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Versatile Op Amps Need No Resistors

by Glen Brisebois and Jon Munson

Introduction

The LT1990, LT1991 and LT1995 are ready-to-use op amps with their own resistors and internal compensation capacitors. Many difference or instrumentation amps offer precisely matched internal components, but such devices are usually designed to solve a specific application problem, and thus have limited versatility. Not the LT1990, LT1991, and LT1995. These are flexible parts that can be configured into inverting, non-inverting, difference amplifiers, and even buffered attenuators, just by strapping their pins (Figure 1).

The internal precisely matched resistors and capacitors make it possible to configure these op amps into hundreds of different application circuits without external components. Simply hook them up for type and gain and move on. By reducing the external components in your design, you simplify inventory, reduce pick and place costs, and make for easy probing.

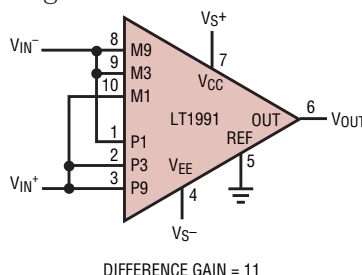


Figure 1. What could be easier? A precision difference amplifier; no resistors in sight. This is only one of hundreds of possible configurations. See Figure 3 for a few more.

What Do You Need: High Precision, High Input Voltage or High Speed?

Figure 2 shows simplified schematics of the three new amplifiers and in general, their comparative performance. The LT1990 is optimized for supporting high input common mode voltages of up to $\pm 250V$. The LT1991 is

**The LT1990, LT1991 and
LT1995 are ready-to-use
op amps with their own
resistors and internal
compensation capacitors.
Just wire them up.**

optimized for gain flexibility and overall precision, and supports common mode ranges up to $\pm 60V$. The LT1995 is designed for high speed applications up to 30MHz.

The LT1991 for the Greatest Flexibility and Precision

The LT1991 is the most flexible and most precise of the three new devices. Its internal resistors guarantee 0.04% ratio-matching and 3ppm/ $^{\circ}C$ MAX matching temperature coefficient. The op amp offers 15 μV typical input offset voltage and 50pA of input offset current. The LT1991 operates on supplies from 2.7V to 36V with rail to rail outputs, and remains stable while driving capacitive loads up to 500pF. Gain bandwidth product is

continued on page 3

Issue Highlights

Our cover article presents three devices that simplify amplifier designs by removing the external resistors: the LT1990, LT1991, and LT1995. Each is a fully integrated functional building block—they may be the last amplifiers you ever have to stock—and because they require *no external resistors* for precision operation, design and testing is a snap. Simply hook them up for type and gain and move on.

Featured Devices

Below is a summary of the other devices featured in this issue.

High Voltage to Low Voltage Converter for High Power CPUs

The LTC3802 is designed to excel in generating **low output voltages from high input voltages**, a common problem for the power supplies of fast CPUs. It is the latest in Linear Technology's family of high speed, voltage feedback, synchronous step-down regulator controllers. It retains the constant frequency architecture and Burst Mode® operation of the LTC1702A, while improving on its performance and adding features. (Page 5)

Switching Controller with SSFM Reduces EMI

One way to knock down the amplitude of noise components of a switching regulator is to spread the operating frequency around. If the frequency of the switcher is modulated using spread spectrum frequency modulation, the energy of the EMI is spread over many frequencies, instead of concentrated at one frequency and its harmonics, thus reducing the peak noise at any given frequency. The LTC3736-1 integrates an **SSFM oscillator with a dual synchronous switching regulator controller** to randomly modulate its clock frequency. (Page 9)

Superfast Video Amplifiers for High Resolution Video

The LT6553 and LT6554 **triple video amplifiers** offer 600MHz performance

in a compact package, requiring no external gain-setting resistors to establish gain of 2 or unity-gain, respectively.

The superfast performance of the LT6553 and LT6554 satisfies the demands of the latest high resolution video equipment. (Page 12)

Sequencing and Tracking for LDOs

The LTC2923 **power supply tracking controller** is specifically designed to work with switching power supplies but it is easily adapted to linear regulators, including popular low-dropout (LDO) types. Summarized here are several techniques for controlling linear regulators with the LTC2923. (Page 14)

Space and Power Saving Low Voltage, Adjustable Reference

The LT6650 is a **0.4V to 18V adjustable voltage reference** that runs from low voltage and consumes only a few μ A. It features a low-dropout (LDO) characteristic, can source or sink current, can be configured in either series or shunt mode and saves space in the tiny 5-lead ThinSOT-23 package. (Page 16)

Intelligent Hot Swap™ Controller with Onboard ADC

The LTC4260 combines a wide input range **Hot Swap controller, ADC voltage monitor and I2C serial communication** in one device. The LTC4260 provides the means for quantitatively measuring the board current and voltages with an on-board ADC and multiplexer. It reports this information using the I²C serial communication bus when polled by a host processor. (Page 18)

Design Ideas and Cameos


Starting on page 22 are six new Design Ideas, including a simple way to implement a redundant 2-wire bus system; a -48V backplane impedance analyzer for clamps and snubbers; a compact solution for powering TFT-LCD panels,

LTC in the News...


On October 12, Linear Technology Corporation announced its financial results for its first quarter of fiscal year 2005, ending September 26, 2004. According to Robert H. Swanson, Chairman of the Board and CEO, "Sales grew 6% sequentially from the June quarter and we continued to be cash flow positive and strongly profitable as evidenced by our 41% return on sales."

Net sales for the quarter were \$253,028,000, an increase of 45% over net sales of \$174,077,000 for the first quarter of the previous year. The Company also reported net income for the quarter of \$103,476,000 or \$0.33 diluted earnings per share, an increase of 49% from the \$69,471,000 diluted earnings per share reported for the first quarter last year. During the quarter, the Company's cash and short-term investments increased by \$48.1 million, net of spending \$54.6 million to purchase 1,500,000 shares of common stock. A cash dividend of \$0.08 will be paid on November 10, 2004 to stockholders of record on October 22, 2004.

Electronica Show

Once again, Linear Technology had a presence at Electronica, the major European electronics show, held in Munich, Germany, November 9-12, 2004. At Electronica, Linear showcased its broad range of high performance analog solutions, including power management, data conversion, signal conditioning, and RF. 

including the LED backlight; and several compact power supply designs.

At the back are four New Device Cameos, including one about BodeCAD—software that greatly simplifies large AC signal analysis in power systems. Visit www.linear.com for complete device specifications and applications information. 

LT1990, LT1991 and LT1995, continued from page 1

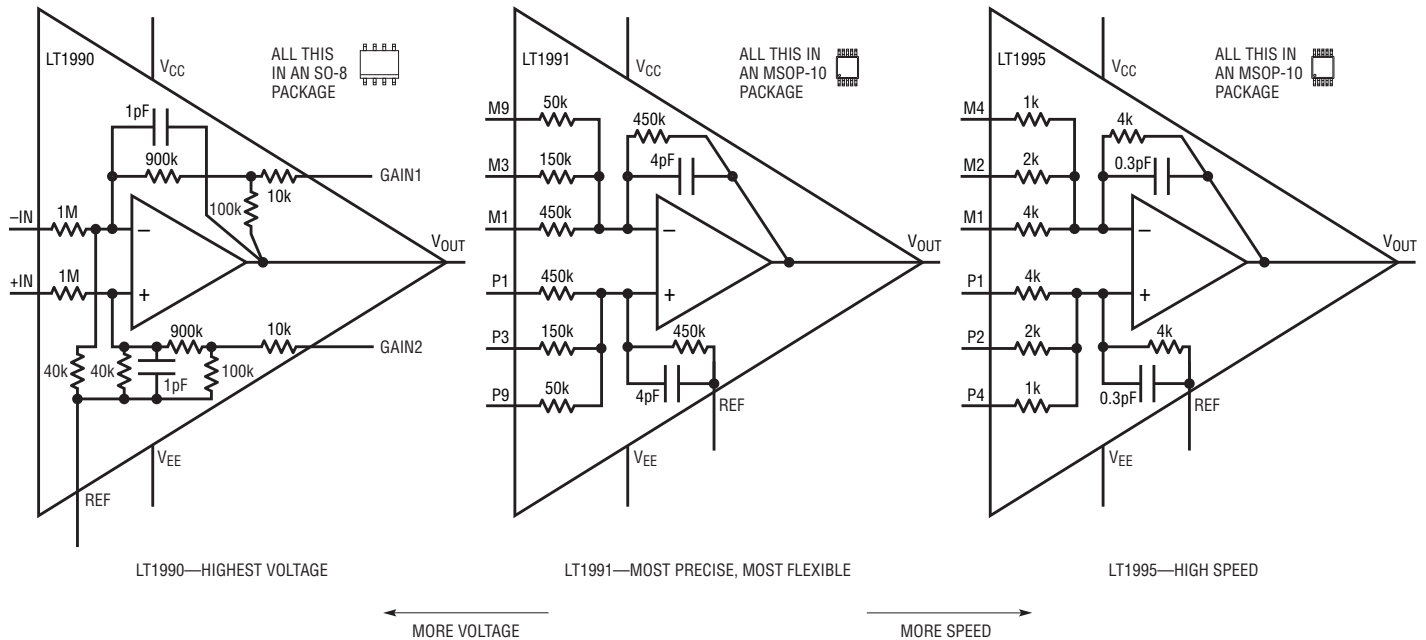


Figure 2. The LT1990, LT1991, and LT1995 are ready-to-use op amps with their own resistors and internal compensation capacitors. Just wire them up.

560kHz, while drawing only 100 μ A supply current.

The resistors are nominally 50k, 150k, and 450k. One end of each resistor is connected to an op amp input, and the other is brought out to a pin. The pins are named “M” or “P” depending on whether its resistor goes to the “minus” or “plus” input, and numbered “M1” M3” or “P9” etcetera

according to the relative admittance of the resistor. So the “P9” pin has 9 times the admittance (or force) of the “P1” pin. The 450k resistors connected to the M1 and P1 inputs are not diode clamped, and can be taken well outside the supply rails, ± 60 V maximum.

To use the LT1991, simply drive, ground or float the P, M, and REF inputs to set the configuration and gain.

There is a whole series of high input common mode voltage circuits that can be created simply by just strapping the pins. Figure 3 demonstrates the flexibility of the LT1991 with just a few examples of different configurations and gains. In fact, there are over 300 unique achievable gains in the non-inverting configuration alone. Gains

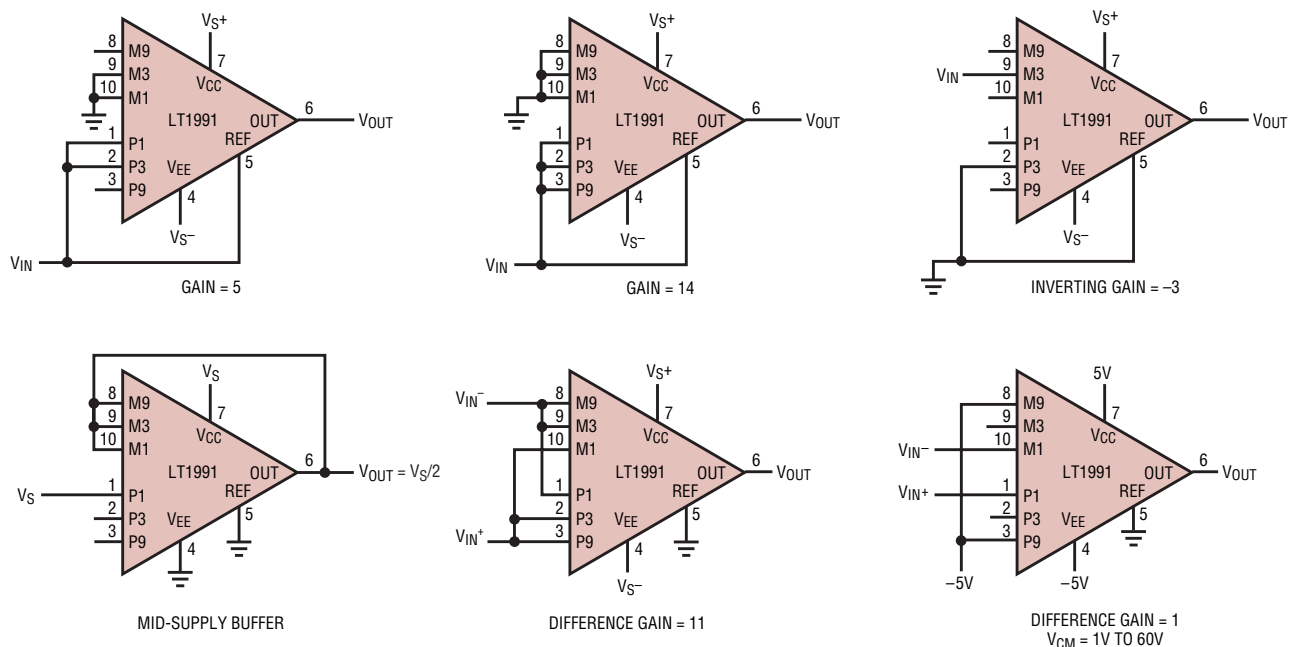


Figure 3. Non-inverting, inverting, difference amplifiers, and buffered attenuators achieved simply by connecting pins. This illustration shows only a small sample of the op amp configuration circuits possible with the LT1991, all without requiring external resistors.

of up to 14 and buffered attenuations down to 0.07 are possible.

The LT1990 for High Input Voltages

The LT1990 has internal components with similar precision to the LT1991, but it is configured for high input voltages, up to $\pm 250\text{V}$. The high input voltage capability is achieved by the $1\text{M}\Omega:40\text{k}\Omega$ attenuation at the inputs, and by careful internal layout and shielding. The LT1990 has two effective gain settings, 1 and 10. A gain of 1 is set by floating the Gain1 and Gain2 pins, and a gain of 10 is set by shorting the Gain1 and Gain2 pins. Bandwidth is 100kHz in a gain of 1, and 6.5kHz in a gain of 10. The op amp operates on supplies from 2.7V to 36V with rail to rail outputs, on $105\mu\text{A}$ supply current. Like the LT1991, it remains stable while driving capacitive loads up to 500pF .

The LT1995 for High Speed

The LT1995 offers high speed with 30MHz bandwidth, 24MHz full power bandwidth and $1000\text{V}/\mu\text{s}$ slew rate. It works on supplies from $\pm 2.5\text{V}$ to $\pm 15\text{V}$ drawing 7mA supply current. Accuracy is unusually good for a high speed amplifier, with input offset typically $600\mu\text{V}$ and guaranteed better than 4mV . It is pinned out identically to the LT1991, but with different resistor ratios and values. The resistors are lower impedance (1k , 2k , and 4k) than those in the LT1991 and LT1990 to support this device's higher speed. They are as high a quality as you should ever need in a high speed application, guaranteeing 0.25% matching, worst case over temperature. As with the LT1991, the

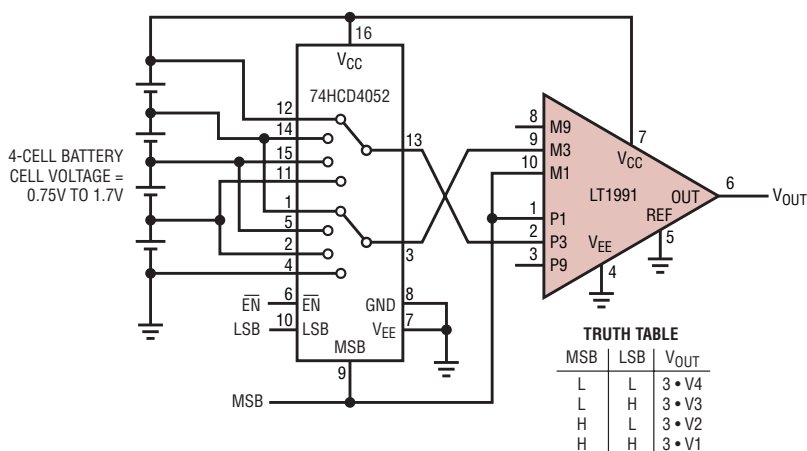


Figure 4. LT1991 applied as an individual battery cell monitor for a 4-cell battery

LT1995 takes its pin names from the relative admittances of their resistors and the amplifier input polarity: hence "M1", "M2", "P4", etcetera. For a difference gain of 6, short the M2 and M4 pins, and short the P2 and P4 pins ($2 + 4 = 6$). In this example, the difference amplifier is formed by the minus input of the shorted M2 and M4 inputs, and the plus input of the shorted P2 and P4 inputs.

Applications

Battery Monitor

Many batteries are composed of individual cells with working voltages of about 1.2V each, as for example NiMH and NiCd. Higher total voltages are achieved by placing these in series. The reliability of the entire battery pack is limited by the weakest cell, so battery designers often like to maintain data on individual cell charge characteristics and histories.

Figure 4 shows the LT1991 configured as a difference amplifier in a gain of 3, applied across the individual

cells of a battery through a dual 4:1 mux. Because of the high valued $150\text{k}\Omega$ resistors on its M3 and P3 inputs, the error introduced by the multiplexer switch ON-resistance is negligible. As the mux is stepped through its addresses, the LT1991 takes each cell voltage, multiplies it by 3, and references it to ground for easiest measurement. Note that worst case combinations of very different cell voltages can cause the LT1991 output to clip. Connecting the MSB line to the M1 and P1 inputs helps reduce the effect of the wide input common mode fluctuations from cell to cell. The low supply current of the LT1991 makes it particularly suited to battery powered applications. Its $110\mu\text{A}$ maximum supply current specification is about the same as that of the CMOS mux!

Single-Supply Video Driver

Most op amps operating from a single supply voltage require several high-quality external resistors to generate a local bias voltage—to optimize the DC

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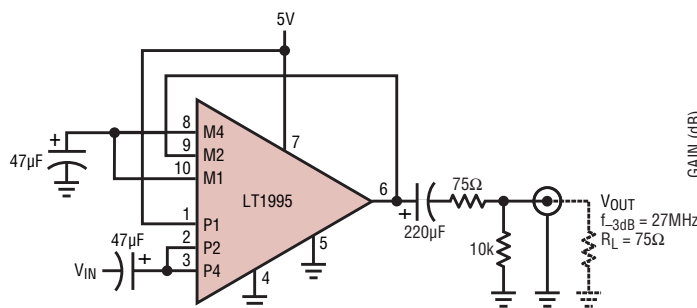
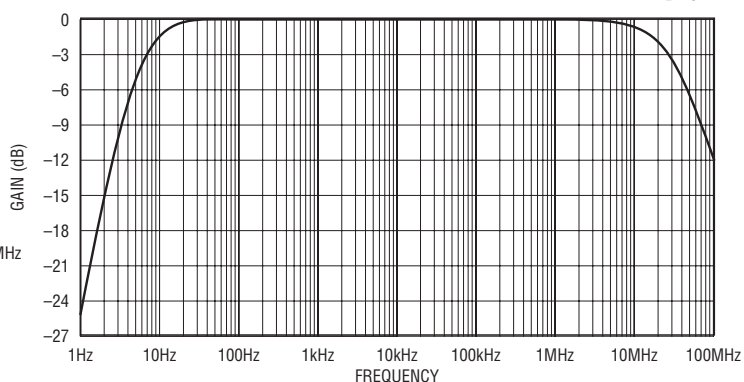


Figure 5. This single-supply composite video output-port driver requires no DC-biasing or gain-setting resistors



30V, Dual Output Regulator Controller is Efficient, Rich in Features, and Saves Space

by Teo Yang Long and Theo Phillips

Introduction

The LTC3802 is designed to excel in generating low output voltages from high input voltages, a common problem for the power supplies of fast CPUs. It is the latest in Linear Technology's family of high speed, voltage feedback, synchronous step-down regulator controllers. It retains the constant frequency architecture and Burst Mode® operation of the LTC1702A, while improving on its performance and adding features (see Table 1).

The input supply operating range is extended from a nominal 5V to the entire 3V–30V range. The internal reference voltage has decreased, allowing the output to go as low as 0.6V. An advanced modulation scheme facilitates these low duty cycles and fast switching frequencies. The two channels are still run 180° out of phase—effectively doubling the frequency of the switching pulses seen by the input bypass capacitor and thereby lowering its RMS current and reducing its required value—but a new PLLIN pin extends these benefits by allowing two LTC3802s to control a

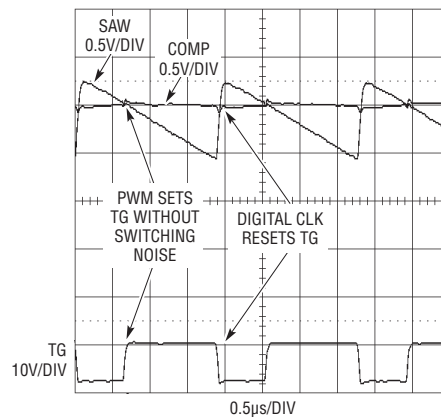


Figure 1. Leading edge modulation architecture PWM switching waveform for $V_{IN} = 5V$, $V_{OUT} = 3.3V$

4-phase converter. This pin also allows external synchronization of the switching frequency from 330kHz–750kHz, rather than a fixed 550kHz. Output voltage tracking governs the 2 channels' output slew rate during power up and power down, to comply with various power sequencing requirements.

Leading Edge Modulation

The LTC3802 uses a high switching frequency and precision voltage

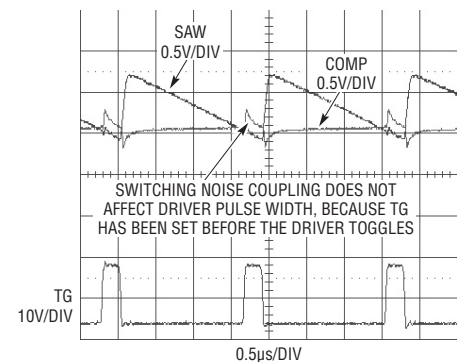


Figure 2. In a 20V to 3.3V buck converter, switching noise couples to the error amplifier output after the top gate (TG) turns on; this would cause unpredictable switching in traditional PWM converters.

feedback architecture to provide exceptional regulation and transient response performance at each of its two outputs. The 10MHz gain-bandwidth feedback op-amps permit loop crossover in excess of one-tenth the switching frequency, whether that frequency is externally synchronized or running at the default 550kHz. Large integrated gate drivers allow the LTC3802 to control multiple MOSFETs efficiently throughout its range of switching frequencies.

A typical LTC3802 application down converts a high input voltage source to two low output voltage supplies and requires the two channels to run at low duty cycles. Such an application presents several challenges to a traditional PWM controller. First, the controller is forced to make a decision about pulse width after the control switch (top MOSFET) turns on. The turn-on of the control switch in the buck converter is the noisiest event in the whole switching cycle. The input supply current jumps from zero current to the loaded current, causing ground bounce; the large voltage swing at the inductor flying node can further induce noise in the controller. Either event can disrupt

Table 1. Comparison of the LTC3802 and the LTC1702/LTC1702A

	LTC3802	LTC1702/LTC1702A
V_{IN}	3V–30V	3V–7V
Switching Architecture	Leading Edge Modulation with Line Feedforward Compensation	Trailing Edge Modulation
Reference	0.6V \pm 1%	0.8V \pm 1%
Phase Lock Loop	330kHz–750kHz PLL Free Run at 550kHz	No Free Run at 550kHz
Tracking	Ratiometric or Coincident Power Up and Power Down Tracking	No
Packages	GN28 and QFN 32	GN24

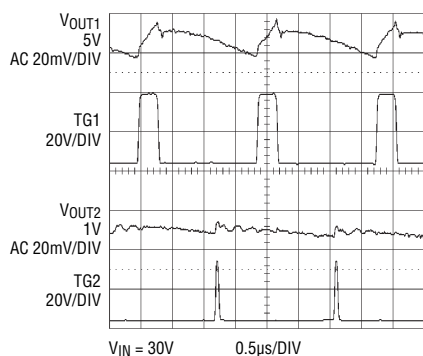


Figure 3. Switching waveform obtained from the LTC3802 dual out of phase buck converter

the operation of the PWM comparator within the first 100ns–200ns after the transition, producing random control pulse width variations and irregular inductor current ripple.

The second challenge to the traditional PWM operating scheme is that the PWM comparator response time limits the controller's minimum pulse width. A typical PWM comparator takes at least 100ns to toggle the output. This sets the minimum top gate on-time for the switcher. Third, traditional trailing edge modulation suffers from slow transient recovery. The internal

clock turns on the control switch at a fixed time interval regardless of output voltage (V_{OUT}). If the load current jumps up after the top gate turns off, the controller must wait for the next clock cycle to charge up the output capacitor. In this situation, controllers with slower switching frequencies can have larger output droops.

