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The LT1304: Micropower DC/DC Converter with Independent Low-Battery Detector

by Steve Pietkiewicz

Introduction

In the expanding world of low power portable electronics, a 2- or 3-cell battery remains a popular power source. Designers have many options for converting the 2V-4V battery voltage to 5V, 3.3V, and other required system voltages using low voltage DC/DC converter ICs. The LT1304 offers users a micropower step-up DC/DC converter featuring Burst Mode[™] operation and a low-battery detector that stays alive when the converter is shut down. The device consumes only 125 μ A when active, yet can deliver 5V at up to 200mA from a 2V input. High frequency operation up to 300kHz allows the use of tiny surface mount inductors and capacitors. When the device is shut down, the low-battery detector draws only 10 μ A. An efficient internal power NPN switch handles 1A switch current with a drop of 500mV. Up to 85% efficiency is obtainable in 2-cell-to-5V converter applications. The fixed-output LT1304-5 and LT1304-3.3 versions have internal resistor dividers that set the output voltage to 5V or 3.3V, respectively.

Operation

The LT1304's operation can best be understood by examining the block diagram in Figure 1 (page 17).

Comparator A1 monitors the output voltage via resistor-divider string R3/R4 at the FB pin. When V_{FB} is higher than the 1.24V reference, A2 and the timers are turned off. Only the reference, A1, and A3 consume current, typically 120 μ A. As V_{FB} drops below 1.24V plus A1's hysteresis (about 6mV), A1 enables the rest of the circuit. Power switch Q1 is then cycled on for 6 μ s, or until current comparator A2 turns off the on-timer, whichever comes first. Off-time is fixed at approximately 1.5 μ s. Q1's switching causes current to alternately build up in inductor L1 and discharge into output capacitor C2 via D1, increasing the output voltage. As V_{FB} increases enough to overcome C1's hysteresis, switching action ceases. C2 is left to supply current to the load until V_{OUT} decreases enough to force A1's output high, and the entire cycle repeats.

If switch current reaches 1A, causing A2 to trip, switch on-time is reduced. This allows continuous-mode operation during bursts. A2 monitors the voltage across 7.2 Ω resistor R1, which is directly related to the switch current. Q2's collector current is set by the emitter-area ratio to 0.5% of Q1's collector

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Eggs on the Wall?

by Richard Markell

I was in the lunch room several weeks ago calmly eating my lunch, when suddenly a large "boom" sounded behind me. (No, it wasn't an underrated tantalum capacitor.) I turned around (slowly) to find that one of our physicists had tried to hard boil an egg in the microwave without first putting a vent hole in the eggshell. My first reaction was to exclaim that any physicist should know that when most things get hot, they expand (go boom!). Sure, we could plot the size of the vent hole versus the "time to go boom." The conservative engineer might do this. Both the analytical approach and trying things to see what happens are useful, but maybe we need more "booms." It seems that, in today's business atmosphere, there is too much conservatism and too few people just trying things to see what happens. At LTC we try to encourage both approaches.

The featured article for this issue highlights the LT1304 micropower DC/DC converter, which includes a low-battery detector that stays alive when the converter is shut down. The LT1304 is a micropower step-up DC/DC converter tailored for operation from two or three cells and featuring Burst Mode operation. The device consumes only 125 μ A when operating and can deliver 5V at up to 200mA from a 2V input. The device features 300kHz operation, allowing the use of tiny surface mount inductors.

A second article features the LT1118 family of fixed voltage regulators that can both source and sink current. These devices are ideal for use as active SCSI terminators. The LT1118 family of regulators features excellent stability and load transient settling with most capacitive loads. The devices can source up to 800mA and sink up to 400mA. Quiescent current is only 600 μ A unloaded.

The LT1510 is a battery charger IC that allows fast charging of all types of batteries, including Li-Ion. The

LT1510 allows users to precisely set battery current and voltage, and to choose the best charging algorithm for a given application. For many situations, such as Li-Ion, no external monitoring of the battery is required. The LT1510 is a current mode PWM battery charger that allows constant-current and/or constant-voltage charging. The internal switch is capable of delivering 1.5A DC current (2A peak current).


Another featured product is the LT1508/LT1509 "combo" power factor correction circuit and PWM controller in a single package. The LT1508/LT1509 integrates the functionality of an LT1248 power factor controller with a 50% maximum duty cycle PWM in a single 20-pin DIP or SOIC.

"Power Factor Correction: Part Three," continues the series that began two issues ago. This installment discusses discontinuity, and the grief it causes for those trying to understand the operation of the PFC.

The LTC1329, a micropower, 8-bit, current output DAC, is the subject of a short article. The LTC1329 is designed to digitally control a power supply's output voltage or to serve as a trimmer pot replacement. This is a versatile part with a 1, 2 or 3 wire interface. Additional information on applying LTC's DAC product line will be found in a feature on 12-bit voltage mode DACs. This article highlights circuits ranging from a computer-controlled 4-to-20mA current loop to an opto-isolated serial interface.

In our "Design Information" section, we present information on electrostatic discharge (ESD) and the IEC-801-2 specification. Our popular RS232 transceiver, the LT1137A, has been upgraded to include on-chip ESD protection compliant with level four of the IEC-801-2 standard. This upgrade allows full ESD protection with no energy absorbing components external to the LT1137A.

Our "Design Ideas" section is chock full of useful circuits. These include

circuits for driving diode-ring mixers, a fully differential 8-channel, 12-bit A/D system, and a circuit for touchscreen interface. 

LTC in the News...


Linear Technology reports record annual and quarter sales!!!

Thanks to all of you for your hard work and constant drive to make Linear Technology the best analog company world wide, the company's annual sales once again increased to a record \$265,023,000 for fiscal year end July 2, 1995, an increase of 32% over net sales from the previous year. The company also reported record net income for the year of \$84,696,000 or \$2.22 per share, an increase of 49% over the \$56,827,000, or \$1.51 per share, reported for fiscal 1994.

Linear Technology also announced that the board of directors approved both an increase in the quarterly dividend to \$0.08 from \$0.07 per share and a two-for-one stock split of its common stock. A cash dividend will be paid on August 23, 1995 to shareholders of record on August 4, 1995. Shareholders of record on August 11, 1995 will be issued certificates reflecting the stock split; these shares will be distributed on September 1, 1995.

Chief Executive Magazine's May, 1995 issue features the prestigious 'Charting The Chief Executive 100 Index.' Robert Swanson Jr. is among the top ten Chief Executives for a third year in a row. In order to make the CE100, the following criteria had to be met. Each CEO had to have been in their position throughout 1992-1994, the company had to be public for that time frame and the market value of the firms public stock had to exceed \$500 million at the end of 1994.

Financial World's July 18, 1995 issue included Linear Technology Corporation on its list of "America's 50 Best Mid-Cap Companies." LTC was ranked number 5 up from 21 the previous year.

Investor's Business Daily June 28, 1995 issue featured an interview with Paul Coghlan, titled ANALOG LIFE, Linear Technology Keeps Pulse Of Light, Heat, Sound, Speed. Mr. Coghlan attributes the company's success to being in the right market at the right time with a good strategy. 

by Chia Wei Liao

Introduction

A sure way to start an argument among electronic engineers is to bring up the subject of battery charging. This relatively simple problem has spawned more solutions than the proverbial mouse trap. This situation seems to be the result of misinformation about the basic electrochemistry of the batteries themselves, the difficulty of quantifying results, and the fertile imaginations of engineers who like to create. At times, battery charging seems to have as much to do with religion as with science.

The LT1510 was created with this situation in mind. It is a basic building block for transferring energy efficiently and accurately from one source to another. Battery current and voltage are precisely defined by the user, who can decide which charging algorithm is best and set up the LT1510 in whatever programmed current or voltage combination is required. For many situations, especially lithium-ion, no extra external battery monitoring is needed.

The LT1510 current mode PWM battery charger is the simplest, most efficient solution to fast charge modern rechargeable batteries, including lithium-ion, nickel-metal-hydride (NiMH), and nickel-cadmium (NiCd), that require constant-current and/or constant-voltage charging. The internal switch is capable of delivering 1.5A DC current (2A peak current). The onboard current sense resistor (0.1 Ω) makes programming the charging current very simple. One resistor (or a programming current from a DAC) is required to set the full charging current (1.5A max.) to within 5% or the trickle-charge current (150mA) to 10% accuracy. With 1% reference-voltage accuracy, the LT1510S16 meets the critical constant-voltage charging requirement for lithium cells.

The LT1510 can charge batteries ranging from 1V to 20V; ground-sensing of current is not required. The battery negative terminal can be tied directly to ground. A saturating switch running at 200kHz gives high charging efficiency and small inductor size. A blocking diode between the chip and the battery is not required because the chip goes into sleep mode and drains only 3 μ A when the wall-adaptor is unplugged. Soft-start and shutdown features are also provided. The LT1510S8 comes in an 8-pin SOIC and the LT1510S16 comes in a 16-pin, fused-lead, power SO package with a thermal resistance of 45°C/W.

Circuit Operation

The LT1510 is a current mode PWM step-down (buck) switcher (see Figure 1). The battery DC charging current is programmed by a resistor R_{PROG} (or a DAC output current) at the PROG pin. Amplifier CA1 converts the charging current through RS1 to a much lower current ($I_{\text{PROG}} = 500\mu\text{A/A}$), which is then fed into the PROG pin. Amplifier CA2 compares the output of CA1 with the programmed current and drives the PWM loop to force them to be equal. High DC accuracy is achieved with averaging capacitor C_{PROG} . Note that I_{PROG} has both AC and DC components. I_{PROG} flows through R1 and generates a ramp signal that is fed to the PWM

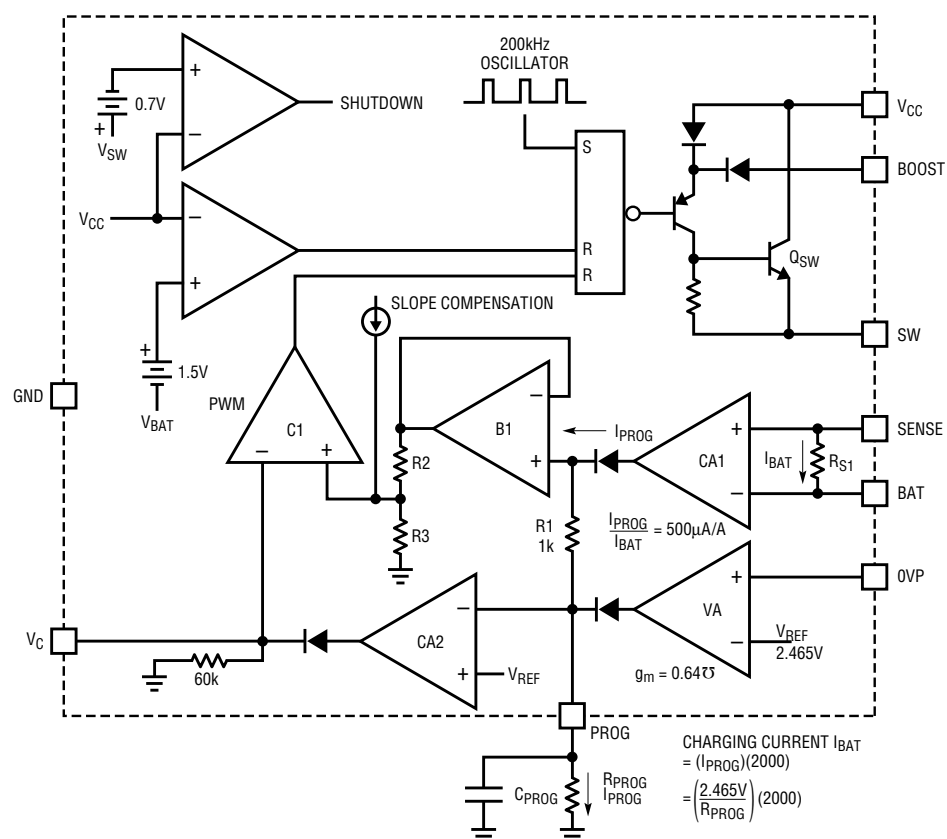


Figure 1. LT1510 block diagram

control comparator C1 through buffer B1 and level-shift resistors R2 and R3, forming the current mode inner loop. The BOOST pin drives the NPN switch (Q_{SW}) into saturation and reduces power loss. For batteries such as lithium, which require both constant-current and constant-voltage charging, the 1% 2.5V reference and the amplifier VA reduce the charging current when battery voltage reaches the preset level. For NiMH and NiCd, VA can be used for overvoltage protection. When an input voltage is not present, the charger goes into low current (3 μ A typical) sleep mode as input drops to 0.7V below the battery voltage. To shut down the charger, pull the V_C pin low with a transistor.

Typical Charging Algorithms

The following algorithms are representative of current techniques:

Lithium-Ion — charge at constant voltage with current limiting set to protect components and to avoid overloading the charging source. When the battery voltage reaches the programmed voltage limit, current will automatically decay to float levels. The accuracy of the float voltage is critical for long battery life. Be aware that lithium ion batteries in series suffer from “walk away,” because of the required constant-float-voltage charging technique. “Walk away” is a condition where the batteries in the series string wind up in different states of the charge/discharge cycle. They may need to be balanced by redistrib-

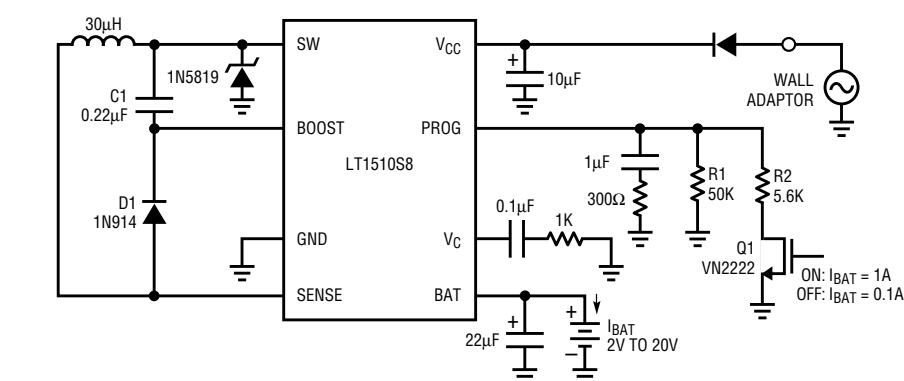


Figure 2. Charging NiCd or NiMH batteries

uting charge from one battery to another. This phenomenon is minimized by carefully matching the batteries within a pack.

Nickel-Cadmium — charge at a constant current, determined by the power available or by a maximum specified by the manufacturer. Monitor battery charge state using voltage change with time (dV/dt), second derivative of voltage (d^2V/dt), battery pressure, or some combination of these parameters. When the battery approaches full charge, reduce the charging current to a value (top-off) that can be maintained for a long time without harming the battery. After the top-off period, usually set by a simple time out, reduce the current further to a trickle level that can be maintained indefinitely, typically 1/10 to 1/20 of the battery capacity.

Nickel-Metal-Hydride — same as NiCd, except that some NiMH batteries cannot tolerate a *continuous* low level trickle charge. Instead, they require a pulsed current of moderate

value, with a low duty cycle, so that the average current represents a trickle level. A typical scenario would be one second “on” and thirty seconds “off,” with the current set to thirty times desired trickle level.

Recharging NiMH or NiCd Batteries

The circuit in Figure 2 will charge battery cells with voltages up to 20V at a full charge current of 1A (when Q1 is on) and a trickle charge current of 100mA (when Q1 is off). If the third charging level is needed, simply add a resistor and a switch. The basic formula for charging current is:

$$\frac{2.465}{R1 \parallel R2} \times 2000 \text{ (when Q1 on)}$$

$$\frac{2.465}{R1} \times 2000 \text{ (when Q1 off)}$$

For NiMH batteries, a pulsed trickle charge can be easily implemented with a switch in series with R1; switch Q1 at the desired rate and duty cycle. If a micro-controller is used to control the charging, connect the DAC current sink output to the PROG pin.

Recharging Lithium-Ion Batteries

The circuit in Figure 3 will charge lithium ion batteries at a constant current of 1.5A until battery voltage reaches 8.4V, set by R3 and R4. It then goes into constant-voltage charging and the current slowly tapers off to zero. Q3 can be added to disconnect R3 and R4 so they will not drain the battery when the wall-adaptor is unplugged.

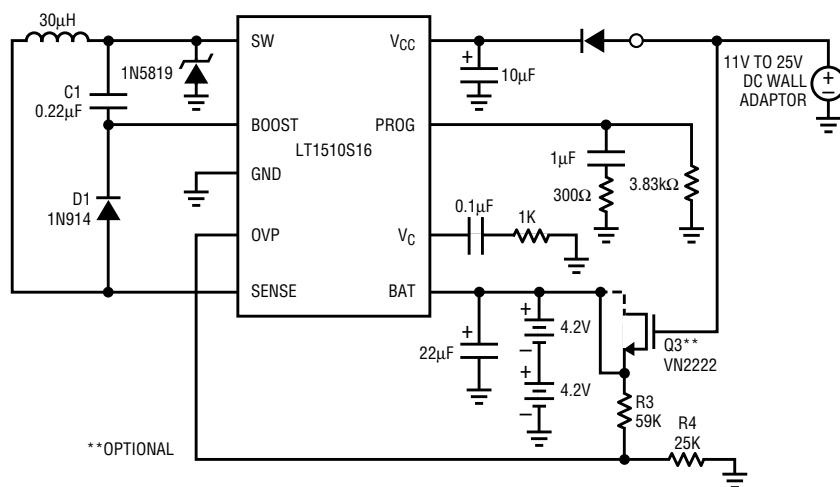


Figure 3. Charging Li-Ion batteries

LT1118 Terminates Active SCSI and More

by Gary Maulding

The new LT1118 family of positive, fixed voltage regulators is unique in maintaining output voltage regulation while both sinking and sourcing current. Conventional positive voltage regulators lose regulation if the load circuit attempts to return current to the regulator output. Capable of sourcing up to 800mA and sinking 400mA, LT1118 family regulators feature excellent stability and load-transient settling with any capacitive load greater than 0.22 μ F. Their insensitivity to capacitive loading makes the regulators easy to use in any application within their power level. Quiescent current is only 600 μ A when unloaded. 5V, 2.5V, and 2.85V versions of the LT1118 are available in 3-pin SOT-223 or fused-lead SO-8 packages. The SO-8-packaged devices include an enable pin that allows the device to be shut down with the output at high impedance. Quiescent current drops to zero when the enable pin is low.

The ability to maintain regulation while both sourcing and sinking current is valuable in several regulator applications. Appropriate applications are data-bus termination, power-supply splitting, and any application where fast settling of large load transients is needed. The 2.85V version is ideal for use as an active termination for SCSI cables. The 2.5V version makes a convenient low power supply splitter in 5V systems. The 5V version is ideal where fast settling of

load transients is required. The shut-down pin on the SO-8-packaged devices is valuable for power management and supply sequencing.

Active SCSI Terminator

The SCSI parallel bus requires cable terminations at each end of the cable. The termination serves the dual purposes of impedance matching and providing the pull-up to the 2.8V negated logic level. The open-drain or open-collector drivers on the bus establish the low level and make multiplexing of equipment simple, due to the "wired-OR" nature of the bus signals.

Passive terminators consisting of a resistor divider from the termpower supply (Figure 1) were commonly used on early SCSI networks. Passive terminators have mostly disappeared from use in recent systems due to the high power dissipation and lack of power supply rejection of the resistor divider. The Boulay or active termination has replaced the passive terminations in today's equipment. A Boulay active termination uses a 2.85V voltage regulator to establish the negated logic level and 110 Ω resistors to provide the impedance match to each data and control line (Figure 2). Power consumption when all lines are in the high (negated) state is dominated by the low quiescent current of the regulator, a 0.225W saving over resistor terminators in a 27-line wide SCSI bus.

SCSI-2 introduced the use of active-negation drivers in SCSI bus systems. Active-negation drivers have three states: active (low), negated (high), and high impedance. By raising the data line above the 2.85V termination voltage when negated, the effects of cable reflections and crosstalk between lines are overcome, improving the noise margin in high speed data busses. Active negation may be used on all but three SCSI data and control lines (three lines, BSY, SEL, and RST, must be wire-OR'd and cannot use active negation drivers). Each negated driver can source up to 7mA when in the high state. Most voltage regulators will lose regulation if any current is returned to their outputs, causing the termination voltage to rise.

The LT1118-2.85, with current sinking, is the perfect solution for this application. The 800mA sourcing capability and 400mA current-sink capability provide ample margin to terminate a 27-line wide SCSI bus in the fully asserted or negated state. Even though SCSI standards restrict the active negation drivers to a 3.24V and 7mA negated signal level, many driver circuits in use exceed the allowed SCSI drive levels. If the maximum of 24 lines are negated in a Wide-SCSI bus, 168mA of current is forced back into the terminator. The 400mA sink capability of the LT1118 gives more than twice the required

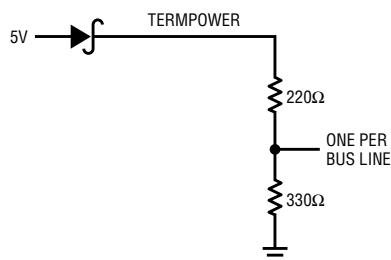


Figure 1. SCSI passive terminator

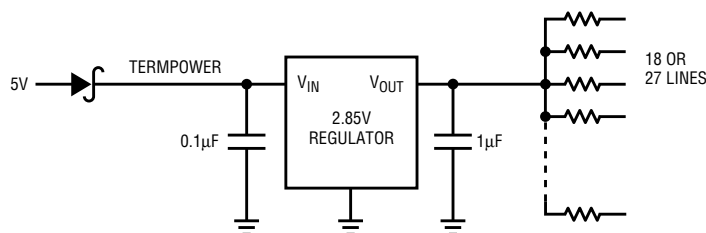
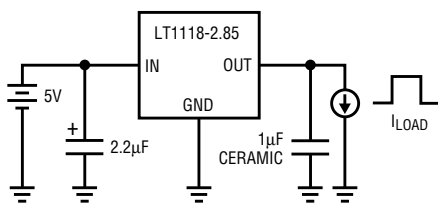


Figure 2. Boulay active terminator

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Figure 3. LT1118 typical application

current handling without damage or loss of regulation.

