Linear Technology Magazine Circuit Collection, Volume II

Power Products

Richard Markell, Editor

INTRODUCTION

Application Note 66 is a compendium of "power circuits" from the first five years of *Linear Technology*. The objective is to collect the useful circuits from the magazine into several applications notes (another, AN67, will collect signal processing circuits into one Application Note) so that valuable "gems" will not be lost. This Application Note contains circuits that can power most any system you can imagine, from desktop computer systems to micropower systems for portable and handheld equipment. Also included here are circuits that provide 300W or more of power factor corrected DC from a universal input. Battery chargers are included, some that charge several battery types, some that are optimized to charge a single type. MOSFET drivers, high side switches and H-bridge driver circuits are also included, as is an article on simple thermal analysis. With these introductory remarks, I'll stand aside and let the authors describe their circuits.

ARTICLE INDEX

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Regulators—Switching (Buck)

High Power (>4A)

BIG POWER FOR BIG PROCESSORS: THE LTC1430 SYNCHRONOUS REGULATOR

by Dave Dwelley

The LTC1430 is a new switching regulator controller designed to be configured as a synchronous buck converter with a minimum of external components. It runs at a fixed switching frequency (nominally 200kHz) and provides all timing and control functions, adjustable current limit and soft start, and level shifted output drivers designed to drive an all N-channel synchronous buck converter architecture. The switch driver outputs are capable of driving multiple paralleled power MOSFETs with submicrosecond slew rates, providing high efficiency at very high current levels while eliminating the need for a heat sink in most designs. The LTC1430 is usable in converter designs providing from a few amps to over 50A of output current, allowing it to supply 3.3V power to the most current-hungry arrays of microprocessors.

A Typical 5V to 3.3V Application

The typical application for the LTC1430 is a 5V to 3.xV converter on a PC motherboard. The output is used to power a Pentium® processor, Pentium® Pro processor or similar class processor and the input is taken from the system 5V ±5% supply. The LTC1430 provides the precisely regulated output voltage required by the processor without the need for an external precision reference or trimming. Figure 1 shows a typical application with a 3.30V ±1% output voltage and a 12A output current limit. The power MOSFETs are sized so as not to require a heat sink under ambient temperature conditions up to 50°C. Typical efficiency is above 91% from 1A to 10A output current and peaks at 95% at 5A (Figure 2).

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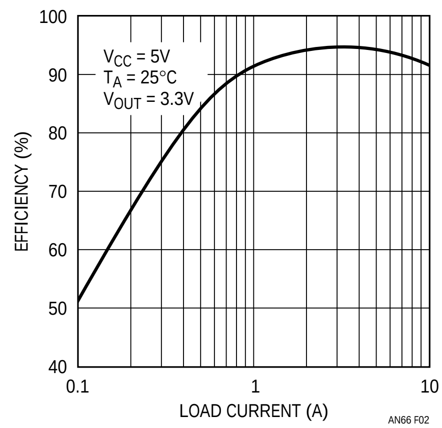


Figure 2. Efficiency Plot for Figure 1's Circuit. Note That Efficiency Peaks at a Respectable 95%