The LTC3802 uses a leading edge modulation architecture to overcome these three obstacles. In a typical LTC3802 switching cycle, the PWM comparator turns on the top MOSFET; the internal master clock turns it off. The comparator makes a decision in a quiet interval before the MOSFETs toggle, avoiding pulse width jitter.

Figure 1 shows the leading edge modulation architecture PWM switching waveform. Figure 2 shows the noise at the error amplifier output due to relatively high input supply voltage—even with this noise, the LTC3802 maintains a stable switching waveform. At even lower duty cycles, the comparator's propagation delay no longer limits the minimum pulse width of the top gate; the switching

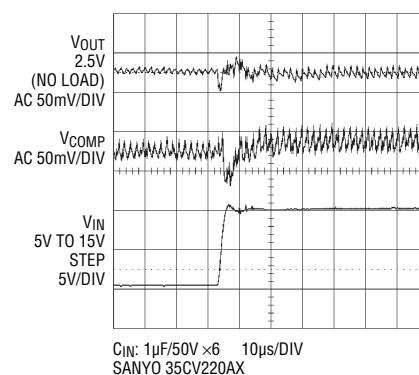


Figure 4. A large swing in V_{IN} produces a very small disturbance at V_{OUT} .

feedback loop adjusts the duty cycle to give the correct output voltage. Figure 3 shows the narrow TG pulse generated from a 30V to 1V buck converter. With a 550kHz switching frequency converter, the TG pulse width is only 60ns! The comparators in traditional PWM converters are not sensitive enough to permit such a narrow pulse width; otherwise they would be easily triggered by noise.

Leading edge modulation also yields fast load transient response. Once the output is loaded, the error

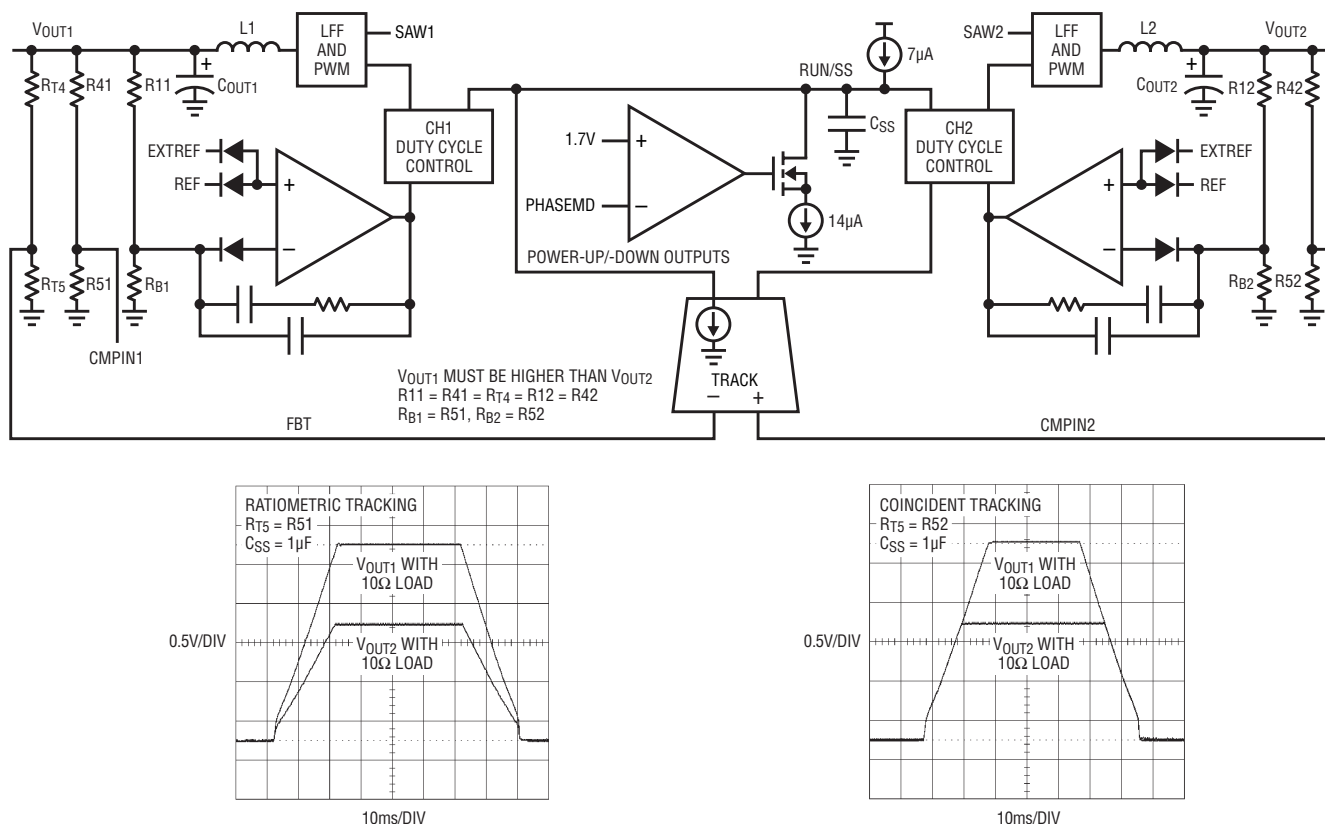


Figure 5. Simplified tracking schematic and associated power-up and power-down waveforms for ratiometric and coincident tracking

amplifier senses the output droop, and the controller immediately turns on the top MOSFET to replenish the output capacitor. The LTC3802 does not need to wait for the next clock cycle to enable the top gate. When the load is removed, the undershoot recovery time is determined by the error amplifier frequency compensation network. In either case, recovery times of well under 20 μ s are easily attained at a switching frequency of 550kHz. This fast transient response, combined with the low output ripple current produced at high switching frequencies, reduces the amount of output capacitance required to support the output voltage during a load transient.

The LTC3802 includes compensation for line transients. The line feedforward compensation input monitors the power supply (V_{IN}), immediately modulating the input to the PWM comparator and changing the pulse width in an inversely proportional manner. Instead of waiting for a droop in output voltage, feedforward compensation bypasses the feedback loop and provides excellent regulation during line transients (Figure 4).

Programmable Power Up, Power Down Tracking

Next generation power modules use power up, power down tracking to reduce the amount of external circuitry required to power up modern digital semiconductors, such as DSPs, microprocessors, FPGAs and ASICs. Such devices require at least two supply voltages, one to power the high speed core logic and another to power the I/O interface. These voltages must be applied in a well-controlled sequence.

During power-up and power-down, variations in the starting points and ramp rates of the supplies may cause current to flow between the isolation structures. When prolonged and excessive, these currents can shorten the life of the semiconductor devices, or trigger latch-up leading to device failure.

To meet these sequencing requirements, power system designers can

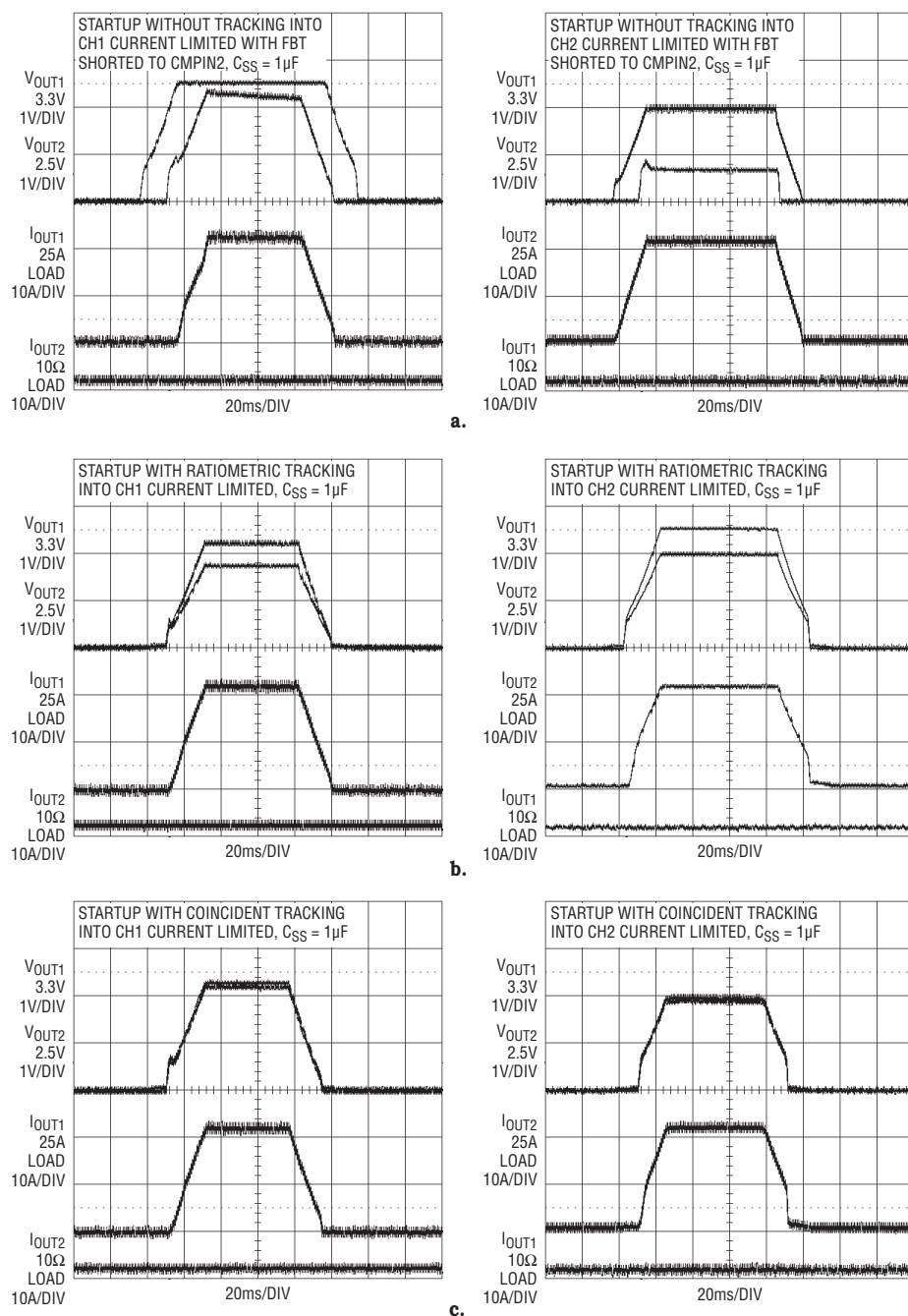


Figure 6. Power up and power down waveforms with one of the channels current limited. Results are shown without tracking (a), with ratiometric tracking (b), and with coincident tracking (c).

avoid adding extra circuitry by using the LTC3802's easily programmable power up, power down tracking. The LTC3802 can adhere to two different schemes: ratiometric and coincident tracking.

With a ratiometric configuration, the LTC3802 produces two different output slew rates (with $V_{OUT1} > V_{OUT2}$). Because each channel's slew rate is proportional to its corresponding output voltage, the two outputs

simultaneously reach their steady-state values.

The coincident configuration produces the same slew rate at both outputs, so that the channel with the lower V_{OUT} reaches its steady state value first.

Figure 5 shows the simplified schematic of how tracking is implemented. During power up or power down, the tracking amplifier, TRACK, servos the tracking feedback loop and forces

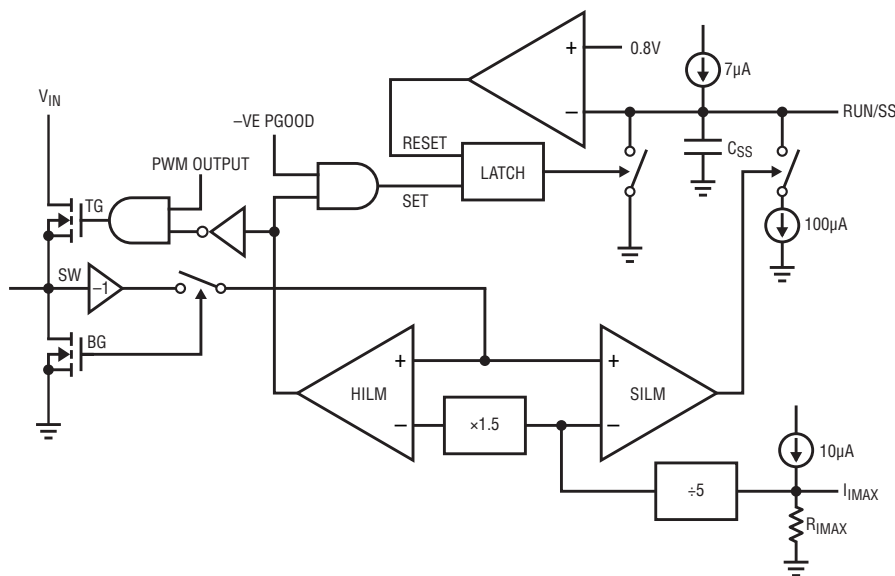


Figure 7. Simplified LTC3802 current limit circuitry

FBT to be at the same potential as CMPIN2. Setting $R_{T5} = R_{51}$ creates the ratiometric startup, and setting $R_{T5} = R_{52}$ produces the coincident start-up. The tracking function can be easily disabled by disconnecting the FBT resistive divider and shorting FBT to CMPIN2.

To have the proper power-down sequence, ground the PHASEMD pin. This turns on an internal current source that slowly discharges the soft-start capacitor. Once the RUN/SS potential is low enough to control the duty cycle, the tracking amplifier takes control and servos the tracking feedback loop to produce the selected output ramp. Note that in this tracking scheme, there is no master and slave assignment; if either output goes low, the other channel's output follows. Figure 5 includes the ratiometric and coincident tracking waveforms with 10Ω loads.

Figures 6a to 6c show the power up and power down waveforms with one of the channels current limited. Figure 6a shows that when FBT is shorted to CMPIN2, the tracking function is disabled. The first waveform shows that when channel 1 is current limited, channel 2's output potential is lowered due to the lower RUN/SS voltage (both channels share the same RUN/SS pin). The second photo shows that when channel 2 is current limited, channel 1's 3.3V output voltage is lower than

nominal. Figures 6b and 6c show the output waveforms with ratiometric and coincident tracking. Figure 6b shows that for ratiometric tracking, if either output is current limited, the other output is pulled low such that both outputs maintain their voltage ratio. On the other hand, for the coincident Tracking configuration shown in Figure 6c, both channels have the same output voltages even if only one channel is current limited.

Current Limit

The LTC3802 bottom MOSFET current sensing architecture not only eliminates the external current sense resistors and the corresponding power losses in the high current paths, but also allows a wide range of output

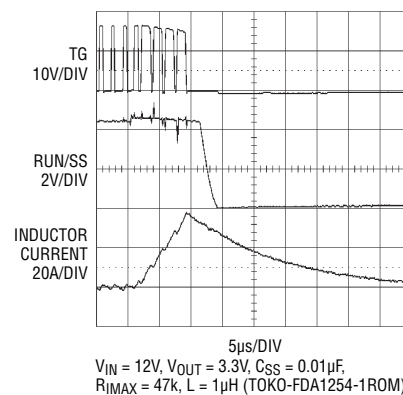


Figure 8. LTC3802 short circuit waveform

voltages, even at extremely low duty cycles.

The LTC3802's current limit scheme improves on that of the LTC1702A by employing a user-programmable current limit level. It works by sensing the V_{DS} drop across the bottom MOSFET when it is on and comparing that voltage to a programmed voltage at I_{MAX} .

The I_{MAX} pin includes a trimmed 10µA current, enabling the user to set the I_{MAX} voltage with a single resistor, R_{IMAX} , to ground. The current comparator reference input is equal to V_{IMAX} divided by 5 (see Figure 7). The current comparator begins limiting the output current when the voltage across the bottom MOSFET is larger than its reference. The current limit detector is connected to an internal 100µA current source.

Once current limit occurs, this current source begins to discharge the soft-start capacitor at RUN/SS,

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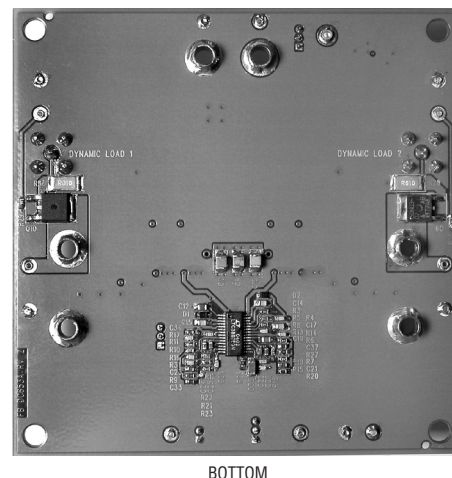
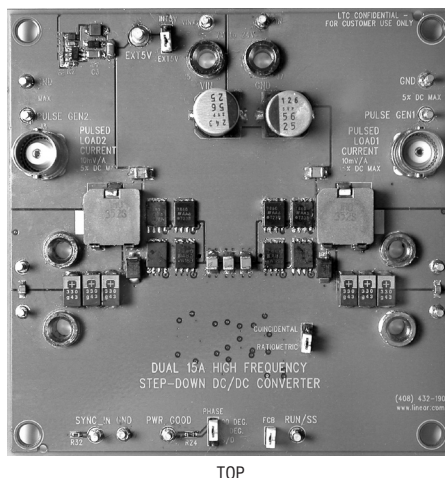


Figure 9. An 87W, LTC3802 application circuit occupies less than 6in²

Dual Switcher with Spread Spectrum Reduces EMI

by Jason Leonard

Introduction

Switching DC/DC power supplies are increasingly popular in modern electronic devices because of their high efficiency, which reduces heat dissipation and increases battery run time. Nevertheless, the rapid switching of current makes them a potential source of radiated and conducted electromagnetic interference (EMI). EMI can cause a variety of problems, from the relatively benign addition of noise to a television picture or radio receiver to the more serious impairment of the operation of electronic devices in critical applications.

Unfortunately, the amount of EMI generated, and whether it will produce significant interference, is not easily quantifiable and is often not known until the late stages of the development. Therefore, it is wise to proactively minimize the potential sources of EMI to save troubleshooting time later on. There are many

techniques to significantly reduce EMI, but few are as simple as using Spread Spectrum Frequency Modulation (SSFM) in the clocking of a switching power supply.

Switching regulators operate on a cycle-by-cycle basis to transfer power to an output. In most cases, the frequency of operation is either fixed or is a constant based on the output load. This method of conversion creates high amplitude noise components at the

frequency of operation (fundamental) and at the multiples of the operating frequency (harmonics).

One way to knock down the amplitude of the fundamental and harmonic noise components is to spread the operating frequency around. If the frequency of the switcher is modulated using spread spectrum frequency modulation, the energy of the EMI is spread over many frequencies, instead of concentrated at one frequency and

Table 1. LTC3736-1's Switching Frequency

SSDIS pin	FREQ pin	Switching Frequency
GND	Filter Capacitor (e.g., 2200pF)	Spread Spectrum (450kHz to 580kHz)
V _{IN}	Floating	Constant 550kHz
V _{IN}	V _{IN}	Constant 750kHz
V _{IN}	GND	Constant 300kHz

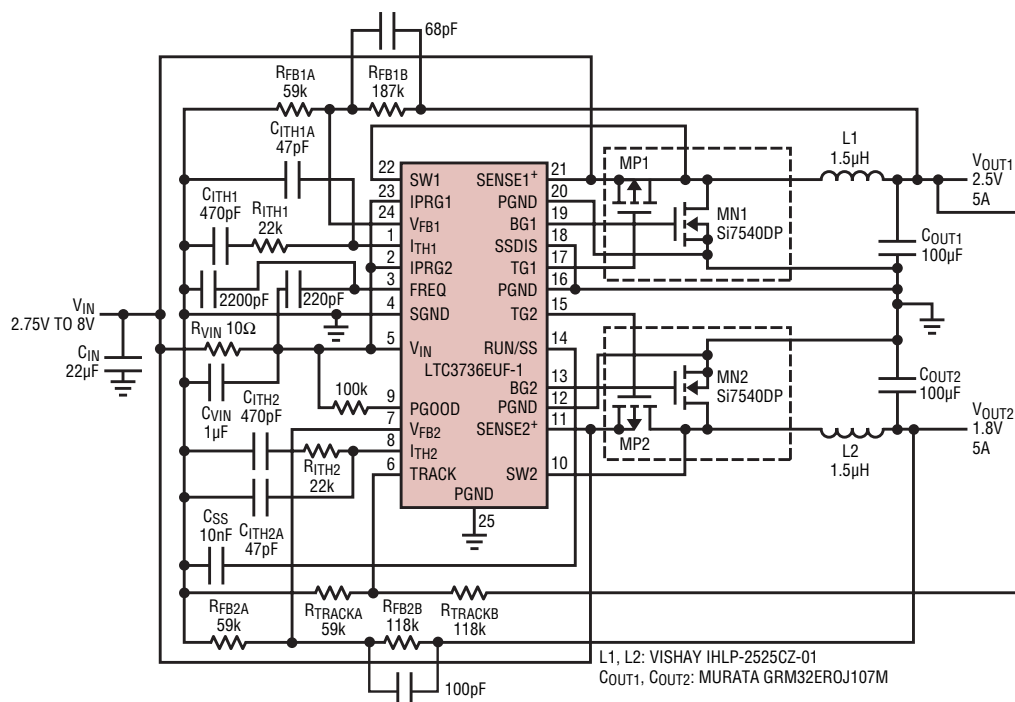


Figure 1. 3.3V to 2.5V and 1.8V dual DC/DC converter with spread spectrum frequency modulation (SSFM). The circuit uses only ceramic capacitors and requires no current sense resistors or Schottky diodes.

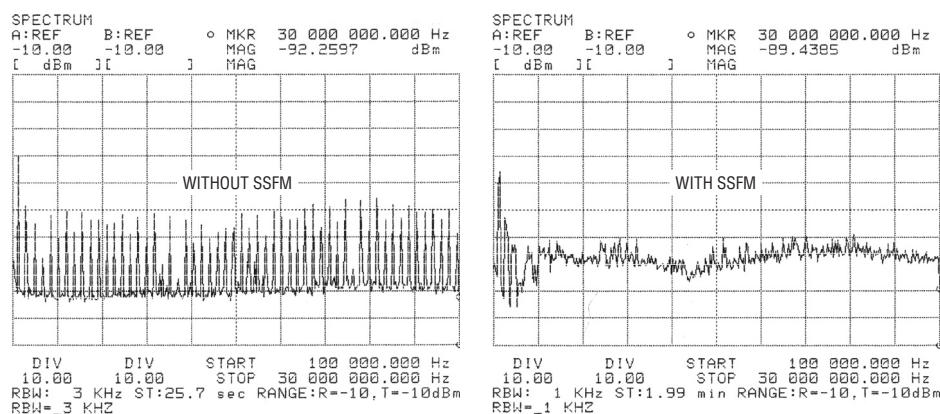


Figure 2. Output voltage spectra for circuit of Figure 1 with and without SSFM enabled. Notice the diminished harmonic peaks with SSFM enabled.

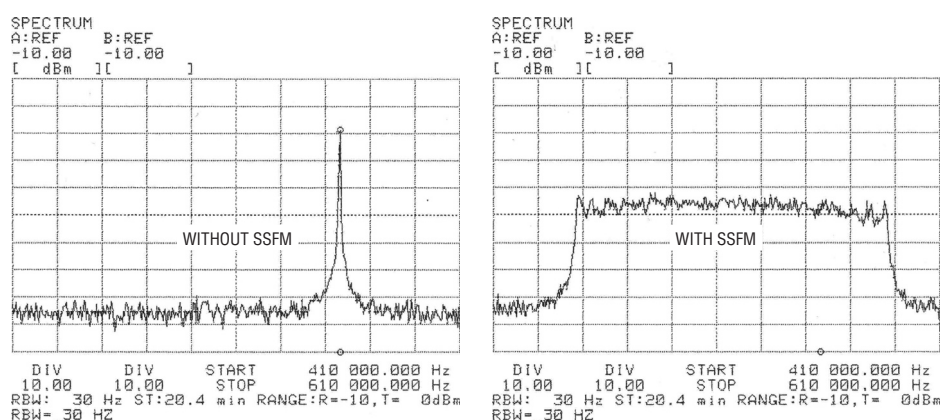


Figure 3. Zoom-in of output voltage spectra showing fundamental frequency. Notice the >20dB reduction in peak noise with SSFM enabled.

its harmonics, thus reducing the peak noise at any given frequency. The LTC3736-1 achieves this by integrating an SSFM oscillator with a dual synchronous switching regulator controller to randomly modulate its clock frequency.