The fast settling time and stability of the LT1118 allow the regulator output capacitor to be reduced from a typical 10μF to a surface mount 1μF ceramic capacitor saving board space and cost. The low, 600μA quiescent current of the regulator allows the terminator to be connected to the SCSI term power line when not in use.

The maximum voltage spikes for Figure 3's circuit are less than 0.3V and settle out in 5μs. These tests were under maximum load transient conditions: 800mA sourcing to 400mA sinking. This transient is far more severe than would be seen in a SCSI bus termination, yet the voltage output is always within acceptable SCSI voltage levels. The amplitude of the output voltage spikes may be reduced further by use of larger output capacitors.

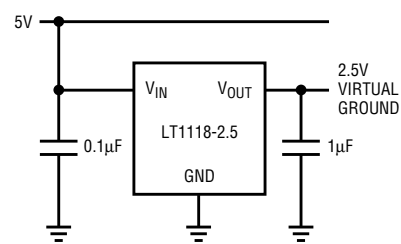
Several competitors offer active SCSI terminators that use on-chip resistors. These terminators, while offering some convenience, are limited in performance and are not as economical a solution as the LT1118-2.85. Capacitance on the output pins of the terminators cause impedance mismatches at high frequencies, which increase cable reflections and

degrade the noise margin of the bus signals. The capacitance inevitably comes from the pin-to-pin capacitance of the IC package, the ESD protection devices on the chip, and the distributed capacitance of the diffused resistors on the chip. The reduced noise margin of the SCSI bus increases error rates and limits the length of the bus and the number of nodes that can reliably be connected. The external surface mount resistors used with the LT1118-2.85 reduce the capacitance to the stray PC board capacitance.

Power consumption further limits the capabilities of SCSI terminators with on-chip resistors. The worst-case power consumption for a SCSI terminator occurs when all data and control lines are active (low). For a 27-line wide SCSI bus, the regulator dissipates 1.2W and the resistors dissipate 2.0W. The surface mount IC packages used for SCSI terminators cannot dissipate this total power of 3.2W. This fundamental limit restricts the devices with on-chip resistors to nine lines. Three terminator chips with integral resistors must be used, versus one LT1118-2.85 with inexpensive surface mount resistors. The cost of three SCSI terminator ICs with integral resistors is more than three times the cost of an LT1118-2.85 and the external 110Ω resistors.

Power Supply Splitter

Often a system has only a single 5V power supply, but has analog sections that require positive and negative supplies. The current source and sink capabilities of the LT1118-2.5 are ideal for generating a virtual ground

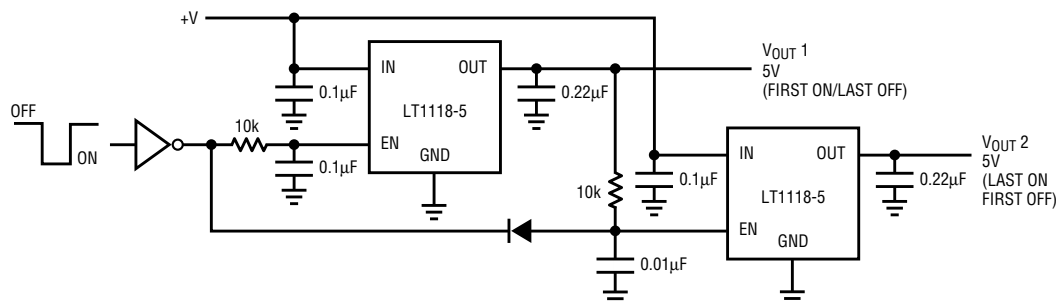

Figure 4. Supply splitter

to allow the use of analog components without a negative supply. Figure 4 shows the LT1118-2.5 used as a 5V supply splitter. The LT1118 can absorb large DC or transient ground currents that would force high power consumption if a resistive supply splitter were used. The regulator's fast settling time allows the use of small capacitors, and its stability with any large capacitive load makes it easy to apply to almost any splitter application. Competing supply splitters are known to suffer from stability problems with capacitive loading.

Power Management

The enable control on the SO-8 packaged LT1118 provides easy power management solutions. A logic low on the enable pin causes the regulator's output to go high impedance and its quiescent current to drop to zero. A logic high brings the regulator into normal regulation. Intelligent power controllers can make use of this capability to selectively power subcircuits as needed (see Figure 5). The small output capacitor requirements of the LT1118 minimize turn-on time and the supply current spikes that result from charging large filter capacitors.

continued on page 15


Figure 5. Power sequencing using the LT1118

LT1580 Low Dropout Regulator Uses New Approach to Achieve High Performance

by Craig Varga

Introduction

Low dropout regulators have become more common in desktop computer systems as microprocessor manufacturers have moved away from 5V-only CPUs. A wide range of supply requirements exist today, with new voltages just over the horizon. In many cases the input-output differential is very small, effectively disqualifying many of the low dropout regulators on the market today. Several manufacturers have chosen to achieve lower dropout by using PNP-based regulators. The drawbacks of this approach include much larger die size, inferior line rejection, and poor transient response.

Enter the LT1580

The new LT1580 NPN regulator is designed to make use of the higher supply voltages already present in most systems. The higher voltage source is used to provide power for the control circuitry and supply the drive current to the NPN output transistor. This allows the NPN to be driven into saturation, thereby reducing the dropout voltage by a V_{BE} compared to a conventional design. Applications for the LT1580 include 3.3V to 2.5V conversion with a 5V control supply, 5V to 4.2V conversion with a 12V control supply, or 5V to 3.6V conversion with a 12V control supply. It is easy to obtain dropout voltages as low as 0.4V at 4A, along with excellent static and dynamic specifications.

The LT1580 is capable of 7A maximum, with approximately 0.8V input-to-output differential. The current requirement for the control voltage source is approximately 1/50 of the output load current, or about 140mA for a 7A load. The LT1580 presents no supply-sequencing issues. If the control voltage comes up

first, the regulator will not try to supply the full load demand from this source. The control voltage must be at least 1V greater than the output to obtain optimum performance. For adjustable regulators the adjust-pin current is approximately 60 μ A and varies directly with absolute temperature. In fixed regulators the ground pin current is about 10mA and stays essentially constant as a function of load. Transient response performance is similar to that of the LT1584 fast-transient-response regulator. Maximum input voltage from the main power source is 7V, and the absolute maximum control voltage is 14V. The part is fully protected from overcurrent and over-temperature conditions. Both fixed voltage and adjustable versions are available. The adjustables are packaged in 5-pin TO-220s, whereas the fixed voltage parts are 7-pin TO-220s.

The LT1580 Brings Many New Features

Why so many pins? The LT1580 includes several innovative features that require additional pins. Both the fixed and adjustable versions have remote-sense pins, permitting very accurate regulation of output voltage at the load, where it counts, rather than at the regulator. As a result, the typical load regulation over a range of 100mA to 7A with a 2.5V output is approximately 1mV. The sense pin and the control-voltage pin, plus the conventional three pins of an LDO regulator, give a pin count of five for the adjustable design. The fixed-voltage part adds a ground pin for the bottom of the internal feedback divider, bringing the pin count to six. The seventh pin is a no-connect.

Note that the adjust pin is brought out even on the fixed-voltage parts. This allows the user to greatly improve the dynamic response of the regulator by bypassing the feedback

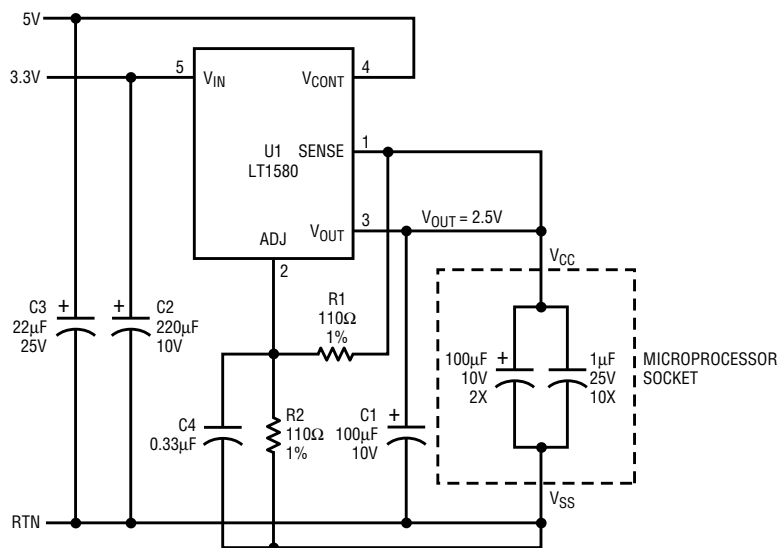


Figure 1. LT1580 delivers 2.5V from 3.3V at up to 6A

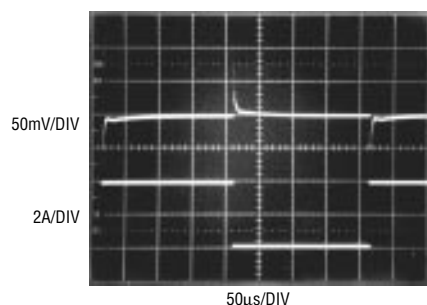


Figure 2. Transient response of Figure 1's circuit with adjust-pin bypass capacitor. Load step is from 200mA to 4A

divider with a capacitor. In the past, using a fixed regulator meant suffering a loss of performance due to lack of such a bypass. A capacitor value of 0.1 μ F to approximately 1 μ F will generally provide optimum transient response. The value chosen depends on the amount of output capacitance in the system.

In addition to the enhancements already mentioned, the reference accuracy has been improved by a factor of two, with a guaranteed 0.5% tolerance. Temperature drift is also very well controlled. When combined with ratiometrically accurate internal divider resistors, the part can easily hold 1% output accuracy over

temperature, guaranteed, while operating with an input/output differential of well under 1V.

In some cases a higher supply voltage for the control voltage will not be available. If the control pin is tied to the main supply, the regulator will still function as a conventional LDO and offer a dropout specification approximately 70mV better than conventional NPN-based LDOs. This is the result of eliminating the voltage drop of the on-die connection to the control circuit that exists in older designs. This connection is now made externally, on the PC board, using much larger conductors than are possible on the die.

Circuit Examples

Figure 1 shows a circuit designed to deliver 2.5V from a 3.3V source with 5V available for the control voltage. Figure 2 shows the response to a load step of 200mA to 4.0A. The circuit is configured with a 0.33 μ F adjust-pin bypass capacitor. The performance without this capacitor is shown in Figure 3. This difference in performance is the reason for providing the adjust pin on the fixed-voltage

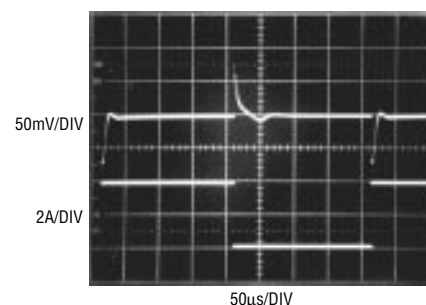



Figure 3. Transient response without adjust-pin bypass capacitor. Otherwise, conditions are the same as in Figure 2

devices. A substantial savings in expensive output decoupling capacitance may be realized by adding a small "1206-case" ceramic capacitor at this pin.

Figure 4 shows an example of a circuit with shutdown capability. By switching the control voltage rather than the main supply, the transistor providing the switch function needs only a small fraction of the current handling ability that it would need if it was switching the main supply. Also, in most applications, it is not necessary to hold the voltage drop across the controlling switch to a very low level to maintain low dropout performance. 

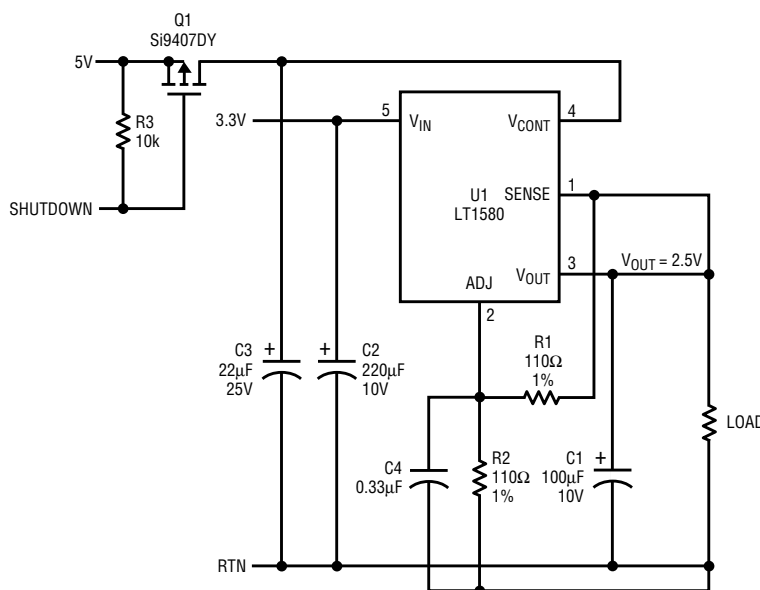


Figure 4. Small FET adds shutdown capability to LT1580 circuit

Power Factor Correction — Part Three: Discontinuity

by Dale Eagar

In Part One of this series (in *Linear Technology* V:1, pp. 17–18, 21), we investigated power factor correction (PFC) by looking at its line-frequency voltage, current, and power waveforms. We developed the concept of a DC variac and demonstrated its utility for performing PFC.

In Part Two (*LT* V:2, pp. 3–7), we developed the boost converter and investigated its properties when operating in the continuous-mode. We showed that a continuous-mode boost converter does a pretty good job of emulating the DC variac.

Part Three continues our investigation of the boost converter by focusing on its properties when operating in the discontinuous-mode. We will see that a discontinuous-mode boost converter doesn't look anything like a DC variac.

There is elegance and simplicity in the concept of the DC variac. There is beauty in its application to PFC, a beauty that is manifest in the portions of line cycle where the boost converter is operating in the continuous-mode.

Insubordination Wreaks Havoc

In analog engineering, we like to view things as continuous and smooth, with all nonlinearities well defined. We look at an AC sine wave and see the voltage go from positive, through zero, to negative with no discontinuity. This explains a good portion of the genuine perplexity experienced by the author when the bank practices alchemy on perfectly legitimate checks—changing the paper into rubber. One would think

the bank would know that it is only two days until payday and the check book is still half full.

Research into the character of banking employees reveals the startling truth that the local bank is staffed by the genetic precursor of the digital engineer, a human so vile, so base, as to actually enjoy inserting discontinuities into an otherwise continuous system. In the boost converter it is the output diode that acts as the banker.

The boost converter shown in Figure 1a is the boost converter from heaven. Its glory is directly attributed to its having a switch, rather than a vile diode, in its output. The switch allows current flow in both directions, and thus inserts no additional discontinuities. The boost converter from heaven always operates in the continuous-mode. Back here on earth we must live with discontinuities.

In a boost converter operating in continuous-mode, the volt-seconds, and thus the current in the inductor, is entirely controlled by:

1. The input voltage
2. The output voltage
3. The duty factor.

The boost converter of Figure 1b is the mortal version. The inductor current (I_L) in the boost converter, driven by the three factors above, goes to zero when the input line voltage swings through zero. Figure 2 shows that the input line current is actually the average of the inductor current over a full switching-frequency cycle. Figure 2a is the inductor current when the input line voltage has reached its peak with the average value representing the line current at the plug. (The ripple is filtered out by the input EMI filter.) Similarly, Figure 2b shows the inductor current and input current when the instantaneous input line voltage is at 40% of its peak value. Finally, Figure 2c shows the

inductor current and input line current as the instantaneous input line voltage is at 10% of its peak value. Notice that on each switching-frequency cycle, the inductor current goes to zero and then tries to reverse. When the inductor current attempts to reverse, the diode commits the disgraceful act of insubordination, destroying the beauty, simplicity, and continuity of the control function of the continuous-mode boost converter. Figure 2c also shows the area gained due to the clipping of the negative portion of the inductor current. The net effect is that the average current is higher than the current that the continuous-mode's duty factor control of this DC variac would have requested. The very act of preventing the inductor current from reversing, a hideous act indeed, necessitates a new set of models to describe the discontinuous-mode operation of the boost converter. Further, a new control law is required to control the discontinuous-mode boost converter.

The Boundary is a Moving Target

The boost converter could be likened to a fish tank, with the surface of the water representing the boundary of continuity, above which the boost converter operates in the continuous-mode and below which it operates in the discontinuous-mode. Folks living outside the fish tank tend to obtain their oxygen from the air, whereas folks living inside the fish tank tend to get their oxygen from the water. Air folk and water folk have great trouble communicating, as

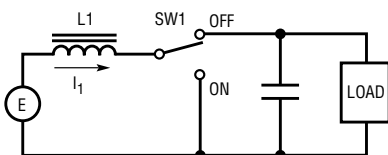


Figure 1a. The boost converter from heaven

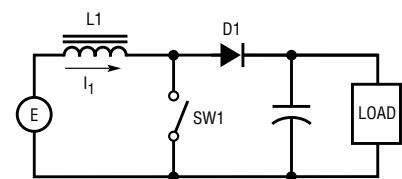


Figure 1b. Mortal boost converter

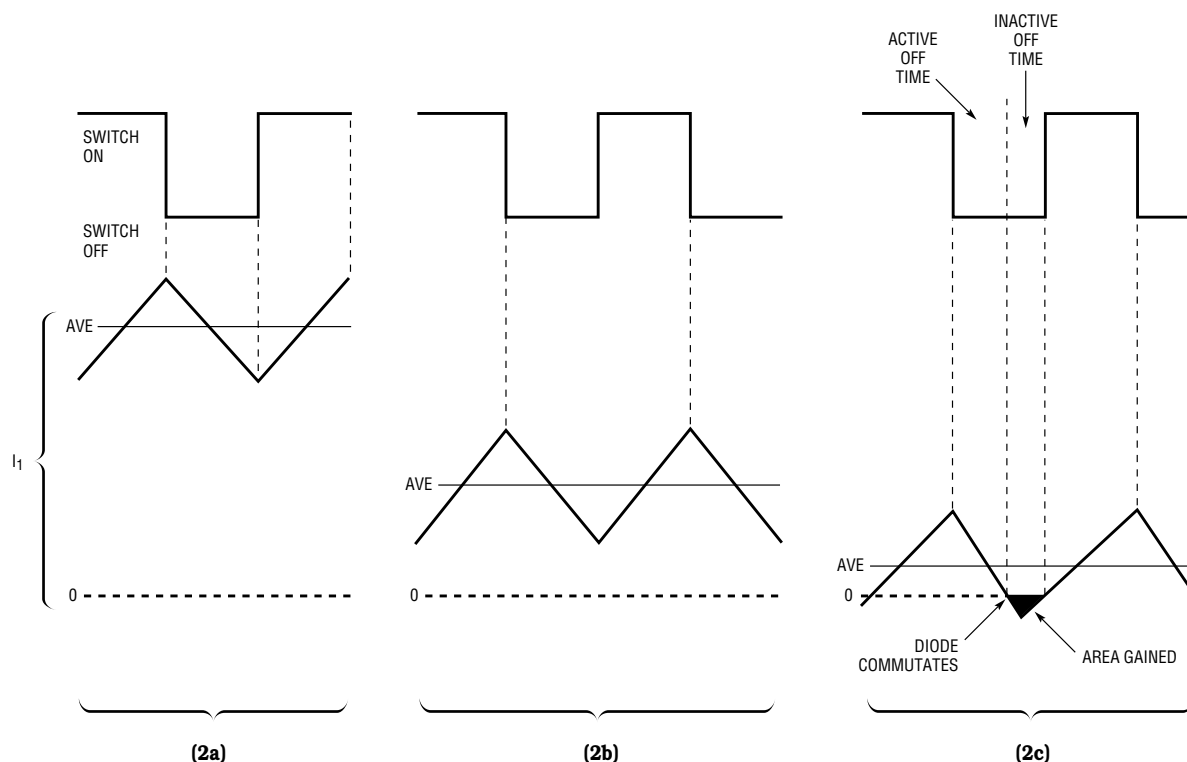


Figure 2. Choke-current waveforms

neither is equipped to speak, let alone breathe, in the other's medium. This lack of communication has had a significant effect on the way the two groups think.

When air folk look at controlling their boost converters, they speak of duty-factor, the all important controlling entity. Water folk control their boost converters with on-time. We, looking at the whole system from the outside, know that the switching-frequency is constant, and thus on-time and duty-factor are describing the same thing.

When air folk plot duty-factor, they see a smooth curve extending down to the surface of the water, and then being refracted at the surface to a different slope as it goes further down to the bottom of the tank. The air folk mathematicians have developed equations that predict the duty-factor plot, equations that exactly describe everything it does in the air. The air folk equations are incapable of predicting what happens when the plot is below the water. The water folk mathematicians have the exact same problem in reverse. We the onlookers

can inspect the equations, and when we do, we see that the respective equations have no hope of being consolidated into one. The air folk equation states that the duty-factor is a function of instantaneous input and output voltage and the rate of change in the current flow, but is altogether independent of the steady-state value of the current flow, the operating frequency, and the inductor's inductance value.

The water folk equation states that the on-time is a function of the steady-state current flow, the inductance of the inductor, the switching frequency, and the input voltage, and is independent of the output voltage and the rate of change in current flow. Fortunately for all of us, both equations converge on the same point at the water's surface.

Air folks use volt-seconds to describe the stretch in their inductors, and water folk use joules to describe the energy stored in their inductors. As onlookers, we can convert between Joules and volt-seconds by using the following equations:

$$J = \frac{1}{2L} (V \bullet S)^2$$

$$V \bullet S = \sqrt{2LJ}$$

J = Energy in Joules

L = Inductance in Henries

$V \bullet S$ = Stretch in Volt Seconds

To further complicate matters, the index of refraction of the water changes and is directly related to the water level. The water level changes due to input voltage, output voltage, switching frequency, output current, and inductance of the inductor.

Memory Lost

In continuous-mode, the choke current, and thus power intercepted from the input line, depends on what the choke current was during the previous cycle. Evaluation of continuous-mode boost is best performed by looking at the overall action of an integral number of switching cycles of a running converter.