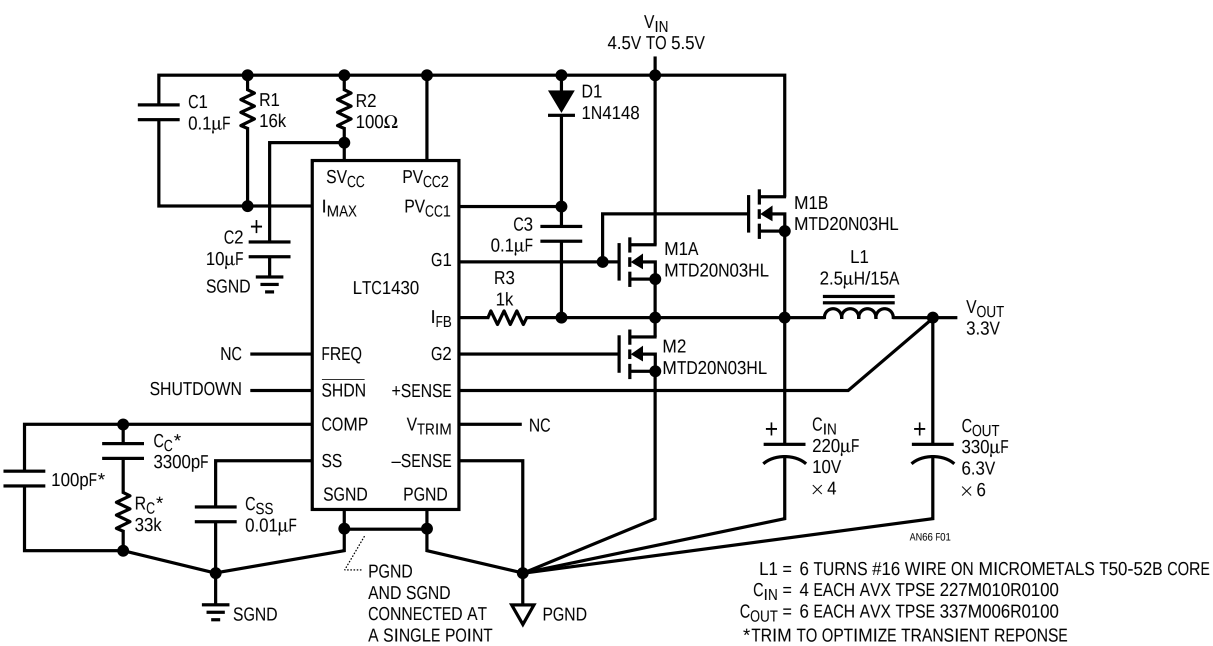


Figure 1. Typical 5V to 3.3V, 10A LTC1430 Application

The 12A current limit is set by the 16k resistor R1 from PV_{CC} to I_{MAX} and the 0.035Ω ON resistance of the MTD20N03HL MOSFETs (M1A, M1B).

The 0.1µF capacitor in parallel with R1 improves power supply rejection at I_{MAX}, providing consistent current limit performance when voltage spikes are present at PV_{CC}. Soft start time is set by C_{SS}; the 0.01µF value shown reacts with an internal 10µA pull-up to provide a 3ms start-up time. The 2.5µH, 15A inductor is sized to allow the peak current to rise to the full current limit value without saturating. This allows the circuit to withstand extended output short circuits without saturating the inductor core. The inductor value is chosen as a compromise between peak ripple current and output current slew rate, which affects large-signal transient response. If the output load is expected to generate large output current transients (as large microprocessors tend to do), the inductor value will need to be quite low, in the 1µH to 10µH range.

Loop compensation is critical for obtaining optimum transient response with a voltage feedback system like the LTC1430; the compensation components shown here give good response when used with the output capacitor values and brands shown (Figure 3). The ESR of the output capacitor has a significant effect on the transient response of the system. For best results use the largest value, lowest ESR capacitors that will fit the design budget and space requirements. Several smaller capacitors wired in parallel can help reduce total output capacitor ESR to acceptable levels. Input bypass capacitor ESR is also important to keep input supply variations to a minimum with $10A_{P-P}$ square wave current pulses flowing into M1. AVX TPS series surface mount tantalum capacitors and Sanyo OS-CON organic electrolytic capacitors are recommended for both input and output bypass duty. Low cost "computer grade" aluminum electrolytics typically have much higher series resistance and will significantly degrade performance. Don't count on that parallel 0.1µF ceramic cap to lower the ESR of a cheap electrolytic cap to acceptable levels.

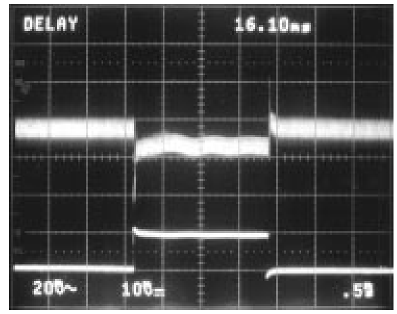


Figure 3. Transient Response: 0A to 5A Load Step Imposed on Figure 1's Output

APPLICATIONS FOR THE LTC1266 SWITCHING REGULATOR

by Greg Dittmer

Figures 4, 5 and 6 show the three basic circuit configurations for the LTC1266. The all N-channel circuit shown in Figure 4 is a 3.3V/5A surface mount converter with the internal MOSFET drivers powered from a separate supply, PWR V_{IN}. The $V_{GS(ON)}$ of the Si9410 N-channel MOSFETs is 4.5V; thus the minimum allowable voltage for PWR V_{IN} is $V_{IN(MAX)}$ + 4.5V. At the other end, PWR V_{IN} should be kept under the maximum safe level of 18V, limiting V_{IN} to 18V – 4.5V = 13.5V. The current sense resistor value is chosen to set the maximum current to 5A according to the formula $I_{OUT} = 100mV/R_{SENSE}$. With V_{IN} = 5V, the 5µH inductor and 130pF timing capacitor provide an operating frequency of 175kHz and a ripple current of 1.25A.

Figure 5 shows an LTC1266 in the charge pump configuration designed to provide a 3.3V/10A output from a single supply. The Si4410s are new logic level, surface mount, N-channel MOSFETs from Siliconix that provide a mere 0.02Ω of on-resistance at V_{GS} = 4.5V and thus provide a 10A solution with minimal components. The efficiency plot shows that the converter is still close to 90% efficient at 10A. Because the charge pump configuration is used, the maximum allowable V_{IN} is 18V/2 = 9V. Due to the high AC currents in this circuit we recommend low ESR OS-CON or AVX input/output capacitors to maintain efficiency and stability.