Circuit Description

The LTC3736-1 is a 2-phase dual synchronous step-down DC/DC controller that requires few external components. Its No R_{SENSE}^{TM} , current mode architecture eliminates the need for current sense resistors and improves efficiency, without requiring a Schottky diode. The two controllers are operated 180 degrees out of phase, reducing the required input capacitance and power loss and noise due to its ESR.

The LTC3736-1 is nearly identical to the LTC3736 (See '2-Phase Dual Synchronous DC/DC Controller with Tracking Provides High Efficiency in

a Compact Footprint' in the August, 2004 issue of *Linear Technology Magazine*), except the LTC3736-1 has a built-in SSFM oscillator that randomly varies its switching frequency.

A tracking input allows the second output to track the first output (or another supply) during startup, allowing the LTC3736-1 to satisfy the power-up requirements of many microprocessors, FPGAs, DSPs and other digital logic circuits. The LTC3736-1 can operate from input voltages between 2.7V and 9.8V and is available in a low profile 4mm × 4mm leadless QFN package and 24-lead narrow SSOP package.

A typical application circuit using the LTC3736-1 is shown in Figure 1. This circuit provides two regulated outputs from a single 3.3V input supply. The 2200pF capacitor connected to the FREQ pin is used to filter and smooth out the abrupt changes in frequency of the LTC3736-1's internal

SSFM oscillator. This allows time for the regulator's feedback control loop to adjust to the frequency changes without adversely affecting output voltage ripple or regulation. The digital input control pin SSDIS is used to disable the SSFM oscillator and force the LTC3736-1 to operate at constant frequency for debugging purposes. Table 1 summarizes how to use the LTC3736-1's SSDIS and FREQ pins.

Figure 2 shows a comparison of the spectra of the LTC3736-1 with and without SSFM enabled. These show a spectrum analyzer view of the output voltage, using a peak measurement technique. Without SSFM, most of the signal energy in the output appears at the 550kHz switching frequency and its harmonics. With SSFM enabled, the energy is spread among many frequencies and the harmonic peaks are diminished or disappear.

Figure 3 shows a zoom-in of the spectra showing the fundamental frequency. With SSFM enabled, the output signal energy is spread nearly uniformly from 450kHz to 580kHz, with a peak energy more than 20dB below the 550kHz peak with SSFM disabled. In other words, with SSFM enabled, the EMI energy at any particular high frequency has an amplitude that is less than one-tenth that of the single fixed frequency with SSFM disabled. These lower amplitude frequency components reduce the amount of potential interference.

No Adverse Effect on Transient Response, Ripple, Efficiency, or Tracking

One of the greatest difficulties in implementing an SSFM switcher is ensuring that the randomly changing frequencies do not cause the regulator's control loop become unstable. This can manifest itself as significantly increased output voltage and inductor current ripples, or worse, total instability and loss of regulation. The LTC3736-1 is proof that these challenges have been overcome, and better yet, all that is required externally is a single capacitor connected to the FREQ pin.

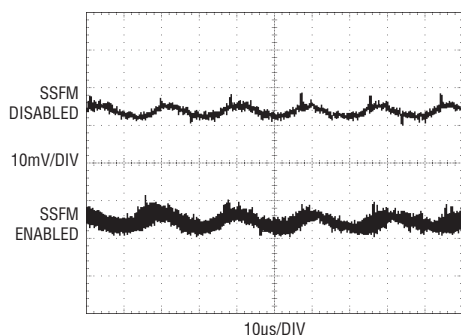


Figure 4. Output voltage ripple for 1.8V output using “envelope” oscilloscope function

Figure 4 shows the output voltage ripple for the circuit of Figure 1 with and without SSFM enabled. Note that since SSFM is constantly changing the LTC3736-1 switching frequency, it is difficult to show the true behavior of SSFM using a still oscilloscope snapshot—a video would be much more informative.

Nonetheless, the scope traces in Figure 4 have been acquired using the “envelope” oscilloscope function, which shows the leading and trailing waveform edges blending in with each other as the frequency is varied. The peak to peak ripple with SSFM enabled does increase slightly, but this is expected since output ripple is inversely proportional to switching frequency, and SSFM introduces some frequencies that are lower than the single fixed 550kHz frequency.

Although it is not easily detected from this still snapshot, note that while the frequency is varying—one can think of SSFM as introducing frequency jitter—the duty cycle is constant. In other words, there is no duty cycle jitter or sub-harmonic instability with SSFM enabled on the LTC3736-1.

Figure 5 compares the efficiency of the circuit in Figure 1 with and without SSFM enabled. Figure 6 shows load step transients and Figure 7 shows tracking startup waveforms with SSFM enabled. In all cases, the behavior of the LTC3736-1 is unaffected by the addition of SSFM.

Conclusion

The LTC3736-1 is an easy-to-use dual synchronous switching DC/DC controller that requires few external components. Additionally, it features

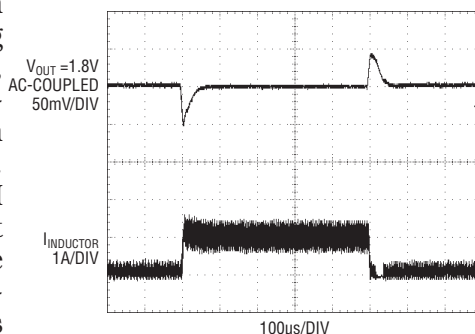


Figure 6. Load step response for circuit of Figure 1 with SSFM enabled

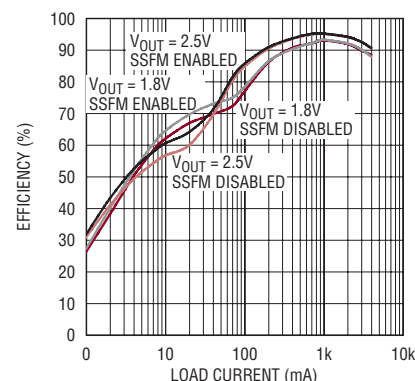



Figure 5. Efficiency for circuit of Figure 1. There is little difference with SSFM enabled

an internal spread spectrum oscillator that randomly varies the controllers’ switching frequency, providing a simple solution to reduce power-supply-induced EMI that otherwise might require significant and costly troubleshooting and redesign. 

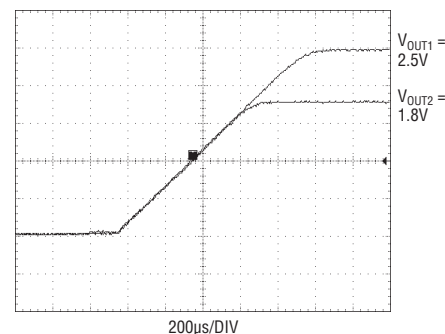


Figure 7. Startup of circuit of Figure 1 showing the two supplies tracking with SSFM enabled


LTC3802, continued from page 8

reducing the duty cycle and hence the output voltage until the current drops below the limit. The soft-start capacitor needs to move a fair amount before it has any effect on the duty cycle, adding a delay until the current limit takes effect. This allows the LTC3802 to experience brief overload conditions while maintaining output voltage regulation.

Nevertheless, at high input voltages, even a small RUN/SS time delay could cause the output current to overshoot badly during a severe short circuit. To avoid that situation, LTC3802 adds a hard current limit circuit.

If the load current is 1.5 times larger than the programmed current limit threshold, the LTC3802 shuts off the top MOSFET immediately. This stops the increase in the inductor current. At this moment, if CMPIN (which samples V_{OUT}) is 10% lower than its nominal value, the LTC3802 hard current-limit latches and discharges the RUN/SS capacitor with a current source of more than 1mA until RUN/SS hits its shutdown threshold. Once RUN/SS is completely discharged, the LTC3802 cycles its soft start cycle again. Figure 8 shows waveforms during a severe short circuit at the output of a 12V–3.3V converter.

Conclusion

The high efficiency LTC3802 is the latest member of Linear Technology’s family of constant frequency, voltage feedback, synchronous N-channel controllers. With its unique set of powerful features and performance improvements (summarized in Table 1), it improves on the LTC1702/LTC1702A, and is ideal for high input voltage and low duty cycle applications. The LTC3802 is available in small 28-Lead SSOP and 32-Lead (5mm × 5mm) QFN Packages, allowing an entire 87W converter to be laid out in less than 6 square inches (Figure 9). 

Superfast Fixed-Gain Triple Amplifiers Simplify Hi-Res Video Designs by Jon Munson

Introduction

The LT6553 and LT6554 triple video amplifiers offer 600MHz performance in a compact package, requiring no external gain-setting resistors to establish gain of 2 or unity-gain, respectively. One may wonder “Why are such super-fast amplifiers are now necessary in video designs—isn’t that overkill?” The answer is a resounding no. The proliferation of high-resolution video displays, both in the professional and consumer markets has markedly increased the analog bandwidth of baseband video signals. The latest demands of video equipment are so far ahead of the last generation that the performance of the LT6553 and LT6554 is not overkill at all, but in fact mandatory.

For example, digital studio equipment for NTSC broadcast television typically uses pixel-rates around 14 million per second, while now ubiquitous XGA computer outputs (1024 x 768) routinely churn out about 80 Megapixels per second. The latest High Definition consumer formats put out a comparable 75Mpixel stream and the increasingly popular UXGA professional graphics format (1600 x 1200) generates a whopping 200Mpixel per second flow. Obviously the accurate video reproduction of these newer formats is placing exceptional demands on the frequency response of the video amplifiers involved. Specifically, pulse-amplitude waveforms like those

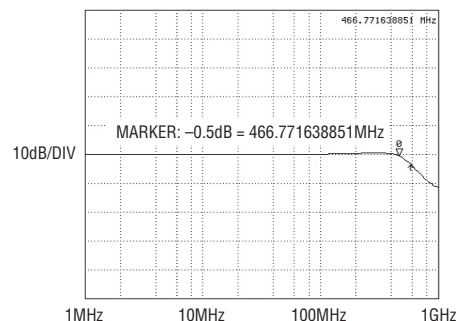


Figure 2. Wide frequency response of circuit in Figure 1

of baseband video generally require reproduction of high-frequency content up to at least the 5th harmonic of the fundamental frequency component, which is 2.5 times the video pixel rate, accounting for the 2 pixels per fundamental cycle relationship. This indicates that for UXGA in particular, flat frequency response to beyond 0.5GHz is required!

Easy Solution for Multi-Channel Video Applications

Baseband video generated at these higher rates is processed in either native red-green-blue (RGB) domain or encoded into “component” luma plus blue-red chroma channels (YPbPr); three channels of information in either case. With frequency response requirements extending to beyond 500MHz, amplifier layouts that require external resistors for gain setting tend to waste valuable real-estate, and frequency response and crosstalk anomalies can plague the printed circuit development process.

The LT6553 and LT6554 conveniently solve all these problems by providing internal factory-matched resistors and an efficient 3-channel flow-through layout arrangement using a compact SSOP-16 package.

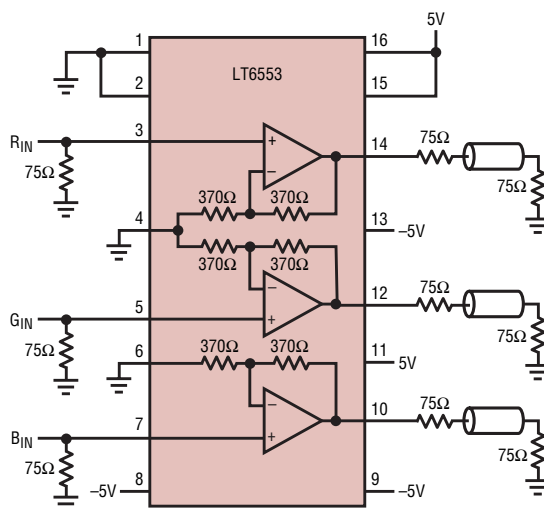


Figure 1. LT6553 RGB cable driver circuit

Figure 1 shows the typical RGB cable driver application of an LT6553, and its excellent frequency and time response plots are shown in Figures 2 and 3. Frequency markers in Figure 2 show the -0.5dB response beyond 450MHz and -3dB response at about 600MHz.

What's Inside

The LT6553 and LT6554 integrate three independent sections of circuitry that form classic current-feedback amplifier (CFA) gain blocks, all implemented on a very high-speed fabrication process. The diagram in Figure 4 shows the equivalent internal circuitry (one CFA section shown).

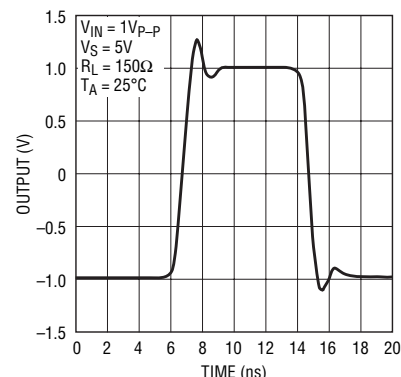


Figure 3. Fast pulse response of circuit in Figure 1

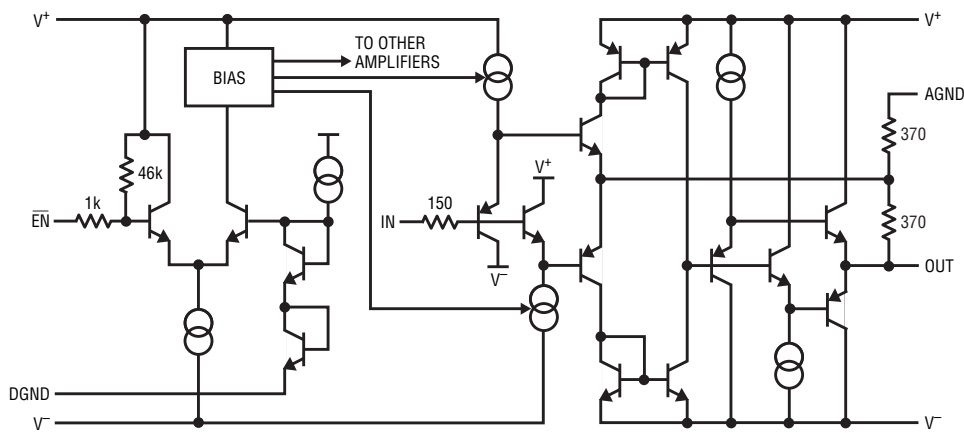


Figure 4. LT6553 & LT6554 simplified internal circuit functionality

The on-chip feedback resistors set the closed-loop gain to unity or two, depending on the part. The nominal feedback resistances are chosen to optimize the frequency response for maximal flatness under the anticipated loading conditions. The LT6553 is intended to drive back-terminated 50Ω or 75Ω cables (for effective loading of 100Ω to 150Ω respectively), while the LT6554 is useful for driving ADCs or other high impedance loads (characterized with 1kΩ as a reference loading condition).

All three CFAs have a bias control section with a power-down command input. The shutdown function includes internal pull-up resistance to provide a default disable command, which when invoked, reduces power consumption to less than 100μA for an entire three-channel part. During shutdown mode the amplifier outputs become high impedance, though in the case of the LT6553, the feedback resistor string to AGND is still present. The parts come into full-power operation when the enable input voltage is brought

within 1.3V above the DGND pin. The typical on-state supply current of 8mA per amplifier provides for ample cable-drive capacity and ultra-fast slew rate performance of 2.5V per nanosecond!

MUXing Without Switches

RGB and YPbPr video signals are commonly multiplexed (selections made on an occasional basis) to reduce I/O connector count or otherwise re-use various high-value video signal-processing sections when selecting various modes of operation in the end use of the product. This has often been accomplished with the use of FET switches and buffer amps to route the various video channel signals, but can alternatively be performed by use of the power-down functionality included in the LT6553 and LT6554. Figure 5 shows an example circuit using LT6554 units cross-controlled to allow a single video path to be enabled at any particular time. This might be the situation at the input side of a video display or AV receiver

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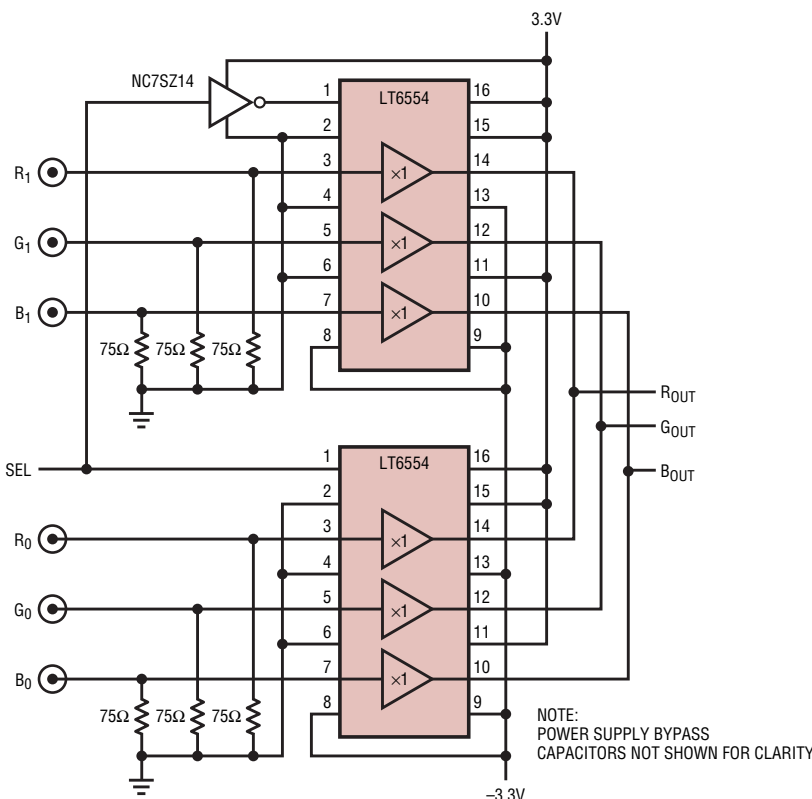
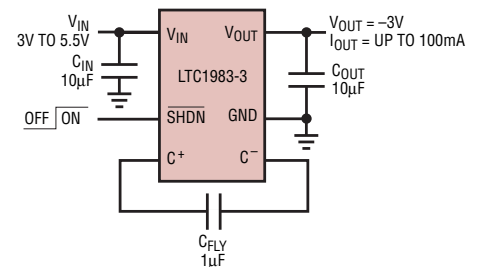


Figure 5. Video input multiplexer using LT6554 shutdown feature



C_{FLY}: TAIYO YUDEN LMK212BJ105
C_{IN}, C_{OUT}: TAIYO YUDEN JMK316BJ106ML

Figure 6. Generating a local -3V supply with 4 tiny components

Power Supply Tracking for Linear Regulators

by Dan Eddleman

Introduction

The LTC2923 provides simple and versatile control over the power-up and power-down behavior of switching power supplies. It allows several supplies to track the voltage of a master supply, so that their relative voltages meet the stringent specifications for the power up of modern digital semiconductors, such as DSPs, microprocessors, FPGAs and ASICs. The LTC2923 is specifically designed to work with switching power supplies

(see “Versatile Power Supply Tracking without MOSFETs” from *Linear Technology Magazine*, February, 2004) but it is easily adapted to linear regulators, including popular low-dropout (LDO) types. Summarized here are several techniques for controlling linear regulators with the LTC2923.

Monolithic Regulators

Table 1 lists three popular monolithic linear regulators that have been tested with the LTC2923. Using these three

monolithic LDOs with the LTC2923 is generally very simple:

- The LTC3020 is a 100mA low dropout regulator (LDO) that operates with input supply voltages between 1V and 10V. Since its ADJ pin behaves like the feedback pin on most switching regulators, tracking the LTC3020's output using the LTC2923 is simple. The standard circuits and design procedures shown in the LTC2923 data sheet require no modification when used with the LTC3020 (Figures 1 and 2).
- The LTC3025 is a 300mA monolithic CMOS LDO that regulates input supplies between 0.9V and 5.5V, while a bias supply between 2.5V and 5.5V powers the part. Similar to the LT3020, the LTC3025's ADJ pin is operationally identical to common switchers. For that reason, the LTC3025 combined with an LTC2923 provides a simple supply tracking solution for loads less than 300mA (Figures 1 and 2).
- The LTC1844 CMOS LDO drives loads up to 150mA with input supply voltages between 1.6V and 6.5V. When used in conjunction with the LTC2923, a feedforward capacitor should be included as described in the “Adjustable Operation” section of the LTC1844 data sheet. Otherwise, no special considerations are necessary.

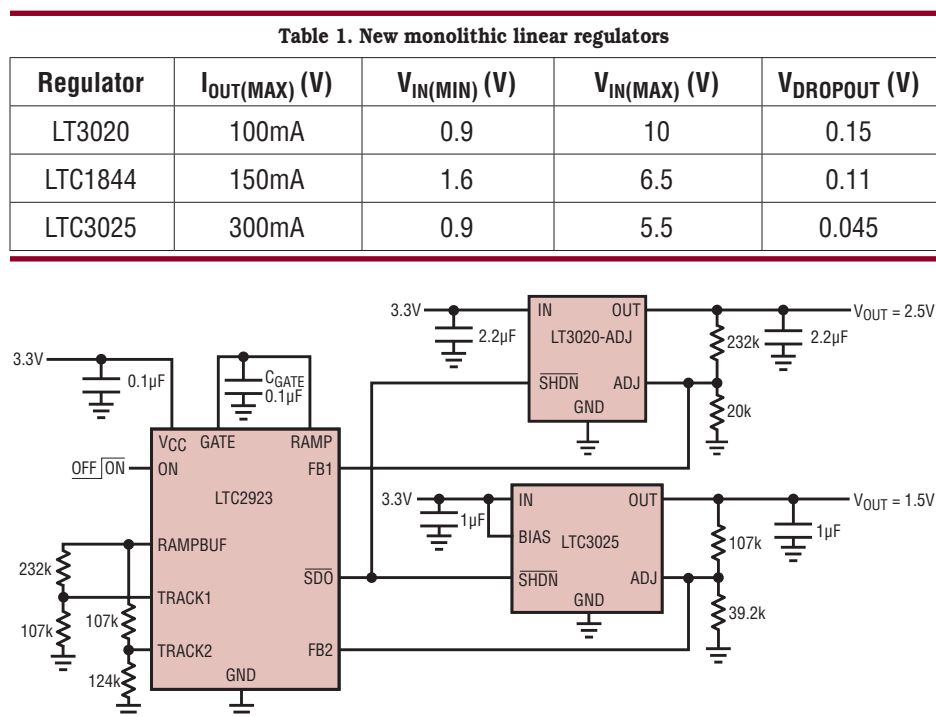


Figure 1. An LTC2923 causes the outputs of the LT3020 and LTC3025 to track during power-up and power-down.

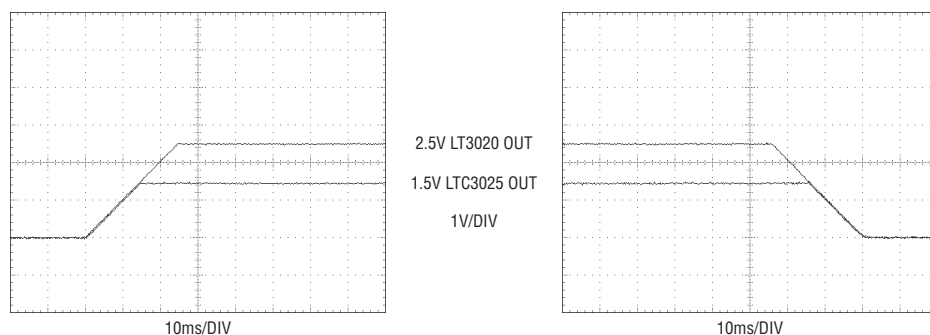


Figure 2. The outputs of the LT3020 and LTC3025 low-dropout linear regulators ramp-up and ramp-down together. (Output of circuit in Figure 1.)

The LTC1761 Family of Monolithic, Bipolar Regulators

Table 2 shows the LTC1761 family of monolithic, bipolar low dropout regulators. These regulators cover a wide range of load currents and offer outstanding transient response and low noise, making them a popular choice for applications with loads less than 3A.