In discontinuous-mode, the choke current goes to zero during each

continued on page 15

The New LT1508/LT1509 Combines Power Factor Correction and a PWM in a Single Package

by Kurk Mathews

Introduction

When a supply requires power factor correction (PFC), designers commonly add a PFC controller, such as the LT1248, to the front end of a traditional off-line DC/DC converter. Although this is a viable solution, additional circuitry is usually required to ensure that the two controllers work well together. The LT1508/LT1509 eliminate the need for the additional circuitry and PWM by integrating the functionality of a LT1248 power factor controller and a 50% maximum-duty-cycle PWM in a single 20-pin DIP or SOIC. The LT1508 retains all the pins of the LT1248 except the EN/SYNC, and adds quiet-ground, soft-start, control-voltage, current-limit and gate-driver pins to form a voltage-mode PWM (see Figure 1). In the LT1509, the current limit becomes the ramp pin (PWM current sense), forming a current-mode PWM (see Figure 2).

Full Featured PFC

All features of the LT1248 remain, requiring no compromises when implementing PFC with the LT1508/LT1509. Continuous-boost average current mode reduces magnetics size while operating at a fixed frequency, without the need for slope compensation. Built-in overvoltage protection reduces the external parts count. Current amplifier offset voltage is dealt with internally, minimizing crossover distortion. Full differential input current sense is available, as are 1.5A gate drivers for both the PFC and PWM stages. In addition to these familiar performance features there are a few new ones incorporated in the PWM section.

Hand-Holding

Whether the application involves voltage mode or current mode control, smooth interaction between the PFC and PWM stages requires some hand-holding. The first thing to consider is start-up. The PFC preregulator turns on with soft start as usual. During this period the PWM's soft-start pin is held low, preventing the second stage converter from turning on until the intermediate bus voltage reaches its preset value (typically 380VDC to 400VDC for a universal input). In addition to delaying the PWM soft start during start-up, the same comparator employs hysteresis to shut down the PWM stage in the event the intermediate bus voltage drops below 73% of the preset voltage (typically 280VDC). Once enabled, the PWM runs somewhat independently of the PFC controller, with a couple of exceptions. The oscillator frequency is set by the values of the capacitor and resistor to ground on the C_{SET} and R_{SET} pins. Both the PFC and PWM stages are synchronized to this oscillator and operate at the same switching frequency. Synchronization is accomplished by delaying the turn-on of the PWM stage through 50% of the switching period. In addition to making oscilloscope measurements easier, these features ensure that both switches don't turn on simultaneously, causing bus voltage spikes, noise and problems with current mode in the PWM stage. Single frequency and synchronization also ensure the absence of beat frequencies in the conducted emissions. A preset duty cycle limit of 50% and leading-edge blanking further ease the design of the PWM section. The PWM stage is quickly disabled by pulling the control voltage low. Finally, the LT1509's

current mode comparator has a 1.2 VDC offset to accommodate opto-feedback, and both the LT1508 and LT1509 contain independent over-current comparators.

Typical Application

Figure 3 shows a 24VDC, 300W power-factor-corrected, universal-input supply. The continuous, current mode boost PFC preregulator minimizes the differential-mode input filter size required to meet European low-frequency conducted emission standards while providing a high power factor. The 2-transistor forward converter offers many benefits including low peak currents, a non-dissipative snubber, 500VDC switches and automatic core reset guaranteed by the LT1509's 50% maximum duty-cycle limitation. An LT1431 and inexpensive opto-isolation is used to close the loop conservatively at 3kHz with excess margin (see Figure 4). Figure 5 shows the output voltage's response to a 2A to almost 10A current step. Regulation is maintained to within 0.5V. Efficiency curves for output powers of 30, 150, and 300W are shown in Figure 6. The PFC preregulator alone has efficiency numbers of between about 87% and 97% over line and load.

Start up of the circuit begins with the LT1509's V_{CC} bypass capacitors trickle charging through 91kΩ to 16VDC, overcoming the chip's .25mA typical start-up current (V_{CC} ≤ lockout voltage). PFC soft start is then released, bringing up the 382VDC bus with minimal overshoot. As the bus voltage reaches its final value, the forward converter comes up, powering the LT1431, and closing the feedback loop. A 3-turn

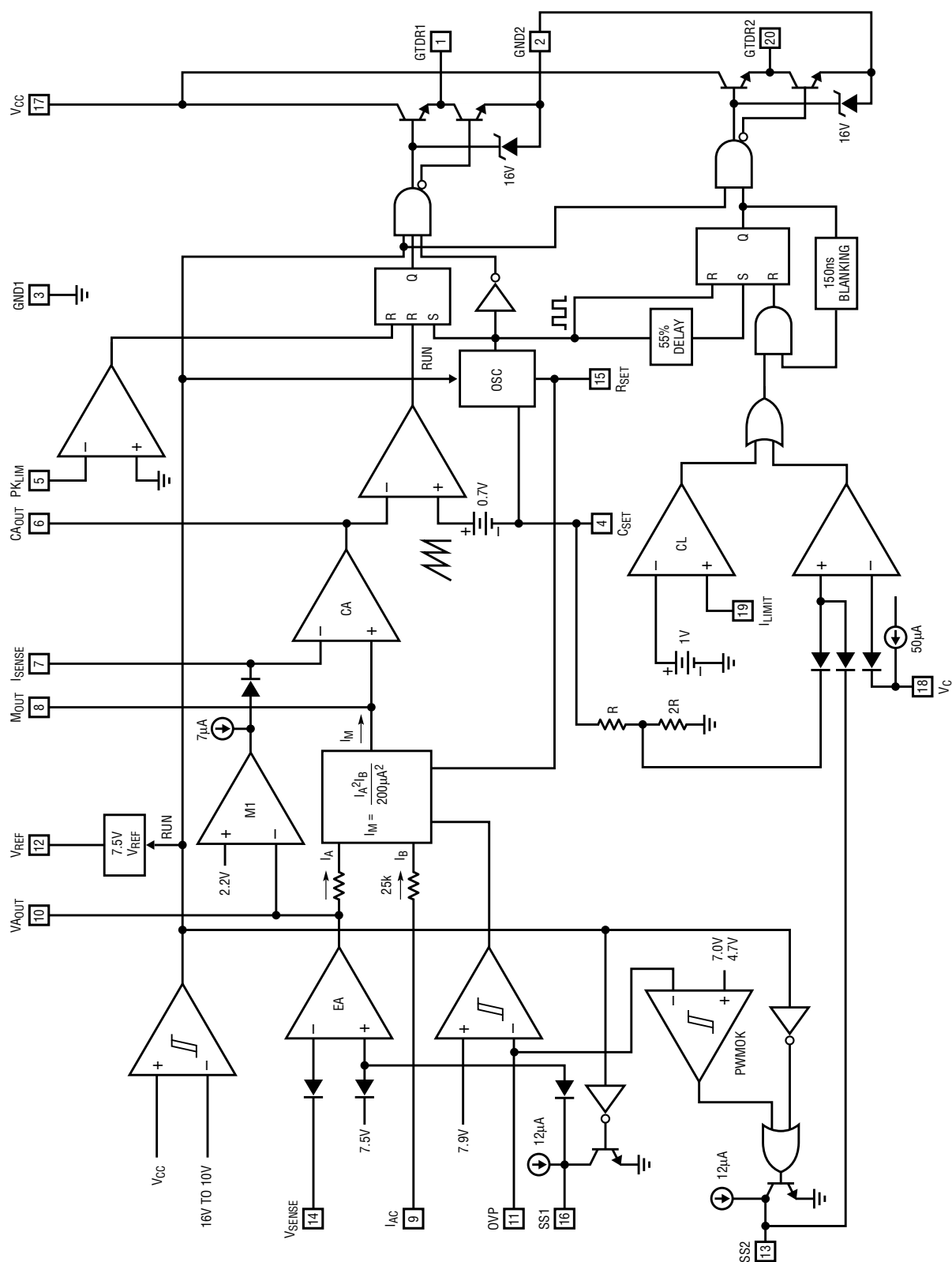


Figure 1. Block diagram: LT1508

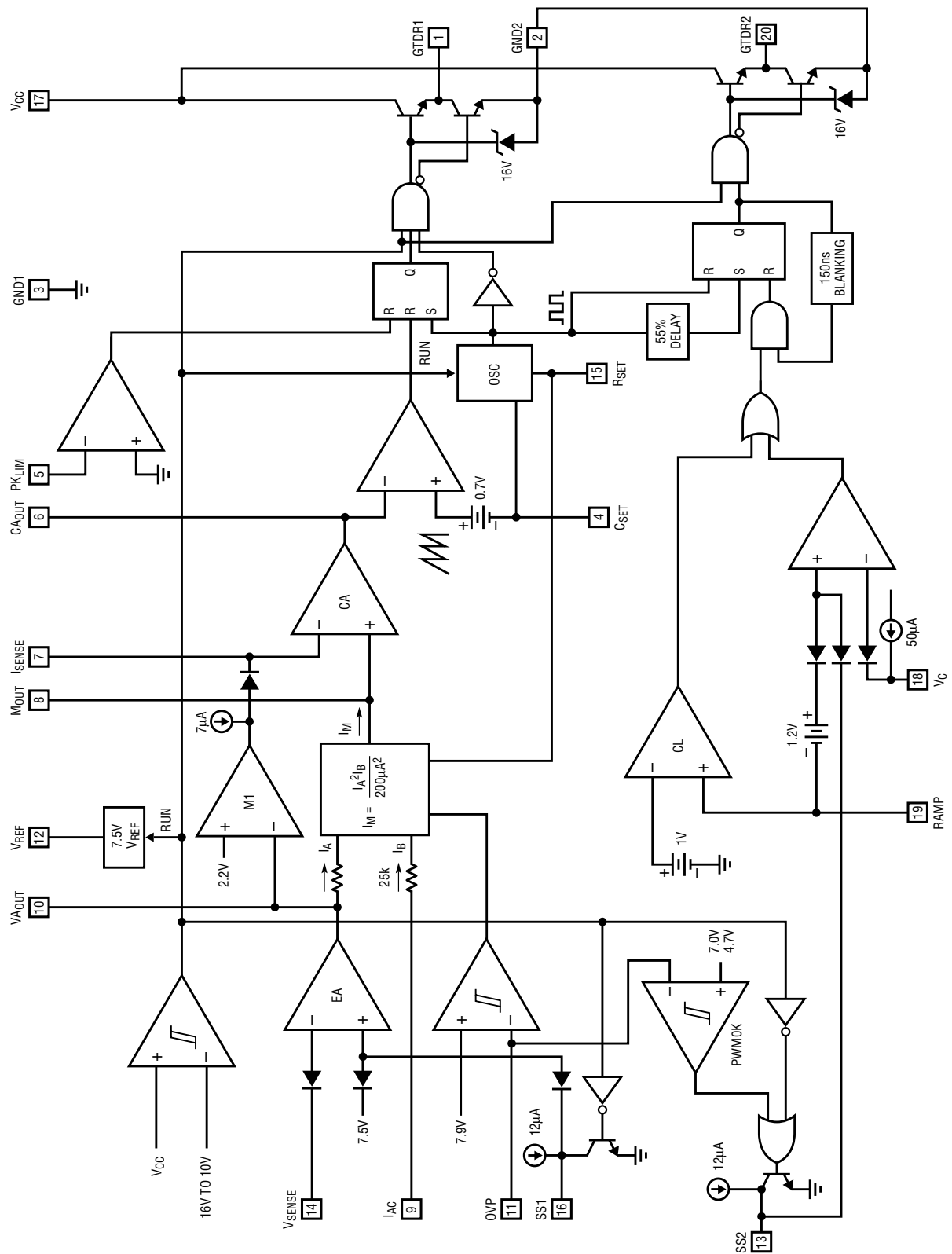


Figure 2. Block diagram: LT1509

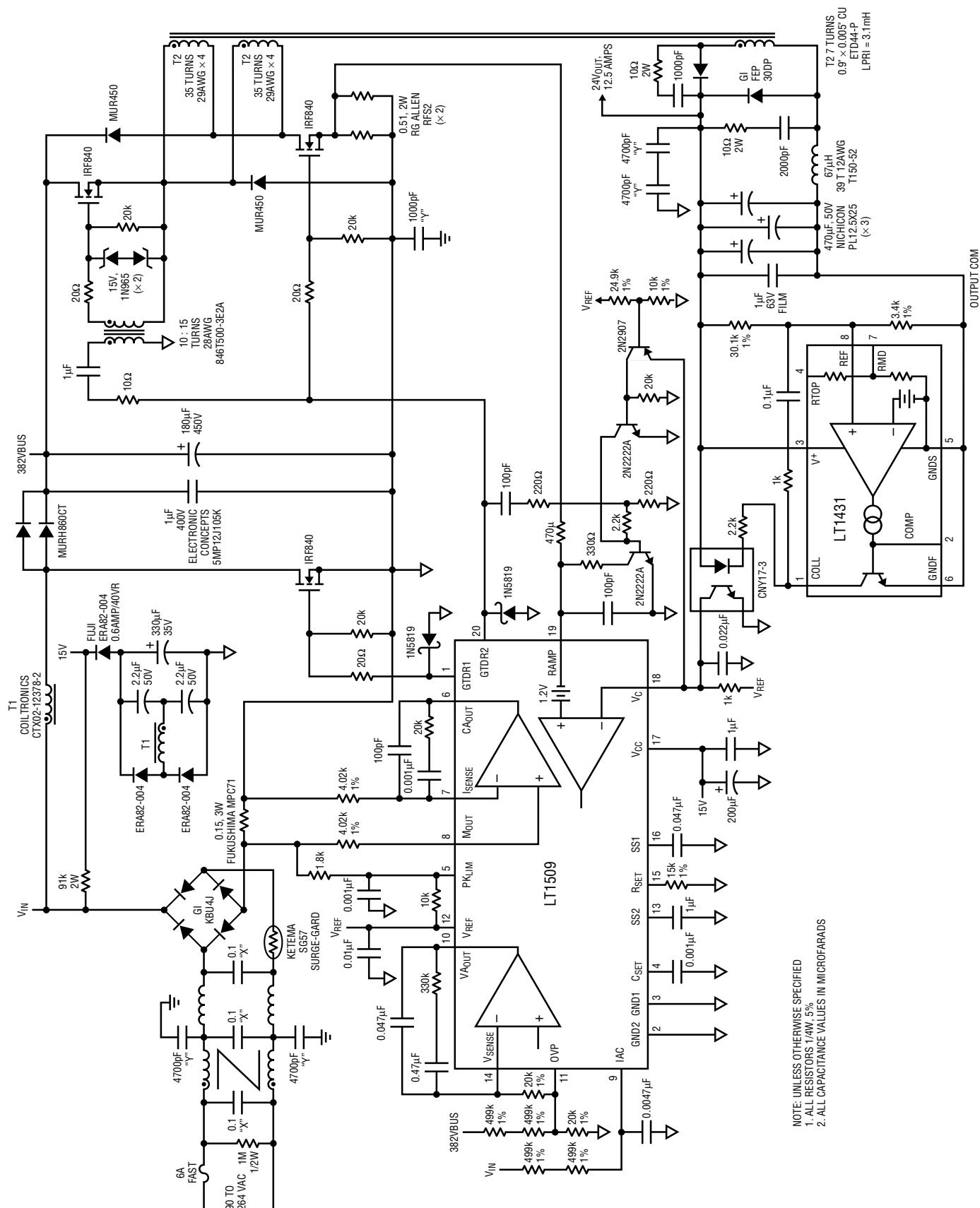


Figure 3. Schematic diagram of 300W, 24V DC output, power-factor corrected, universal-input supply

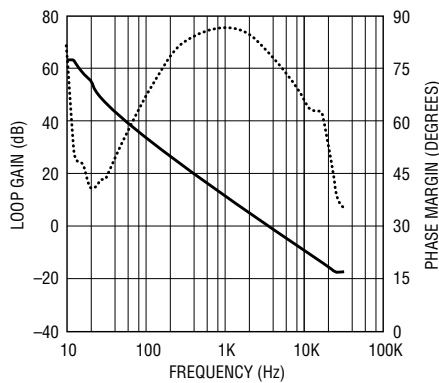


Figure 4. Bode plot of the circuit shown in Figure 3

secondary added to the 70-turn primary of T1, bootstraps V_{CC} to about 15VDC, supplying the chip's 13mA requirement as well as about 25mA to cover the gate current of the three FETs and high-side transformer losses. A 0.15 Ω sense resistor senses input current and compares it to a reference current (I_M) created by the outervoltage loop and multiplier. Thus the input current follows the input line voltage and changes, as necessary, in order to maintain a constant

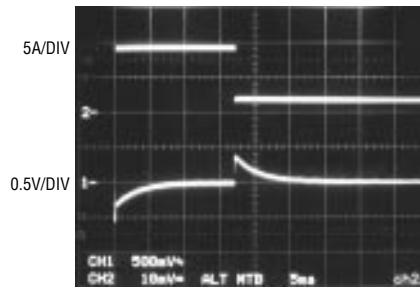


Figure 5. Figure 3's response to a 2A to approximately 10A load step

bank voltage. The forward converter sees a voltage input of 382VDC unless the line voltage drops out, in which case the 180 μ F main capacitor discharges to 250VDC before the PWM stage is shut down. Compared to a typical off-line converter, the effective input voltage range of the forward converter is smaller, simplifying the design. Additionally, the higher bus voltage provides greater hold-up times for a given capacitor size. The high-side transformer effectively delays the

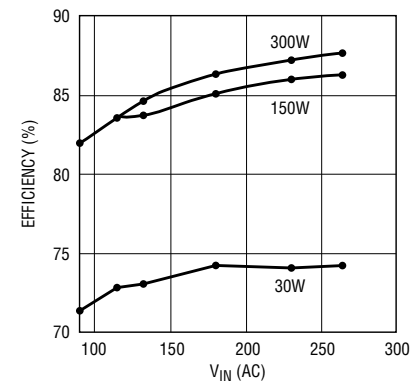


Figure 6. Efficiency curves for Figure 3's circuit

turn-on spike to the end of the built-in blanking time, necessitating the external blanking transistor.

Conclusion

The LT1508 and LT1509 offer complete solutions for low-to-high power, isolated, off-line power-factor-corrected supplies. The many built-in features ensure a simplified, low-parts-count design for a variety of applications. **LT**

1118, continued from page 6

The enable pin provides many other useful capabilities to the resourceful designer. Figure 5 shows two LT1118-5 regulators configured to provide sequential turn-on and turn-off from logic control. A logic high level turns on the first regulator; the second regu-

lator turns on only after the first output voltage is stable. Turn-off is in the reverse order, due to the rapid discharge of capacitor C3 through the diode.

These and other applications will benefit from the LT1118 family's

unique performance attributes. Any application that requires fast transient load response, unconditional stability under capacitive loading, or logic control of the power output is a potential application for the LT1118 family. **LT**

PFC, continued from page 10

off-time. This allows us to model the discontinuous-mode boost converter as a sum of discrete events. Each event consists of an on-time and an off-time, with the off-time being longer than the time it takes to empty all stored energy from the inductor.

Figure 2c shows that the off-time consists of two separate times:

- ❑ the time when the diode is conducting, the active off-time.
- ❑ the time after the diode has committed, yet before the next on-time, the inactive off-time.

The Energy Bucket

We look at a discontinuous-mode boost converter as an energy bucket that gets filled and dumped with each cycle. During each cycle the bucket takes a certain amount of energy (J). Over a given number of cycles the bucket will intercept many bucketsful of energy and deliver them to the load. A boost converter operating at a switching frequency of f and intercepting an amount of energy of J during each cycle, will be intercepting $f \times J$ watts of power from the input

power line. Because this converter is very efficient, essentially all of the intercepted power is delivered to the load.

This concludes part three of our series introducing the power factor correction conditioner. **LT**

LTC1329 Micropower, 8-Bit, Current Output DAC

by K.S. Yap

The LTC1329 is a micropower, 8-bit, current output DAC that provides precision current output over a wide supply range. This part is ideal for portable and battery-powered applications.

Description

The LTC1329 can communicate with external circuitry by means of one of three interface modes: standard, 3-wire serial-data mode, 1-wire pulse mode, and 2-wire pulse mode. In serial-data mode, the system microprocessor can serially transfer the 8-bit data to and from the LTC1329. In pulse mode, the upper 6 bits of DAC output can be programmed for increment-only (1-wire interface) or increment/decrement (2-wire interface) operation. In increment-only mode, when the count increases beyond full scale, the counter rolls over and sets the DAC to zero. In increment/decrement mode, the counter stops incrementing at full scale and stops decrementing at zero, and will not roll over.

The LTC1329 can operate over a wide supply range from 2.5V to 6.5V, with a very low quiescent current of only 50 μ A, which drops to only 0.2 μ A in shutdown mode. The DAC current output can be biased from -20V to +2.5V. The full-scale DAC current output is 10 μ A \pm 2.5% and 50 μ A \pm 2.5% for the -10 and -50 versions respectively. On power-up, the DAC current output is automatically set at mid-range.

The LTC1329's micropower operation makes it ideal for portable and battery-powered applications like LCD contrast control, backlight brightness control, power supply voltage adjustment, battery charger voltage/current adjustment, and trim pot replacement. The LTC1329 is available in 8-pin surface mount and DIP packages.

Power Supply

Voltage Adjustment

Figure 1 is a schematic of a digitally controlled power supply voltage adjustment circuit using a 2-wire interface. The LT1107 is configured as a step-up DC/DC converter, with the output voltage (V_{OUT}) determined by the values of the feedback resistors. The LTC1329's DAC current output is connected to the feedback node of these resistors, and an 8051 microprocessor is used to interface to the LTC1329. By simply clocking the LTC1329, the DAC current output is decreased or increased (decreased if $D_{IN} = 0$, increased if $D_{IN} = 1$), causing V_{OUT} to change accordingly.