Figure 6 shows the conventional P-channel topside switch circuit configuration for implementing a 3.3V/3A regulator. The P-channel configuration allows the widest possible supply range of the three basic circuit configurations, 3.5V to 18V, and provides extremely low dropout, exceeding that of most linear regulators. The low dropout results from the LTC1266's ability to achieve a 100% duty cycle when in P-channel mode. In N-channel mode the duty cycle is limited to less than 100% to ensure proper start-up and thus the dropout voltage for the all N-channel converters is slightly higher.

The three application circuits demonstrate the fixed 3.3V version of the LTC1266. The LTC1266 is also available in fixed 5V and adjustable versions. All three versions are available in 16-pin SO packages.

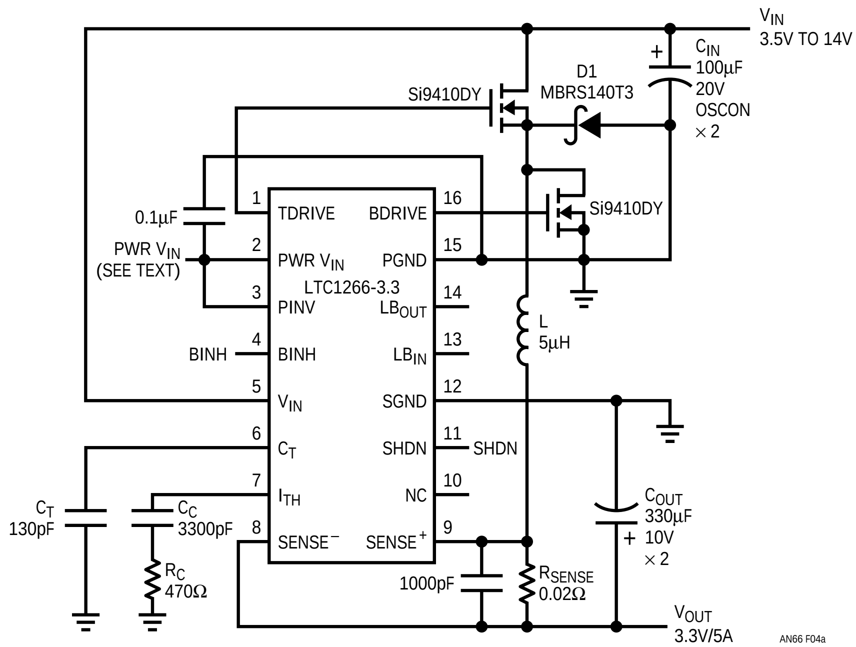


Figure 4a. All N-Channel 3.3V/5A Regulator with Drivers Powered from Seperate Power V_{IN} (PWR V_{IN}) Supply