In these regulators, the ADJ pin draws excess current when the OUT

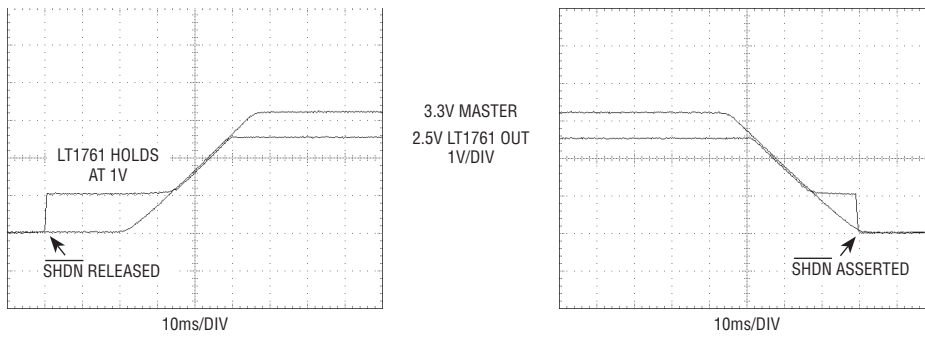


Figure 3. LT1761/LT1962/LT1762/LT1763/LT1963A/LT1764A with adjustable outputs only track above 1V unless modified as discussed in this article. The SHDN pin of the LDO is active before the ramp-up and after ramp-down.

pin drops below about 1V, a region of operation that LDOs do not normally experience. Nevertheless, an LDO which tracks another supply, enters this region when the output tracks below 1V (Figure 3). If this excess current is not accounted for, the output of the LDO will be slightly higher than ideal when it tracks below 1V. Three techniques have been used to successfully track outputs of this LDO family below 1V.

If low dropout voltages are not necessary, simply connect two diodes in series with the OUT pin (Figure 4). In this configuration, the OUT pin remains two diode drops above the circuit's output. As a result, the LDO remains in its normal region of operation even when the output is driven near ground. Since the feedback resistors are connected to the output, the LDO regulates the voltage at the circuit output instead of the LDO's OUT pin. Diode voltage varies with both load current and temperature, so verify that the output is low enough at the minimum diode voltage. Likewise, the input voltage must be high enough to regulate the output when the diode

drops are at their maximum. This solution effectively increases the dropout voltage of the linear regulator by two diode drops. Therefore, applications that require a low dropout voltage are better served by the solutions that follow.

Consider using the LTC1761, LT1962, LT1762, or LT1763 voltage regulators when the load is less than 500mA and a low dropout voltage is necessary. A fixed output part, (such as the LTC1763A-1.5) can be used as an adjustable LDO if the SENSE pin is treated like an ADJ pin with a feedback voltage of 1.5V (Figure 5). The SENSE pin on the fixed output parts draws about 10 μ A regardless of the OUT pin's voltage, unlike the ADJ pin on the adjustable parts. When choosing feedback resistors, minimize the output error by compensating for the extra 10 μ A of current that appears across the upper resistor. Also, use small valued resistors to minimize the error due to the 0 μ A to 20 μ A data sheet limits while avoiding values that are so small that the LTC2923's 1mA I_{FB} will be unable to drive the output to ground. To satisfy these constraints,

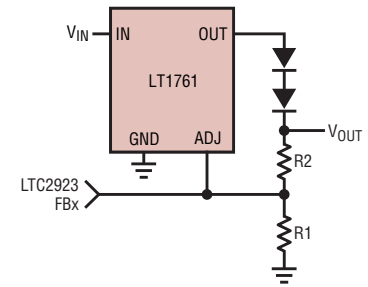


Figure 4. Diodes placed in series with the OUT pin allow the LT1761 to track down to 0V.

ensure that the parallel combination of the two feedback resistors is slightly greater than 1.5k Ω . For most output voltages, this reduces the output error due to the SENSE pin current to about 1%.

For applications that require higher load currents and a low dropout voltage, the LT1963A and LT1764A may be appropriate. These parts are specified for 1.5A and 3A load currents respectively. Unfortunately, the SENSE pins on these fixed output parts draw about 600 μ A.

To use these parts, configure an operational amplifier to buffer the voltage from the feedback resistors to the SENSE pin of the 1.5V fixed output versions (Figure 6). If the op amp is configured with a voltage gain of 2, the 1.5V regulator in combination with the op amp behaves as an adjustable output regulator with a 0.75V reference voltage. The input to the op amp now serves as the ADJ input of the new regulator. This technique allows the use of the high current LT1963A/LT1764A where the voltage loss of series diodes would be unacceptable. It also works for the LT1761, LT1962, LT1762, and LT1763 in cases where the 10 μ A ADJ pin cur-

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Table 2. LT1761 family of low-dropout linear regulators

Regulator	$I_{OUT(MAX)}$ (V)	$V_{IN(MIN)}$ (V)	$V_{IN(MAX)}$ (V)	$V_{DROPOUT}$ (V)
LT1761	100mA	1.8	20	0.30
LT1762	150mA	1.8	20	0.30
LT1962	300mA	1.8	20	0.27
LT1763	500mA	1.8	20	0.30
LT1963A	1.5A	2.1	20	0.34
LT1764A	3A	2.7	20	0.34

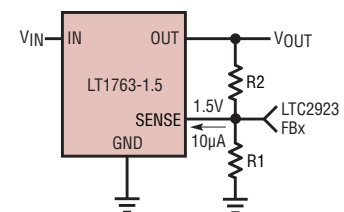


Figure 5. The fixed-output LT1763-1.5 can track down to 0V, has low dropout, and a resistive divider can be used for outputs greater than 1.5V.

Tiny, Resistor-Programmable, μ Power 0.4V to 18V Voltage Reference

by Dan Serbanescu and Jon Munson

Introduction

The LT6650 is a 0.4V to 18V adjustable voltage reference that runs from low voltage and consumes only a few microamps. It features a low-dropout (LDO) characteristic, can source or sink current, can be configured in either series or shunt mode and saves space in the tiny 5-lead ThinSOT-23 package.

Figure 1 shows a block diagram of the reference. Its 400mV internal voltage reference is connected to the non-inverting input of an operational amplifier. The inverting input is brought to a pin, thus making a series-mode reference adjustable to any output voltage from 400mV up to ($V_{\text{SUPPLY}} - 0.35\text{V}$) by using two external resistors. It can also be configured as

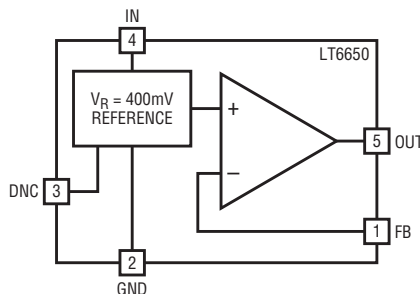


Figure 1. Block diagram of 1% accurate micropower 0.4V to 18V adjustable reference.

a fixed 400mV reference by simply connecting the output to the op amp inverting input. While the LT6650 is designed as a series reference, it can be used as a shunt-mode reference simply by shorting the positive rail to the output pin—it can be programmed

to produce any precision “zener” voltage within the wide supply range (1.4V to 18V) by selection of the two external resistors.

Specifications

Table 1 summarizes the performance of the LT6650. The supply current is only 5.6 μ A and the supply voltage may range from 1.4V to 18V, which permits battery-powered equipment to be plugged into an unregulated wall adapter without the need for peripheral circuitry to limit the voltage input to the reference. The 400mV internal reference is $\pm 1\%$ accurate over the -40°C to 85°C temperature range and is also fully specified from -40°C to 125°C for extended temperature range

Table 1. LT6650 Performance ($T_A = 25^{\circ}\text{C}$, $V_{\text{IN}} = 5\text{V}$, $V_{\text{OUT}} = 400\text{mV}$, $C_L = 1\mu\text{F}$, unless otherwise noted)

Parameter	Conditions	Min	Typ	Max	Units
Input Voltage Range	$-40^{\circ}\text{C} \leq T_A \leq 125^{\circ}\text{C}$	1.4		18	V
Output Voltage	$-40^{\circ}\text{C} \leq T_A \leq 85^{\circ}\text{C}$	396 -1	400	404 1	mV %
Line Regulation	$1.4\text{V} \leq V_{\text{IN}} \leq 18\text{V}$		1	6	mV
Load Regulation	0 to $-200\mu\text{A}$ (Sourcing) 0 to $200\mu\text{A}$ (Sinking)		-0.04 0.24	-0.2 1	mV mV
Output Voltage Temperature Coefficient			12		$\mu\text{V}/^{\circ}\text{C}$
Dropout Voltage	$V_{\text{OUT}} = 1.4\text{V}$ $I_{\text{OUT}} = 0\mu\text{A}$ $I_{\text{OUT}} = 200\mu\text{A}$ sourcing		75	100 250	mV mV
Supply Current	$1.4\text{V} \leq V_{\text{IN}} \leq 18\text{V}$		5.6	12	μA
FB Pin Input Current	V_{FB} shorted to V_{OUT}		1.2	10	nA
Turn-On Time			0.5		ms
Output Voltage Noise	0.1Hz to 10Hz		20		$\mu\text{V}_{\text{P-P}}$
Thermal Hysteresis	-40°C to 85°C		100		μV

applications. The rail-to-rail output delivers 200µA in both sourcing and sinking modes of operation.

How it Works Inside

Figure 2 shows the simplified schematic of the reference. Transistors Q1–Q7 form the band-gap-derived 400mV reference that is fed to the non-inverting input of the error amplifier formed by Q8–Q12. The resistors R1–R3 set the correct current flow into the internal reference, while R4 provides for post-package trimming capability. Transistors Q20 and Q21 form the rail-to-rail output stage and are driven by Q13–Q19. Resistors R5–R8 and the I_2 current generator establish the gain and quiescent operating current of the output stage. In conjunction with the minimum recommended output capacitance of 1µF, stabilization is assured through Miller compensation inside error amplifier Q8–Q12. Pins are ESD protected by diodes D1–D3.

Applications

Battery Powered Pocket Reference

The unique pocket reference shown in Figure 3 can operate for years on a pair of AAA alkaline cells or a single Lithium coin-cell, as the circuit draws just 10µA supply current. An input capacitor of 1µF as shown should be used when the LT6650 is operated from small batteries or other sources

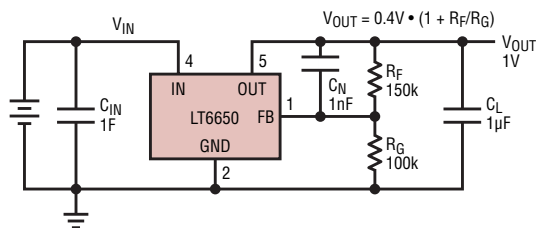


Figure 3. Battery powered pocket voltage reference runs for years on a coin cell.

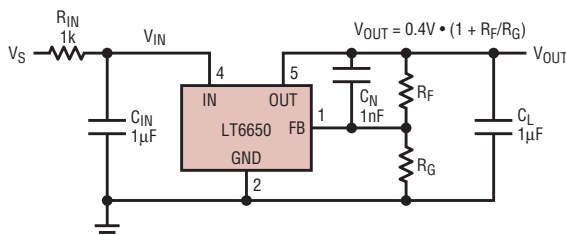


Figure 4. Simple input network for improved supply rejection

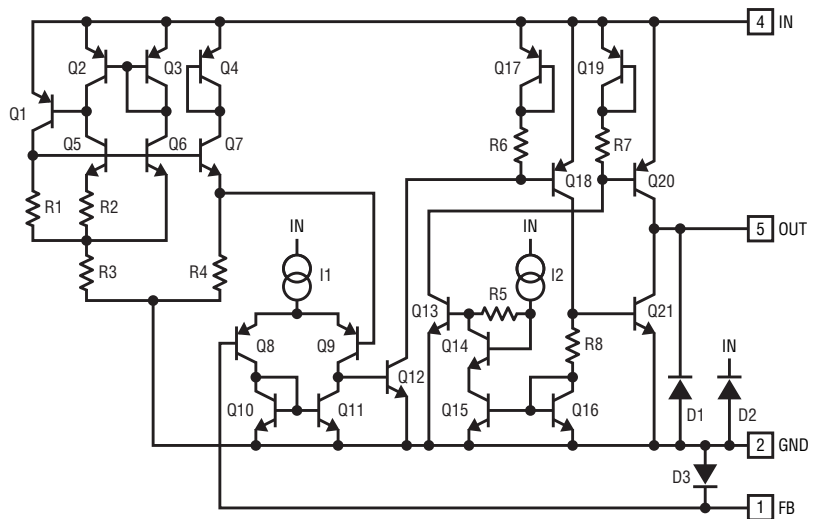


Figure 2. LT6650 simplified schematic showing detail of low-dropout topology

with impedance over about 50Ω. The output is adjustable from 0.4V up to the battery supply by selecting two feedback resistors (or setting a trimmer potentiometer position) to configure the non-inverting gain of the internal operational amplifier. A feedback resistor R_F is connected between the OUT pin and the FB pin and a gain resistor R_G is connected from the FB pin to GND. The resistor values are related to the output voltage by the following relationship:

$$R_F = 2.5 \cdot (V_{OUT} - 0.4) \cdot R_G$$

The worst-case FB pin bias current (I_{BIAS}) can be neglected with an R_G of 100kΩ or less. In ultra-low-power applications where current in the voltage programming resistors might

be reduced to where the 1.2nA typical I_{BIAS} becomes relatively significant loading, the relationship between the resistors then becomes:

$$R_F = R_G \cdot \frac{V_{OUT} - 0.4}{0.4 - (I_{BIAS} \cdot R_G)}$$

The minimum allowable gain resistor value is 2kΩ established by the 400mV FB pin voltage divided by the maximum guaranteed 200µA output current sourcing capability. In applications that scale the reference voltage, intrinsic noise is amplified along with the DC level. To minimize noise amplification, a 1nF feedback capacitor (C_N) as shown in Figure 3 is recommended. Any net load capacitance of 1µF or higher assures amplifier stability.

Automotive Reference

In the presence of high supply noise, such as in automotive applications or DC-DC converters, an RC filter can be used on the V_{IN} input as shown in Figure 4. Due to the exceptionally low supply current of the LT6650, the input resistor (R_{IN}) of this filter can be 1kΩ or higher, depending on the difference in V_{IN} and V_{OUT} . Figure 5 shows supply rejection better than 30dB over a wide frequency spectrum, for a minimum sourcing output current of 40µA and an input filter comprising $R_{IN} = 1k\Omega$ and $C_{IN} = 1\mu F$. If even higher rejection is necessary, the input filter structure presented in Figure 6 effectively eliminates any supply transients

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Hot Swap for High Availability Systems

by David Soo

Introduction

Critical computer, mass storage and communication systems are designed to operate with zero down time, or to at least approach that ideal. Such high availability systems must continue functioning even when system upgrades and maintenance are performed. Often this requires circuit boards be inserted into, and removed from, a live powered system.

Hot swapping requires a power switch to initially isolate and then control inrush current via a controlled ramp up of power, which prevents any disturbance to the backplane and adjacent circuits. Because the Hot Swap circuit is the gateway for all board power, it is a natural place to monitor and collect power supply data. Such data reveals the health of the board and the integrity of the power path.

With this in mind, the LTC4260 combines a wide input range Hot Swap controller, ADC voltage monitor, I²C™ serial communication, and other features in one device (see Table 1). The LTC4260's Hot Swap circuit uses an external N-channel pass transistor

to isolate the hot swapped board from the backplane when it is first inserted. After a de-bounce time the controller can begin to apply power to the board or wait for a turn-on command from a host processor. Power is ramped

gradually to minimize any backplane disturbance. After the power-up process is complete, the LTC4260 continues to monitor for faults in the power path.

I²C is a trademark of Philips Electronics N.V.

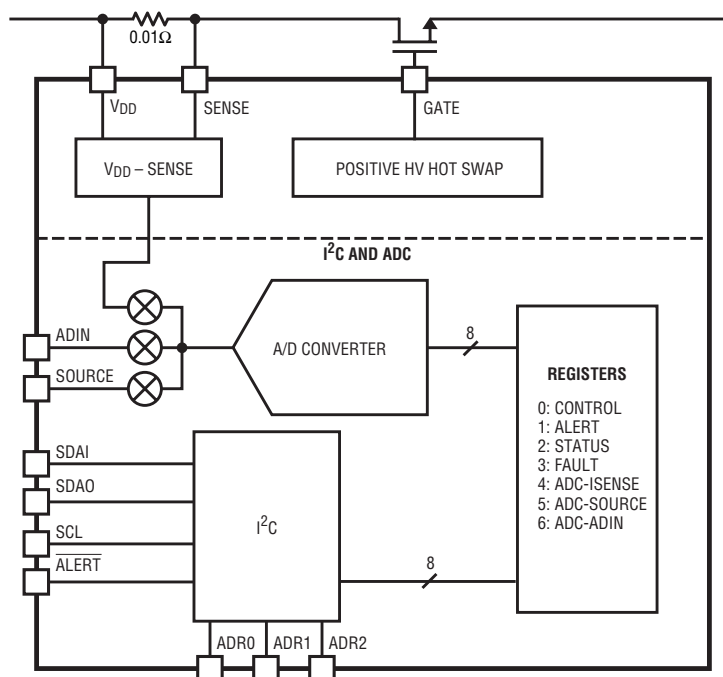


Figure 1. Block diagram of the LTC4260

Table 1. Some LTC4260 features

Feature	Benefits
Wide Input Voltage Range: Operates from inputs of 8.5V to 80V, with 100V absolute maximum	<ul style="list-style-type: none"> □ Suitable for 12V, 24V and 48V systems □ Simplifies design because part functions on a semi-regulated supply. □ Large overvoltage transient range eases design tolerances for transient protection.
8-Bit ADC: ADC monitors current, output voltage and external pin voltage and measures off-state current in the FET to determine FET failures	<ul style="list-style-type: none"> □ Increases reliability. □ Board power information provides an early warning of board failure. □ Verify board is staying within its allotted power □ Allows integrity check of redundant supply paths
I²C/SMBus: Communicates as a read-write slave device using a 2-wire serial interface.	<ul style="list-style-type: none"> □ Improves integration with the host system. Interface allows the host to configure the part, determine which faults are present or have occurred, and read back ADC measurements
Fast Short Circuit Response: Fast (<1μs) current limit response to shorts	<ul style="list-style-type: none"> □ Protects connector from overcurrent. □ Limits the short circuit caused glitch on the input supply.
Alerts Host after Faults: When configured (using I²C), faults activate an active pull-down on the ALERT pin	<ul style="list-style-type: none"> □ Interrupting the host for immediate fault servicing limits system damage.

The LTC4260 provides the means for quantitatively measuring the board current and voltages with an onboard ADC and multiplexer. It reports this information using the I²C serial communication bus when polled by a host processor. The device will interrupt the host for specific fault conditions, if configured to do so.

The LTC4260 works in applications from 80V (with transients to 100V) to 12V battery systems where the operating range could drop to 8.5V.

An Inside Look

The block diagram of the LTC4260 is shown in Figure 1. The lower section of the block diagram shows the ADC voltage monitoring, the registers and the I²C interface.

The ADC monitors the current via the sense resistor voltage, V_{DD}-SENSE. The SOURCE pin and the external ADIN pin are also multiplexed to the ADC. The registers allow the user to configure the part and to read back useful information on the status of the part and if any faults have occurred.

The I²C block uses a 2-wire serial interface using the SCL and SDA signals. To facilitate communications across two isolated grounds, the SDA is split into SDAI and SDAO pins to allow the part to drive optoisolators with a minimum number of external components. For normal I²C communications sharing a common ground these two pins are shorted together.

The $\overline{\text{ALERT}}$ pin is used for interrupts. The upper block diagram

Table 2. LTC4260 register address and contents

Register	Description
CONTROL	Register turns-on or turns-off the pass transistor and controls whether the part will Auto-Retry or lathoff after a fault. It also configures the behavior of the GPIO pin
ALERT	Alert register enables which faults interrupt the host using the $\overline{\text{ALERT}}$ pin. At power-up the default is to not alert on faults.
STATUS	Status register provides pass transistor (on/off), $\overline{\text{BD_PRST}}$ (high/low) and GPIO (high/low) conditions. It also lists five fault present conditions.
FAULT	Fault register logs overcurrent, overvoltage, undervoltage, power-bad, FET short and $\overline{\text{BD_PRST}}$ changed state faults.
SENSE	ADC data for the V _{DD} -SENSE voltage measurement
SOURCE	ADC data for the SOURCE pin voltage measurement
ADIN	ADC data for the ADIN pin voltage measurement

contains the Hot Swap blocks required to monitor the input supply, and when appropriate to turn on the gate of the external pass transistor.

Measure Real-Time Board Power with Integrated ADC

Collecting and compiling information on the voltage and current flowing into each card is a useful way to measure the health of the card. Operating data can be compared to historical data to discern whether a card was actually using its allotted power or if it was operating abnormally. An abnormally operating card could be flagged for service, perhaps even before it failed. The LTC4260 includes an 8-bit data converter that continuously monitors three voltages: the ADIN pin, the

SOURCE pin and the amplified difference between the V_{DD} and the SENSE pins. The ADIN pin is an uncommitted ADC input. This pin allows the user to monitor any available voltages.

The ADIN pin is monitored with a 2.56V full scale direct connection to the converter. The SOURCE pin uses a 1/40 divider at the input which gives a 102.4V full scale. The V_{DD}-SENSE voltage amplifier has a voltage gain of 33.33 which results in a 76.8mV full scale.

The results from each conversion are stored in three ADC registers (see Table 2) and updated 10 times a second. Setting the test mode control register bit halts the data converter so that registers can be written to and read from for software testing.

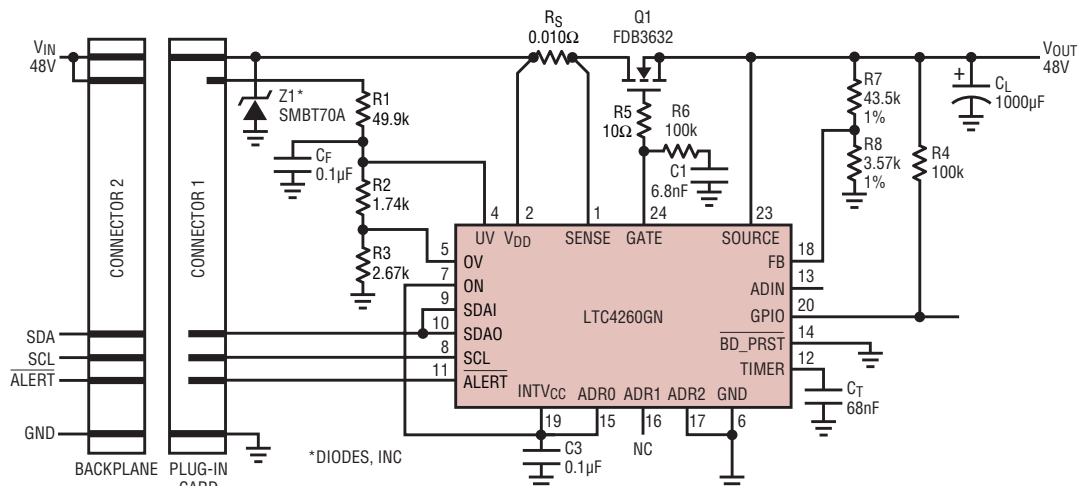


Figure 2. This 3A, 48V Hot Swap application resides on the plug-in card.

Typical Hot Swap Application

An N-channel pass transistor Q1 in the power path, as shown in Figure 2, controls power to the board. The sense resistor R_S detects current for overcurrent faults and ADC measurements. Capacitor C1 controls the GATE slew rate while resistor R6 compensates the current control loop. Resistor R5 suppresses self-oscillations in Q1. Resistors R1, R2 and R3 provide undervoltage and overvoltage sensing at the input while R7 and R8 provide output power-good monitoring.

The staggered pins of the male connector ensure all power supplies are physically connected before output power is allowed to ramp. The following is a typical board insertion sequence:

- ❑ Long power and ground pins make contact and the internal 5.5V supply (INTVCC) becomes active. The internal registers are reset after a power-on-reset pulse. The pass-transistor (Q1) is off.
- ❑ The medium length pins, SDA, SCL and ALERT make contact. This allows I²C communication to begin.
- ❑ The short pin connects resistor R1 to the supply voltage bringing the UV and OV pins to the adjusted level. The UV, OV and BD_PRST pins must remain in the acceptable range for 100ms to ensure that any contact bounce

during insertion has ended. After 100ms the ON pin is tied high. If it is high, then the external switch turns on. If it is low, the external switch turns on when the ON pin is brought high or if a serial bus turn-on command is received.

Power-Up Sequence

The pass transistor is turned on by charging up the GATE with a 18μA current source. The voltage at the GATE pin rises with a slope equal to 18μA/C1 and the supply inrush current is set at

$$I_{INRUSH} = \frac{C_L}{C_1} \cdot 18\mu A$$

When the GATE voltage reaches the Q1 threshold voltage, the switch begins to turn on and the SOURCE voltage follows the GATE voltage as it increases.