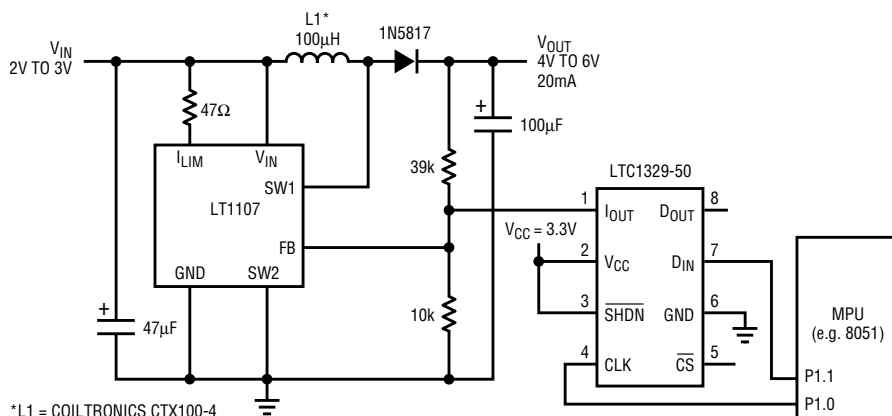


Figure 1. LTC1329 digitally controls the output voltage of a power supply

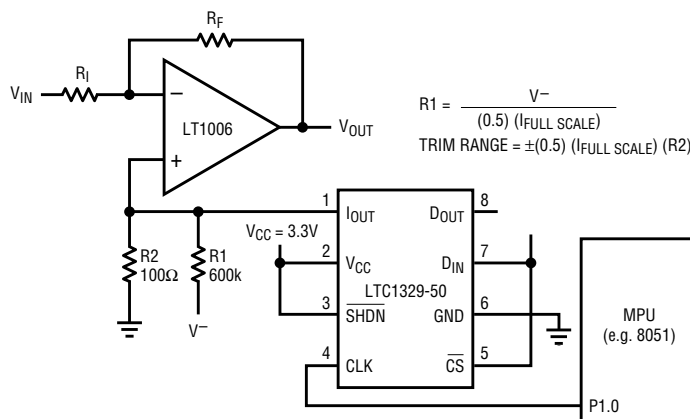


Figure 2. LTC1329 used to null op amp's offset voltage

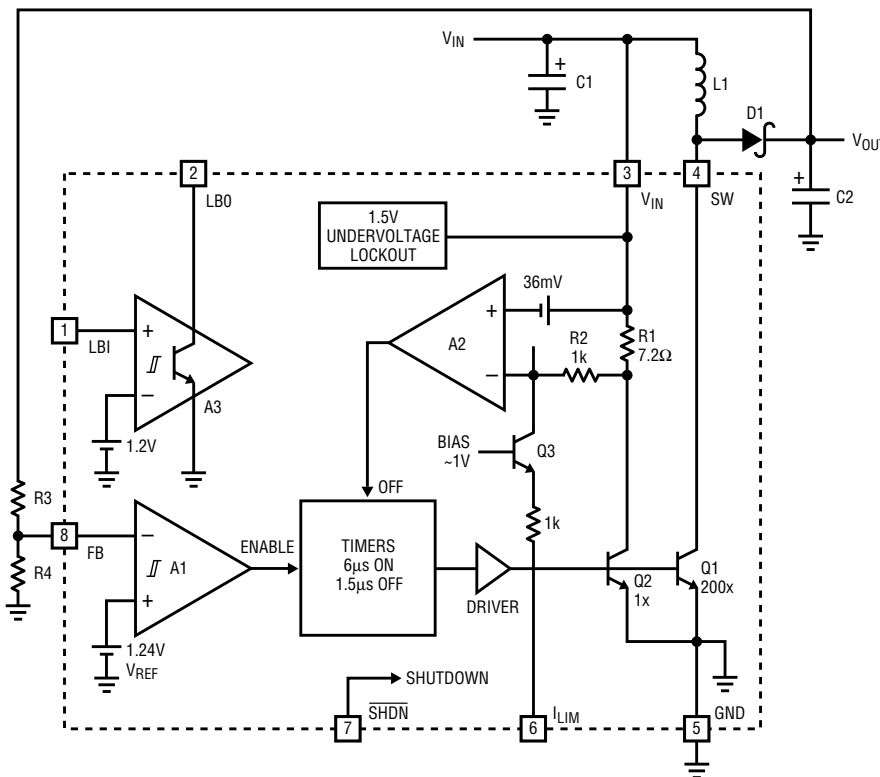


Figure 1. LT1304 Block diagram. Independent low-battery detector A3 remains alive when device is in shutdown

LT1304, continued from page 1

current. When R1's voltage drop exceeds 36mV, corresponding to 1A switch current, A2's output goes high, truncating the on-time part of the switch cycle. The 1A peak current can be reduced by tying a resistor between the I_{LIM} pin and ground, causing a voltage drop to appear across R2. The drop offsets some of the 36mV reference voltage, lowering peak current. A 22k resistor limits current to approximately 550mA. A capacitor connected between I_{LIM} and ground provides soft start. Shutdown is accomplished by grounding the \overline{SHDN} pin.

The low-battery detector A3 has its own 1.2V reference and is always on. The open collector output device can sink up to 500μA. Approximately 35mV of hysteresis is built into A3 to reduce "buzzing" as the battery voltage reaches the trip level.

A 2-Cell-to-5V Converter

A compact 2-cell-to-5V converter can be constructed using the circuit of Figure 2. Using the LT1304-5 fixed-

output device eliminates the need for external voltage-setting resistors, lowering component count. As the battery voltage drops, the circuit continues to function until the LT1304's undervoltage lockout disables the part at approximately $V_{IN} = 1.5V$. 200mA is available at a battery voltage of 2.0V. As the battery voltage decreases below 2V, cell impedance starts to quickly increase. End-of-life is usually assumed to be around 1.8V, or 0.9V per cell. Efficiency is detailed in Figure 3. Burst Mode micropower operation keeps efficiency above 70%, even for load current below 1mA. Efficiency reaches 85% for a 3.3V input. Load transient response is illustrated in Figure 4. Since the LT1304 uses a hysteretic comparator in place of the traditional linear feedback loop, the circuit responds immediately to changes in load current. Figure 5 details start-up behavior without soft-start circuitry (R1 and C1 in Figure 2). Input current rises to 1A as the device is turned on, which can cause the input supply voltage to

sag, possibly tripping the low-battery detector. Output voltage reaches 5V in approximately 1ms. The addition of R1 and C1 to Figure 2's circuit limits inrush current at start-up, providing for a smoother turn-on, as indicated in Figure 6.

A 4-Cell-to-5V Converter

A 4-cell-to-5V converter is more complex than a simple boost converter because the input voltage can be either above or below the output voltage. The single-ended primary inductance converter (SEPIC) shown in Figure 7 accomplishes this task, with the additional benefit of output isolation. In shutdown conditions, the converter's output will go to zero, unlike the simple boost converter, where a DC path from input to output through the inductor and diode remains. In this circuit, peak current is limited to approximately 500mA by the addition of 22k resistor R1. This allows very small, low-profile components to be used. The 100μF capacitors are D-case size, with a height of 2.9mm, and the inductors are 3.2mm high. The circuit can deliver 5V at up to 100mA. Efficiency is relatively flat across the 1mA-to-100mA load range.

Super Burst™ Mode Operation: 5V/100mA DC/DC with 15μA Quiescent Current

The LT1304's low-battery detector can be used to control the DC/DC converter. The result is a reduction in quiescent current by almost an order of magnitude. Figure 9 details this Super Burst circuit. V_{OUT} is monitored by the LT1304's LBI pin via resistor divider R1/R2. When LBI is above 1.2V, LBO is high, forcing the LT1304 into shutdown mode and reducing current drain from the battery to 10μA. When V_{OUT} decreases enough to overcome the low battery detector's hysteresis (about 35mV), LBO goes low. Q1 turns on, pulling \overline{SHDN} high and turning on the rest of the IC. R3 limits peak current to 500mA; it can be removed for higher output power. Efficiency is illustrated in Figure 10.

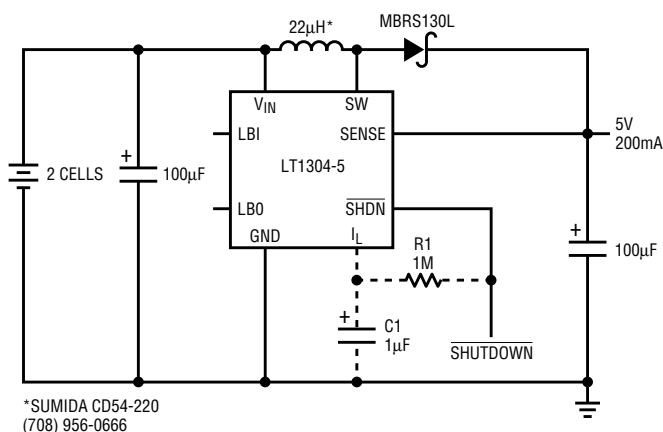


Figure 2. 2-cell-to-5V/200mA boost converter takes four external parts. Components with dashed lines are for soft-start (optional)

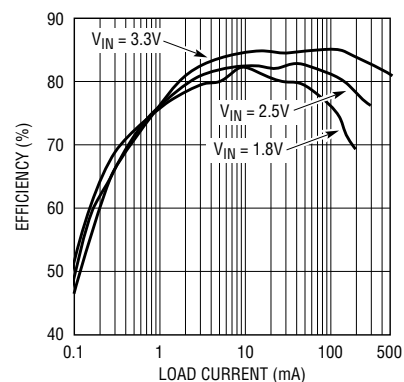


Figure 3. 2-cell-to-5V converter efficiency

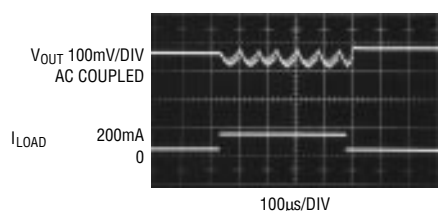


Figure 4. Boost converter load-transient response with $V_{IN} = 2.2V$

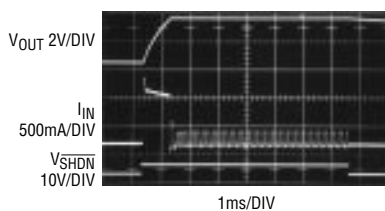


Figure 5. Start-up response. Input current rises quickly to 1A. V_{OUT} reaches 5V in approximately 1ms. Output drives 20mA load

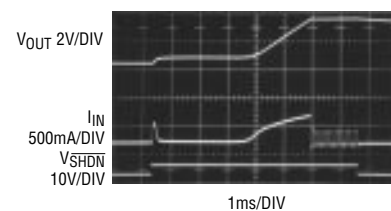


Figure 6. Start-up response with 1µF/1MΩ components in Figure 2 added. Input current is more controlled. V_{OUT} reaches 5V in 6ms. Output drives 20mA load

The converter is approximately 70% efficient at a 100µA load, 20 points higher than the circuit of Figure 2. Even at a 10µA load, efficiency is in the 40%-50% range, equivalent to 100µW–120µW total power drain from the battery. In contrast, Figure 2's circuit consumes approximately 300µW–400µW unloaded.

An output capacitor charging cycle or "burst" is shown in Figure 11, with the circuit driving a 50mA load. The slow response of the low-battery detector results in the high number of individual switch cycles or "hits" within the burst.

Figure 12 depicts output voltage at the modest load of 100µA. The burst

repetition rate is around 4Hz. With the load removed, the repetition rate drops to approximately 0.2Hz, or one burst every 5 seconds. Systems that spend a high percentage of operating time in sleep mode can benefit from the greatly reduced quiescent power drain of Figure 9's circuit.

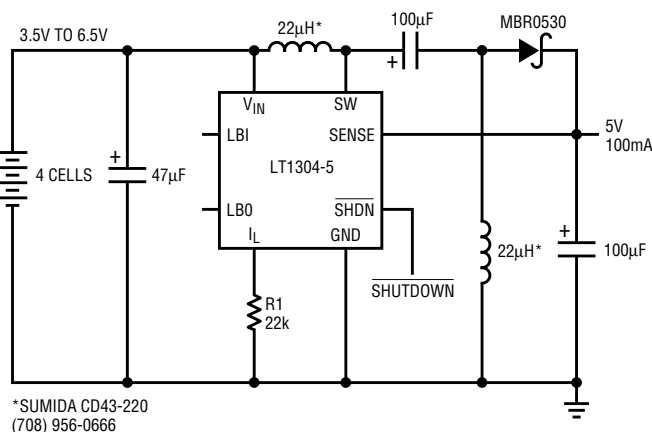


Figure 7. 4-cell-to-5V step-up/step-down converter, also known as SEPIC (single-ended primary inductance converter). Low profile components are used throughout

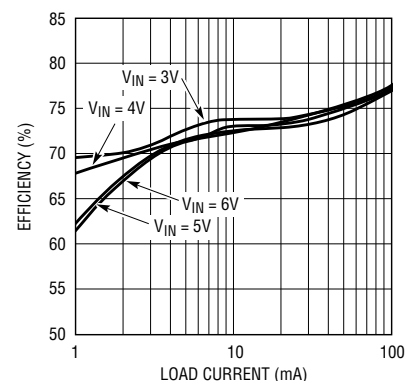


Figure 8. Efficiency plot of SEPIC converter shown in Figure 7

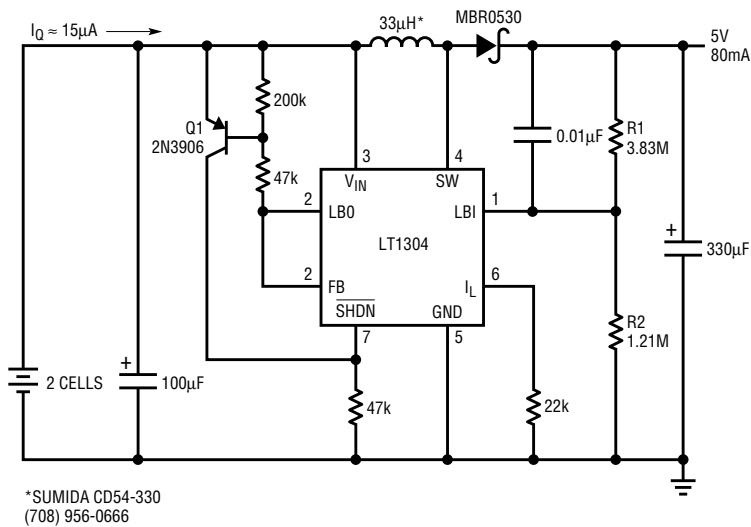


Figure 9. 2-cell-to-5V DC/DC converter using Super Burst Mode Operation draws only 15µA unloaded. Two AA Alkaline cells will last for years

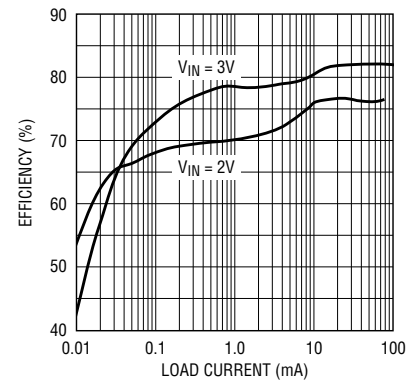


Figure 10. DC/DC converter efficiency of Figure 9

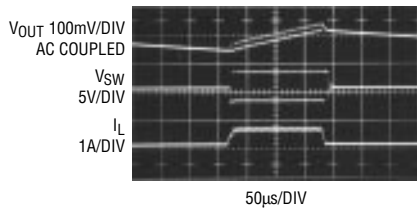


Figure 11. Super Burst Mode Operation in action. $V_{IN} = 2.5V$, $I_{LOAD} = 50mA$

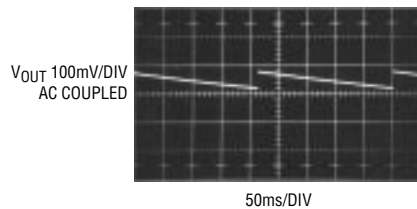


Figure 12. Circuit using Super Burst Mode Operation, 100µA load. Burst occurs approximately once every 240ms

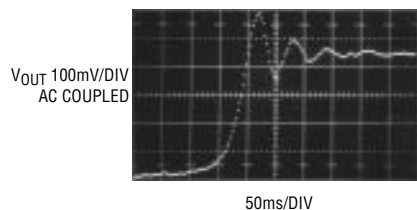


Figure 14. LT1304 Switch rise time is in the 5ns range. These types of edges emphasize the need for proper PC board layout

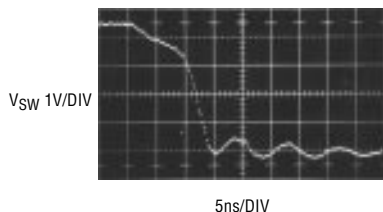


Figure 15. Switch fall time. Lower slope in second and third gratitudes shows effect of lead and bond wire inductance

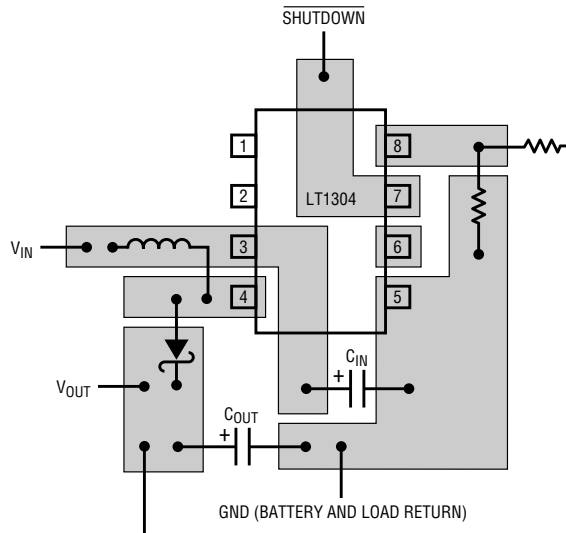



Figure 13. Suggested layout for best performance. Input capacitor placement as shown is highly recommended. Switch trace (pin 4) copper area is minimized

Layout

The LT1304 switch turns on and off very quickly. For best performance we suggest the component placement in Figure 13. Improper layouts will result in poor load regulation, especially at heavy loads. Parasitic lead inductance must be kept low for proper operation. Switch turn-off is detailed in Figure 14¹. A close look at the rise time (5ns) will confirm the need for good PC board layout. The 200MHz ringing of the switch voltage

is attributable to lead inductance, switch and diode capacitance, and diode turn-on time. Switch turn-on is shown in Figure 15. Transition time is similar to that of Figure 14. Adherence to the layout suggestions will result in working DC/DC converters with a minimum of trouble. 

1. Instrumentation for oscillographs of Figures 14 and 15 include Tektronix P6032 active probe, Type 1S1 sampling unit and type 547 mainframe.

RS232 Transceiver Protected to $\pm 15\text{kV}$ IEC-801-2 ESD Standards

by Gary Maulding

Linear Technology's popular LT1137A RS232 transceiver has been upgraded to include on-chip ESD protection compliant with IEC-801-2 level 4 standards. Device pins connected to the RS232 line can absorb up to $\pm 15\text{kV}$ air-gap or $\pm 8\text{kV}$ contact-mode transients using the IEC-801-2 ESD model, with no damage or interruption of operation. The 28-lead SOIC or SSOP packaged circuit and five inexpensive ceramic capacitors are the only components required to implement a 5V powered AT serial port that is fully compliant with RS232 or V.28 electrical specifications and IEC-801-2 static-discharge standards.

ESD Effects on Electronic Equipment

Electrostatic discharge (ESD) is a well known killer of semiconductor devices. Unless protective measures are taken, ESD will result in various forms of device degradation. Discharge damage ranges from increased leakage currents on device inputs and outputs, to increased offset or shifted threshold voltages, to failure of the affected circuit.

To minimize the monetary loss due to damaged IC's and to improve the reliability of equipment, the electronics industry has adopted many standard safeguard procedures against ESD. These safeguards include the use of grounded work surfaces, equipment, and personnel;

conductive plastic shipping materials, on-chip protection circuits at the input and output pins, and testing of the ESD sensitivity of circuits.

The "Human Body Model," defined by Mil-Std-883B Method 3015, is a commonly used ESD test standard for rating integrated circuits for ESD-damage immunity. This test method is aimed directly at reducing the incidence of failed devices installed on equipment circuit boards. The test discharges a charged 100pF capacitor through a 1.5k Ω resistance directly to the IC's pins. Both positive and negative voltages are tested. Rarely is the device powered when testing is conducted, and only permanent damage to the circuit is considered.

The widespread adoption of environmental static control and the design of integrated circuits with on-chip protection devices has dramatically reduced the incidence of circuit failures due to ESD damage during circuit manufacturing and handling. Most semiconductor devices enter a much safer environment once installed in a PC board. Installation of the PC board into a chassis further shields the IC from the environment. No longer are most IC's input and output pins exposed to direct ESD strikes, but ESD is still an operational and reliability hazard for electronic equipment. Everyone has experienced the shock caused by walking across a nylon carpet on a dry winter day and touching a piece of electrical equipment. This static discharge can damage internal circuitry of the equipment if the manufacturer does not provide a safe discharge path for the energy. Input/output ports on personal computers are especially vulnerable. The

ports are available so that users can attach cabling to printers, mice, or networks. Portable computers are particularly vulnerable since the port connections are made and broken much more often than for desktop machines. Most consumers would not like to wear static control ground straps when using their personal computers. Hence, it is important that ESD protection be built in to the end user equipment.

What is IEC-801-2?

The International Electrotechnical Commission (IEC) has defined a test and rating standard for electronic equipment exposed to ESD transients. IEC-801-2 defines two alternate test methodologies. The preferred method is a contact-mode test, but the alternate air-gap test is more easily implemented and is more commonly used. Equipment certified for IEC-801-2 testing consists of a charging circuit to apply the desired test voltage to a 150pF capacitor. A discharge switch discharges the energy into the equipment under test through a 330 Ω resistance (see Figure 1). IEC-801-2 certified test equipment must be able to demonstrate that the applied test

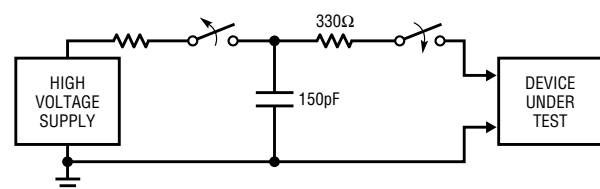


Figure 1. IEC-801-2 ESD test system

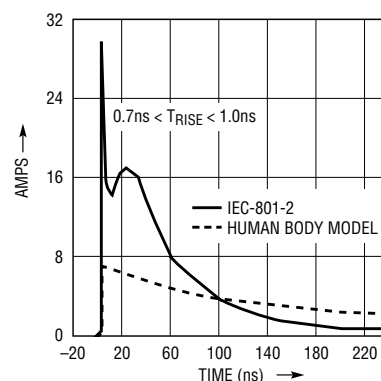


Figure 2. ESD Transient current waveforms

voltage and current waveform meet the prescribed rise and decay wave-shape shown in Figure 2. Linear Technology uses a Keytek MiniZap for IEC-801-2 ESD testing.