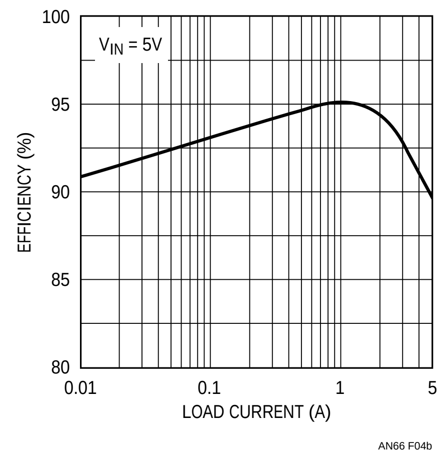


Figure 4b. Efficiency for Figure 4a's Circuit

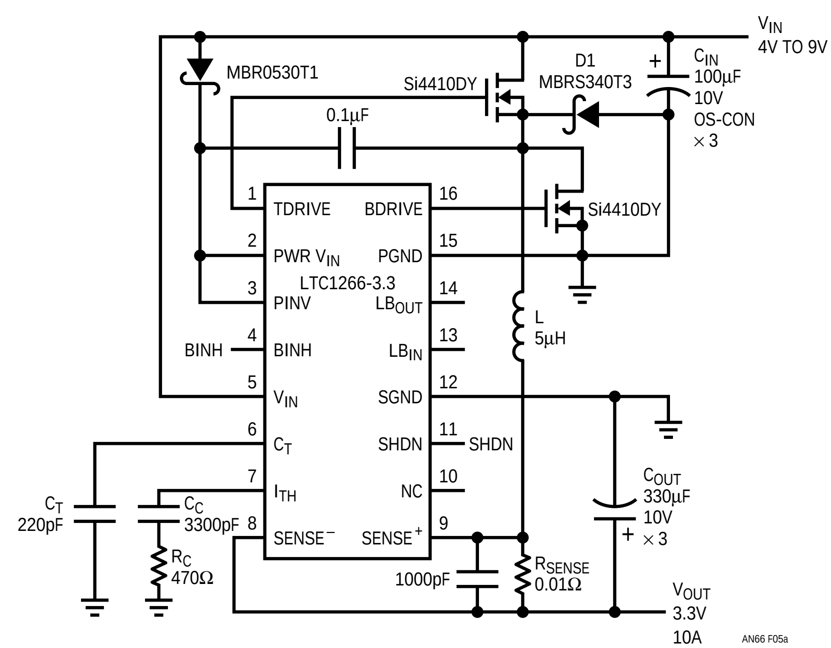


Figure 5a. All N-Channel Single Supply 5V to 3.3V/10A Regulator

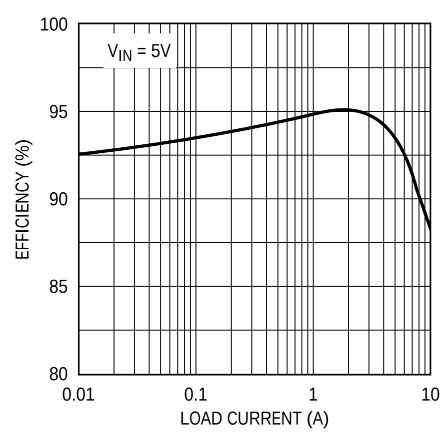


Figure 5b. Efficiency for Figure 5a's Circuit

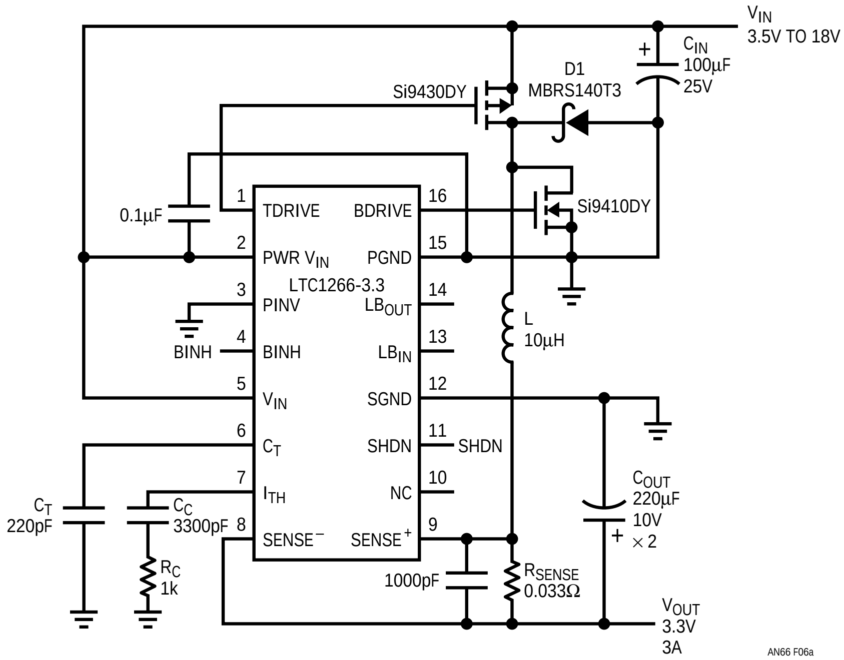


Figure 6a. Low Dropout 3.3V/3A Complementary MOSFET Regulator

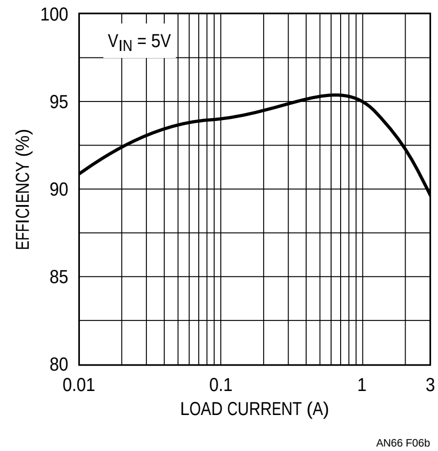


Figure 6b. Efficiency for Figure 6a's Circuit

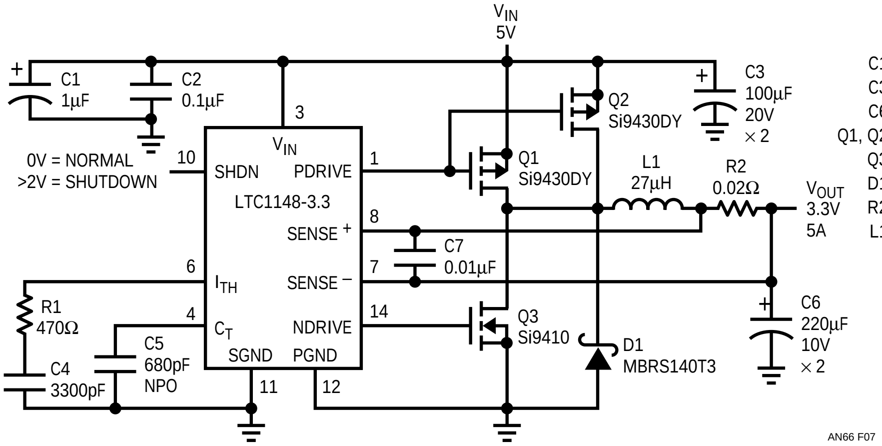
A HIGH EFFICIENCY 5V TO 3.3V/5A CONVERTER

by Randy G. Flatness

The next generation of notebook and desktop computers is incorporating more 3.3V ICs alongside 5V devices. As the number of devices increases, the current requirements also increase. Typically, a high current 5V supply is already available. Thus, the problem is reduced to deriving 3.3V from 5V efficiently in a small amount of board space.