The LTC4260 uses 3.5V reference, a precision voltage comparator and an external resistive divider to monitor the output supply voltage. When the voltage at the FB pin rises above the 3.5V threshold, the GPIO pin, in its default configuration ceases to pull low, indicating that the power is now good. Figure 3 shows a typical Hot Swap, 100ms delay and power-up event.

Controlled Turn-Off

Controlling the GATE pin slew rate during turn-off prevents inductor driven voltage spikes on the drain and source of the pass transistor due to the rapid change in current. The controlled turn-off of the switch uses a 1mA current pulling the GATE pin to ground. Normally the turn-off is initiated by the ON pin going low or a serial bus turn-off command. Additionally, several fault conditions

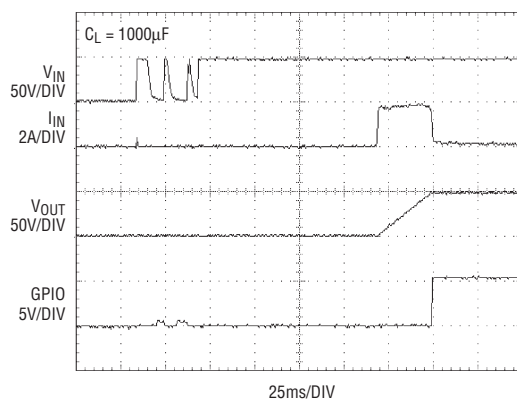


Figure 3. Power-up waveforms for the 48V Hot Swap application

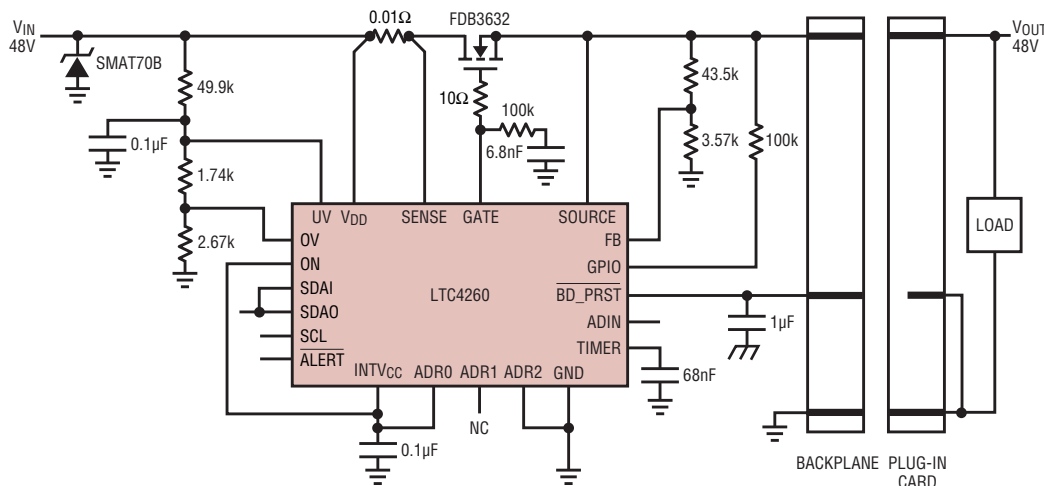


Figure 4. This 3A, 48V Hot Swap application resides on the backplane or motherboard

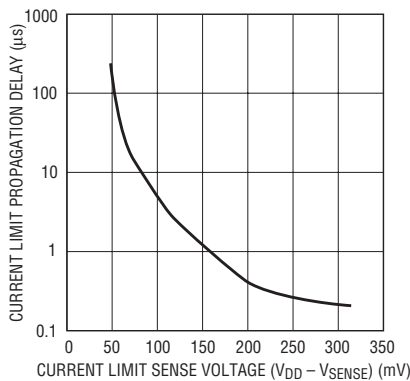


Figure 5. The response time to an overcurrent depends on the sense voltage. In the case of a short circuit in the load, the current is brought under control in less than 1μs.

will turn off the switch. These include an input overvoltage (OV pin), input undervoltage (UV pin), overcurrent circuit breaker (SENSE pin) or $\overline{\text{BD_PRST}}$ going high.

LTC4260 Resides on Either Side of the Connector

In Figure 2 the LTC4260 is located on the plug-in board side of the connector. The backplane side of the connector contains power and signal routing. Some designers choose to place the Hot Swap controller on the backplane or motherboard side of the connector along with host processing of the data.

A typical backplane resident application is shown in Figure 4. The plug-in card is inserted into an unpowered slot with ground and power pins mating first. Next the connection sensing pin directly ties the $\overline{\text{BD_PRST}}$ pin to ground. This signals the Hot

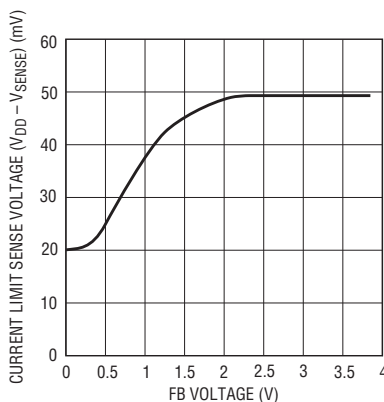


Figure 6. To protect against excessive power dissipation in the switch, the current limit folds back or drops as a linear function of the output voltage, which is sensed at the FB pin.

Swap controller to begin a power-up sequence.

If the LTC4260 shuts down due to a fault, it may be restarted by simply removing and reinserting the card. There is an internal 10μA pull-up current source on the $\overline{\text{BD_PRST}}$ pin. When the card is removed and re-inserted the $\overline{\text{BD_PRST}}$ pin is pulled high then low which clears the offending fault and begins a new power-up sequence.

Fast Current Limiting Isolates Faults and Protects Backplane Voltage

The LTC4260 features an adjustable current limit with foldback that protects against excessive power dissipation in the switch during active current limit. The current limit level is set by the value of the sense resistor located between the V_{DD} pin and the SENSE pin. When the load current exceeds the current limit, the LTC4260 regulates the GATE pin voltage to keep the current through the sense resistor at a constant value.

The response time to an overcurrent depends on the sense voltage, as shown in Figure 5. In the case of a short circuit in the load, the current is brought under control in less than 1μs. The GATE pin is pulled down with a 600mA GATE-to-SOURCE current.

To protect against excessive power dissipation in the switch, the current limit folds back or drops as a linear function of the output voltage, which is sensed at the FB pin. The current limit threshold as a function of output voltage is shown in Figure 6.

An overcurrent circuit breaker limits the time the part is in active current limit. While the LTC4260 is

in active current limit the capacitor C_T is charged with a 100μA pull-up current. If the voltage at the TIMER pin reaches 1.235V, the part turns off the pass transistor and records an overcurrent fault. Figure 7 shows output short waveforms.

Control Board Power through I²C Interface

The LTC4260 features seven registers as shown in Table 2. The control register sets the state of the pass transistor and controls whether the part automatically attempts to turn on after certain faults or stays in the latched off state.

One bit in the control register sets the ADC to test mode, where a host processor can write into the ADC registers. The mass write feature, which allows the use of a special I²C address to write to all LTC4260s at once, can be masked using a bit in the control register.

The control register also configures the behavior of the general purpose input/output (GPIO) pin. At power-up the GPIO pin defaults as a powergood indicator. Other uses for the GPIO pin are as a power-bad indicator, general logic input pin or a general logic output pin.

The alert register sets which faults interrupt the host. There are control bits for each specific fault allowing the $\overline{\text{ALERT}}$ pin to pull low when that fault occurs. At power-up the default state is to not alert on faults. After the bus master controller broadcasts the Alert Response Address, the LTC4260 responds with its address on the SDA line and releases $\overline{\text{ALERT}}$.

Collecting Fault Information Aids Diagnosis

After a board fault occurs, diagnosing the problem is simplified by checking the LTC4260's onboard fault information. The fault and status registers contain a record if faults are present or have occurred and can be accessed through the I²C interface.

Three major faults can turn off the pass transistor: overcurrent, undervoltage and overvoltage. An

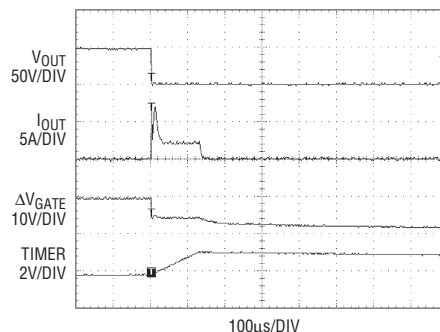


Figure 7. Short circuit waveforms

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1.2MHz, 2A, Monolithic Boost Regulator Delivers High Power in Small Spaces

by Kevin Ohlson

Introduction

Even as cell phones, computers and PDAs shrink, they require an increasing number of power supply voltages. The challenge, of course, is how to squeeze more voltage converter circuits into less space—without sacrificing power or efficiency. Boost converters, in particular, are becoming more prevalent, as main supply voltages are lowered to accommodate core logic circuits, while many components require a higher supply voltage. The LTC3426 boost converter meets the challenge with converter-shrinking features, including a low $R_{DS(ON)}$ monolithic switch, internal compensation and a 3mm × 3mm × 1mm ThinSOT package. The LTC3426 operates at high frequency and therefore works with small, low cost inductors and tiny ceramic capacitors.

The LTC3426 incorporates a constant frequency current mode architecture, which is low noise and provides fast transient response. With

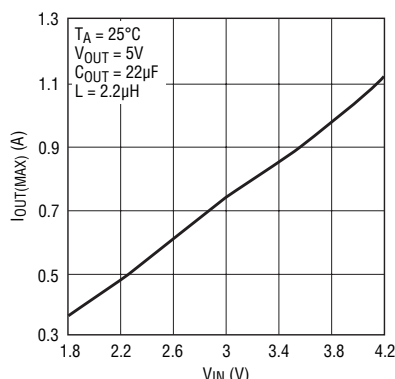


Figure 1. High current outputs are attainable with minimum 2A switch limit.

a minimum peak current level of 2A, the LTC3426 delivers up to 900mA of output current. Figure 1 shows the converter's output current capability at 5V as a function of V_{IN} with peak inductor current at 2A. An input supply range of 1.6V to 4.3V makes the LTC3426 ideal for local supplies ranging from 2.5V to 5V. Efficiencies above 90% are made possible by its low 0.11 Ω (typ.) $R_{DS(ON)}$ internal switch.

There is no need for an external compensation network because the LTC3426 has a built-in loop compensation network. This reduces size, lowers overall cost and greatly simplifies the design process. Figure 2 shows the V_{OUT} response to a 250mA-to-500mA load step in a 1.8V to 3.3V application.

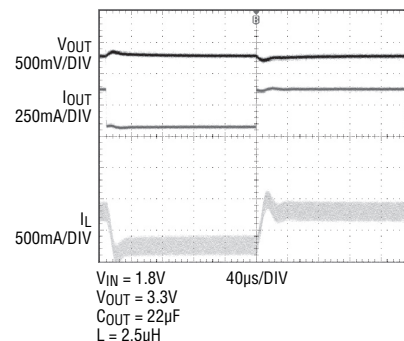


Figure 2. Fast transient response to load step of 250mA to 500mA

The Shutdown input can be driven with standard CMOS logic above either V_{IN} or V_{OUT} (up to 6V maximum). Quiescent current in shutdown is less than 1 μ A. A simple resistive pull-up to V_{IN} configures the LTC3426 for continuous operation when V_{IN} is present.

3.3V Output 800mA Converter

Some applications require local 3.3V supplies which are utilized periodically yet are required to deliver high currents. The LTC3426 is an ideal solution which requires minimal board space and, when in shutdown, draws less than 1 μ A quiescent current. Figure 3 shows a circuit which delivers up to 800mA at 3.3V from a 2.5V input. This circuit also works with V_{IN} down to 1.8V with 750mA output. The output voltage is easily programmed by changing the feedback ratio of R1 and R2 according to the formula:

$$V_{OUT} = 1.22V \cdot \left(1 + \frac{R1}{R2}\right)$$

Lithium-Ion 5V Boost Converter

Some portable applications still require a 5V supply. Figure 4 shows a circuit which operates from a single Lithium-Ion battery and delivers at

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DESIGN IDEAS

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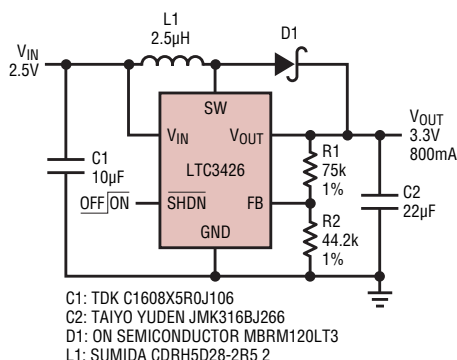


Figure 3. Application circuit for 3.3V output delivers 800mA

3-Phase Buck Controller for Intel VRM9/VRM10 with Active Voltage Positioning

by Xiaoyong Zhang

Introduction

Each new generation of CPUs demands more from power supplies than the last: more power, tighter voltage regulation and faster transient response. Meeting all of the new requirements is a difficult proposition, but the LTC3738 helps power supply designers do just that. It is a 3-phase buck controller with active voltage positioning specifically designed for Intel VRM9 and VRM10 (Figure 1).

High Power and Thermal Management

The LTC3738 can easily work with the 3-phase LTC3731 to form a 6-phase (up to 12-phase interleaved) power supply to deliver more than 100A current to its load. For such high currents, proper thermal management is

crucial. The current mode architecture of the LTC3738 evenly distributes the load, and thus thermal stress, across the channels. This improves the thermal performance and reliability of the entire power solution. The LTC3738 also includes a thermal detector that generates a VR_HOTB warning signal when chip itself gets hot (around 120°C) plus a self-protect thermal detector that shuts down the device when chip becomes extremely hot and endangers the safety of the power supply. The LTC3738 also has a comparator for external thermal detection. Power designers can put thermal detection resistors at the hottest spot on the board and let the LTC3738 send a VR_HOTB signal to the CPU when its thermal comparator trips.

Accurate Load Line Control

The tight load line window of the VRM9/VRM10 specification asks for accurate static and dynamic voltage control. The $\pm 1\%$ DC regulation accuracy and precise programmable active voltage positioning of LTC3738 helps power designers meet the load line window easily. The unique active voltage positioning solution of LTC3738 makes the load line slope control easy and accurate. The slope is programmed by the ratio of two external resistors. LTC3738 senses true load current including ripple current of all three channels and generates an accurate AVP control voltage. The precise regulation of the LTC3738 gives more range for output voltage ripple. Hence power designers can use smaller output capacitor values

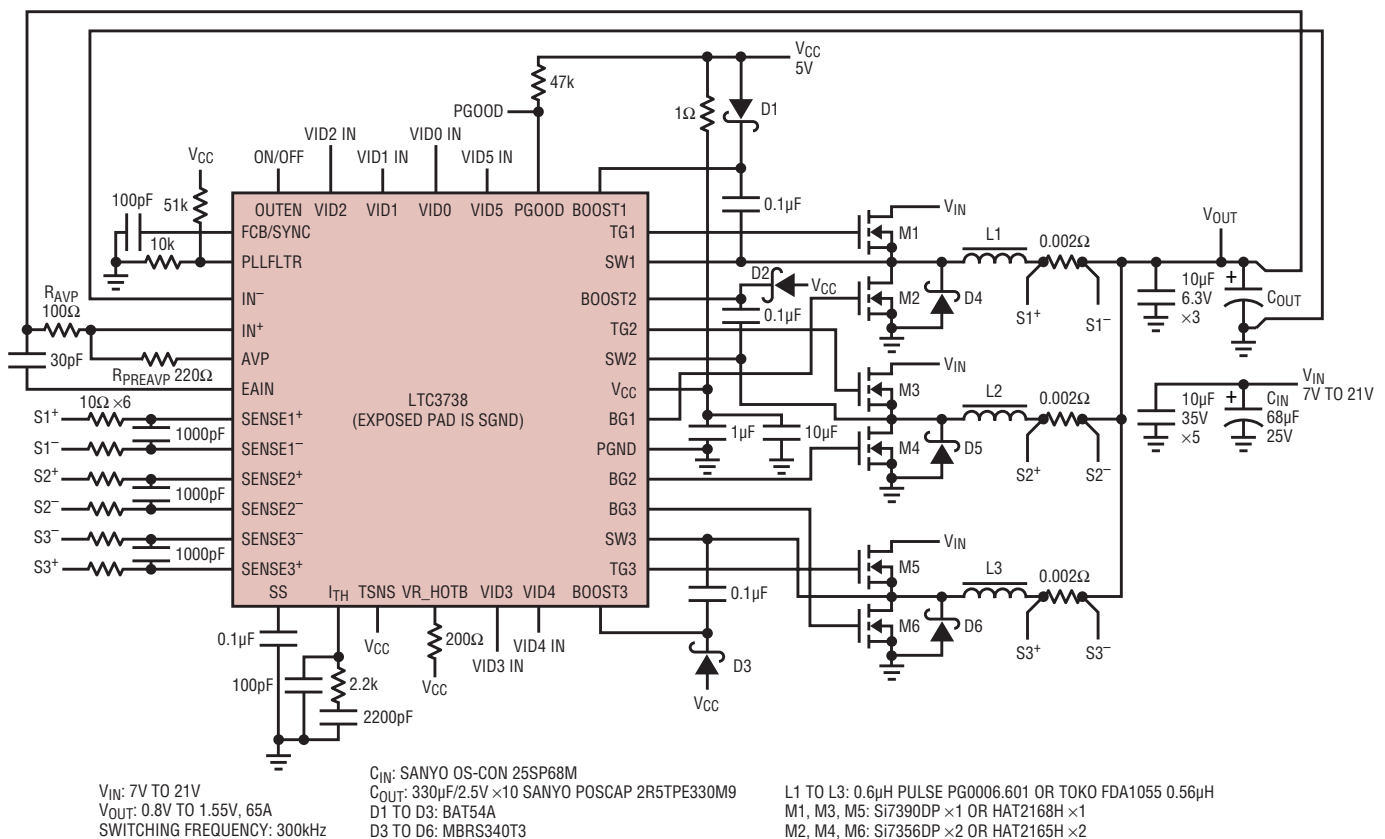


Figure 1. This 3-phase power supply manages the thermal problems inherent in high current Intel VRM10 applications.

and lower their total solution cost. Smaller output capacitor values also speed up the changing of the output voltage when the CPU generates a different VID code.

Other Features

The LTC3738 has a differential amplifier for remote sensing of both the high and low sides of the output voltage.

There is no reverse current during start-up, which allows the LTC3738 to power up into a pre-biased output without sinking current from the output. The LTC3738 also has a defeatable short-circuit shutdown timer. Three operation modes—PWM, pulse skip and Stage Shedding™—allow power supply designers to optimize for efficiency and noise.

Conclusion

LTC3738 is specifically designed to simplify power supply designs for Intel VRM9/VRM10 applications. It is a complete power supply solution with essential thermal management features, accurate load line control, precise output voltage sensing, and comprehensive fault protection.

LT6650, continued from page 17

from affecting the output by the inclusion of a pre-regulating Zener diode. With this extra input decoupling and the LT6650 circuitry operating from a 12V bus, 50V transients induce less than 0.5% V_{OUT} perturbation.

To obtain the micropower performance of the LT6650, quiescent currents of the internal circuitry are minute, which by nature, results in a higher output impedance than traditional references. Since output impedance is inversely related to the output stage operating current, a modest additional load current can easily reduce the output impedance by an order of magnitude from the unloaded case. Thus in applications where the output impedance and noise must be minimized, a light DC loading of the output provides enhanced performance. This loading can exist naturally in the application, or the feedback resistors can be designed to provide it. For example, setting the gain resistor value to 10k Ω establishes a moderate $I_{OUT} = -40\mu A$ and decreases the output peak resistance value from

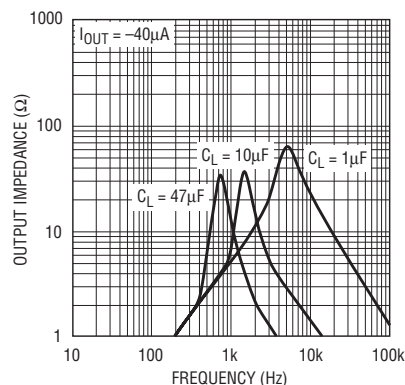


Figure 7. Output impedance is reduced while sourcing moderate current (40 μA).

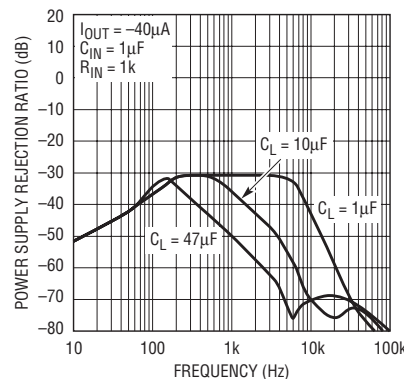


Figure 5. Improved supply noise rejection of Figure 4 reference circuit

hundreds of ohms to the tens of ohms shown in Figure 7.

Shunt-Mode Reference

When the output voltage is tied to the input voltage, the high side of the rail-to-rail buffer amplifier is effectively disabled and only the low side remains active. In this mode of operation the LT6650 operates as a shunt reference as shown in Figure 8. Any shunt reference voltage from $\pm 1.4V$ up to $\pm 18V$ can be established by the feedback resistor selection. The noise and load capacitors have the same functions as in the series mode of operation. A 10 μF minimum load capacitance is recommended for best stability and transient response. In shunt mode, an external biasing resistor R_B is connected from

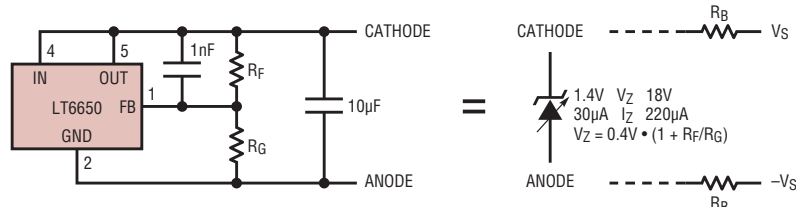


Figure 8. Create your own adjustable micropower “zener” 2-terminal reference.

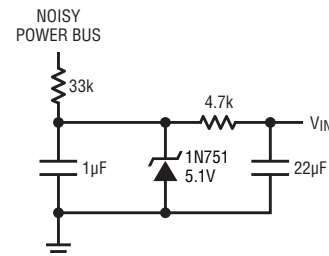


Figure 6. High noise-immunity input network allows 50V transients on automotive power bus.

the power supply to the output, and delivers all the current required for supplying the LT6650 and the load current. R_B is selected to ensure the operating current of the reference (I_Z in the Figure 8 zener-diode analogy) is in the range of 30 μA to 220 μA under all loading conditions.

Conclusion

The LT6650 voltage reference incorporates a unique blend of low voltage, micropower operation and functional versatility. With the additional features of series and shunt mode configurability, source and sink output current, wide output voltage range, adjustability, and a tiny ThinSOT-23 package, the LT6650 provides an excellent solution to the many design challenges in both portable and industrial voltage control.

Redundant 2-Wire Bus for High Reliability Systems

by John Ziegler

Introduction

The effort to achieve high reliability in data processing, data storage and communication systems has necessitated the use of circuitry to monitor parameters such as temperature, fan speed, and system voltages. These circuits often communicate through 2-wire serial buses, such as SMBus or I²C. Redundant subsystems are important in high reliability systems, and the 2-wire bus subsystem is no exception. High reliability 2-wire bus systems incorporate two master controllers in a redundant configuration, to maintain system operation if one master fails or is removed. In a redundant configuration, each master

is connected to its own 2-wire bus, while all of the slaves are connected to a single downstream redundant bus. Either master can take control of the redundant bus at any time.

Figure 1 shows a circuit using two LTC4302's, each dedicated to a master, to allow either master to take control of a redundant 2-wire bus. The LTC4302's GPIO pins default to a high impedance state at power-up, so that 10K pull-up resistors R5, R6 and R13 set each GPIO voltage high. With each LTC4302's GPIO1 pin connected to the CONN pin of the other, both LTC4302's are active at power-up and can be accessed via their SDAIN and SCLIN pins.