The test results on equipment are classified by the voltage level successfully applied to the equipment and the affect of test discharges on the equipment. The IEC-801-2 standard defines the following performance criteria:

1. Normal performance within specification limits
2. Temporary degradation or loss of function or performance that is self-recoverable
3. Temporary degradation or loss of function or performance that requires operator intervention or system reset
4. Degradation or loss of function that is not recoverable, due to damage of equipment (components) or software, or loss of data

Table 1 details the rated test levels for both contact and air-gap tests. The contact-mode test is the preferred methodology due to its superior repeatability. The air-gap test may be influenced by the test environment's relative humidity, as well as by the approach speed of the tester tip to the device under test.

LT1137A ESD Test Performance

The LT1137A is an RS232 interface transceiver with an integral charge pump for single 5V operation. The complement of three drivers and five receivers matches the requirements for AT serial ports on personal computers. The eight RS232 lines are directly connected to the socket on the computer's case, and are therefore directly exposed to possible ESD transients when cables are installed or removed from the machine. The inclusion of rugged ESD protection on the chip eliminates the need for the user to install costly and space-consuming transient suppression diodes on the RS232 lines.

The new, upgraded LT1137A has been tested to level 4 IEC-801-2 testing. In both the $\pm 8\text{kV}$ contact-mode test and the $\pm 15\text{kV}$ air-gap test, the transceiver suffered no damage and no interruption of normal operation.

Linear Technology provides this enhanced ESD protection on the LT1137A at no extra cost to the user. The improved level of protection requires no increase in die area over the older $\pm 10\text{kV}$ protected devices. The user must only follow the data sheet and use low ESR capacitors (ceramic chip caps are perfect) located close to the device pins to obtain the benefits of the on-chip protection (see Figure 3). The capacitors and the length of

PC board traces are important, since the on-chip protection devices shunt the ESD transient energy through the capacitors and away from active circuitry. The user should take care to provide a short ground return path for the energy in order not to cause secondary damage to other PC board components caused by the high-speed, high-current transients that will flow during an ESD discharge.

Is ESD Protection Effective?

In 1992 Linear Technology introduced the first integrated circuits capable of withstanding $\pm 10\text{kV}$ ESD transients as tested per Mil-Std.-883B Method 3015 (Human Body Model). Since their introduction, millions of units have been installed in equipment, but not one single unit has been returned as a field failure due to ESD damage. Prior to the on-chip 10kV ESD protection, interface circuits would routinely be found to have suffered ESD damage when they were used in systems that did not use a TransZorb[®] for ESD protection. The on-chip ESD protection has proven to be a great enhancement to equipment reliability and a cost saving to computer manufacturers, who no longer need to add protective devices to their I/O ports. The enhanced protection achieved by meeting the IEC-801-2 test standards will prolong this perfect record. **LT**

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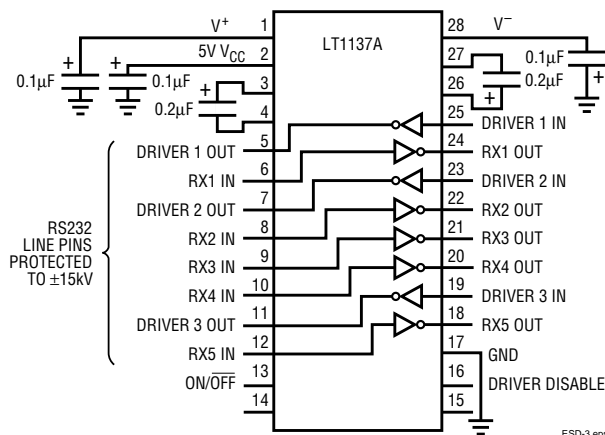


Figure 3. LT1137A ESD test circuit

Table 1. IEC-801-2 severity levels

Level	Test Voltage	
	Contact Discharge (kV)	Air Discharge (kV)
1	2	2
2	4	4
3	6	8
4	8	15

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V.35 Transceivers Allow 3-chip V.35 Port Solution

by Y.K. Sim

Two new LTC interface devices, the LTC1345 and the LTC1346, provide the differential drivers and receivers needed to implement a V.35 interface. When used in conjunction with an RS232 transceiver like the LT1134A, they allow a complete V.35 interface to be implemented with just two transceiver chips and one resistor-termination chip. The signals in a typical V.35 interface are divided into two groups, with five high-speed differential signals for clock and data transfer and five or eight slower, single-ended V.28 (RS232) signals for handshaking. The LTC1345 and LTC1346 provide the three differential drivers and receivers necessary to implement the high-speed path, and the LT1134A provides the four RS232 drivers and receivers required for the handshaking interface. Both the LTC1345 and the LT1134A provide onboard charge pump power supplies, allowing a complete V.35 interface to be powered from a single 5V supply. For systems where $\pm 5V$ supplies are present, the LTC1346 is offered without charge pumps, representing a 30% power savings.

Appendix II of the CCITT V.35 specification describes the differential signal requirements for the high-speed interface lines. The specification stipulates a differential transmitter of 100Ω source termination impedance driving a 100Ω twisted-pair cable and a differential

receiver with 100Ω as load termination. The ground potential between the source and load is allowed to vary up to $\pm 4V$, and the common-mode resistance at both source and load is 150Ω . The specification calls for differential transmitter signals of $0.55V \pm 20\%$ and no more than $\pm 0.6V$ common mode, exclusive of ground-potential differences. Early V.35 implementations used constant-voltage-output drivers for the differential signals, and resistors in series with the outputs to provide the required $\pm 0.55V$ output swing. This technique suffered from a strong dependence on power supply voltage and driver output impedance to control the output voltage. Further, common-mode voltages present at the driver output had a strong effect on output swing, significantly limiting the driver-output common mode range.

The LTC1345/LTC1346 transceivers avoid these problems by using a constant-current, high impedance output driver. Designed to be used with matching external resistor networks to meet the Appendix II specifications, the transmitter output structure consists of two complementary switched current sources that maintain a $0.55V$ signal magnitude over $\pm 4V$ common mode range. These current sources are trimmed to source and sink $11mA$ at the complementary outputs. The differential output voltage of the transmitter is set by the external source and load termination resistor networks. The typical differential source/load resistance of 100Ω gives an output voltage of $0.55V$ ($11mA \times 100\Omega/2$). The common mode voltage at the transmitter is typically less than $0.2V$ with zero ground potential difference. The LTC1345 and LTC1346 require 5% tolerance in the termination resistors to meet the 20% differential voltage variation specification. A termination resistor

network such as BI Technologies part number 627T500/1250 (SOIC) or 899TR50/125 (DIP) meets this specification and is recommended for proper cable termination. Discrete 5% resistors can also be used to terminate the drivers and receivers; either a Y (wye, Figure 1A) or Δ (delta, Figure 1B) connection can be used to give the required 100Ω differential and 150Ω common mode impedances.

The LTC1345/LTC1346 differential input receivers have $30k\Omega$ input impedance and a typical hysteresis of $40mV$. This allows the receivers to meet the V.35 specification with the specified termination resistor network. Additionally, they can be used to detect RS422 signals if the 125Ω V.35 common mode resistor is removed. Both the LTC1345 and LTC1346 transceivers feature four digitally-selectable operating modes: DTE, with two drivers and three receivers active; DCE, with three drivers and two receivers active; all three drivers and receivers active, and shutdown. In shutdown, supply current drops to $1\mu A$ typically. The differential transceivers are capable of operating at data rates above 10MBaud in non-return-to-zero (NRZ) format.

The RS232 handshaking lines can be implemented with standard RS232 transceivers. The LT1134A provides four RS232 drivers and four receivers, enough to implement the extended 8-line handshaking protocol specified in V.35. The LT1134A also includes an onboard charge pump to generate the higher voltages required by RS232 from a single 5V supply, making it an ideal companion to the LTC1345. These two chips, together with the BI Technologies termination resistor network chip, provide a complete, surface mountable, 5V-only V.35 data port. Systems that have multiple power supply

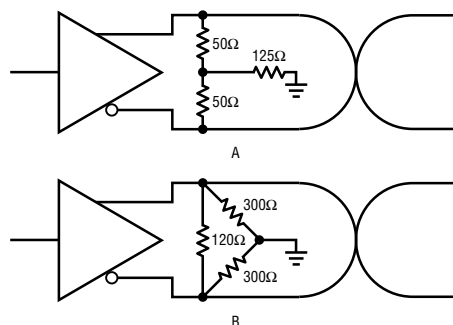


Figure 1. Y and Δ termination topologies

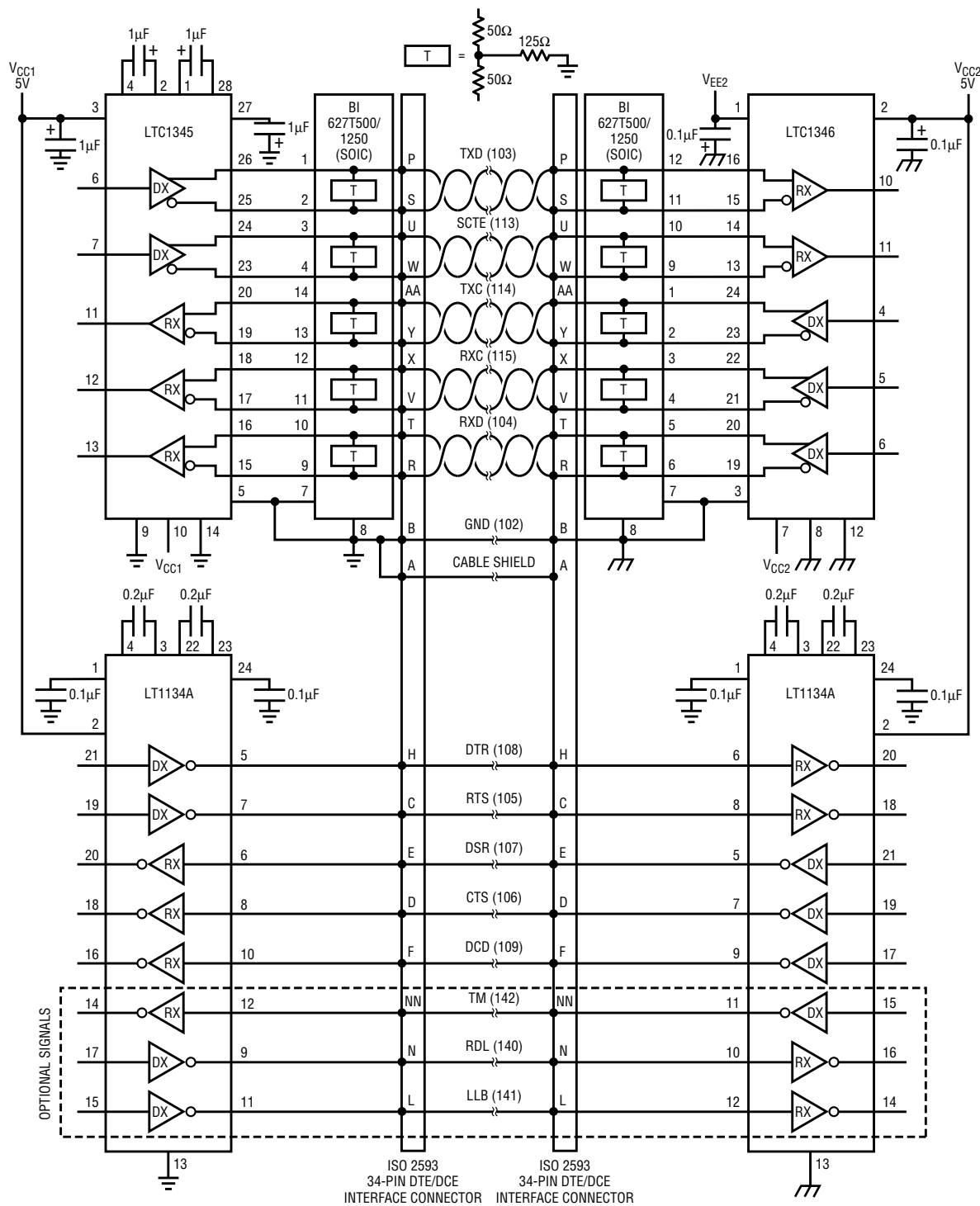


Figure 2. Typical V.35 implementation using LTC1345 and LTC1346

voltages available and use only the simpler 5-signal V.35 handshaking protocol can use the LTC1346 with the LT1135A or LT1039 RS232 trans-

ceivers; this combination provides a complete port while saving board space and complexity. Figure 2 shows a typical LTC1345/LTC1134A V.35

implementation with five differential signals and five basic handshaking signals, with an option for three additional handshaking signals. \blacksquare

1MHz Step-Down Converter Ends 455kHz IF Woes

by Mitchell Lee

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Fully Differential, 8-Channel, 12-Bit A/D System Using the LTC1390 and LTC1410 36


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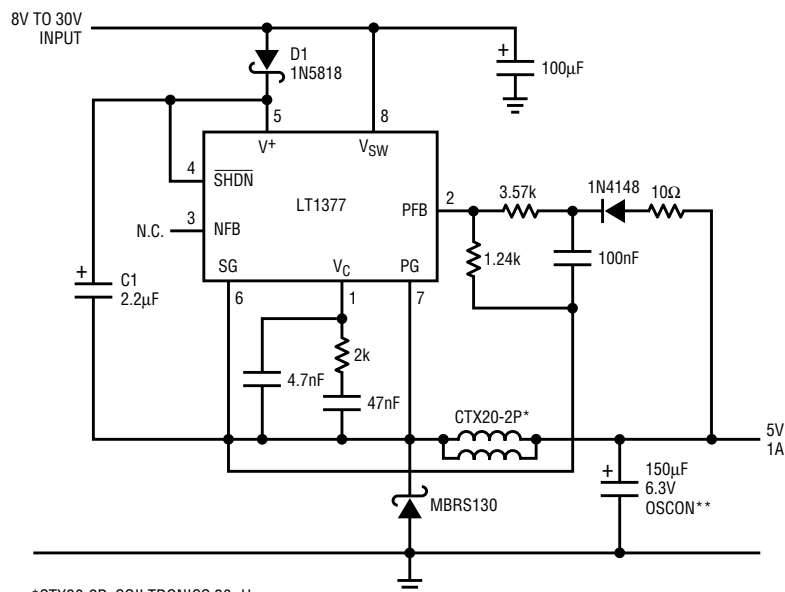
There can be no doubt that switching power supplies and radio IFs don't mix. One-chip converters typically operate in the range of 20kHz–100kHz, placing troublesome harmonics right in the middle of the 455kHz band. This contributes to adverse effects such as "de-sensing" and outright blocking of the intended signals. A new class of switching converter makes it possible to mix high-efficiency power supply techniques and 455kHz radio IFs without fear of interference.

The circuit shown in Figure 1 uses an LT1377 boost converter, operating at 1MHz, to implement a high-efficiency buck-topology switching regulator. The switch is internally grounded, calling for the floating

supply arrangement shown (D1 and C1). The circuit converts inputs of 8V–30V to a 5V/1A output.

The chip's internal oscillator operates at 1MHz for load currents of greater than 50mA, with a guaranteed tolerance of 12% over temperature. Even wideband 455kHz IFs are unaffected, as the converter's operating frequency is well over one octave distant.

Figure 2 shows the efficiency of Figure 1's circuit. You can expect 80%–90% efficiency over an 8V–16V input range with loads of 200mA or more. This makes the circuit suitable for 12V battery inputs (that's how I'm using it), but no special considerations are necessary with adapter inputs of up to 30V. 



*CTX20-2P, COILTRONICS 20µH
**OSCON, SANYO VIDEO COMPONENTS

Figure 1. Schematic diagram: 1MHz LT1377-based boost converter

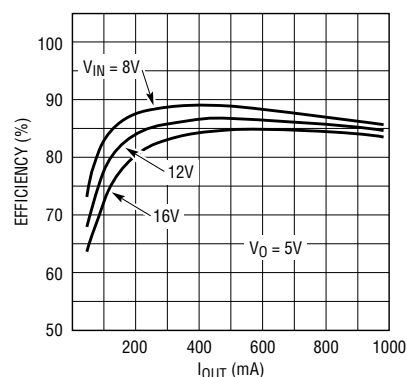


Figure 2. Efficiency graph of the circuit shown in Figure 1

A Circuit That Smoothly Switches Between 3.3V and 5V

by Doug La Porte

Many subsystems require supply switching between 3.3V and 5V to support both low-power and high-speed modes. This back-and-forth voltage switching can cause havoc to the main 3.3V and 5V supply buses. If done improperly, switching the subsystem from 5V to 3.3V can cause a momentary jump on the 3.3V bus, damaging other 3.3V devices. When switching the subsystem from 3.3V to 5V, the 5V supply bus can be pulled down while charging the subsystem's capacitors, and may inadvertently cause a reset.

The circuit shown in Figure 1 allows smooth voltage switching between 3.3V and 5V, with added protection features to ensure safe operation. IC1 is an LTC1470 switch-matrix device. This part has on-chip charge pumps running from the 5V supply to fully enhance the internal N-channel MOSFETs. The LTC1472 also has guaranteed break-before-make switching to prevent cross-conduction between buses. It also features current limiting and thermal shutdown.

When switching the subsystem from 5V to 3.3V, the holding capacitor and the load capacitance are initially charged up to 5V. Connecting these capacitors directly to the main 3.3V bus causes a momentary step to 5V. This transient is so fast that the power supply cannot react in time. Switching power supplies have a particularly difficult time coping with this jump. Switching supplies source current to raise the supply voltage and require the load to sink current to lower the voltage. A switching supply will be unable to react to counter the large positive voltage step. This jump will cause damage to many low-voltage devices.

The circuit in Figure 1 employs a comparator (IC2) and utilizes the high impedance state of the LTC1470 to allow switching with minimal effect on the supply. When the 3.3V output is selected, IC1's output will go into a high impedance state until its output falls below the 3.3V bus. The output capacitors will slowly discharge to 3.3V, with the rate of discharge depending on the current being pulled by the subsystem and the size of the

holding capacitor. The example shown in Figure 1 is for a 250mA subsystem. The discharge time constant should be about 4ms. Once the subsystem supply has dropped below the 3.3V supply, the comparator will trip, turning on the 3.3V switch. The comparator has some hysteresis to avoid instabilities. The subsystem supply will reach a low point of about 3V before the 3.3V switch is fully enhanced.

When switching from 3.3V to 5V, IC1's current limiting prevents the main 5V bus from being dragged down while charging the holding capacitor and the subsystem's capacitance. Without current limiting, the inrush current to charge these capacitors could cause a droop in the main 5V supply.

If done improperly, supply voltage switching leads to disastrous system consequences. The voltage switch should monitor the output voltage and have current limiting to prevent main supply transient problems. A correctly designed supply voltage switch avoids the pitfalls and results in a safe, reliable system. **LT**

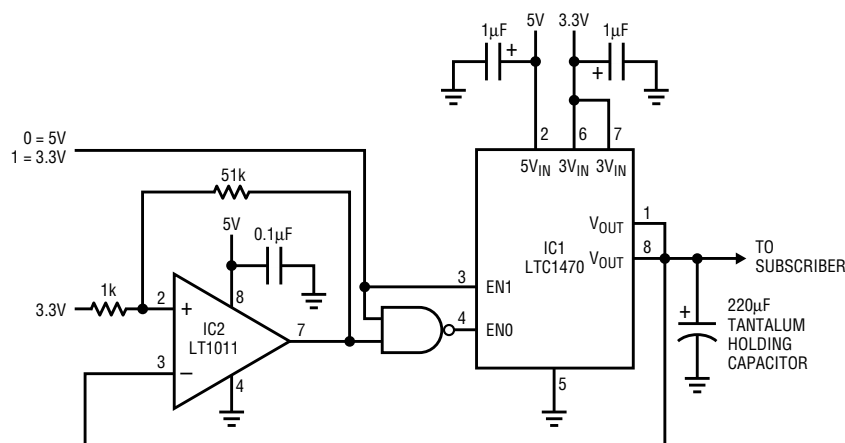


Figure 1. Schematic diagram of 3.3V and 5V switchover circuit

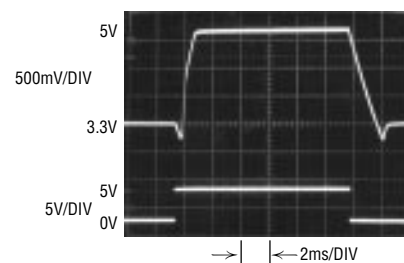


Figure 2. Oscilloscope of the switchover waveform showing smooth transitions

by Kevin R. Hoskins

Introduction

Let's face it, microprocessors and microcontrollers cannot control mechanical potentiometers. They have no fingers, just pins. And they have neither eyes with which to set a wiper's position nor the intuition to know when it's set correctly. They need the organs that provide these functions transplanted into their "operational environment." The "organs" that fill these roles are analog-to-digital converters (ADCs) and the sensors connected to their inputs, and digital-to-analog converters (DACs) and the manipulative circuits or mechanics they drive.

This article concentrates on the “fingers” and “hands” capabilities that DACs bestow upon microprocessors and microcontrollers. Linear Technology has recently introduced 12-bit, serially interfaced, voltage-output DACs packaged in small 8-pin SOs. These parts include the LTC1257, LTC1451, LTC1452, and LTC1453. All of these DACs are micropower, have 12-bit monotonic transfer functions, and operate on 4.75V to 15.75V (LTC1257), 5V (LTC1451), or 3V to 5V (LTC1452 and LTC1453) supply voltages. Table 1 captures the important specifications of the LTC1257, LTC1451, LTC1452, and LTC1453.