High efficiency is mandatory in these applications, since converting 5V to 3.3V at 5A using a linear regulator would require dissipating over 8W. This wastes power and board space for heat sinking.

The LTC1148 synchronous switching regulator controller accomplishes the 5V to 3.3V conversion with high efficiencies over a wide load current range. The circuit shown in Figure 7 provides 3.3V at efficiencies greater than 90%



C1 = TANTALUM
C3 = SANYO (OS-CON) 20SA100M ESR = 0.037Ω I_{RMS} = 2.25A
C6 = AVX (TA) TPSE227K01R0080 ESR = 0.080Ω I_{RMS} = 1.285A
Q1, Q2 = SILICONIX PMOS BV_{DSS} = 20V DCR_{ON} = 0.100Ω Q_g = 50nC
Q3 = SILICONIX NMOS BV_{DSS} = 30V DCR_{ON} = 0.050Ω Q_g = 30nC
D1 = MOTOROLA SCHOTTKY VBR = 30V
R2 = KRL NP-2A-C1-0R020J Pd = 3W
L1 = KOOL Mµ® CORE, 16 GAUGE
COILTRONICS (408)241-7876
KRL BANTRY (603) 668-3210
SILICONIX (800) 554-5565
KOOL Mµ IS A REGISTERED TRADEMARK OF MAGNETICS, INC.

Figure 7. LTC1148-3.3 High Efficiency 5V to 3.3V/5A Step-Down Converter

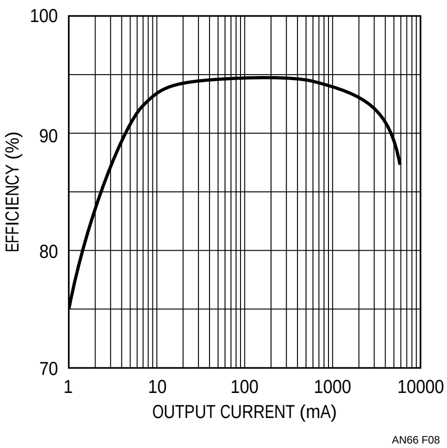


Figure 8. Efficiency for 5V to 3.3V Synchronous Switcher

from 5mA to 5A (over three decades of load current). The efficiency of the circuit in Figure 7 is plotted in Figure 8.

At an output current of 5A the efficiency is 90%; this means only 1.8W are lost. This lost power is distributed among R_{SENSE}, L1 and the power MOSFETs; thus heat sinking is not required.

The LTC1148 series of controllers use constant off-time current mode architecture to provide clean start-up, accurate current limit and excellent line and load regulation. To maximize the operating efficiency at low output currents, Burst Mode™ operation is used to reduce switching losses. Synchronous switching, combined with Burst Mode operation, yields very efficient energy conversion over a wide range of load currents.

The top P-channel MOSFETs in Figure 7 will be on 2/3 of the time with an input of 5V. Hence, these devices should be carefully examined to obtain the best performance. Two MOSFETs are needed to handle the peak currents safely and enhance high current efficiency. The LTC1148 can drive both MOSFETs adequately without a problem. A single N-channel MOSFET is used as the bottom synchronous switch, which shunts the Schottky diode. Finally, adaptive anti-shoot-though circuitry automatically prevents cross conduction between the complementary MOSFETs which can kill efficiency.

The circuit in Figure 7 has a no-load current of only 160µA. In shutdown mode, with Pin 10 held high (above 2V), the quiescent current decreases to less than 20µA with all MOSFETs held off DC. Although the circuit in Figure 7 is specified at a 5V input voltage, the circuit will function from 4V to 15V without requiring any component substitutions.

Burst Mode is a trademark of Linear Technology Corporation.

