the latest zero-drift operational amplifier in the LTC family. Available in the SOT-23 and SO-8, the LTC2050 permits single-supply operation down to 2.7V. The op amp consumes 800 $\mu$ A of current and the 6-lead SOT-23 and SO-8 packages include a shutdown pin that drops supply current below 10 $\mu$ A. Input common mode range has been extended from the negative rail to within 1V (typical) of the positive rail.

The LTC2050's specifications rival those of the other members of the zero-drift op amp family, with typical offset voltages of 1 $\mu$ V, offset drifts of 10nV/ $^{\circ}$ C, DC to 10Hz noise of 1.5 $\mu$ V<sub>P-P</sub>, and a gain-bandwidth of 3MHz. The LTC2050 uses the industry-standard op amp pinout and requires no external components.

## **LTC1563-2/LTC1563-3: Easy-to-Use 3V, Rail-to-Rail, DC Accurate, Active RC Lowpass Filter Family**

The LTC1563-2 and LTC1563-3 are active RC 4th order lowpass filters suitable for systems with resolution of 16 bits or more. They operate on a supply voltage as ranging from 2.7V to  $\pm$ 5V. They support cutoff frequencies from 5kHz to 256kHz with rail-to-rail input and output. Each part comes in the narrow SSOP-16 package (SO-8 footprint).

The LTC1563-2 and LTC1563-3 are extremely easy to use: unlike conventional 4th order, discrete RC active lowpass filters, which require the calculation of a minimum of six different external resistors and four different external capacitors, they require only six equal-valued resistors to produce a unity gain, 4th order Butterworth (LTC1663-2) or Bessel (LTC1563-3) filter. The calculation of the resistor values is trivial—complex algorithms or filter design software is not needed. By simply allowing the six external resistors to be of different values, gain and other transfer functions (for example, Chebyshev, Gaussian transitional and linear-phase equiripple) are achieved. Cascading two devices forms an 8th order lowpass filter. Simple resistor-value tables make the design of any all-pole lowpass filter elementary. Complicated design algorithms are a thing of the past.

The LTC1563-2 and LTC1563-3 also feature excellent DC offset (typically less than 1mV); they are DC accurate and their broadband noise ranges from 30 $\mu$ V<sub>RMS</sub> to 60 $\mu$ V<sub>RMS</sub> depending on the cutoff frequency. For cutoff frequencies below 25.6kHz, the parts have a low power mode where the supply current is typically 1mA. For higher frequencies, the supply current is 10mA typically. A shutdown mode is also provided to limit the supply current to less than 10 $\mu$ A. 

**For further information on any of the devices mentioned in this issue of *Linear Technology*, use the reader service card or call the LTC literature service number:**

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**Noise Disk** — This IBM-PC (or compatible) program allows the user to calculate circuit noise using LTC op amps, determine the best LTC op amp for a low noise application, display the noise data for LTC op amps, calculate resistor noise and calculate noise using specs for any op amp. Available at no charge

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

**SwitcherCAD™** — The SwitcherCAD program is a powerful PC software tool that aids in the design and optimization of switching regulators. The program can cut days off the design cycle by selecting topologies, calculating operating points and specifying component values and manufacturer's part numbers. 144 page manual included. \$20.00

SwitcherCAD supports the following parts: LT1070 series: LT1070, LT1071, LT1072, LT1074 and LT1076. LT1082. LT1170 series: LT1170, LT1171, LT1172 and LT1176. It also supports: LT1268, LT1269 and LT1507. LT1270 series: LT1270 and LT1271. LT1371 series: LT1371, LT1372, LT1373, LT1375, LT1376 and LT1377.

**Micropower SwitcherCAD™** — The MicropowerSCAD program is a powerful tool for designing DC/DC converters based on Linear Technology's micropower switching regulator ICs. Given basic design parameters, MicropowerSCAD selects a circuit topology and offers you a selection of appropriate Linear Technology switching regulator ICs. MicropowerSCAD also performs circuit simulations to select the other components which surround the DC/DC converter. In the case of a battery supply, MicropowerSCAD can perform a battery life simulation. 44 page manual included. \$20.00

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*continued on page 40*

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