Figures 1 and 2 show the block diagrams of the LTC1257 and the LTC1451, LTC1452, and LTC1453. These devices share many common circuit elements. With a logic low applied to the $\overline{\text{CS}}$ input, serial data is applied to a 12-bit serial shift register. Each bit is latched on the rising

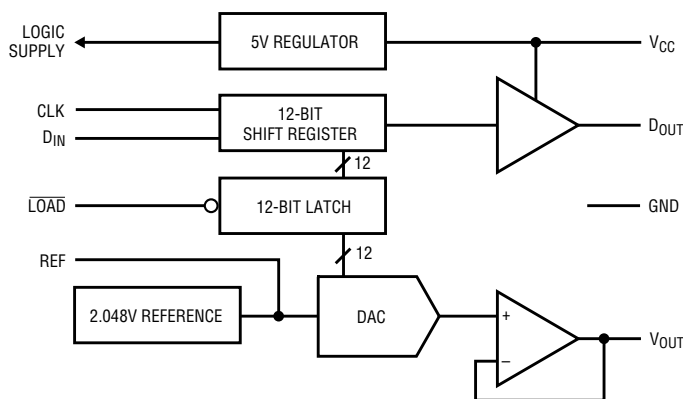


Figure 1. LTC1257 block diagram

edge of the CLK input signal. The shift register's contents are transferred to the DAC register on the rising edge of $\overline{\text{CS}}$, which in turn updates the output voltage's magnitude. Each of these DACs also has a data output, D_{OUT} , making it easy to daisy-chain multiple DACs on a single serial data line. The LTC1257 and LTC1451 have an on-chip 2.048V bandgap reference. The LTC1453 has a 1.22V bandgap reference. The LTC1257 has an internal 5V regulator that powers all onboard digital circuitry.

System Autoranging

System auto-ranging, adjusting an ADC's full-scale range, is an application area for which these DACs are appropriate. Auto-ranging is especially useful when using an ADC with multiplexed inputs. Without auto-ranging, only two reference values are used: one to set the full-scale magnitude and another to set the

zero-scale magnitude. Since it is common to have input signals with different zero-scale and full-scale magnitude requirements, fixed reference voltages present a problem. Although the ranges selected for some of the inputs may take advantage of the full range of ADC output codes, inputs that do not span the same range will not generate all codes, reducing the ADC's effective resolution. One possible solution is to match the reference voltage span to the need of each multiplexer input.

The circuit shown in Figure 3 uses two LTC1257s to set the full-scale and zero operating points of the LTC1296 12-bit, 8-channel ADC. The ADC shares its serial interface with the DACs. To further simplify bus connections, the DACs' data is daisy-chained. Two chip selects are used, one to select the LTC1296 when programming its multiplexer, and the other to select the DACs when setting their output voltages.

Table 1. Key specifications of the LTC1257, LTC1451, LTC1452, and LTC1453 8-pin, SO packaged, voltage-out DACs

Part Number	Supply Voltage Range	Supply Current (Typ)	Differential Nonlinearity (Max)	Integral Linearity Error (Max)
LTC1257	4.75V to 15.75V	350µA at 5V Supply	±0.5 LSB	±3.5 LSB
LTC1451	4.5V to 5.5V	400µA at $4.5V \leq V_{CC} \leq 5.5V$	±0.5 LSB	±3.5 LSB
LTC1452	2.7V to 5.5V	225µA at $2.7V \leq V_{CC} \leq 5.5V$	±0.5 LSB	±3.5 LSB
LTC1453	2.7V to 5.5V	250µA at $2.7V \leq V_{CC} \leq 5.5V$	±0.5 LSB	±3.5 LSB

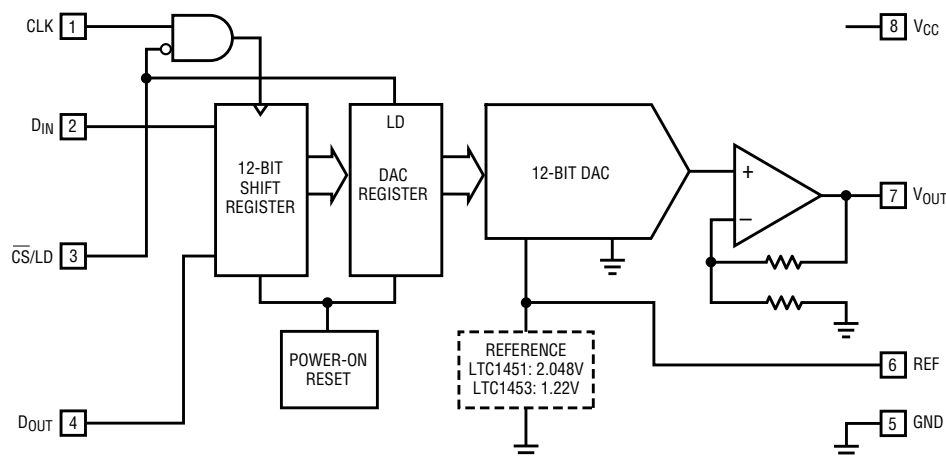


Figure 2. Block diagram of the LTC1451, LTC1452, and LTC1453

During the conversion process, U2 and U3 receive the full- and zero-scale codes, respectively, that correspond to a selected multiplexer channel. For example, let channel 2's span begin at 2V and end at 4.5V. When a host processor wants a conversion of channel 2's input signal, it first sends code that sets the output of U2 to 2V and U3 to 4.5V, fixing the span to 2.5V. The processor sends data to the LTC1296, selecting channel 2. The processor next reads data that was generated during the conversion of the 3.5V_{P-P} signal applied

to channel 2 from the LTC1296. As other multiplexer channels are selected, the DAC outputs are changed to match their spans.

Computer Controlled 4mA-to-20mA Current Loop

A common and useful circuit is the 4mA-to-20mA current loop. It is used to transmit information over long distances using varying current levels. The advantage of using current over voltage is the absence of IR losses and the transmissions errors and signal losses they can create.

The circuit in Figure 4 is a computer controlled 4mA-to-20mA current loop. It is designed to operate on a single supply over a range of 3.3V to 30V. The circuit's zero output reference signal, 4mA, is set by R1 and calibrated using R2, and its full-scale output current is set by R3 and calibrated using R4. The zero- and full-scale output currents are set as follows: with a zero input code applied to the LTC1453, the output current, I_{OUT} , is set to 4 mA by adjusting R2; next, with a full-scale code applied to the DAC, the full-scale output current is set to 20mA by adjusting R4.

The circuit is self regulating, forcing the output current to remain stable for a fixed DAC output voltage. This self regulation works as follows: starting at $t = 0$, the LTC1453's fixed output (in this example, 2.5V) is applied to the left side of R3; instantaneously, the voltage applied to the LT1077's input is 1.25V; this turns on Q1 and the voltage across R_S starts increasing beginning from 1.25V; as the voltage across R_S increases, it lifts the LTC1453's GND pin above 0V; the voltage across R_S continues to increase until it equals the DAC's output voltage.

Once the circuit reaches this stable condition, the constant DAC output voltage sets a constant current through $R3 + R4$ and $R5$. This constant current fixes a constant voltage across $R5$ that is also applied to the LT1077's noninverting input. Feedback from the top of R_S is applied to the inverting input. As the op amp forces its inputs to the same voltage, it will fix the voltage at the top of R_S . This in turn fixes the output current to a constant value.

Opto-Isolated Serial Interface

The serial interface of the LTC1451 family and the LTC1257 make opto-isolated interfaces very easy and cost effective. Only three opto-isolators are needed for serial data communications. Since the inputs of the LTC1451, LTC1452, and LTC1453 have generous hysteresis, the switching speed of the opto-isolators is not

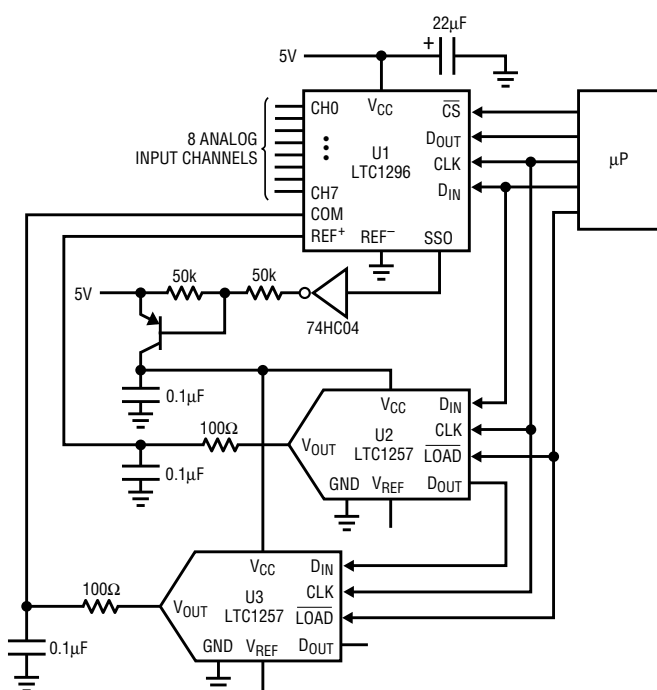


Figure 3. Using two LTC1257 12-bit voltage-output DACs to set the input span of the 12-bit, 8-channel LTC1296

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Wideband RMS Noise Meter

by Mitchell Lee

Recently, I needed to measure and optimize the wideband RMS noise of a power supply over about a 40MHz bandwidth. A quick calculation showed that the $12\text{nV}/\sqrt{\text{Hz}}$ noise floor of my spectrum analyzer would come up short—my circuit was predicted to exhibit a spot noise of perhaps $8\text{nV}/\sqrt{\text{Hz}}$. In fact, I didn't have a single instrument in my lab that would measure $50\mu\text{V}$ – $60\mu\text{V}$ RMS.

For the 40MHz bandwidth, the HP3403C RMS voltmeter is a good choice, but its most sensitive range is 100mV, about 66dB shy of my requirement. This obsolete instrument today carries a hefty price on the used market. The fact that here in the Silicon Valley HP3403Cs are a common sight at flea markets is of little consolation to most customers wishing to reproduce my measurements. We have several of these meters in the LTC design lab, but they are in constant use and closely guarded by "The Keepers of the Secret RMS Knowledge." I resolved to build my own

meter using an LT1088 thermal RMS converter.

Full scale on the LT1088 is 4.25V RMS. To measure $50\mu\text{V}$ full-scale, I'd need an amplifier with a gain of 100,000. At 40MHz bandwidth, this didn't sound like it would have a good chance of working first-time—built by hand and without benefit of a custom casting.

Rather than build a circuit with 40MHz bandwidth and a gain of 100dB, I decided to use just enough gain to put my desired noise performance around twice minimum scale. Aside from gain, this amplifier would also need less than $5\text{nV}/\sqrt{\text{Hz}}$ input noise, and the output stage would have to drive the 50 ohm load presented by the LT1088.

It wasn't hard to find an appropriate output stage. The LT1206 (see Figure 1) can easily drive the required 120mA peak current into the LT1088 converter, and there's plenty left over for handling noise spikes. To preserve 40MHz bandwidth, the LT1206 was set to run at a gain of 2.

The front end was harder to solve. I needed a low noise, high speed amplifier that could give me plenty of gain. Here I selected the LT1226. This is a 1GHz GBW op amp with only $2.6\text{nV}/\sqrt{\text{Hz}}$ input noise. It has a minimum stable gain of 25, but in this circuit, high gain is an advantage.

Cascading two LT1226s on the front end gives a gain of 625, a little shy of the 5,000 to 10,000 required. Another gain of 5, plus the gain of 2 in the LT1206 adds up to a gain of 6,250—just about right.

There are several ways to get 40MHz bandwidth at a gain of 5, including the LT1223 and LT1227 current-feedback amplifiers, but I settled on the LT1192 voltage amplifier because it is the lowest cost solution. This brings the gain up to 6250, for a minimum scale sensitivity of $34\mu\text{V}$ RMS, and a full-scale sensitivity of $680\mu\text{V}$ RMS.

My advice-filled coworkers assured me that there was no way I could build a wideband amplifier with a gain of 6250, and make it stable. Nevertheless, I built my amplifier on a

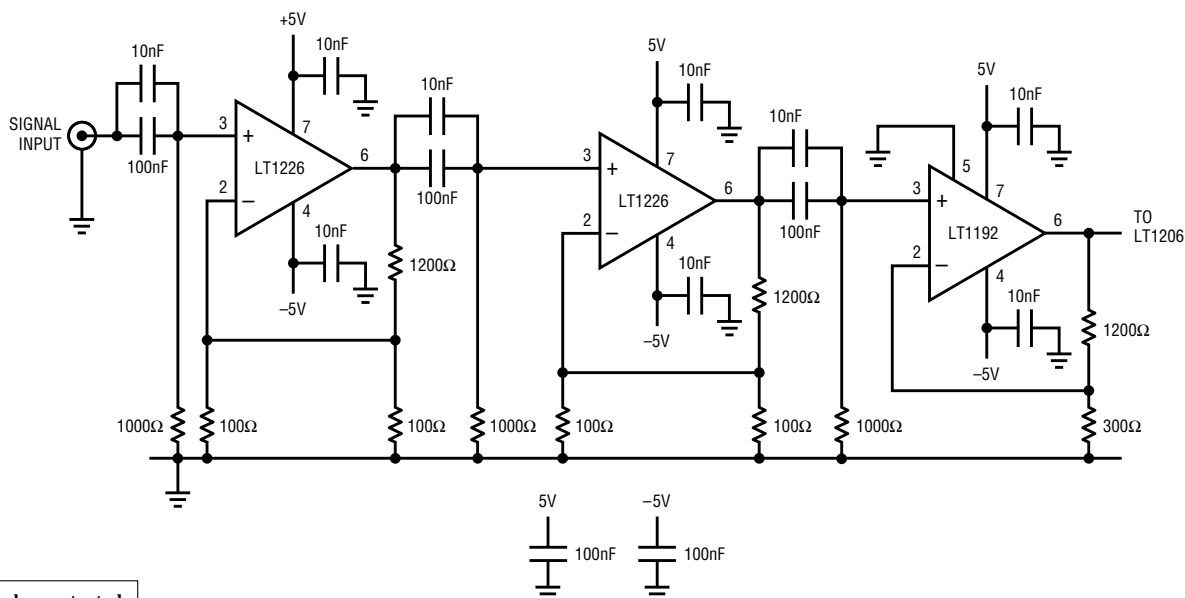


Figure 1. Noise meter gain stage

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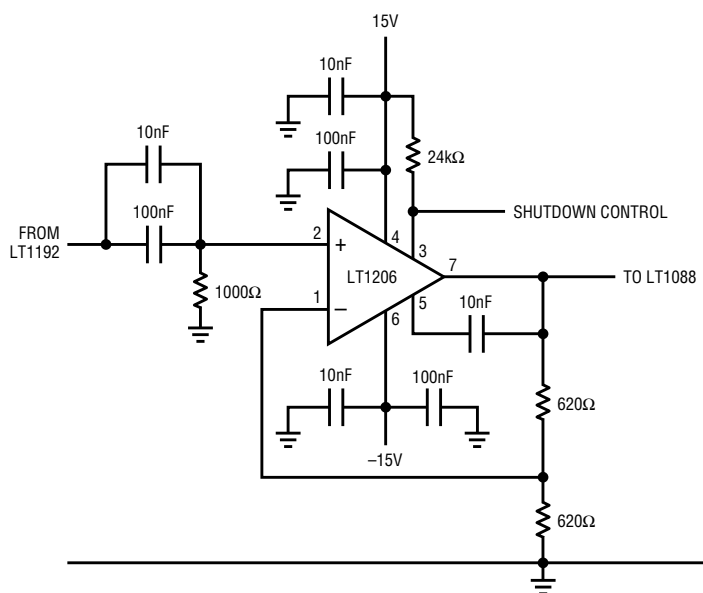


Figure 2. LT1206 buffer/driver section

1.5" × 6" copperclad board, taking care to maintain a linear layout. The finished circuit was stable, provided that a coaxial connection was made to the input. The amplifier was flat, with 3dB points at 4kHz and 43MHz, and some peaking at high frequencies.

The performance of the amplifier and thermal converter can be optimized by adjusting the value of the feedback and gain setting resistors around the LT1206. Slightly more bandwidth can be achieved at the expense of higher peaking by reducing the resistor values 10%. Reducing the resistor values will decrease peaking effects at the expense of bandwidth. A good compromise value is 680Ω.

I've shown the LT1226 amplifiers operating from ±5 supplies, which puts their bandwidth on the edge at

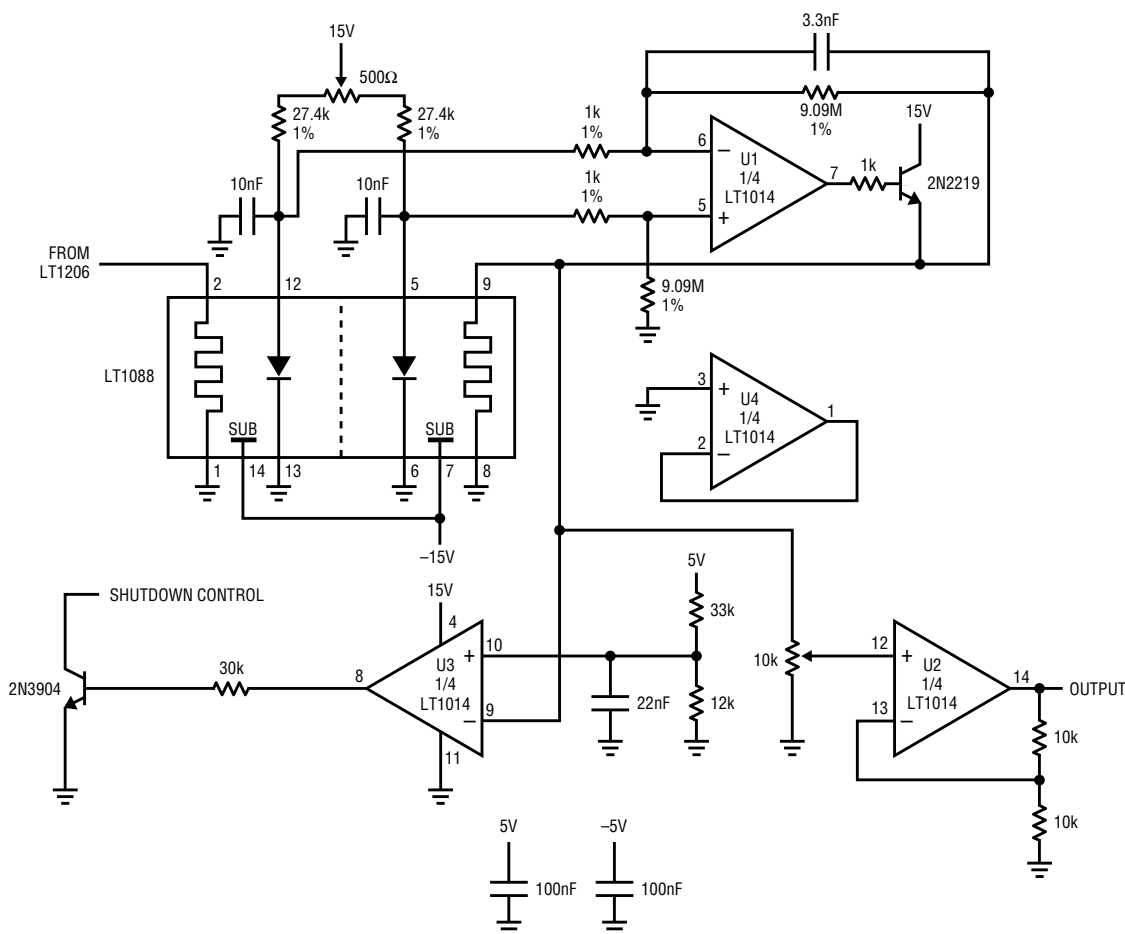


Figure 3. LT1088 RMS detector section

Coaxial Measurements

When measuring low-level signals, it is difficult to get a clean, accurate result. Scope probes have two problems. 10× probes attenuate the already small signal, and both 1× and 10× probes suffer from circuitous grounds. Coaxial adapters are a partial solution, but these are expensive. They make for a lot of wear and tear on the probes, and, without a little forethought, they can be a bear to attach to the circuit under test. My favorite way to get clean measurements of small signals is to directly attach a short length of coaxial cable as shown in Figure 4.

I use the good part of a damaged BNC cable, cutting away the shorter portion to leave at least 18" of RG-58/U and one good connector. At the cut, or as I call it, "real world" end of the cable, I unbraid, twist, and tin a very small amount of outer conductor to form a stub 1/4" to 3/8" long. Next, I cut away the dielectric, exposing a similar length of center conductor, which I also tin. Now the probe is ready for use. It can be soldered directly to a circuit or breadboard, eliminating any lead length that might otherwise pick up stray noise, or worse, act as an antenna in a sensitive, high gain circuit.

Small signals aren't the only beneficiaries of this technique. This works great for looking at ripple on the outputs of switching supplies. Ripple measurements are simplified because the large voltage swings associated with the switch node are completely isolated and no loop is formed where di/dt could inject magnetically coupled noise.

In some instances, I've found it important to retain the 50Ω termination impedance on the cable, but it is rarely possible to place a terminator at the "BNC" end of the cable, since this creates a DC path directly across the circuit under test. There is, however, another way as shown in Figure 5. Here, a technique known as back-termination is used. No termination is used at the far end of the cable, but a 51Ω resistor is connected in series with the measurement end. Signals sent down the cable reach the BNC connector without attenuation, and fast edges that bounce off the unterminated end are absorbed back into the 51Ω source resistor. I've found this especially useful for measuring fast switch signals, or, when measuring the RMS value of small signals, for ensuring that the amplifier input sees a properly terminated source. The resistor trick does not work if the

node under test is high impedance; an FET probe is a better choice for high-impedance measurements.

While I'm at it, I might as well give away my only other secret. We've all encountered ground loop problems, giving rise to 60Hz (50Hz for my friends overseas) injection into sensitive circuits. Every lab is replete with isolation transformers and "controlled substance" line cords with missing ground prongs for battling ground loops. There are similar problems at high frequencies, but the victim is wave fidelity, not AC pick up. To determine whether or not high frequency grounding, ground loops, or common-mode rejection is a problem for your oscilloscope, simply clamp a small ferrite E core around the probe lead while observing any effects on the waveform (see Figure 6). Sometimes the news is bad; the waveform really is messed up and there is some work to be done on the circuit. But occasionally the circuit is exonerated, the unexplainable aberrations disappear, proving that high frequency gremlins are at work. If necessary, several passes of the probe cable can be made through the E-Core, and it can be taped together for as long as needed.



Figure 4. BNC cable used as a "probe"



Figure 5. BNC cable probe back terminated

40MHz. Their bandwidth can be improved by operating at $\pm 15V$.

Because the LT1206 operates on 15V rails, it is possible to overdrive the LT1088 and possibly cause permanent damage. One section of the

LT1014 (U3) is used to sense an over-drive condition on the LT1088 and shut down the LT1206. Sensing the feedback heater instead of the input heater allows the LT1088 to accommodate high-crest-factor waveforms,

shutting down only when the average input exceeds maximum ratings.