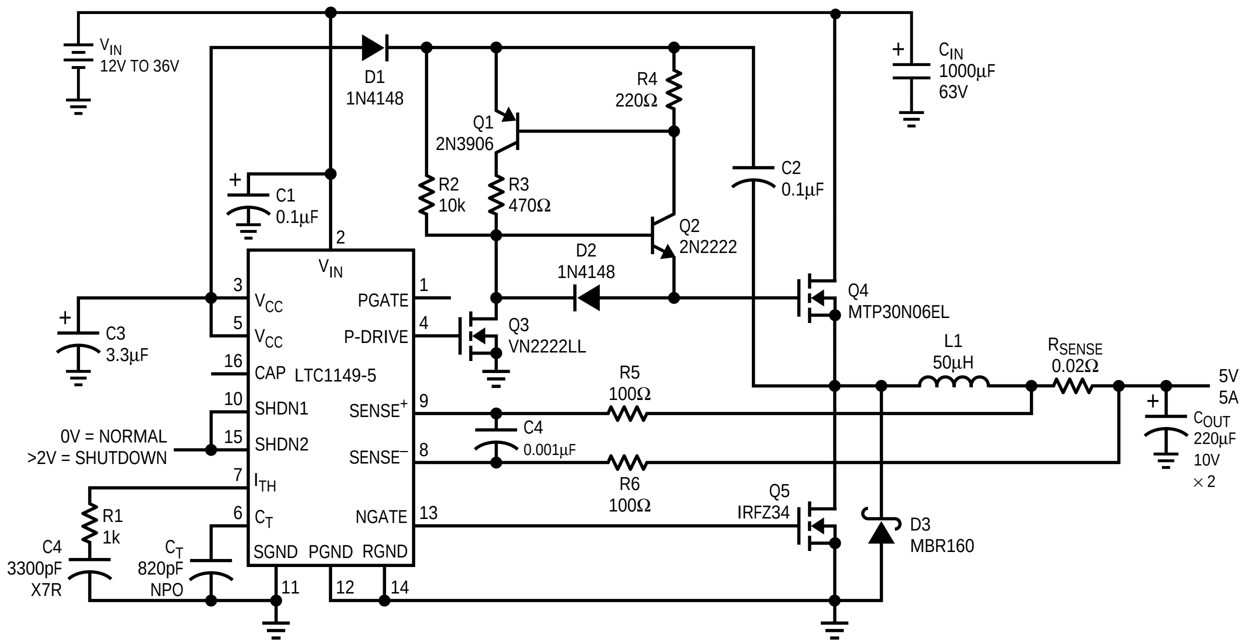
HIGH CURRENT, SYNCHRONOUS STEP-DOWN SWITCHING REGULATOR

by Brian Huffman

The LTC1149 is a half-bridge driver designed for synchronous buck regulator applications. Normally a P- and N-channel output stage is employed, but the P-channel device ON resistance becomes a limiting factor at output currents above 2A. N-channel MOSFETs are better suited for use in high current applications, since they have a substantially lower ON resistance than comparably priced P-channels. The circuit shown in Figure 9 adapts the LTC1149 to drive a half-bridge consisting of two N-channel MOSFETs, providing efficiency in excess of 90% at an output current of 5A.

The circuit's operation is as follows: the LTC1149 provides a P-drive output (Pin 4) that swings between ground and 10V, turning Q3 on and off. While Q3 is on, the N-channel MOSFET (Q4) is off because its gate is pulled low by Q3 through D2. During this interval, the Ngate output (Pin 13) turns the synchronous switch (Q5) on creating a low resistance path for the inductor current.

Q4 turns on when its gate is driven above the input voltage. This is accomplished by bootstrapping capacitor C2 off the drain of Q4. The LTC1149 V_{CC} output (Pin 3) supplies a regulated 10V output that is used to charge C2 through D1 while Q4 is off. With Q4 off, C2 charges to 5V during the first cycle in Burst Mode operation and to 10V thereafter.



C3 (TA) LOW ESR
C_{IN} NICHICON (AL) UPL1J102MRH, ESR = 0.027Ω, I_{RMS} = 2.370A
C_{OUT} SANYO (OS-CON) 10SA220M, ESR = 0.035Ω, I_{RMS} = 2.360A
Q1 PNP, BVCEO = 30V
Q2 NPN, BVCEO = 40V
Q3 SILICONIX NMOS, BVDSS = 60V, RDS_{ON} = 5Ω
Q4, Q5 NMOS, BVDSS = 60V, RDS_{ON} = 0.05Ω
D1, D2 SILICON, VBR = 75V
D3 MOTOROLA SCHOTTKY, VBR = 60V
R_{SENSE} = KRL NP-2A-C1-0R020J, P_D = 3W
L1 = COILTRONICS CTX50-5-52, DCR = 0.21Ω, IRON POWDER CORE
ALL OTHER CAPACITORS ARE CERAMIC

Figure 9. LTC1149-5 (12V-36V to 5V/5A) Using N-Channel MOSFETs

When Q3 turns off, the N-channel MOSFET is turned on by the SCR-connected NPN/PNP network (Q1 and Q2). Resistor R2 supplies Q2 with enough base drive to trigger the SCR. Q2 then forces Q1 to turn on, supplying more base drive to Q2. This regenerative process continues until both transistors are fully saturated. During this period, the source of Q4 is pulled to the input voltage. While Q4 is on, its gate source voltage is approximately 10V, fully enhancing the N-channel MOSFET.

Efficiency performance for this circuit is quite impressive. Figure 10 shows that for a 12V input the efficiency never drops below 90% over the 0.6A to 5A range. At higher input voltages efficiency is reduced due to transition losses in the power MOSFETs. For low output currents efficiency rolls off because of quiescent current losses.