In this configuration, each master can take control of the downstream redundant bus with two Write Byte operations to its dedicated LTC4302. In the first operation, the master activates the connection to the downstream redundant bus, and writes both of its GPIO pins low. With the GPIO1 pin low, the other master is disconnected from the redundant bus and is also prevented from communicating with its LTC4302. In the second operation, the master writes a logic high to its GPIO1 pin, so that the other master is again free to communicate with its LTC4302. Using this technique, the common GPIO2 pin is low whenever one of the masters is connected to the

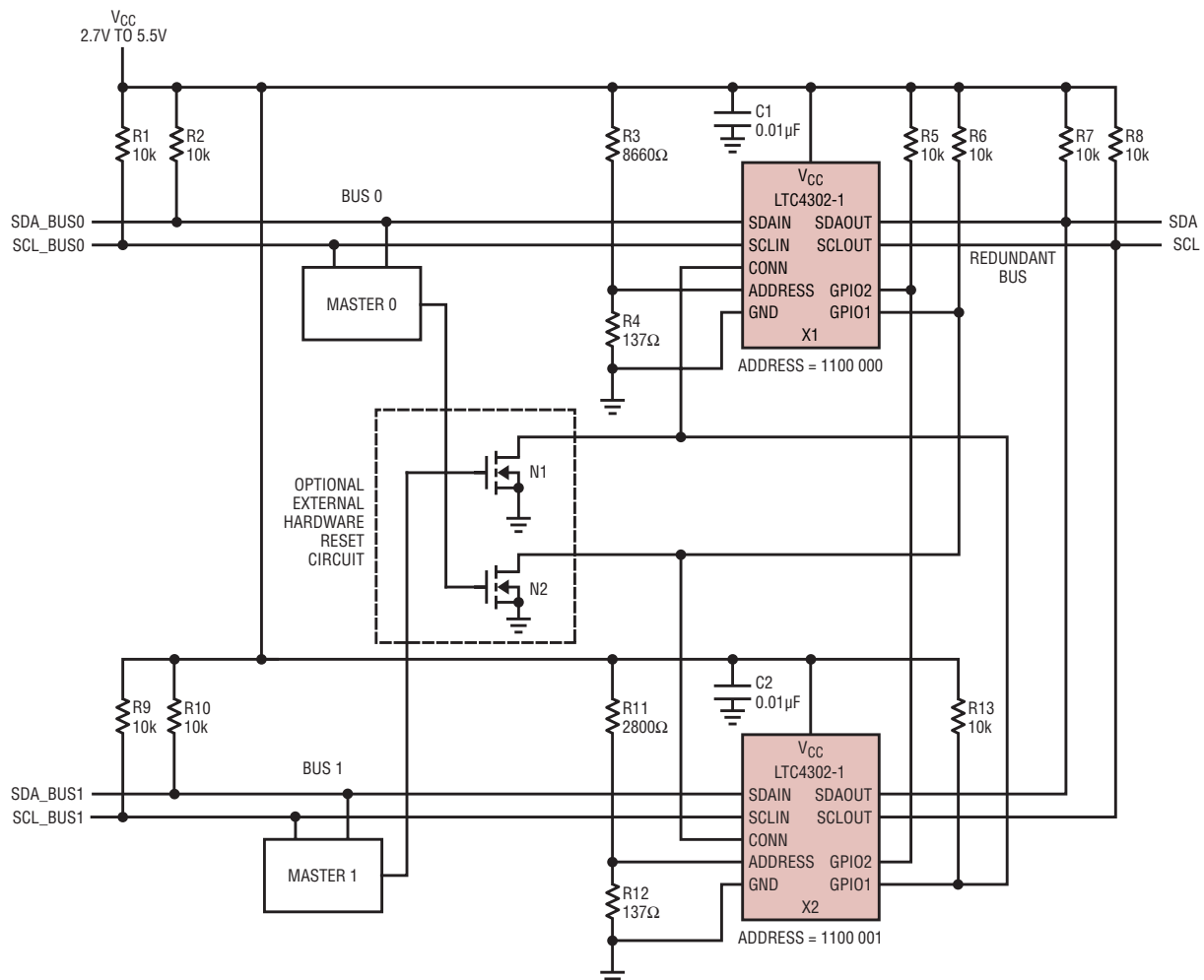


Figure 1. Two LTC4302s in a redundant bus application, with a hardware reset on the CONN pins

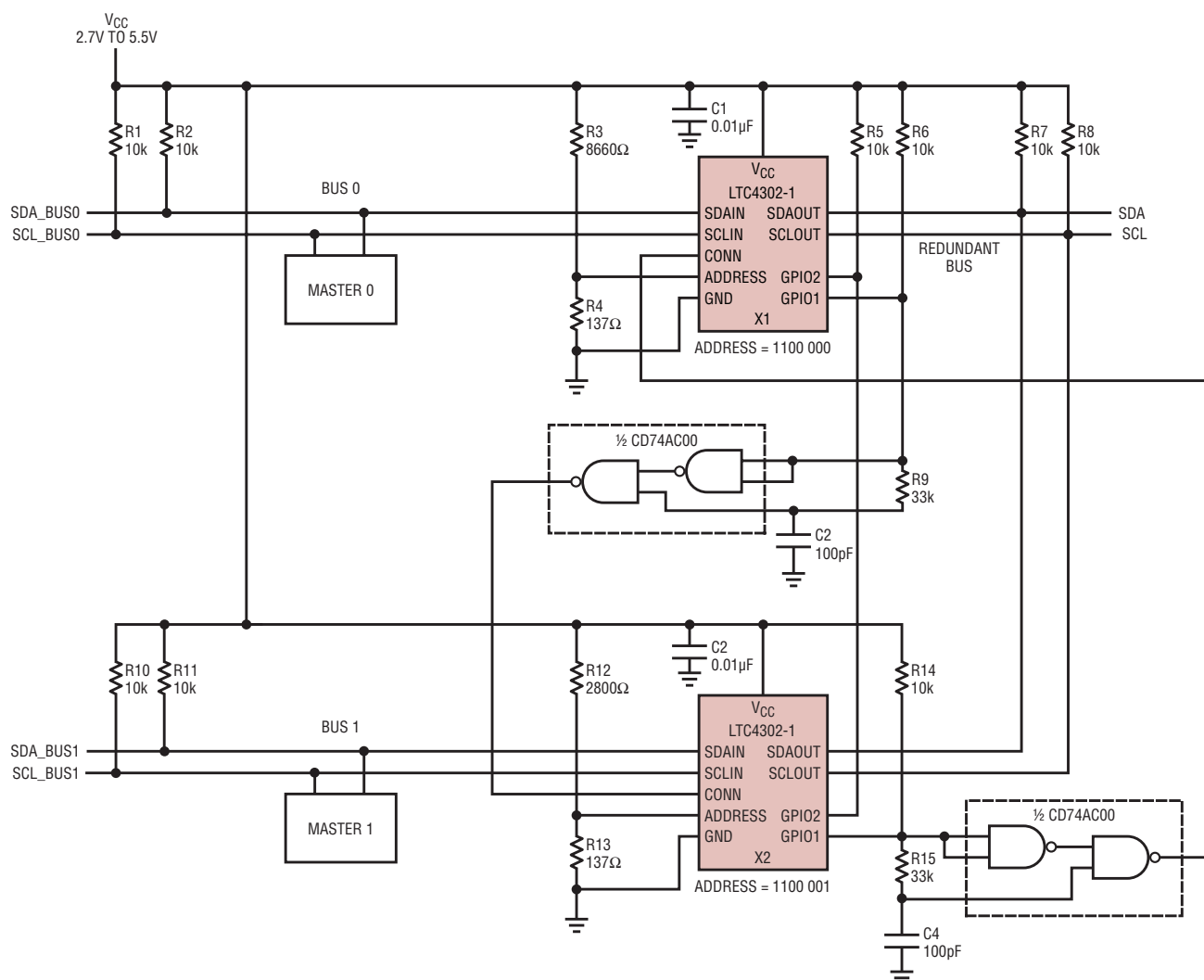


Figure 2. Alternate implementation of two LTC4302s in a redundant bus application, with lock-up prevention circuitry.

redundant bus, so that each master can read its LTC4302 to determine whether the other master has control of the redundant bus.

Either master can take control of the redundant bus at any time except under two conditions. First, if a master tries to access its LTC4302 and receives no Acknowledge signal, it knows that the other master has completed the first Write Byte operation, but has not yet re-written its GPIO1 pin back high. Second, if both masters try to connect to the redundant bus within 100ns of each other, both are connected to the bus temporarily, and are then disconnected.

A disadvantage of this scheme is that two separate write operations are required for a master to take control of the downstream bus properly. After the first operation, the new master has

control of the redundant bus, and the other master cannot access its own LTC4302 because its CONN pin is low. If the new master is removed from the system, or if its 2-wire bus locks up before it can complete the second write operation to write a logic high to its GPIO1 pin, then the other master is permanently prevented from taking control of the redundant bus through the 2-wire interface. An externally controlled pull-down device would have to be used to pull the CONN pin of the new master low, as shown by N-Channel MOSFET transistors N1 and N2 in Figure 1.

Figure 2 shows an alternative approach to solve this problem. Each master can take control of the redundant bus using a single Write Byte operation. For example, Master 0 commands its LTC4302 to connect

to the redundant bus and also to force logic lows on both of its GPIO pins. When its GPIO1 pin transitions high-to-low, the circuit formed by R9, C2 and the two two-input NAND gates generates a negative pulse on the other LTC4302's CONN pin. The duration of the pulse is set by the $R9 \cdot C2$ time constant and is roughly 3.3 μ s. Pulsing CONN low resets the registers of the LTC4302 to their default states, thereby disconnecting Master 1 from the redundant bus. After 1 μ s, Master 1's CONN pin returns high, and Master 1 is again free to take control of the redundant bus.

The LTC4302 also provides bi-directional buffering, keeping the capacitances of the master buses and the redundant bus isolated from each other. Rise time accelerator circuitry

continued on page 32

-48V Backplane Impedance Analyzer Takes the Guesswork Out of Sizing Clippers and Snubbers

by Mitchell Lee

It comes as something of a surprise to most engineers that the -48V power distributed on a backplane exhibits a decidedly inductive impedance. Considering bypass capacitors are often excluded from the backplane, coupled with the lengthy path back to the -48V battery or power supply source, it seems unavoidable. The consequences of an inductive driving

point impedance are twofold: first, for reasons entirely cosmetic, the ringing associated with insertion and other transient events are undesirable. Second, input reaction to high di/dt conditions presents correspondingly high input voltage surges, placing the Hot Swap MOSFET as well as the operation of the Hot Swap controller at risk.

To mitigate these effects a network comprising a clamping element in parallel with a snubber is often found in successful circuit implementations, as seen in Figure 1. D3 serves to clamp input reaction and the R8-C8 snubber eliminates ringing¹. Figure 2 shows the before-and-after results of adding clamping and snubbing, under conditions of insertion and circuit breaker action.

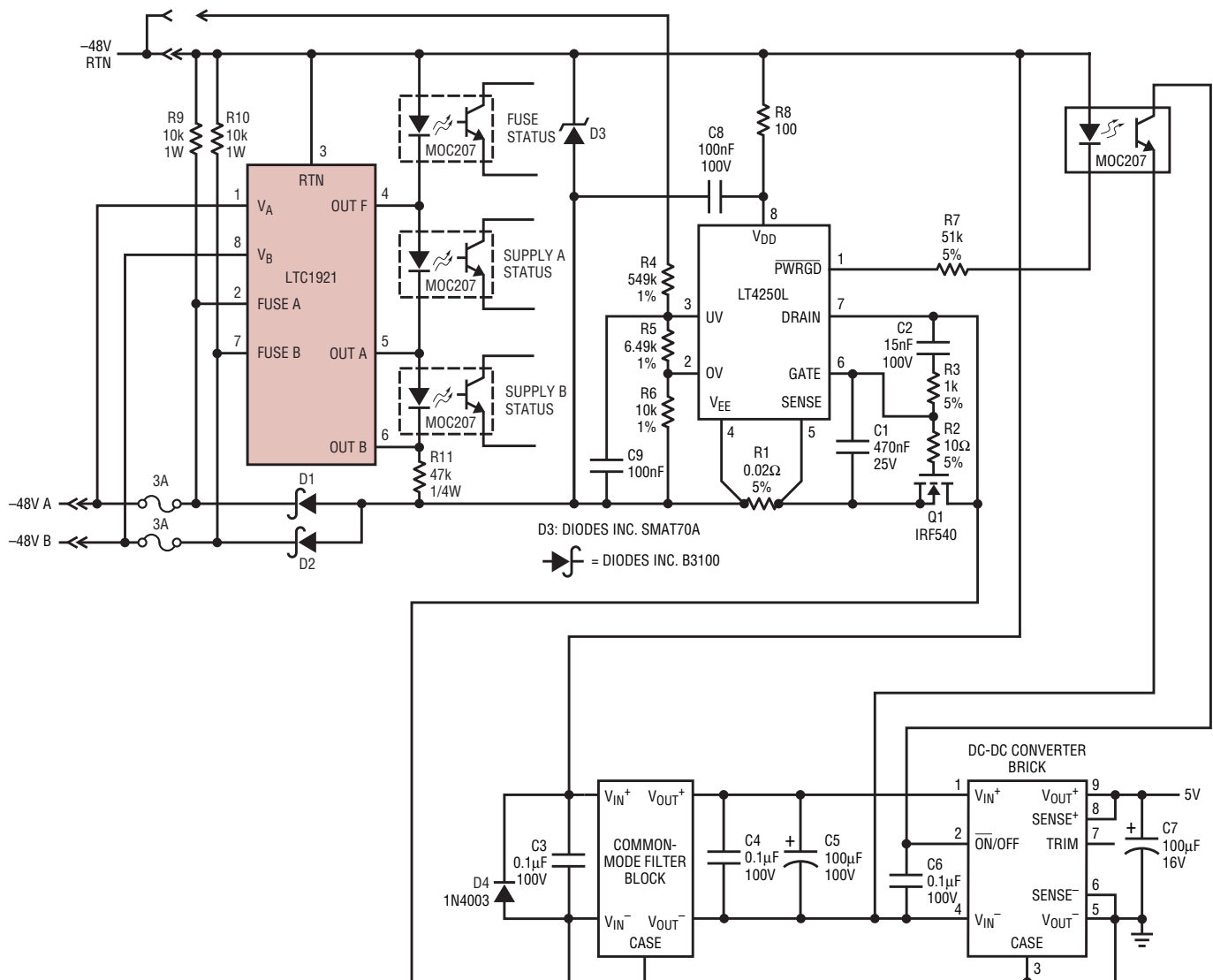


Figure 1. This 75W, -48V telecom supply monitor and Hot Swap controller includes a clamp (D3) to control high voltage surges, and a snubber (R8-C8) to eliminate voltage ringing after transient events, like card insertion. It is important to take the backplane impedance into account when sizing these components.

Before

Without a clamp and a snubber, MOSFET Q1's drain-source capacitance C_{OSS} resonates with little loss against the inductance of the -48V backplane distribution bus. The presence of Schottky diodes D1 and D2 complicate matters, but at best the diode in line with the lowest magnitude input voltage adds capacitance in parallel with C_{OSS} , and at worst the active diode peak detects the input ring, storing the energy (and high voltage) on C_{OSS} . Because C_{OSS} exhibits a strong voltage dependency, the peak ring voltage at insertion can avalanche the MOSFET or the LT4250. The 200V transient input rating of the LTC1921 generally keeps it out of harm's way. The energy available at the peak voltage is limited, and rarely is the source of destruction.

If the circuit breaker function of the LT4250 is invoked by a sustained overload, the inductance of the -48V wiring is loaded with $\frac{1}{2}Li^2$, which represents a potentially destructive energy. The energy is high enough to drive something, usually the MOSFET, into avalanche as shown by the flattened portion of the waveform. Once the input current drops to zero, the remaining energy rings off in a manner not dissimilar to the insertion phase of operation.

After

The addition of a clamping diode and R-C snubber eliminates the aforementioned high voltage transients. At insertion, ringing is eliminated and overshoot controlled by the R8-C8 snubber of Figure 1. Input reaction during a circuit breaker event is clamped to a safe level by D3, a transient suppression diode. Subsequent ring-off and attendant noise burst is again controlled by the snubber.

To quantify the stored energy and to optimally size the snubber and clamping components, one must know something about the magnitude of inductance in the -48V feed. Measuring this impedance is problematic, given the risk inherent in connecting a sensitive, costly piece of test equipment such as an HP4815A to a multi-kW

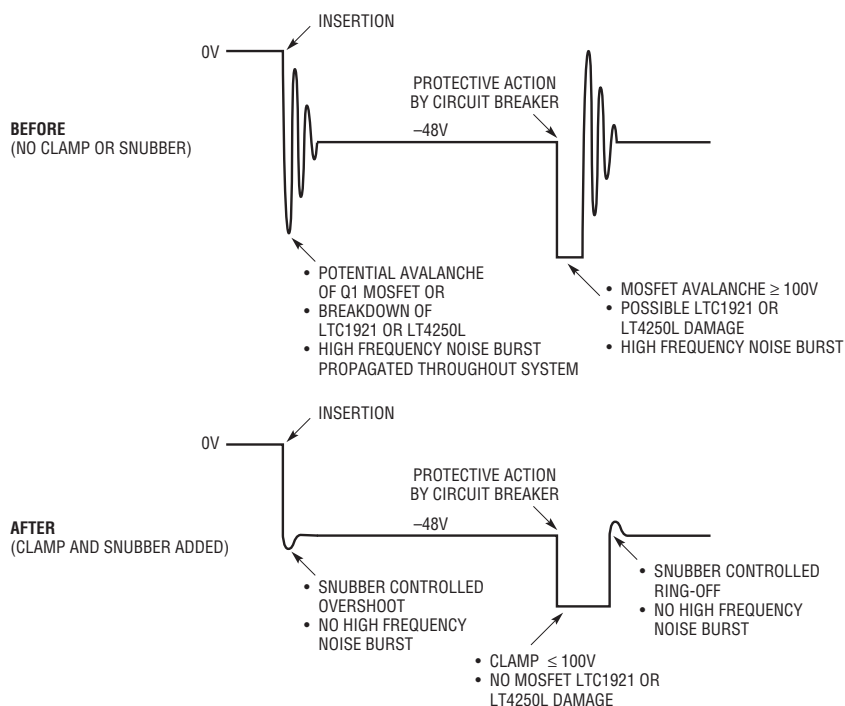


Figure 2. The before-and-after of adding clamping and snubbing to a hot swapped card, plugged into a -48V power distributed bus, under conditions of insertion and circuit breaker action.

-48V supply bus. Fortunately there is an easier risk-free way to get the required information, using a simple oscillator circuit where the unknown inductance resonates with a known capacitance. In all but extreme cases this method gives results adequate for quantifying the inductance of the -48V feed.

Simple Test Oscillator

Figure 3 shows a test circuit that, with the aid of a frequency meter², can measure the inductance of the -48V supply line. The circuit is essentially a Colpitts oscillator, where both the resonating inductance and power are furnished by the -48V bus. The capacitive arm

of the oscillator comprises C1 and C2, with the tap at the junction of C1 and C2 feeding the emitter of Q1. Coupling is set to accommodate inductances down to $\approx 100\text{nH}$. Base components provide bias and bypassing, while R3 and R4 establish an emitter current of approximately 11mA, operating the transistor in a region of favorable frequency. Two resistors are utilized in the emitter circuit to distribute dissipation and permit use of common quarter-watt units. A tiny, off-the-shelf current transformer couples signal to a 50 Ω termination at a frequency counter.

Measurements are made by plugging the test circuit into a -48V

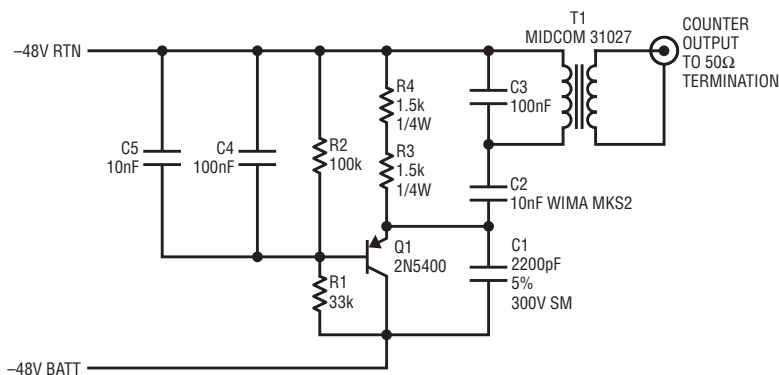


Figure 3. Test oscillator for evaluating -48V driving point inductance

backplane, picking up -48V BATT and -48V RTN and measuring the oscillator frequency. The loop inductance between these two points together with the circuit capacitance determines the frequency of oscillation.

It is important to transformer couple the output signal so that the frequency counter is not grounded to -48V RTN. First, there is concern about DC ground loops since -48V RTN is not earth or chassis ground. Second, if the -48V RTN is contaminated with noise, it could contaminate the oscillator frequency measurement. Third, -48V RTN contributes its own share of inductance, and this would be disturbed by the introduction of the frequency counter's ground at that point. Transformer coupling eliminates these issues.

Use

The oscillator circuit is most usefully constructed on a small circuit board complete with a backplane power connector and a BNC for frequency counter attachment. This assembly is then plugged into the backplane to measure the inductance of the -48V feed. Characterization of various slots and backplanes proceeds quickly as the test circuit is moved from one connector to the next and the frequency logged. The measured inductance varies widely depending on the presence of adjacent cards or noise filters, distance to the power source, backplane and bus bar construction and so on.

Inductance is calculated from the measured frequency of oscillation using the basic relation

$$\frac{1}{\omega^2} = L \cdot C_0 \quad (1)$$

where ω is the radian frequency of oscillation and C_0 is the oscillator's total equivalent capacitance at the collector of Q1.

The capacitance C_0 is roughly

$$C_0 = \frac{C_1 \cdot C_2}{(C_1 + C_2)} \quad (2)$$

$$\text{or } C_0 = \frac{2.2 \cdot 10}{(2.2 + 10)} = 1.803 \text{ nF}$$

For example, a measurement taken on the author's test oscillator produced the following results (a rearrangement of equation (1)):

$$L = \frac{1}{1.803 \text{ nF} \cdot (2 \cdot \pi \cdot 2.9376 \text{ MHz})^2} \quad (3)$$

$$L = 1.63 \mu\text{H}$$

Note that for the purposes of designing snubbers and selecting transient clamps, a value of $C_0 = 1.8 \text{ nF}$ yields acceptable results when calculating L.

Calibration

Accumulated tolerances in oscillator components, as well as the performance of the transistor, affect the value of C_0 and therefore the accuracy of the previous calculations. While the approximate value of 1.8 nF is entirely adequate for snubber designs, a potentially more exacting figure for the "correlation" capacitance is easily computed (without the need for a standard inductor) using the following method.

First, attach a $1 \mu\text{H}$ to $10 \mu\text{H}$ inductor between the collector of Q1 and a

-48V bench supply (see Figure 4). To eliminate erratic readings caused by test lead inductance, bypass the -48V supply at the inductor. Measure the resulting frequency, f_1 . Now add a capacitor C_X of 1 nF to 4.7 nF to Q1's collector and measure the new frequency, f_2 . The two operating conditions are related by manipulating equation (1) to eliminate inductance. Thus

$$\frac{1}{\omega_1^2 C_0} = \frac{1}{\omega_2^2 (C_0 + C_X)} \quad (4)$$

$$C_0 = \frac{C_X}{(\omega_1/\omega_2)^2 - 1} \quad (5)$$

The author's setup measured $f_1 = 2.9376 \text{ MHz}$ and $f_2 = 1.5663 \text{ MHz}$ ($C_X = 4.7 \text{ nF}$); from equation (5) C_0 was apparently 1.866 nF , or about 3.5% higher than calculated from equation (2) and the components' marked values.

This calibration method is independent of the test inductor, but limited by the accuracy of the extra capacitor, C_X . A 5% silver mica unit is sufficient to give verification of equation (2). This figure improves if C_X is first measured with an accurate capacitance meter to establish a more exacting value.

Calibration for Advanced Users

A series of measurements made with several C_X calibration "standards" can help statistically improve the accuracy of C_0 , or at least increase the user's faith in the perceived value. Again no "standard" inductor is necessary, only a fixed unit that doesn't change value between readings.

A series of such measurements taken by the author are shown in Table 1, and the data are plotted in Figure 5. It is easy to see the straight-line relationship between total oscillator capacitance ($C_0 + C_X$) and $1/\omega^2$, and it is that relationship which allows us to graphically deduce C_0 from the x-axis intercept.