By the way, my power supply noise measured 200μV; filtering brought it down to less than 60μV. **LT**

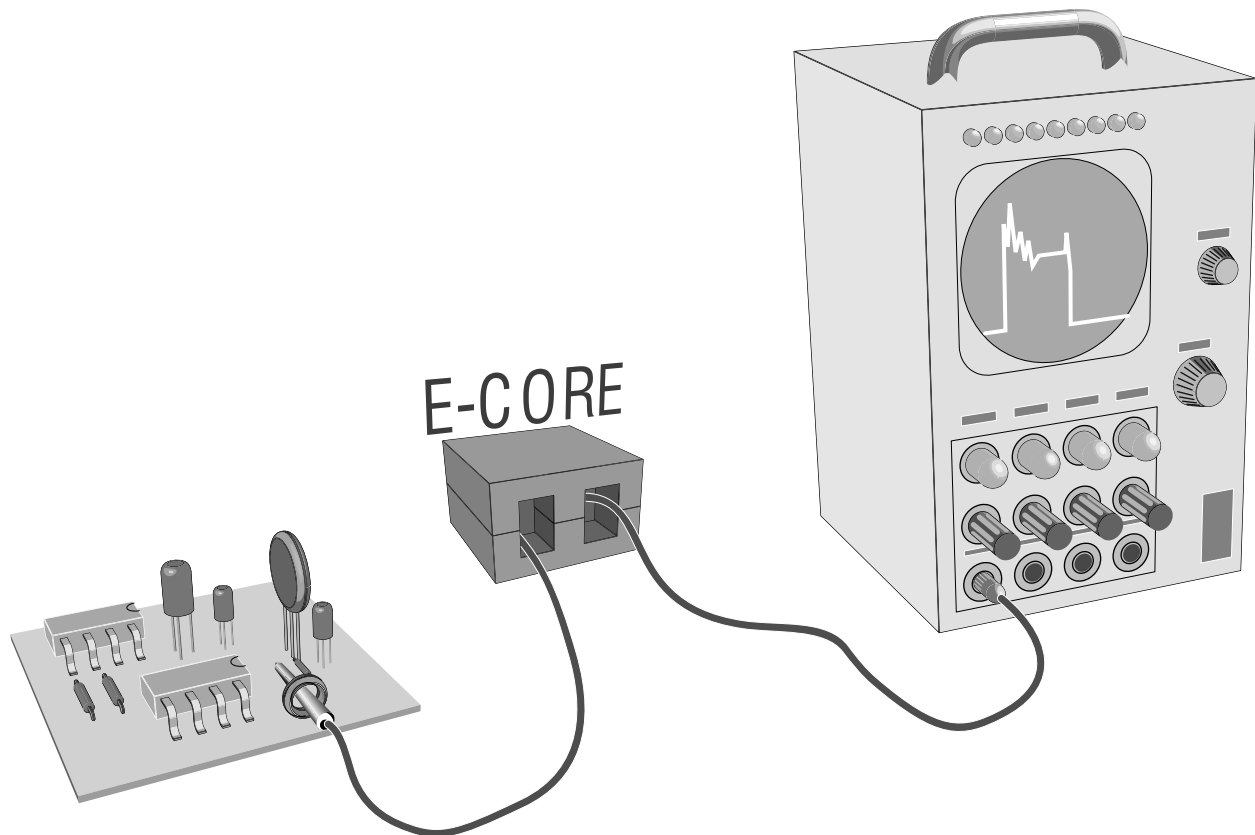


Figure 6. An E-Core serves to choke off high frequency common mode currents from a 'scope probe

12-Bit DACs, continued from page 27

critical. Further, because each of these DACs can be daisy-chained to others, only three opto-isolators are required.

Control Right Where It's Needed

Two major benefits of these DACs are their small SO-8 size and serial interface. These features make it very easy to place an LTC1257 or an LTC1451, LTC1452, or LTC1453 on a PCB exactly where needed, next to its associated circuitry. Unlike the compromises that exist with placing a multi-DAC package and the difficulty in optimizing analog-signal-trace length and routing, the placement flexibility of the LTC1257 and LTC1451, LTC1452, or LTC1453 reduces analog-signal-path lengths to a minimum.

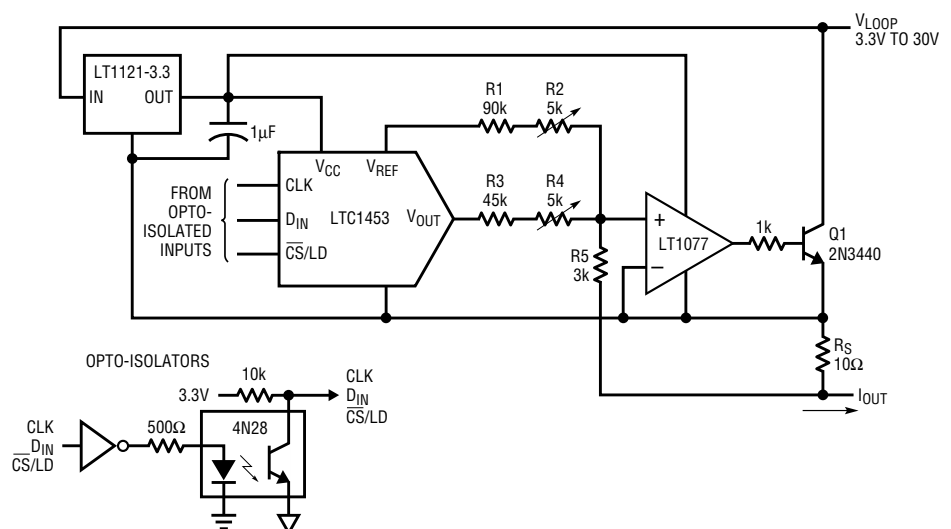



Figure 4. The LTC1453 forms the heart of this isolated 4mA-to-20mA current loop

Conclusion

The LTC1257 and the LTC1451, LTC1452, and LTC1453 serially programmed 12-bit DACs are ideal solutions for applications that

require 12-bit functionality in small SO-8 packages and the ease and economy of a serial interface. 

LTC1477: 70mΩ Protected High-Side Switch Eliminates “Hot Swap” Glitching

by Tim Skovmand

When a printed circuit board is “hot swapped” into a live 5V socket, a number of bad things can happen.

First the instantaneous connection of a large, discharged, supply-bypass capacitor may cause a glitch to appear on the power bus. The current flowing into the capacitor is limited only by the socket resistance, the card-trace resistance, and the equivalent series resistance (ESR) of the supply bypass capacitor. This supply glitch can create real havoc if the other boards in the system have power-on RESET circuitry with thresholds set at 4.65V.

Second, the card itself may be damaged due to the large inrush of current into the card. This current is sometimes inadvertently diverted to sensitive (and expensive) integrated circuits that cannot tolerate either over-voltage or over-current conditions even for short periods of time.


Third, if the card is removed and then reinserted in a few milliseconds, the glitching of the supply may “confuse” the microprocessor or peripheral ICs on the card, generating erroneous data in memory or forcing the card into an inappropriate state.

Fourth, a card may be shorted, and insertion may either grossly glitch the 5V supply or cause severe physical damage to the card.

Figure 1 is a schematic diagram showing how an LTC1477 protected high-side switch and an LTC699 power-on RESET circuit reduce the chance of glitching or damaging the socket or card during “hot swapping.”

The LTC1477 protected high-side switch provides extremely low $R_{DS(ON)}$ switching (typically 0.07Ω) with built-in 2A current limiting and thermal shutdown, all in an 8-pin SO package.

As the card is inserted, the LTC699 power-on RESET circuit holds the enable pin of the LTC1477 low for approximately 200ms. When the enable pin is asserted high, the output is ramped on in approximately 1ms. Even if a very large supply bypass capacitor (for example, over 100μF) is used, the LTC1477 will limit the inrush current to 2A and ramp the capacitor at an even slower rate. Further, the board is protected against short-circuit conditions by limiting the switch current to 2A.

The 5V card supply can be disabled via Q1. The only current flowing is the standby quiescent current of the LTC1477, which drops below 1μA, the 600μA quiescent current of the LTC699, and the 10μA consumed by R1. 

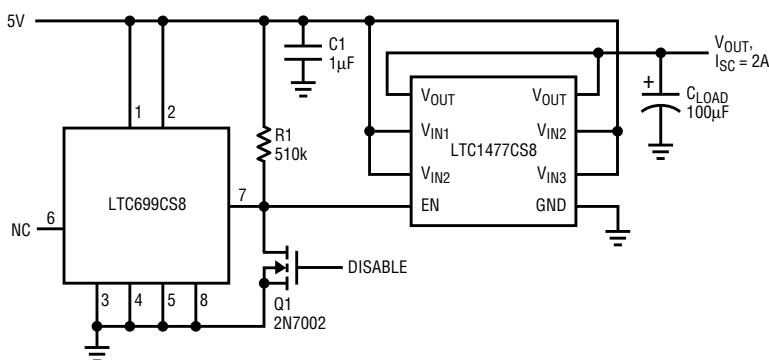


Figure 1. “Hot swap” circuit featuring LTC1477 and LTC699

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Driving a High-Level Diode-Ring Mixer with an Operational Amplifier

by Mitchell Lee

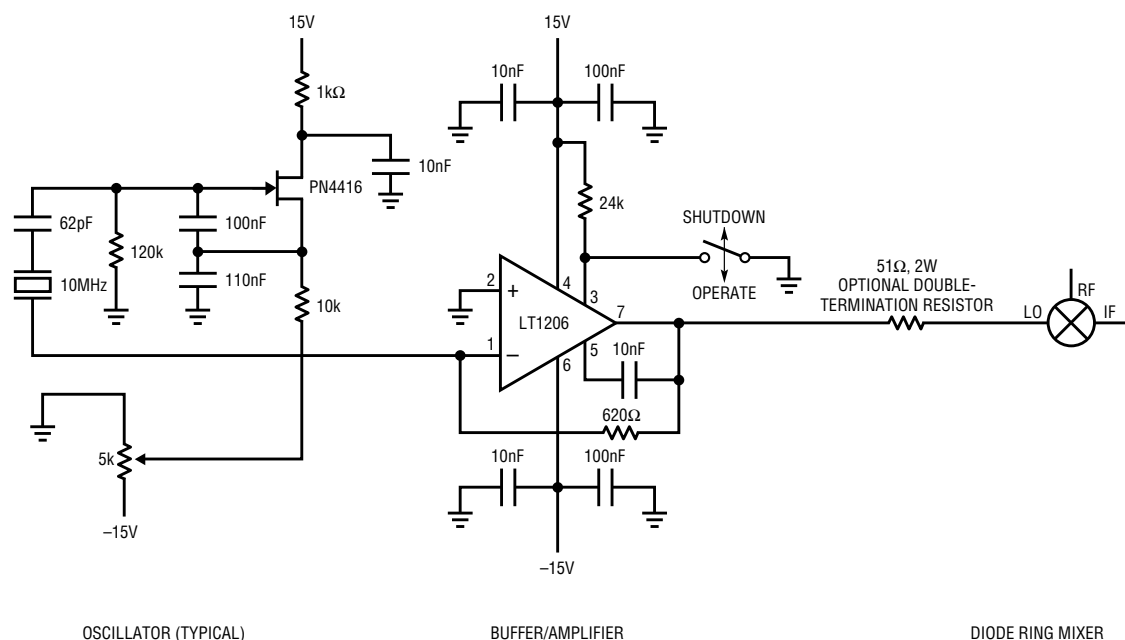


Figure 1. Oscillator buffer drives +17 to +27dBm double balanced mixers

One of the most popular RF building blocks is the diode-ring mixer. Consisting of a diode ring and two coupling transformers, this simple device is a favorite with RF designers anywhere a quick multiplication is required, as in frequency conversion, frequency synthesis, or phase detection. In many applications these mixers are driven from an oscillator. Rarely does anyone try building an oscillator capable of delivering +7dBm for a "minimum geometry" mixer, let alone one of higher level. One or more stages of amplification are added to achieve the drive level required by the mixer. The new LT1206 high-speed amplifier makes it possible to amplify an oscillator to +27dBm in one stage.

Figure 1 shows the complete circuit diagram for a crystal oscillator, LT1206 op amp/buffer, and diode-ring mixer. Most of the components

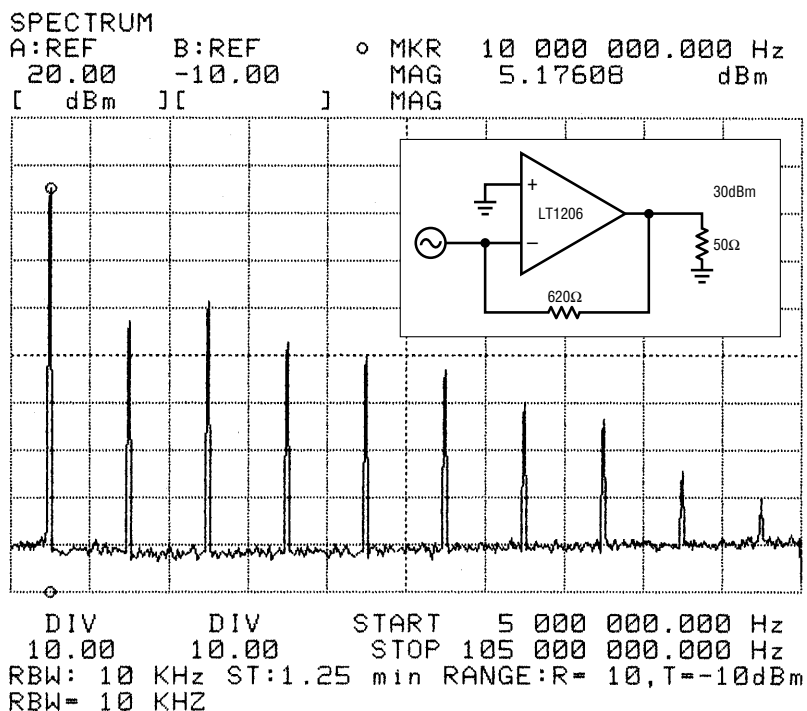


Figure 2. Spectrum plot of Figure 1's circuit driving +30dBm into a 50Ω load (single termination)

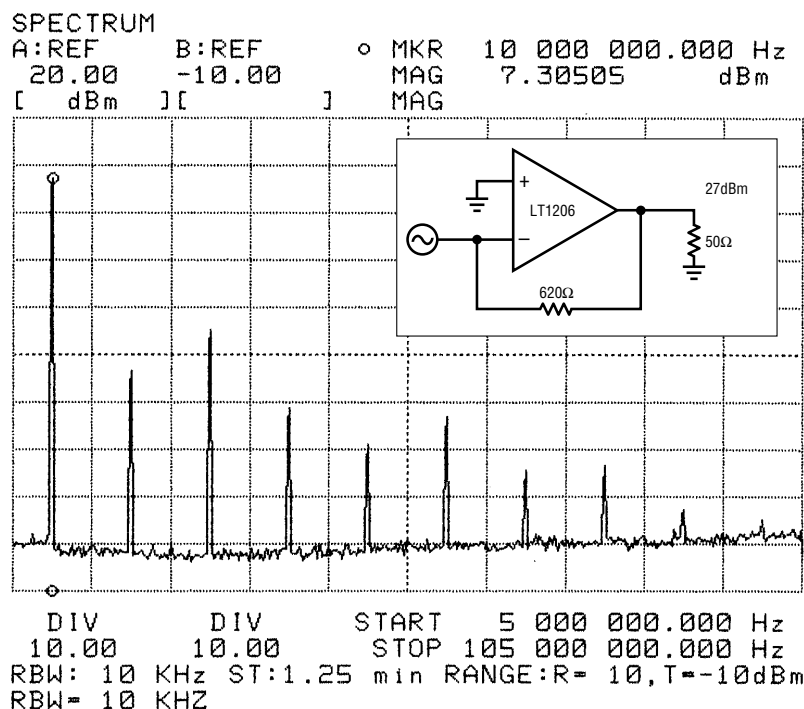


Figure 3. Spectrum plot of Figure 1's circuit driving +27dBm into a 50Ω load (single termination)

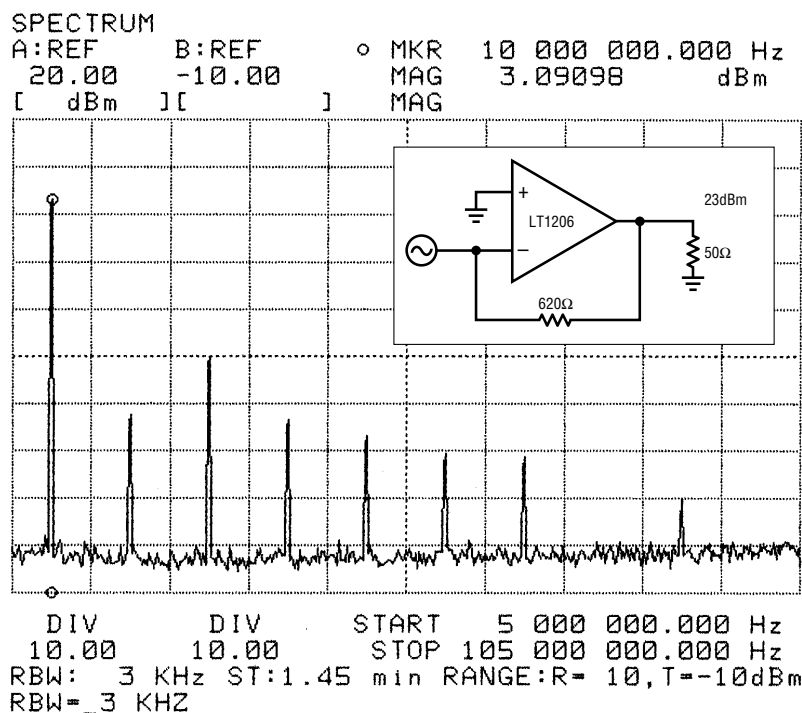


Figure 4. Spectrum plot of Figure 1's circuit driving +23dBm into a 50Ω load (single termination)

are used in the oscillator itself, which is of the Colpitts class. Borrowing from a technique I first discovered in Hewlett Packard's Unit Oscillator, the current of the crystal is amplified, rather than the voltage. There are several advantages to this method, the most important of which is low distortion. Although the voltages present in this circuit have poor wave shape and are sensitive to loading, the crystal current represents essentially a filtered version of the voltage waveform and is relatively tolerant of loading effects.

One of the most popular RF building blocks is the diode-ring mixer. Consisting of a diode ring and two coupling transformers, this simple device is a favorite with RF designers anywhere a quick multiplication is required, as in frequency conversion, frequency synthesis, or phase detection


The impedance, and therefore the voltage at the bottom of the crystal is kept low by injecting the current into the summing node of an LT1206 current-feedback amplifier. Loop gain reduces the input impedance to well under one ohm. Oscillator bias is adjustable, allowing control of the mixer drive. This also provides a convenient point for closing an output power servo loop.

Operating from $\pm 15\text{V}$ supplies, the LT1206 can deliver +32dBm to a 50Ω load, and, with a little extra headroom (the absolute maximum supply voltage is ± 18), it can reach 2W output power into 50Ω . Peak guaranteed output current is 250mA.

Shown in Figures 2–6 are spectral plots for various combinations of single and double termination at power levels ranging from +17 to +27dBm—not bad for an inductorless circuit. Double termination may be used to present a 50Ω source impedance to the mixer, or to isolate two or more mixers driven simultaneously from one LT1206 amplifier.

Although a 10MHz example has been presented here, the LT1206's 65MHz bandwidth makes it useful in circuits up to 30MHz. In addition, the shutdown feature can be used to gate drive to the mixer. When the LT1206 is shut down, the oscillator will likely stop, since the crystal then sees a series impedance of 620Ω and the mixer itself. Upon re-enabling the LT1206 there will be some time delay before the oscillator returns to full power. The circuit works equally well with an LC version of the oscillator.

Note that the current feedback topology is inherently tolerant of stray capacitive effects at the summing node, making it ideal for this application. Another nice feature is the LT1206's ability to drive heavy capacitive loads while remaining stable and free of spurious oscillations.

For mixers below +17dBm, the LT1227 is a lower cost alternative, featuring 140MHz bandwidth in combination with the shutdown feature of the LT1206. 