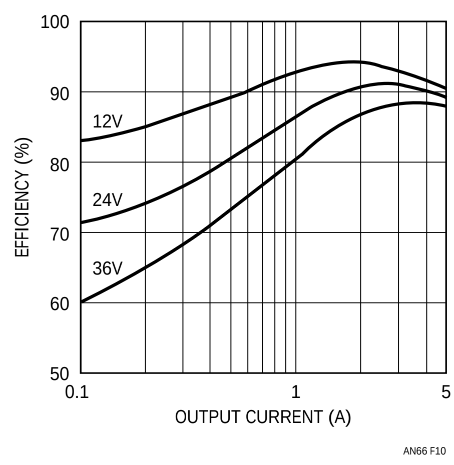


Figure 10. LTC1149-5 (12V-36V to 5V/5A) High Current Buck

Regulators—Switching (Buck)

Medium Power (1A to 4A)

1MHz STEP-DOWN CONVERTER ENDS 455kHz IF WOES

by Mitchell Lee

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 to 100kHz, placing troublesome harmonics right in the middle of the 455kHz band. This contributes to adverse effects such as "desensing" 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 11 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 through 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 12 shows the efficiency of Figure 11's circuit. You can expect 80% to 90% efficiency over an 8V to 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.

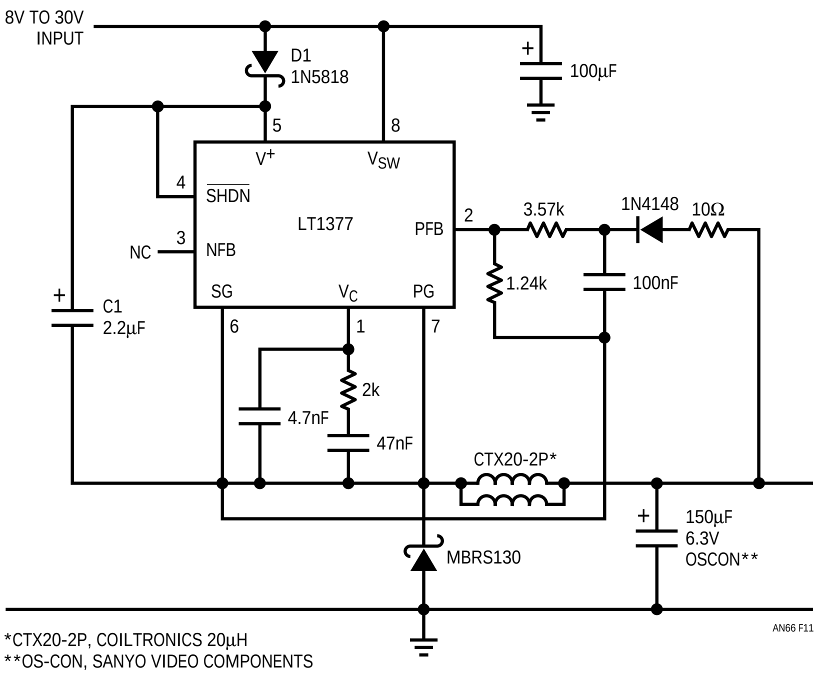


Figure 11. Schematic Diagram: 1MHz LT1377-Based Boost Converter

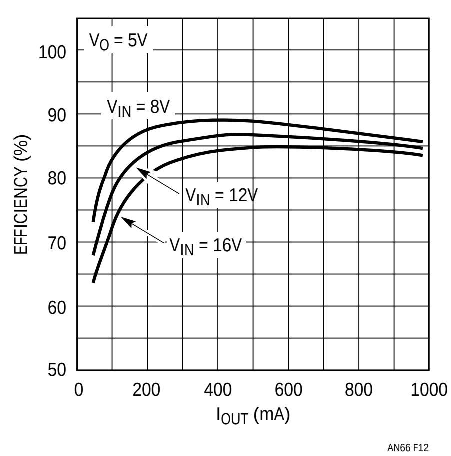


Figure 12. Efficiency Graph of the Circuit Shown in Figure 3

HIGH OUTPUT VOLTAGE BUCK REGULATOR

by Dimitry Goder

High efficiency step-down conversion is easy to implement using the LTC1149 as a buck switching regulator controller. The LTC1149 features constant off-time, current mode architecture and fully synchronous rectification. Current mode operation was selected for its well-known advantages of clean start-up, accurate current limit and excellent transient response.

Inductor current sensing is usually implemented by placing a resistor in series with the coil, but the common mode voltage at the LTC1149's Sense pins is limited to 13V. If a higher output voltage is required, the current sense resistor can be placed in the circuit's ground return to avoid common mode problems. The circuit in Figure 13 can be used in applications that do not lend themselves to this approach.

Figure 13 shows a special level shifting circuit (Q1 and U2) added to a typical LTC1149 application. The LT1211, a high speed, precision amplifier, forces the voltage across R5 to equal the voltage across current sense resistor R8. Q1's drain current flows to the source, creating a voltage across R6 proportional to the inductor current, which is now referenced to ground. This voltage can be directly applied to the current sense inputs of U1, the LTC1149. C12 and C4 are added to improve high frequency noise immunity. Maximum input voltage is now limited by the LT1211; it can be increased if a Zener diode is placed in parallel with C12.