In this case there is fair graphical agreement with the values calculated from equations (2) and (5), as the line appears to cross zero at $\approx 1.8 \text{ nF}$. A curve-fitting utility in the graphing

Table 1. Data taken with a series of 5%, silver mica capacitors

C_X (nF)	f (MHz)	$1/\omega^2$ (Radians ⁻²)
0	2.9376	$2.9353 \cdot 10^{-15}$
1	2.4004	$4.3962 \cdot 10^{-15}$
2.2	2.0132	$6.2498 \cdot 10^{-15}$
3.3	1.7742	$8.0470 \cdot 10^{-15}$
4.7	1.5663	$10.325 \cdot 10^{-15}$

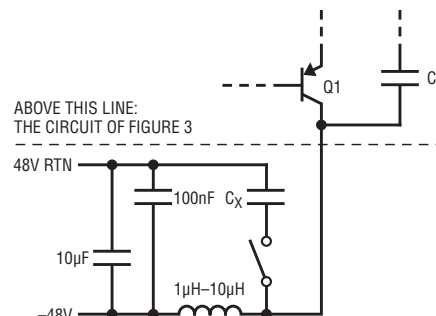


Figure 4. Attaching C_X to the circuit of Figure 3 for testing oscillator capacitance.

program predicted an intercept of 1.813nF.

This method is really just an extension of the calculation made in equation (5); it's just that equation (5) was a 2-point approximation, while here we have extended it to 5 points.

Measuring Inductance and Capacitance

A second purpose for plotting oscillator capacitance against $1/\omega^2$ is that it also resolves distributed capacitance inherent in the backplane and wiring harness.

Using the method of Table 1 and Figure 5 as a starting point, suppose the oscillator is now connected to a backplane and the same sequence of measurements made as C_X varies. To facilitate measurements, several C_X capacitors are mounted on the test oscillator card and selected with a switch or jumpers. A new set of data as plotted in Figure 6 results.

Again, with the aid of a straight edge or curve-fitting utility, the x-axis intercept is found to be 3.081nF. This value is the sum of the oscillator's built-in capacitance C_O , plus the capacitance contributed by the backplane and wiring harness. Removing C_O , we find that

$$C_{\text{DISTRIBUTED}} = 3.081 - 1.813 = 1.268\text{nF}$$

$$\text{and } L = \frac{4.94 \cdot 10^{-15}}{3.081\text{nF}} = 1.6\mu\text{H}$$

Note that the inductance calculation uses the frequency value found at $C_X = 0$, but uses the projected capacitance of 3.081nF at $1/\omega^2 = 0$.

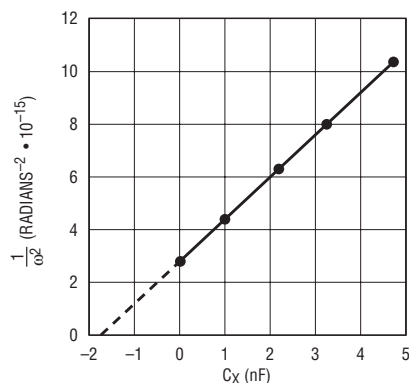


Figure 5. Graphical extrapolation of oscillator capacitance and oscillator capacitance from a series of frequency measurements with different external capacitors.

Snubber Design

For Hot Swap controller circuits such as shown in Figure 1, it is well to size the snubbing capacitance C_8 to be 10 times all other circuit capacitances combined. In Figure 1, capacitance is contributed by circuit board traces (small, usually neglected), D_3 (400pF), and perhaps one of the input diodes (100pF in D_2 , for example). The largest contributor is Q_1 , weighing in at 1500pF under zero bias, and 250pF at 50V. Assuming 500pF as the effective value, the total capacitance to be snubbed in Figure 1 is approximately 1nF, leading us to a value of 10nF for the snubbing capacitor. If we include the backplane capacitance measured in Figure 6, a value of $10 \cdot 2.268\text{nF}$ or $\approx 22\text{nF}$ is adequate.

The snubbing resistor R_8 is sized so that the circuit Q is a conservative 0.1 and the effects of circuit capacitance are nullified. Q is given by the equation

$$Q = \frac{1}{R_8} \sqrt{\frac{L}{C_8}} \quad (6)$$

Setting $Q = 0.1$ and rearranging equation (6) for our special case with $C_8 = 22\text{nF}$ and $L = 1.6\mu\text{H}$ gives

$$R_8 \approx \frac{1}{0.1} \sqrt{\frac{1.6 \cdot 10^{-6}}{22 \cdot 10^{-9}}} \quad (7)$$

$$\text{or } R_8 \approx 85\Omega$$

This is where our measurement of "L" comes in handy—to compute the damping necessary to control Q . With standard values of $R_8 = 82\Omega$ and $C_8 = 22\text{nF}$, ringing is eliminated

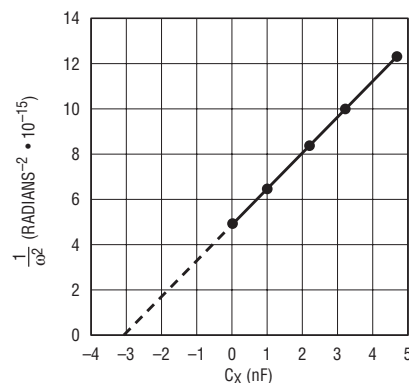


Figure 6. Graphical extrapolation of backplane capacitance and oscillator capacitance from a series of frequency measurements with different external capacitors.

and overshoot is limited to less than 100V_{PK} during initial insertion on a 48V supply.

Nevertheless, R_8 and C_8 have different values in Figure 1. C_8 has been increased in value to serve as a hold up capacitor in the event the input supply collapses, thereby guaranteeing operation of the LT4250 circuit breaker and MOSFET shut-off. The operating Q of C_8 and R_8 in Figure 1 is ≈ 0.04 .

Input Clamp, D3

Again referring to Figure 1, D_3 is sized to handle the energy stored in the backplane and wiring harness inductance. Sticking with 1.6μH, suppose the peak input current reached 50A during a zero-ohm failure of C_3 . The energy stored in the -48V input inductance is given by

$$E = \frac{1}{2} L i^2$$

$$\text{or } E = 0.5 \cdot 1.6\mu\text{H} \cdot 50^2 = 2\text{mJ}$$

Examination of the SMAT70A data sheet reveals that this device handles in excess of 200mJ; thus it is adequate for this application.

The presence of distributed capacitance on the backplane and in the -48V wiring harness plays an interesting role. First, the snubber must be oversized to account for the hindrance of this extra capacitance as we saw in earlier calculations (equation (7)). Second, the distributed capacitance helps the clamp D_3 by absorbing some of the inductive energy, although 1.268nF absorbs less than 5μJ in this example. From this we can conclude that any distributed "parasitic" capacitance affects the snubber design long before there is any need to account for it in the selection of a clamp.

Conclusion

The test oscillator described here is suitable for measuring backplane and wiring harness inductance in -48V systems in the range of 100nH to 100μH or more. Parasitic capacitance can be measured as well, over a range of less than 100pF to 5nF or more. If the circuit refuses to oscillate you can

continued on page 32

Compact Power Supply Drives TFT-LCD and LED Backlight

by Dongyan Zhou

Introduction

The LT1942 is a highly integrated, 4-output switching regulator designed to power small to medium size TFT panels. Three of the switching regulators provide the TFT bias voltages. The fourth regulator is designed to drive backlight LEDs.

The TFT supply includes two boost converters and one negative output DC/DC converter. Since different types of panels may require different bias voltages, all three output voltages are adjustable for maximum flexibility.

The LED driver is a boost converter that has built-in precise dimming control. The user can choose to drive a single string or two strings of LEDs. A built-in ballast circuit helps to match the LED currents precisely if two strings are used.

All four regulators are synchronized to a 1MHz internal clock, allowing the use of small, low cost inductors and ceramic capacitors. Programmable soft-start capability is available for both the primary TFT supply and LED driver to control the inrush current. The LT1942 is available in a tiny 4mm × 4mm QFN package.

Li-Ion to 4-Inch or 5-Inch TFT-LCD

Figure 1 shows a complete power supply for three TFT bias voltages (AV_{DD} , V_{ON} , and V_{OFF}) and a white LED driver. A typical application of this design is a 4- or 5-inch amorphous silicon TFT-LCD panel powered by a single cell Li-ion input. Two boost converters are used to supply AV_{DD} and V_{ON} , while the negative output converter generates V_{OFF} .

The LT1942 has built-in power sequencing to properly power up the TFT panel. When the shutdown pin is driven above 1V, the AV_{DD} switcher is enabled first. After its output reaches 97% of the set value, the $PGOOD$ pin is driven low, which enables both the V_{OFF} and V_{ON} switchers. A built-in PNP separates the V_{ON} bias supply from its boost regulator output. The PNP is not turned on until the programmable delay set by the CT pin has elapsed. The panel is not activated and stays in a low current state until V_{ON} is present. This delay gives the column drivers and the digital part of the LCD panel time to get ready before the panel is turned on.

The fourth switcher in the LT1942 is a boost regulator designed to drive up to 20 LEDs (in two strings) to power the backlight. Built-in current ballast circuitry keeps the current into LED1 and LED2 actively matched, regardless of the difference in the LED voltage drops. Figure 2 demonstrates the current matching between the two LED strings. The LED regulator has a control pin (CTRL4), which provides both shutdown and dimming functions. If any LED fails open, the output of the LED regulator (D4) is clamped

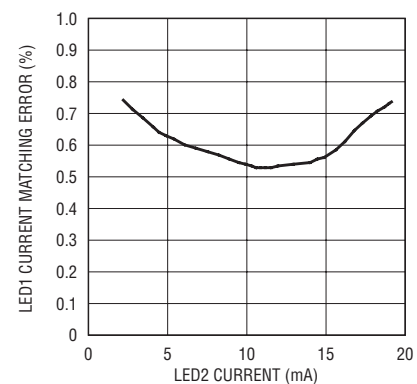


Figure 2. Typical current matching between LED1 and LED2

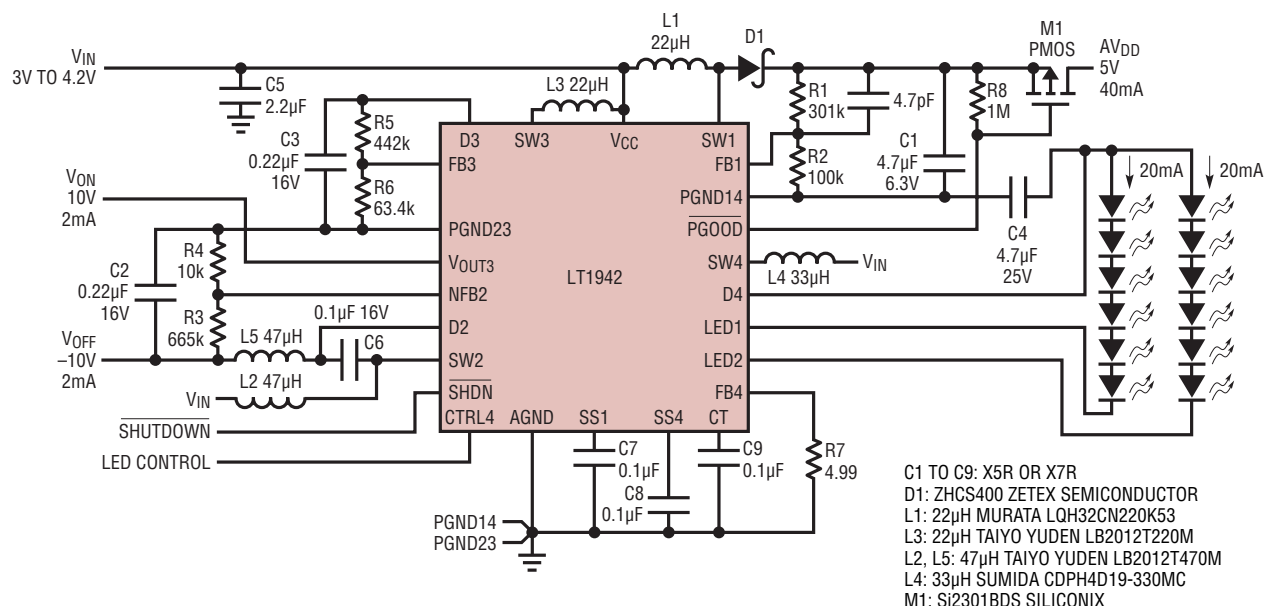


Figure 1. TFT bias voltages and LED backlight power supply from single Lithium-Ion battery input

at around 42V to protect the internal power devices.

Layout Considerations

Proper layout is important to achieve the best performance. Paths that carry high switching current should be kept short and wide to minimize the parasitic inductance. In the boost regulator, the switching loop includes the internal power switch, the Schottky diode (internal or external), and the

output capacitor. In the negative output regulator, the switching loop includes the internal power switch, the flying capacitor between the SW2 and D2 pins, and the internal Schottky diode.

Connect the output capacitors of the AV_{DD} and LED switchers directly to the PGND14 pin before returning to the ground plane. Connect the output capacitor of the V_{ON} switcher to the PGND23 pin before returning to the

ground plane. Also connect the bottom feedback resistors to the AGND pin. Connect the PGND14, PGND23 and AGND pins to the top layer ground and AGND pins to the top layer ground pad underneath the exposed copper ground on the backside of the IC. The exposed copper helps to reduce thermal resistance. Multiple vias into ground layers can be placed on the ground pad directly underneath the part to conduct the heat away from the part. **LT**

LTC3426, continued from page 22

least 750mA from a V_{IN} as low as 3V. When fully charged to 4.2V, over 1A can be supplied. The photograph of a demonstration board in Figure 5 shows just how small the board area is for this application, 10mm × 12mm. Tiny ceramic bypass capacitors and surface mount inductors keep the design small.

Figure 6 shows efficiency exceeding 90% and remaining greater than 85% over a load range from 10mA to 900mA with a fully charged battery.

Component Selection

The LTC3426 requires just a few external components to accommodate various V_{IN} and V_{OUT} combinations. Selecting the proper inductor is important to optimize converter performance and efficiency. An inductor with low DCR increases efficiency and reduces self-heating. Since the inductor conducts the DC output current plus half the peak-to-peak switching current, select an inductor with a minimum DC rating of 2A. To minimize V_{OUT} ripple, use low ESR X5R ceramic capacitors. The average Schottky diode forward current is equal to the DC output current therefore the diode average

current should be greater than 1A. A low forward voltage Schottky diode reduces power loss in the converter circuit.

Conclusion

The addition of the LTC3426 to Linear Technology's high performance boost converter family allows the designer to deliver high current levels with minimal board space. An on chip switch and internal loop compensation reduces component count to provide an inexpensive solution for spot regulation applications. **LT**

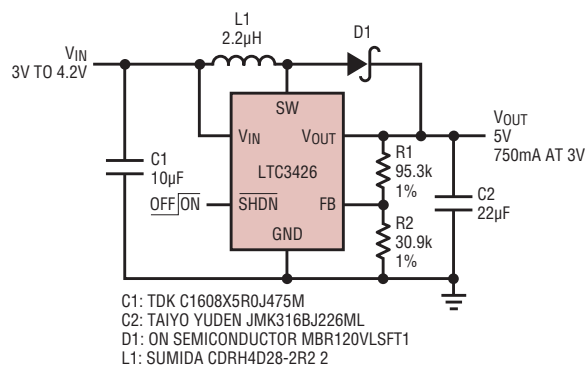


Figure 4. Compact application circuit for V_{OUT} at 5V

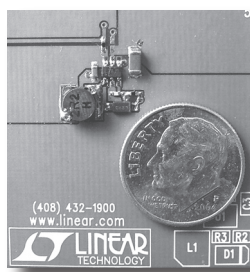


Figure 5. Photograph of demo board of circuit in Figure 4—board area is 10mm × 12mm

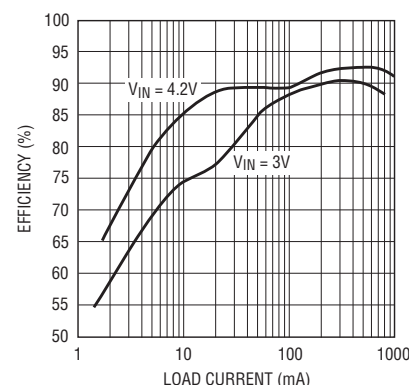


Figure 6. Up to 92% efficiency in Lithium-Ion battery to 5V output applications

LTC4302, continued from page 26

further eases the burden of heavy capacitive loads by providing strong pull-up currents during rising edges to reduce the rise time. Thanks to these two features, the LTC4302 enables the implementation of much larger 2-wire bus systems than are possible with a simple unbuffered multiplexer. **LT**

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:

1-800-4-LINEAR

Ask for the pertinent data sheets and Application Notes.

Impedance Analyzer, continued from page 30

assume that either the inductance is well damped, or it is shunted by large value capacitances. **LT**

Notes

1. This subject is treated in some detail in the LTC1647 data sheet, Figures 9, 10, and 11 inclusive.
2. An hp 5210A Frequency Meter or any common counter gives adequate accuracy for most measurements.

Tiny, Low Noise Boost and Inverter Solutions

by Eric Young

Introduction

The LT3461 and LT3461A are current mode boost converters which combine a 40V rated, 1Ω NPN power switch with a power Schottky diode in a 6-lead ThinSOT package. This level of integration is unmatched by any currently available boost converter. The LT3462 and LT3462A are current mode inverters that offer the same level of integration.

Converters with outputs up to ±38V can be built on a very small footprint, making these parts ideal for compact display or imaging applications.

Everything about these devices focuses on squeezing high performance into the smallest spaces. The 'A' parts operate at high frequency—LT3461A boost switches at 3MHz; the LT3462A inverter at 2.7MHz—which allows the use of tiny, low profile components. For noise sensitive communication applications, the high, constant switching frequency results in low output voltage ripple and easily filtered switching harmonics. The non-'A' parts run at 1.3MHz (LT3461) and 1.2MHz (LT3462) and are intended for applications which require high efficiency or high conversion ratios.

Furthermore, the internally-compensated current-mode PWM architecture minimizes the size and number of external parts, maximizes available output current and optimizes transient response.

Simple, Accurate Negative Regulators

It is easy to set the negative output voltage with the LT3462 and LT3462A inverting converters, because there is no need to compensate for a variable FB input bias current. The LT3462 and LT3462A feature a high impedance ground referenced FB input and a 2% accurate 1.262V reference output. An external resistor (R1) between the reference and the FB pin sets the current in the feedback divider. A second

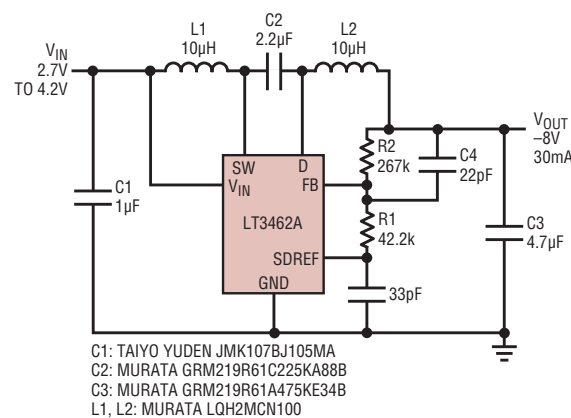


Figure 1. Low profile, 3.3V to -8V, 30mA inverting converter in 50mm²

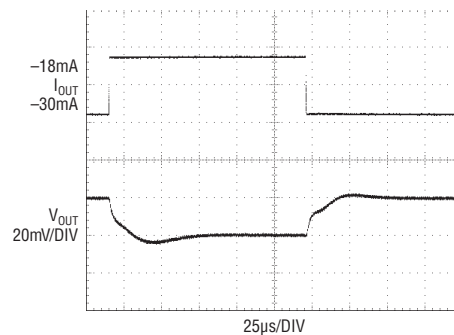


Figure 2. Transient response of the 3.3V-to-(-8V) converter showing less than 0.25% total deviation with a 50% load step

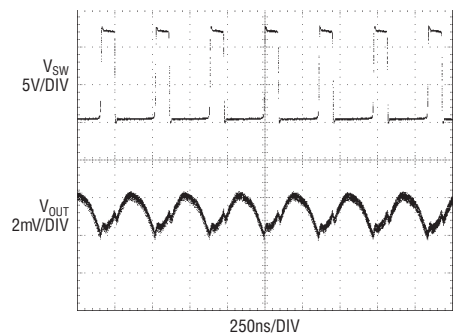


Figure 3. Output ripple of the 3.3V-to-(-8V) inverter at 30mA is only 2.2mV_{P-P}

external resistor (R2) from the FB pin to the negative output sets the output voltage within 2% plus resistor tolerances. By eliminating the untrimmed current sourced by the negative FB (NFB) pin of other inverting regulators, calculation of the feedback divider has been simplified as follows.

$$V_{OUT} = -1.262V \cdot \frac{R2}{R1}$$

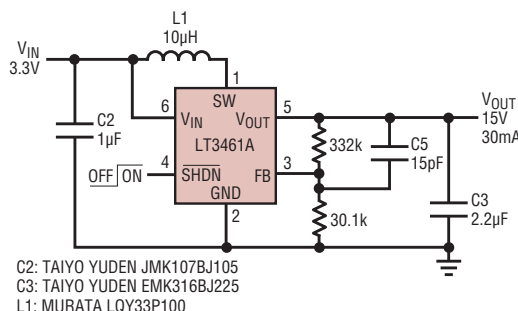


Figure 4. Low profile, 3.3V to 15V, 30mA step-up converter occupies as little as 50mm².

The resulting output voltage is more accurate with less current flowing in the feedback divider.

-8V at 30mA in 50mm²

The 2.7MHz switching frequency of the LT3462A allows the use of tiny low profile inductors and low profile ceramic capacitors. Figure 1 shows a bias supply useful for CCD and OLED applications that produces a well regulated -8V supply at up to 30mA from

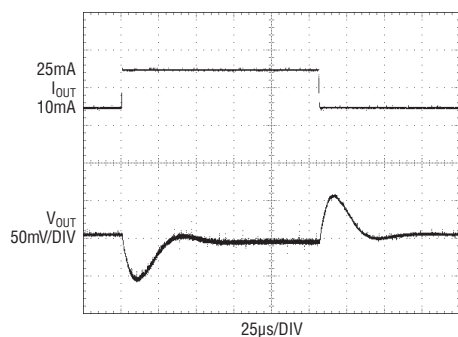


Figure 5. The transient response of the 3.3V-to-15V converter showing less than 120mV total deviation with a 50% load step

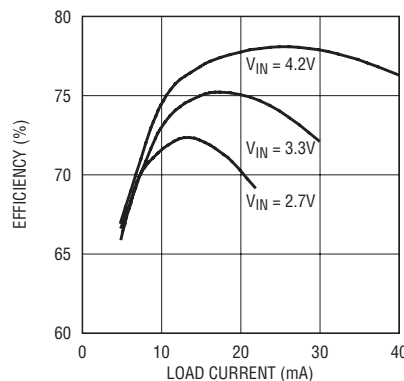


Figure 6. Efficiency of the 3.3V to 15V converter

3.3V using as little as 50mm² of board area. All components in this design are less than 1mm in height.

Board area and profile are usually dominated by the inductor, which is usually the tallest component in the regulator and can occupy more area than the IC. Converters designed with the LT3462A do not have this limitation because the LT3462A works well with tiny, low profile inductors such as the Murata LQH2 series—with little

impact on output power capability, and minimal impact on efficiency.

The -8V converter circuit also uses small (0805) low profile ceramic capacitors for the input, output and flying capacitors. An oscilloscope trace of the half load step on the output (Figure 2) shows these capacitors are sufficient to provide a well-damped transient response. The output voltage remains within 0.25% of the nominal value during the transient steps. Figure 3

shows the output voltage ripple of the -8V converter at 30mA is 2.2mV_{P-P}.

15V at 30mA in 50mm²

The 3MHz switching frequency of the LT3461A also allows the use of tiny, low profile components. Figure 4 shows a circuit that produces a well regulated 15V supply for CCD bias applications at up to 30mA from 3.3V using as little as 50mm² of board area. All components in this design are also less than 1mm in height.

This circuit uses a low profile 2.2µF ceramic output capacitor for well-damped half load step transient response (Figure 5). The output voltage remains well within 1% of the nominal value during these transient steps. The choice of capacitor also impacts output voltage ripple. The output ripple of the circuit in Figure 4 at full load of 30mA is 10mV_{P-P}, or less than 0.07% of the nominal 15V output.

Figure 6 shows that efficiency is better than 70% over a wide range of supply voltages and load currents.

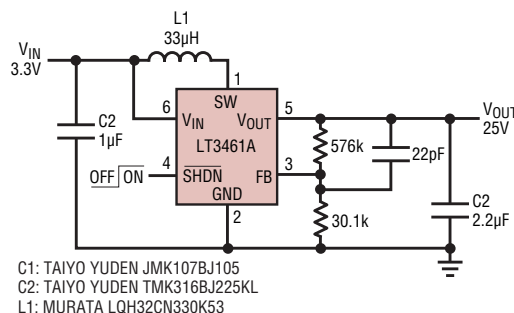


Figure 7. High conversion ratio, 3.3V to 25V step up converter occupies as little as 50mm².