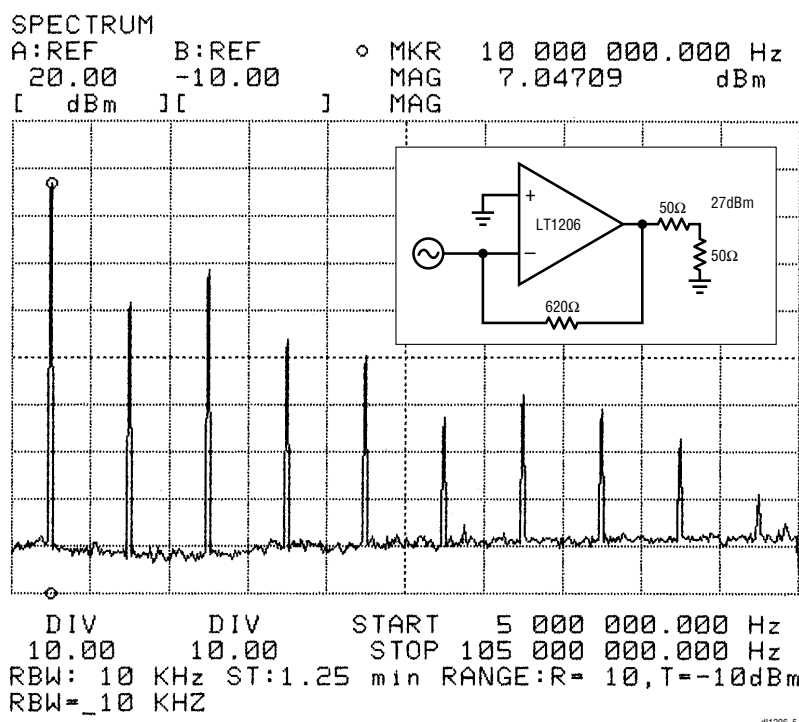


Figure 5. Spectrum plot of Figure 1's circuit driving +27dBm into a 100Ω load (double termination)

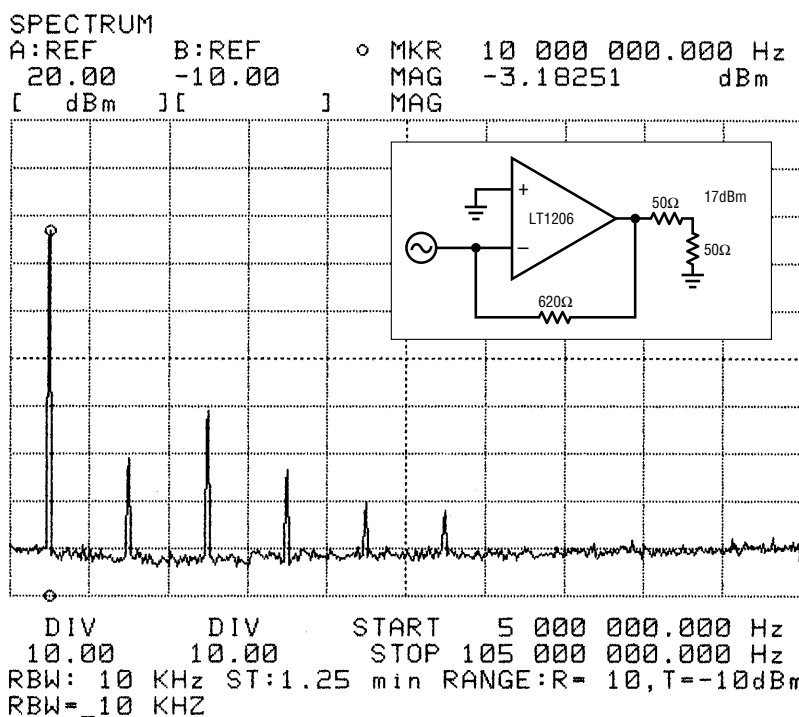


Figure 6. Spectrum plot of Figure 1's circuit driving +17dBm into a 100Ω load (double termination)

Fully Differential, 8-Channel, 12-Bit A/D System Using the LTC1390 and LTC1410

by Kevin R. Hoskins

The LTC1410's fast 1.25MSPS conversion rate and differential $\pm 2.5V$ input range make it ideal for applications that require multichannel acquisition of fast, wide-bandwidth signals. These applications include multitransducer vibration analysis, race-vehicle telemetry data acquisi-

tion, and multichannel telecommunications. The LTC1410 can be combined with the LTC1390 8-channel serial-interfaced analog multiplexer to create a differential A/D system with conversion throughput rates up to 625kSPS. This rate applies to situations where the

selected channel changes with each conversion. The conversion rate increases to 1.25MSPS if the same channel is used for consecutive conversions.

Figure 1 shows the complete differential, 8-channel A/D circuit. Two LTC1390s, U1 and U2, are used

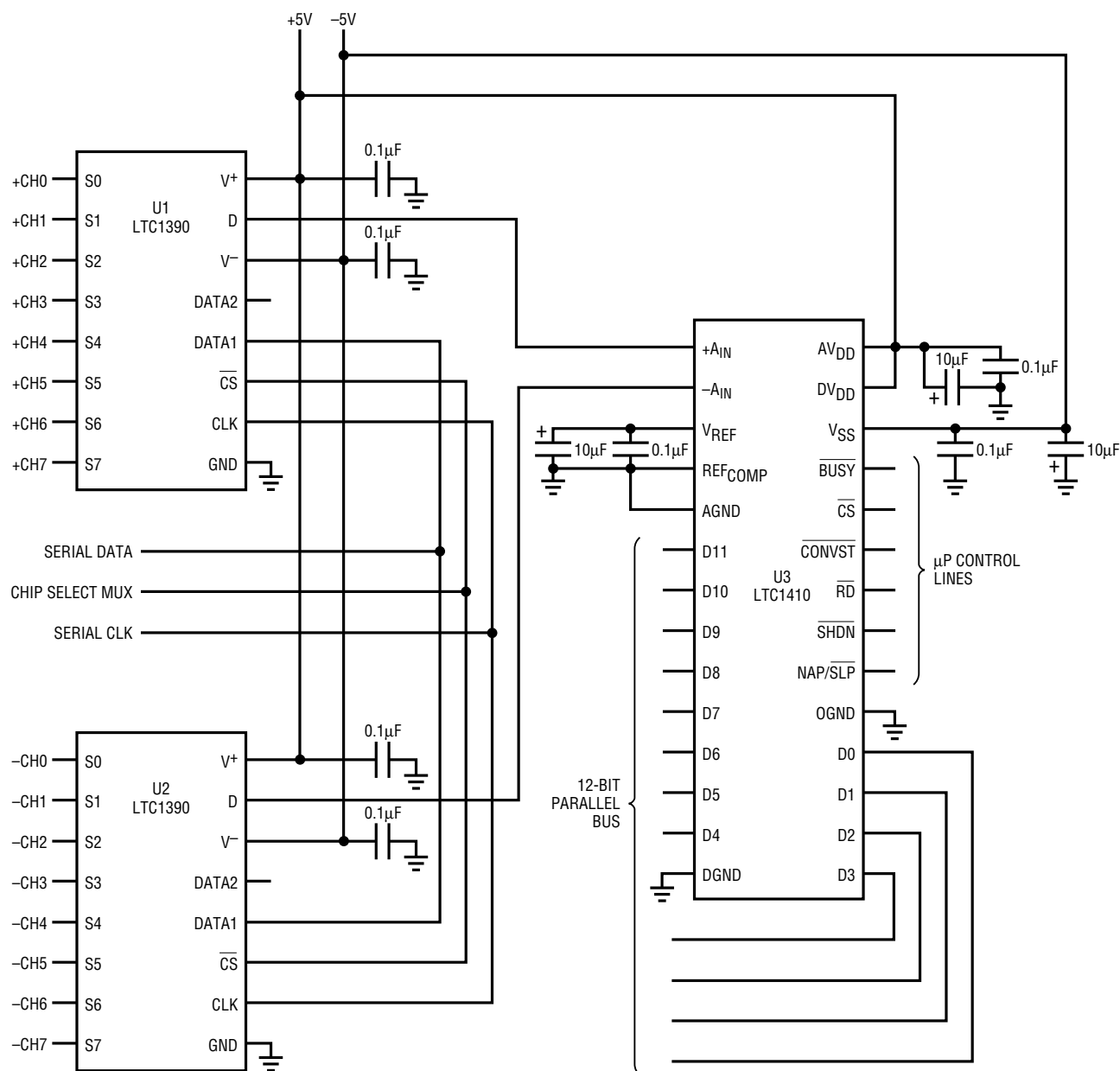


Figure 1. Fully differential 8-channel data acquisition system achieves 625kSPS throughput

as noninverting and inverting input multiplexers. The outputs of the non-inverting and inverting multiplexers are applied to the LTC1410's $+A_{IN}$ and $-A_{IN}$ inputs, respectively. The LTC1390 share the CHIP SELECT MUX, SERIAL DATA and SERIAL CLOCK control signals. This arrangement simultaneously selects the same channel on each multiplexer: S0 for both $+CH0$ and $-CH0$, S1 for both $+CH1$ and $-CH1$, and so on.

As shown in the timing diagram (Figure 2), MUX channel selection and A/D conversion are pipelined to maximize the converter's throughput. The conversion process begins with selecting the desired multiplexer channel-pair. With a logic high applied to the LTC1390's \overline{CS} input, the channel-pair data is clocked into each DATA 1 input on the rising edge of the 5MHz clock signal. CHIP SELECT MUX is then pulled low, latching the

channel-pair selection data. The signals on the selected MUX inputs are then applied to the LTC1410's differential inputs. CHIP SELECT MUX is pulled low 700ns before the LTC1410's conversion-start input, \overline{CONVST} , is pulled low. This accounts for the maximum time needed by the LTC1390's MUX switches to fully turn-on. This ensures that the input signals are fully settled before the LTC1410's S/H captures its sample.

The LTC1410's S/H acquires the input signal and begins conversion on \overline{CONVST} 's falling edge. During the conversion, the LTC1390's \overline{CS} input is pulled high and the data for the next channel-pair is clocked into DATA 1. This pipelined operation continues until a conversion sequence is completed. When a new channel-pair is selected for each conversion, the sampling rate of each channel is 78ksps, allowing an input-signal

bandwidth of 39kHz for each channel of the LTC1390/LTC1410 system.

To maximize the throughput rate, the LTC1410's \overline{CS} input is pulled low at the beginning of a series of conversions. The LTC1410's data output drivers are controlled by the signal applied to \overline{RD} . The conversion's results are available 20ns before the rising edge of \overline{BUSY} . The rising edge of the \overline{BUSY} output signal can be used to notify a processor that a conversion has ended and data is ready to read.

This circuit takes advantage of the LTC1410's very high 1.25Msps conversion rate and differential inputs and the LTC1390's ease of programming to create an A/D system that maintains wide input-signal bandwidth while sampling multiple input signals. **LT**

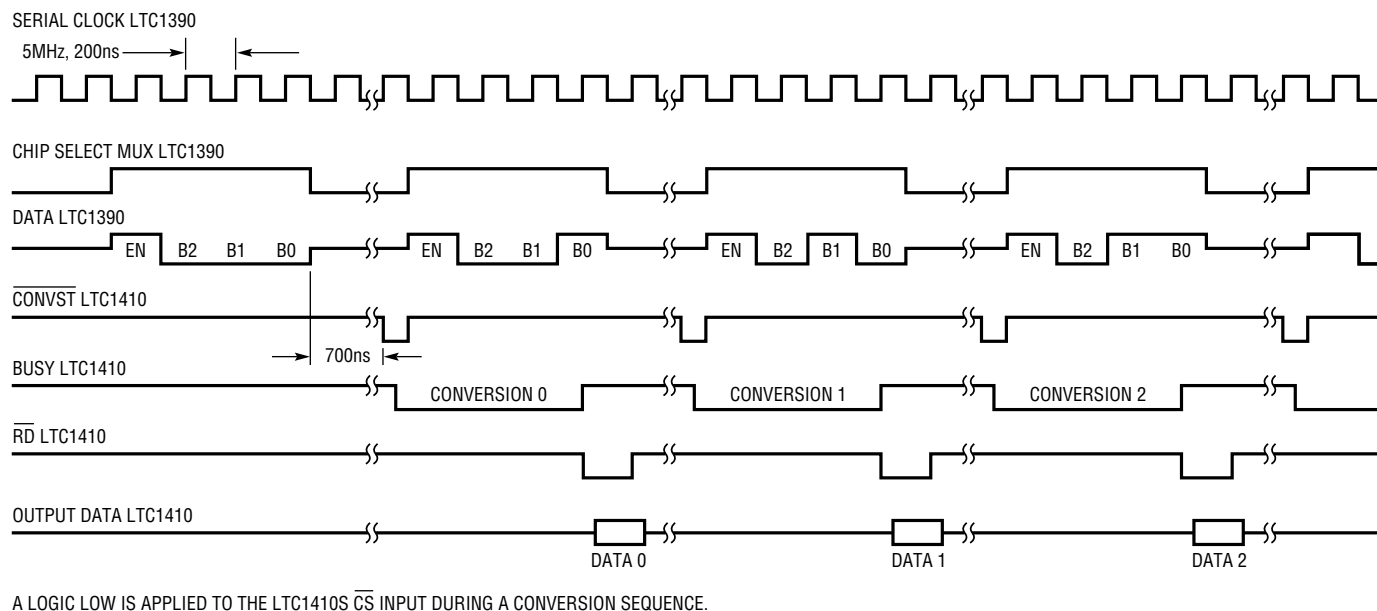


Figure 2. Timing diagram of circuit in Figure 1

New Device Cameos

LTC1266/LTC1266-3.3/ LTC1266-5 Synchronous Controllers Feature Precision Output Regulation

The LTC1266, LTC1266-3.3, and LTC1266-5 synchronous controllers for N- or P-channel MOSFETs provide a precision 1% internal reference for tighter regulation of output voltage. This precision reference, combined with the LTC1266's tight load regulation and ability to drive two N-channel MOSFETs for high currents, provides the accuracy and current capability necessary to meet the stringent demands of many Pentium® microprocessor applications.

The LTC1266 is a synchronous, current mode, switching regulator controller with automatic Burst Mode operation (Burst Mode operation can be disabled), available in the 16-lead narrow SO package. Two user-programmable pins make it possible to drive an N-channel MOSFET for high current or step-up operation, or a P-channel MOSFET for low dropout operation. With an N-channel top-side MOSFET for higher efficiency at high currents and Burst Mode operation enabled for high efficiencies at low currents, the LTC1266 can provide 90% or better efficiency for loads from 10mA to 10A.

Other features of the LTC1266 include a shutdown mode, which reduces supply current to 40µA, and an on-chip comparator for use as a "power-good" monitor, which stays active in shutdown. The maximum operating frequency of 400kHz allows the use of small, surface mount inductors and capacitors.

LT1368 and LT1369 Precision Rail-to-Rail Op Amps with Excellent High- Frequency Supply Rejection

The LT1368 dual and LT1369 quad are the latest members of the LT1366 rail-to-rail op amp family. Designed

to operate with a 0.1µF output-compensation capacitor, these devices feature improved high frequency supply rejection and lower high frequency output impedance. The output capacitance acts as a filter, reducing noise pickup and ripple from the supply. This additional filtering is helpful in mixed analog/digital systems with common supplies, or in systems employing switching supplies.

Like the LT1366, the LT1368 and LT1369 combine rail-to-rail input and output operation with precision specifications. These devices maintain their characteristics over a supply range of 1.8V to 36V. Operation is specified with +3V, +5V, and ±15V supplies. The input offset voltage is typically 150µV, with a minimum open loop gain (A_{VOL}) of 1 million while driving a 2k load. The typical input-bias and offset currents are 10nA and 1nA respectively. The supply current per amplifier is typically 375µA.

Common mode rejection is typically 90dB over the full rail-to-rail input range, and supply rejection is 110dB (DC). The filter formed with the amplifier's output impedance and the external compensating capacitor provides approximately 40dB of supply rejection above 30kHz.

The LT1368 dual is available in plastic 8-pin DIP and 8-lead SO packages. The LT1369 quad is available in a plastic 16-lead SO package.

LT1186 DAC-Programmable CCFL Switching Regulator (Bits-to-Nits™)

The LT1186 is a fixed frequency, current mode switching regulator that provides the control function for cold-cathode fluorescent lighting. The LT1186 supports grounded-lamp and floating-lamp configurations. The IC includes an efficient, high-current switch, an oscillator, output drive logic, control circuitry, protection circuitry, and a micropower 8-bit, 50µA, full-scale current output DAC.

The DAC provides simple "bits-to-lamp-current control," and communicates externally in two interface modes: standard SPI mode and pulse mode. On power-up, the DAC counter is reset to half-scale and the chip is configured to SPI or pulse mode, depending on the CS signal level. In SPI mode, the system microprocessor serially transfers the present 8-bit data and reads back the previous 8-bit data from the DAC. In pulse mode, the upper six bits of the DAC implement increment-only (1-wire interface) or increment/decrement (2-wire interface) operation. The operational mode (increment-only or increment/decrement) is selected by the DIN signal level.

The LT1186 operates from a logic supply voltage of 3.3V or 5V. The IC also has a battery supply voltage pin that operates from 4.5V to 30V. The LT1186 draws 6mA typical quiescent current. An active low shutdown pin typically reduces total supply current to 35µA for standby operation while the DAC retains its last setting. A 200kHz switching frequency minimizes the size of required magnetic components and the use of current-mode switching techniques gives high reliability and simple loop-frequency compensation. The LT1186 is available in a 16-pin narrow body SO.

LTC1477/LTC1478: Single and Dual 70mΩ, 2A, Protected High-Side Switches in SO Packaging

The LTC1477/LTC1478 single and dual protected high-side switches provide extremely low $R_{DS(ON)}$ switching with built in protection against short-circuit and thermal-overload conditions. The LTC1477 single is available in 8-lead SO packaging and the LTC1478 dual is available in 16-lead SO packaging. Both parts operate from 2.7V to 5.5V.

The LTC1477/LTC1478 provide two levels of protection. The first level

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Pentium is a registered trademark of Intel Corporation.

is short-circuit current limiting, which is set at 2A. The short-circuit current can be reduced to as low as 0.85A by disconnecting portions of the power device. The second level of protection is thermal-overload protection, which limits the die temperature to approximately 130°C. A built in charge pump produces gate drive from the 2.7V-to-5.5V power supply input for controlled ramping of the internal ultra-low $R_{DS(ON)}$ NMOS switch. No higher gate voltage power supply is required. Rise time is set internally at 1ms, to reduce inrush currents to negligible levels; therefore, extremely large load capacitors can be ramped without glitching the system power supply. The switch resistance is typically 0.07Ω at 5V and only rises to 0.08Ω at 3.3V. Output leakage current drops to 0.01μA when the output is switched off.

The NMOS switch has no parasitic body diode; therefore, no current flows through the switch when it is turned off and the output is forced above the input supply voltage. (DMOS switches have parasitic body diodes that become forward biased under these conditions.)

All operating modes are micropower and the quiescent current automatically drops below 1μA when the switch is turned off.

The combination of micropower operation and full protection make these devices ideal for battery powered applications such as palmtop computers, notebook computers, personal digital assistants, instruments, handi-terminals, and bar-code readers.

LTC1550/LTC1551: Low-Noise, Switched-Capacitor, Regulated Voltage Inverters

The LTC1550 and LTC1551 are switched-capacitor voltage inverters with onboard linear post regulators, designed to generate a low noise, regulated negative supply for use as bias voltage for GaAs transmitters FETs in portable-RF and cellular-telephone applications. They operate from a single 4.5V to 7V supply and provide a fixed -4.1V or adjustable output

voltage, with low output ripple (1mV typical) and a typical quiescent current of 5mA at $V_{CC} = 5V$. An on-chip oscillator sets the charge pump frequency at 900kHz, away from sensitive 400kHz-600kHz IF bands. The LTC1550-4.1 and the LTC1551-4.1 provide fixed -4.1V outputs, whereas the LTC1550 offers adjustable output voltage. Both devices include a TTL-compatible shutdown pin that drops supply current to 0.2μA typical. The LTC1550 shutdown pin is active low (SD), and the LTC1551 shutdown pin is active high (SD).

Only three external components are required for the fixed-voltage parts: two 0.1μF charge pump capacitors and a 10μF filter capacitor at the linear regulator output. Each version of the LTC1550/LTC1551 will supply up to 20mA output current with guaranteed output regulation of ±5%. All except the 8-pin versions of the LTC1550/LTC1551 include an open-drain REG output, which pulls low to indicate that the output is within 5% of the set value. This allows external circuitry to monitor the output without additional components. The adjustable LTC1550 requires two additional resistors to set the output voltage.

The LTC1550-4.1 and LTC1551-4.1 are available in 8-pin SO, 14-pin SO, and 16-pin SSOP packages. The LTC1550 adjustable-output voltage version is available in 14-pin SO and 16-pin SSOP packages.

Low-Cost, High-Accuracy, 12-Bit Multiplying DAC Family

Linear Technology has introduced the LTC8043, LTC8143, LTC7543, and LTC7541A, four 12-bit, current output, 4-quadrant multiplying digital-to-analog converters (DACs). They are superior pin-compatible replacements for the industry standard DAC-8043, DAC-8143, AD7543, and AD7541A. Low cost, tight accuracy specs, low power, small size, and outstanding flexibility make these DACs excel in process control, remote or isolated applications, and software-programmable gain, attenuation, and filtering. Linear Technology's parts

have improved accuracy and stability, tighter timing specs, reduced sensitivity to external amplifier V_{OS} , lower output capacitance, and reduced cost.

The LTC8043 comes in a small SO-8 or PDIP package and has an easy-to-use serial interface. The LTC8143 and LTC7543 come in 16-Pin SO or PDIP packages and feature a more flexible serial interface that includes an asynchronous Clear pin. The LTC8143 also has a serial data output so users can daisy-chain multiple DACs on a 3-wire bus. The LTC7541A has a 12-bit wide parallel interface and comes in 18-pin SO and PDIP packages.

These DACs boast unbeatable accuracy and stability, with INL and DNL of 1/2LSB MAX, 1/8LSB TYP over temperature. Gain Error is 1LSB MAX, eliminating adjustments in most applications. Gain and nonlinearity temperature coefficients are typically better than 1 ppm/°C and 0.1 ppm/°C respectively.

All four parts are offered in the extended industrial temperature range and the LTC7541A is also available in commercial grades. **LT**

For further information on the above or any of the other 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.

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Information furnished by Linear Technology Corporation is believed to be accurate and reliable. However, Linear Technology makes no representation that the circuits described herein will not infringe on existing patent rights.

DESIGN TOOLS

Applications on Disk

NOISE DISK

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

SPICE MACROMODEL DISK

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

Technical Books

1990 Linear Databook, Volume I — This 1440 page collection of data sheets covers op amps, voltage regulators, references, comparators, filters, PWMs, data conversion and interface products (bipolar and CMOS), in both commercial and military grades. The catalog features well over 300 devices. \$10.00

1992 Linear Databook Supplement — This 1248 page supplement to the 1990 Linear Databook is a collection of all products introduced since then. The catalog contains full data sheets for over 140 devices. The 1992 Linear Databook Supplement is a companion to the 1990 Linear Databook, which should not be discarded. \$10.00

1994 Linear Databook, Volume III — This 1826 page supplement to the 1990 Linear Databook and 1992 Linear Databook Supplement is a collection of all products introduced since 1992. A total of 152 product data sheets are included with updated selection guides. The 1994 Linear Databook Volume III is a supplement to the 1990 and 1992 Databooks, which should not be discarded. \$10.00

Linear Applications Handbook • Volume I — 928 pages full of application ideas covered in depth by 40 Application Notes and 33 Design Notes. This catalog covers a broad range of "real world" linear circuitry. In addition to detailed, systems-oriented circuits, this handbook contains broad tutorial content together with liberal use of schematics and scope photography. A special feature in this edition includes a 22 page section on SPICE macromodels. \$20.00

1993 Linear Applications Handbook • Volume II — Continues the stream of "real world" linear circuitry initiated by the 1990 Handbook. Similar in scope to the 1990 edition, the new book covers Application Notes 41 through 54 and Design Notes 33 through 69. Additionally, references and articles from non-LTC publications that we have found useful are also included. \$20.00

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