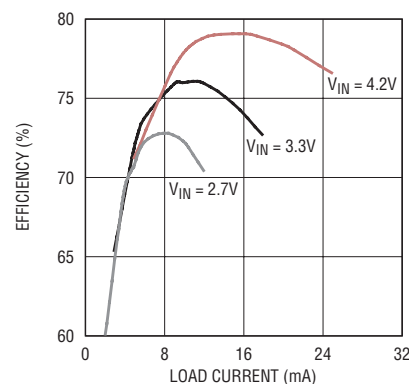


Figure 8. Efficiency of the circuit in Figure 7 at 25V output

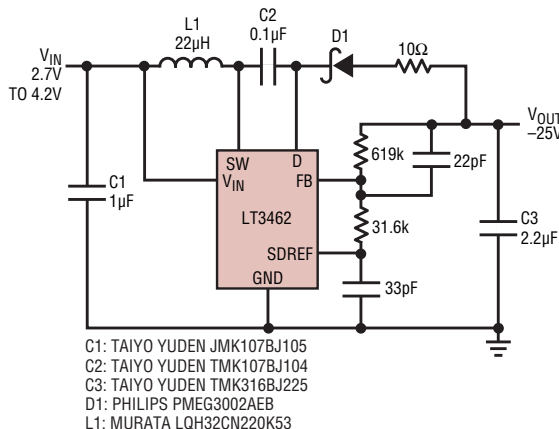


Figure 9. High conversion ratio, 3.3V to 25V inverting converter occupies as little as 55mm².

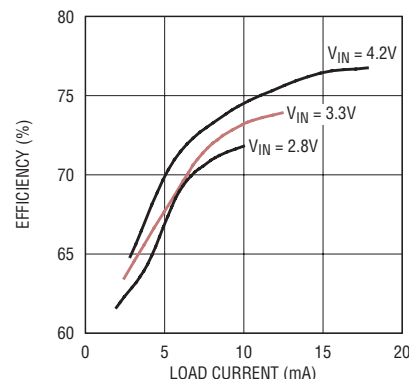


Figure 10. Efficiency of the circuit in Figure 9 at -25V output

Optimizing for Efficiency


While the LT3461A (boost) and LT3462A (inverting) are optimized for small size, the LT3461 (boost) and LT3462 (inverting) are intended for applications requiring higher efficiencies or high conversion ratios. The lower switching frequencies translate to higher efficiencies because of a reduction in switching losses.

The LT3461 (boost) is guaranteed to a maximum switch duty cycle of 92% in continuous conduction mode, and the LT3462 (inverting) is guaranteed to a maximum switch duty cycle of 90%, which enables high conversion ratios at relatively high output currents.

Although high conversion ratios can also be obtained using discontinuous conduction mode (DCM)—where current in the inductor is allowed to go to zero each cycle—the DCM technique requires higher switch currents and larger inductors/rectifiers than a system operating in continuous conduction mode at the same load current. Because the LT3461 can switch at 1.3MHz in continuous conduction mode with up to 92% switch duty cycle, and the LT3462 at 1.2MHz, 90% duty, they are the most compact solutions available for outputs 5 to 10 times the supply voltage. For example, the LCD bias circuit of Figure 7 provides

18mA at 25V from a 3.3V supply and occupies as little as 50mm² of board space. Figure 8 shows that the efficiency of the 25V converter is quite good, peaking at 79% for a 4.2V supply. Figure 9 shows a 3.3V to -25V, 14mA inverter with efficiency above 70% (Figure 10).

Conclusion

The LT3461, LT3461A, LT3462 and LT3462A provide very compact boost and inverter solutions for a wide input voltage range of 2.5V to 16V, and outputs to ± 38 V, making these devices a good fit in a variety of applications. 

LTC2923, continued from page 15

rent produces an unacceptable output voltage error.

Drivers for External, High Current Pass Devices

Table 3 summarizes the characteristics of the LT1575 and LT3150 low dropout regulators. These devices drive external N-channel MOSFET pass devices for high current/high power applications. The LTC3150

additionally includes a boost regulator that generates gate drive for the external FET.


The LTC2923 tracks the outputs of the LT1575 and LT3150 without any special modifications. Because these linear regulators only pull the FET's gate down to about 2.6V, low-threshold FETs may not allow the output to fall below a few hundred millivolts. This is acceptable for most applications. 

Table 3. Drivers for external, high current pass devices

Regulator	I _{OUT(MAX)} (V)	V _{IN(MIN)} (V)	V _{IN(MAX)} (V)	V _{DROPOUT} (V)
LT3150	10A*	1.4	10	0.13
LT1575	*	N/A	22	*

*Depends on selection of external MOSFET

LT1990/91/95, continued from page 4

operating-point—and resistors to set gain. High quality resistors consume precious printed circuit board real estate, and test time. In contrast, the LT1995 provides on-chip resistors for voltage division and gain setting in a highly integrated video-speed op amp.

Figure 5 shows a simple way to drive AC-coupled composite video signals over 75 Ω coaxial cable using minimum component count. In this circuit, the input resistors form a supply splitter

for biasing and a net attenuation of 0.75. The feedback configuration provides an AC-coupled gain of 2.66, so that the overall gain of the stage is 2.0. The output is AC-coupled and series back-terminated with 75 Ω to provide a match into terminated video cable and an overall unity gain from signal input to the destination load. An output shunt resistor (10k Ω in this example) is always good practice in AC-coupled circuits to assure nominal biasing of the output coupling capacitor.

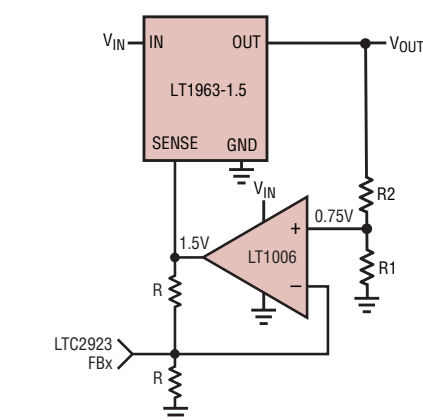


Figure 6. Using an op amp with the LT1963-1.5 allows lower output voltages and removes error due to the SENSE pin current.

Authors can be contacted
at (408) 432-1900


Full Bridge Load Current Monitor

Many new motor-drive circuits employ an H-bridge transistor configuration to provide bidirectional control from a single-voltage supply. The difficulty with this topology is that both motor leads "fly," so current sensing becomes problematic. The LT1990 offers a simple solution to the problem by providing an integrated difference amp structure with an unusually high common-mode voltage rating, up to ± 250 VDC.

Figure 6 shows a solution with an optimization to provide a wide asymmetric common-mode range (–12V to 73V) as might be encountered in an automotive environment. The amplifier is biased from just a single +5V power supply. The asymmetry of the common-mode window is controlled by the applied V_{REF} voltage, provided here by a versatile LT6650 resistor-programmable reference (see

article in this issue: ‘Tiny, Resistor-Programmable, μ Power 0.4V to 18V Voltage Reference’). The LT1990 is shown strapped to produce a gain of ten and outputs a bidirectional signal referenced around V_{REF} . The excellent CMRR of the LT1990 keeps output ripple from the H-bridge PWM activity at a low level so that simple filtering (not shown) can accurately recover the desired low-frequency motor current information.

Conclusion

These three new amplifiers are so versatile and easy to use, it is possible to stock one of them and use it for many varied applications. No external components are needed to achieve hundreds of gains in non-inverting, inverting, difference and attenuator configurations. Just strap the pins and go. It’s a great way to reduce inventory, ease manufacturing, and simplify a bill of materials. 

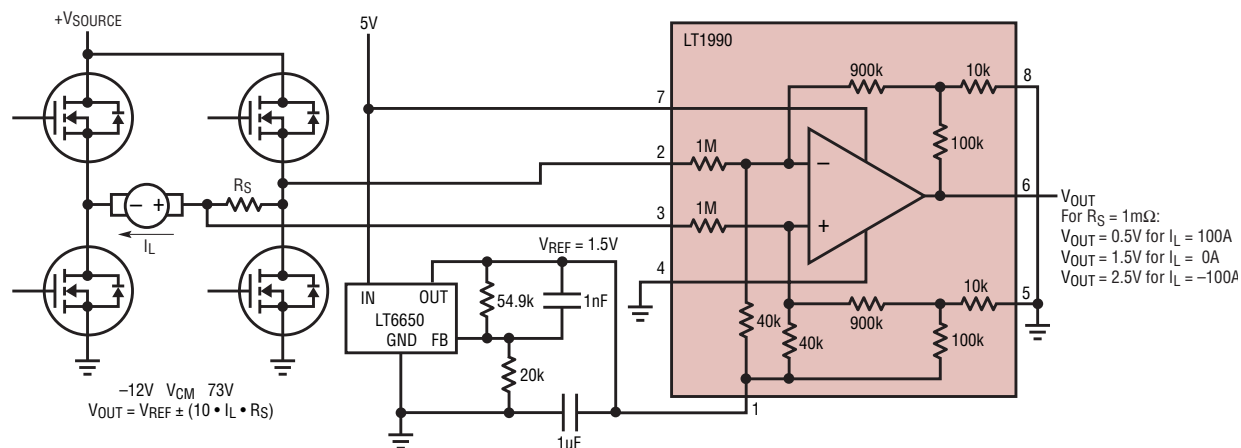


Figure 6. Sensing current in a bidirectional full bridge motor

LT6553/4, continued from page 13

that requires selecting between a set of RGB or component video sources. A similar circuit using LT6553s provides a means of output selection as might be the case in a video recorder where switching between live feed and playback would be needed.


Operating With the Right Power Supplies

The LT6553 and LT6554 require a total power supply of at least 4.5V, but depending on the input and output swings required, may need more to avoid clipping the signal. The LT6554, having unity gain, makes the analysis simple—the output swing is about $(V^+ - V^-) - 2.5\text{V}$ and only governed by the output saturation voltages. This means a total supply of 5V is adequate for standard video (1V_{P-P}). For the LT6553, extra allowance is required for load-driving, so the output swing

is $(V^+ - V^-) - 3.8\text{V}$. This means a total supply of about 6V is required for the output to swing 2V_{P-P}, as when driving cables. For best dynamic range along with reasonable power consumption, a good choice of supplies would be $\pm 3\text{V}$ for the LT6554 and 5V/–3V for the LT6553. Since many systems today lack a negative supply rail, a small LTC1983-3 solution can be used to generate a simple –3V rail for local use, as shown in Figure 6. The LTC1983-3 solution is more cost effective and performs better than AC-coupling techniques that might otherwise be employed.

Demo Circuits Available

Demonstration boards that use the LT6553 and LT6554 are available to simplify evaluation of these parts. To evaluate the LT6553 ask for DC714A or DC743A. DC714A is a DC-coupled circuit that is intended for split supply

operation. DC743A includes biasing and AC-coupling components with the LT6553 in a single supply configuration. DC794A is identical to the DC714A except it has the LT6554 installed. All three of these demo circuits have high-quality 75 Ω BNC connections for best performance and include a calibration trace to allow connector effects to be removed from network analyzer sweeps of the amplifier under evaluation. The demo circuits also illustrate high-frequency layout practices that are important to realizing the most performance from these super-fast parts. 



New Device Cameos

BodeCAD Simplifies Large AC Signal Analysis for DC-DC Power System Stability

BodeCAD is a free software package that simplifies DC-DC power system stability analysis—it can be downloaded from www.linear.com. BodeCAD works with SwitcherCAD to automatically perform large AC signal analysis and produce the bode plot for an entire power system. At present, commercial SPICE programs cannot perform large AC signal analysis and Bode analysis is difficult to simulate. Some simulators use a linear analysis to generate phase and gain plots, but this requires special circuits for each configuration and can have large errors.

BodeCAD is easy to use. Simply insert a voltage source in the feedback path and BodeCAD invokes SwitcherCAD to do Bode analysis over the desired frequency range. Depending on computer speed and the circuit complexity, the Bode summary can be obtained in minutes.

BodeCAD is applicable to a wide range of schemes—it is not tied to a preconfigured schematic. The simulation works with either user-designed SwitcherCAD schematics or any of 170 included example circuits. Enter the input voltage, output voltage and output current, and a list of example circuits appears. V_{IN} , V_{OUT} , and I_{OUT} and other parameters in an example schematic can be tuned and saved for future use. The advantage of using the example circuits is that the user doesn't need to know any schematic or SPICE command. Everything is contained in BodeCAD—results are a few clicks away. If the predefined examples are not suitable, any SwitcherCAD schematic can be used. New devices and topologies are added regularly.

Old Watchdogs Get New Bark

LTC2901-3 and LTC2901-4 are upgrades to the popular LTC2901-1 and LTC2901-2 programmable quad supply supervisors. The important new feature in both is that watchdog

output indication is now at the RESET pin, allowing for a microprocessor reset to occur upon an under-voltage error and/or watchdog error. By comparison, the LTC2901-1 and LTC2901-2 assert an independent watchdog output (WDO) pin upon a watchdog timeout, where the WDO pin is typically connected to a non-maskable interrupt. The LTC2901-3 and LTC2901-4 have also added an under-voltage tolerance select pin to allow for 5% or 10% under-voltage supervisor thresholds.

The reset and watchdog delay times are adjustable using external capacitors. Tight voltage threshold accuracy and glitch immunity ensure reliable reset operation without false triggering. The RESET output is guaranteed to be in the correct state for V_{CC} down to 1V. The LTC2901-1 and LTC2901-3 feature an open-drain RESET output, and the LTC2901-2 and LTC2901-4 have a push-pull RESET drive.

All of the LTC2901 devices are single-pin-programmable with 32 user-selectable quad-supply combinations of 5V, 3.3V, 3V, 2.5V, 1.8V and 1.5V, with $\pm 5\%$, 10% or adjustable-voltage monitor thresholds. The one-pin programming feature eliminates the need to qualify, source and stock different part numbers for different combinations of supply voltages. The LTC2901 is available in a 16-lead narrow SSOP package.

2-Wire Bus Buffer Provides Level Translation, Connection Control

The LTC4300A-3 allows card insertion into a live backplane without corruption of the clock and data lines for I²C, SMBus and IPMB systems. The LTC4300A-3 typically is used on the edge of a peripheral card. The SDAOUT and SCLOUT pins are connected to the card's clock and data buses. V_{CC2} is connected to the card side supply. When the card is inserted, the SDAIN, SCLIN, and V_{CC} pins are connected to the backplane. Control circuitry provides a glitch free connection by

preventing the backplane buses from being connected to the card buses until both card and backplane have completed all data transactions.

The level translating feature allows a card at one supply voltage to communicate with a backplane operating at a different supply voltage. Both the backplane and card may be powered with supplies ranging from 2.7V to 5.5V, with no constraint as to which supply voltage is higher. Having an ENABLE pin allows the LTC4300A-3 card's clock and data buses to be disconnected from the backplane buses, without physically removing the card. By using a shorter pin on the card to connect the ENABLE to the backplane, the device can be held in a disabled and disconnected state until the V_{CC} , V_{CC2} , GND, SDAIN, and SCLIN pins are all properly connected. The ENABLE pin also allows the user to select between multiple devices connected to a bus, creating a mux-like function. Finally the ENABLE pin puts the device into a low current mode, allowing the user to conserve power when the device is not activated.

Another feature of the LTC4300A-3 is electrical isolation, which provides capacitance buffering for the card and backplane buses. Other features of the device include rise time accelerators and pre-charge circuitry. The rise time accelerator circuitry provides pull up current during rising edges, to allow large capacitively loaded systems to meet rise time requirements. The pre-charge circuitry presets both the clock and data ports of the device to 1V, minimizing bus disturbances when a peripheral card is plugged into the backplane connector. The LTC4300A-3 is available in 8-lead MSOP and 3mm \times 3mm DFN packages.

The Cadillac of Power Supply Trackers

The LTC2925 offers coincident tracking, offset tracking, ratiometric tracking, and supply sequencing—all without requiring series MOSFETs. The LTC2925 controls up to four supplies; three without series FETs and an optional fourth supply with a series FET.

The series FET is only required if a power supply does not allow access to its feedback node. An electronic circuit breaker features a current threshold and a short-circuit timeout, adjustable by a resistor and capacitor, respectively. It also contains a remote sense switch so the power supply can regulate the voltage at the load, not at the output of the power supply. This prevents a voltage drop across the series FET from causing problems.

The LTC2925 has a power-good timeout feature. If the supply monitor ever indicates that a supply has left regulation after an adjustable timeout, all supplies shut off and the FAULT pin asserts.

If slave power supplies turn on with input supply voltages below 2.9V, a shutdown feature holds off the slave supplies until the LTC2925 is fully powered.

Configuring each slave supply is as simple as choosing a pair of resistors—no messy I²C buses to worry


about or software engineers to clean up after. The data sheet outlines an easy “3-Step Design Procedure” for choosing these resistor values. By configuring the voltage offset and ramp-rate, a supply can be set up for coincident tracking, offset tracking, ratiometric tracking, and supply sequencing.

Quad Power Supply Monitor with Three Adjustable Inputs in a 6-lead SOT-23

LTC2903-D1 and LTC2903-E1 expand Linear’s family of low current, 6-lead precision quad supply monitors. The LTC2903-D1 provides the user with three adjustable voltage monitor inputs and a fixed 3.3V monitor input. The LTC2903-E1 also has three adjustable inputs and a fixed 5V voltage monitor input.

Options D1 and E1 are configured for 5% under-voltage monitoring, while the A1, B1 and C1 thresholds remain at 10%. The adjustable threshold in-

puts are compared against a precision internal 0.5V reference. All thresholds are guaranteed to $\pm 1.5\%$ of the monitored voltage.

The LTC2903 supervisors incorporate a novel low voltage pull-down circuitry that can hold the $\overline{\text{RST}}$ line low with as little as 200mV of input power supply. This $\overline{\text{RST}}$ pull-down circuitry helps maintain a low impedance path to ground, reducing the risk of a floating the $\overline{\text{RST}}$ node to undetermined voltages, which could trigger external logic to generate an erroneous reset. 

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:

1-800-4-LINEAR

Ask for the pertinent data sheets and Application Notes.

LTC4260, continued from page 21

undervoltage fault occurs when the UV pin drops below 3.12V while an overvoltage fault occurs when the OV pin rises above 3.5V. Each of these major faults has an auto-retry control bit. If a fault occurs and its auto-retry bit is set, then once the fault is removed, the LTC4260 turns on the pass transistor. Otherwise the part is latched off until the fault register is cleared.

There are three minor faults recorded by the fault register that do not turn off the external FET. They include the power bad, $\overline{\text{BD_PRST}}$ changed state and FET short.

A power bad fault is reported if the FB pin drops below the 3.41V threshold while the FET is on. The board present feature allows detection when downstream cards are inserted or removed. This fault is labeled as $\overline{\text{BD_PRST}}$ changed state. The last minor fault, the FET short, indicates

current is flowing in the sense resistor when the pass transistor is turned off. A FET short fault is reported if the data converter measures a current sense voltage greater than or equal to 2mV when the FET is off.

The Status register contains useful information regarding the FET’s on or off condition, all the major and minor fault present conditions and the logic level of the GPIO pin. The Fault register can be regarded as a running log of past faults.


Clearing the Fault Lets the Output Turn-On

As mentioned earlier, the overcurrent, undervoltage and overvoltage faults, once written into the fault register, will keep the pass-transistor off if auto-retry is not selected for these faults. This remains true even when the original recorded fault condition is no longer present. The fault register must be cleared to turn on the output.

The fault register is cleared with any of the following ways:

- ☐ Writing zeroes into the fault register bits using I²C bus
- ☐ An ON pin high to low transition crossing the 1.235V threshold
- ☐ Writing a high-to-low transition in the FET on bit (control register)
- ☐ UV pin brought below 1.235V
- ☐ V_{DD} brought below 7.45V
- ☐ INTV_{CC} brought below 3.8V
- ☐ $\overline{\text{BD_PRST}}$ high-to-low transition crossing the 1.235V threshold clears all faults except $\overline{\text{BD_PRST}}$ changed state fault.

Conclusion

The LTC4260 is a smart power gateway for hot swappable circuits. It provides inrush control and fault isolation while it closely monitors the power through its gates. It logs faults and can interrupt the host if necessary, all while monitoring board power using an internal 8-bit ADC. 

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

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World Headquarters

Linear Technology Corporation

1630 McCarthy Blvd.
Milpitas, CA 95035-7417
Phone: (408) 432-1900
FAX: (408) 434-0507
www.linear.com

U.S. Sales Offices

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FAX: (440) 239-1466

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Cary, NC 27513
Phone: (919) 677-0066
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Asia Sales Offices

CHINA

Linear Technology Corp. Ltd.

Unit 2108, Metroplaza Tower 2
223 Hing Fong Road
Kwai Fong, N.T., Hong Kong
Phone: +852 2428-0303
FAX: +852 2348-0885

Linear Technology Corp. Ltd.

Room 1610, Central Plaza
No. 227 Huangpi Bei Lu
Shanghai, 200003, PRC
Phone: +86 (21) 6375-9478
FAX: +86 (21) 6375-9479

Linear Technology Corp. Ltd.

Room 707, 7th Floor
Beijing Canway Building
66 Nan Li Shi Lu
Beijing, 100045, PRC
Phone: +86 (10) 6801-1080
FAX: +86 (10) 6805-4030

Linear Technology Corp. Ltd.

Rm. 2109, Shenzhen Kerry Centre
2008 Shenzhen Renminnan Lu
Shenzhen, China
Phone: +86 755-8236-6088
FAX: +86 755-8236-6008

JAPAN

Linear Technology KK

8F Shuwa Kioicho Park Bldg.
3-6 Kioicho Chiyoda-ku
Tokyo, 102-0094, Japan
Phone: +81 (3) 5226-7291
FAX: +81 (3) 5226-0268

Linear Technology KK

6F Kearny Place Honmachi Bldg.
1-6-13 Awaza, Nishi-ku
Osaka-shi, 550-0011, Japan
Phone: +81 (6) 6533-5880
FAX: +81 (6) 6543-2588

Linear Technology KK

3F Sakae Members Office Bldg.
4-16-8 Sakae, Naka-ku
Nagoya, 460-0008, Japan
Phone: +81 (52) 269-9510
FAX: +81 (52) 269-9520

KOREA

Linear Technology Korea Co., Ltd.

Yundang Building, #1002
Samsung-Dong 144-23
Kangnam-Ku, Seoul 135-090
Korea
Phone: +82 (2) 792-1617
FAX: +82 (2) 792-1619

SINGAPORE

Linear Technology Pte. Ltd.

507 Yishun Industrial Park A
Singapore 768734
Phone: +65 6753-2692
FAX: +65 6752-0108

TAIWAN

Linear Technology Corporation

8F-1, 77, Nanking E. Rd., Sec. 3
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Linear Technology S.A.R.L.

Immeuble "Le Quartz"
58, Chemin de la Justice
92290 Chatenay Malabry
France
Phone: +33 (1) 41 07 95 55
FAX: +33 (1) 46 31 46 13

Linear Technology

"Le Charlemagne"
140, cours Charlemagne
69286 Lyon Cedex 2
France
Phone: +33 (4) 72 41 63 86
FAX: +33 (4) 72 41 62 99

GERMANY

Linear Technology GmbH

Oskar-Messter-Str. 24
D-85737 Ismaning
Germany
Phone: +49 (89) 962455-0
FAX: +49 (89) 963147

Linear Technology GmbH

Haselburger Damm 4
D-59387 Ascheberg
Germany
Phone: +49 (2593) 9516-0
FAX: +49 (2593) 951679

Linear Technology GmbH

Zettachring 12
D-70567 Stuttgart
Germany
Phone: +49 (711) 1329890
FAX: +49 (711) 7285055

ITALY

Linear Technology Italy Srl

Via Giovanni da Udine, 34
I-20156 Milano
Italy
Phone: +39 (02) 38093656
FAX: +39 (02) 38093659

SWEDEN

Linear Technology AB

Electrum 204
Isafjordsgatan 22
164 40 Kista
Sweden
Phone: +46 (8) 623 16 00
FAX: +46 (8) 623 16 50

UNITED KINGDOM

Linear Technology (UK) Ltd.

3 The Listons, Liston Road
Marlow, Buckinghamshire SL7 1FD
United Kingdom
Phone: +44 (1628) 477066
FAX: +44 (1628) 478153

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