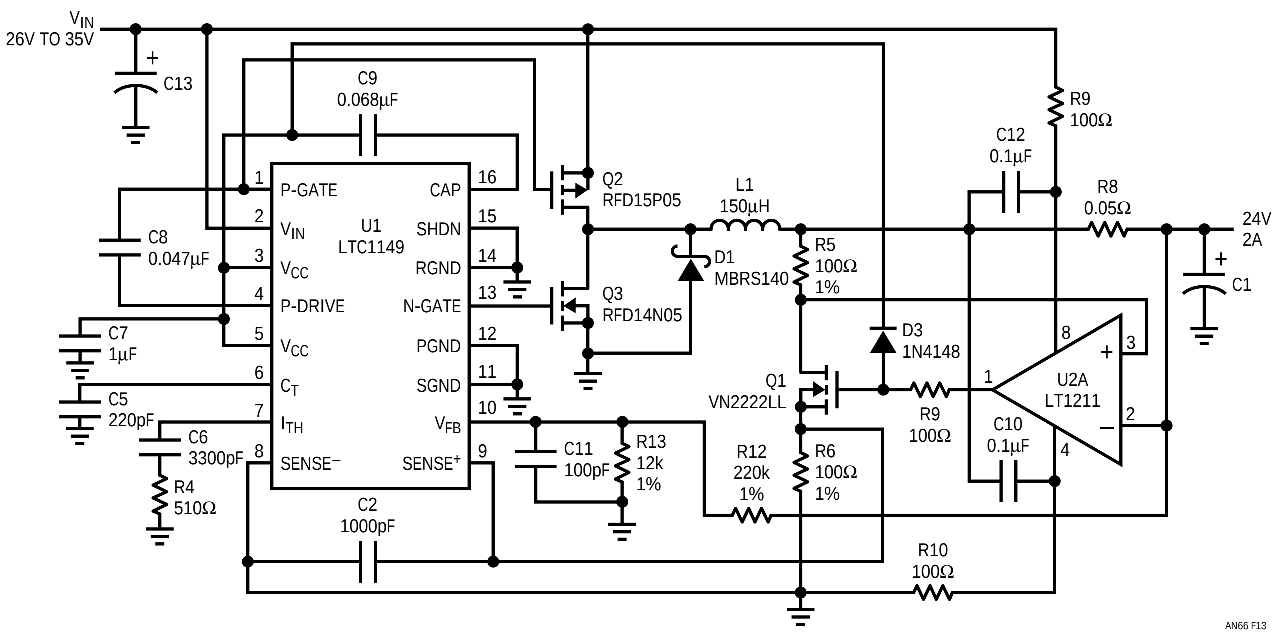


Figure 13. High Output Voltage Buck Regulator Schematic Using LTC1149

THE LTC1267 DUAL SWITCHING REGULATOR CONTROLLER OPERATES FROM HIGH INPUT VOLTAGES

by Randy G. Flatness

Fixed Output 3.3V and 5V Converter

A fixed LTC1267 application circuit creating 3.3V/2A and 5V/2A is shown in Figure 15. The operating efficiency shown in Figure 14 exceeds 90% for both the 3.3V and 5V sections. The 3.3V section of the circuit in Figure 15 comprises the main switch Q1, synchronous switch Q2, inductor L1 and current shunt R_{SENSE3}.

The 5V section is similar and comprises Q3, Q4, L2 and R_{SENSE5}. Each current sense resistor (R_{SENSE}) monitors the inductor current and is used to set the output current according to the formula $I_{OUT} = 100mV/R_{SENSE}$. Advantages of current control include excellent line and load transient rejection, inherent short-circuit protection and controlled start-up currents. Peak inductor currents for L1 and L2 are limited to $150mV/R_{SENSE}$ or 3.0A. The EXT V_{CC} pin is connected to the 5V output increasing efficiency at high input voltages. The maximum input voltage is limited by the MOSFETs and should not exceed 28V.

Adjustable Output 3.6V and 5V Converter

The adjustable output LTC1267-ADJ shown in Figure 16 is configured as a 3.6V/2.5A and 5V/2A converter. The resistor divider composed of R1 and R2 sets the output voltage according to the formula $V_{OUT} = 1.25V (1 + R2/R1)$. The input voltage range for this application is 5.5V to 28V.

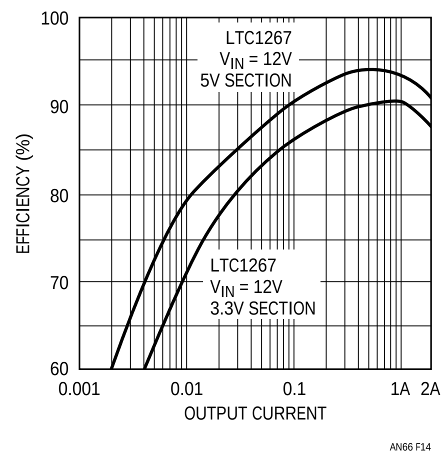


Figure 14. LTC1267 Efficiency vs Output Current of Figure 15 Circuit

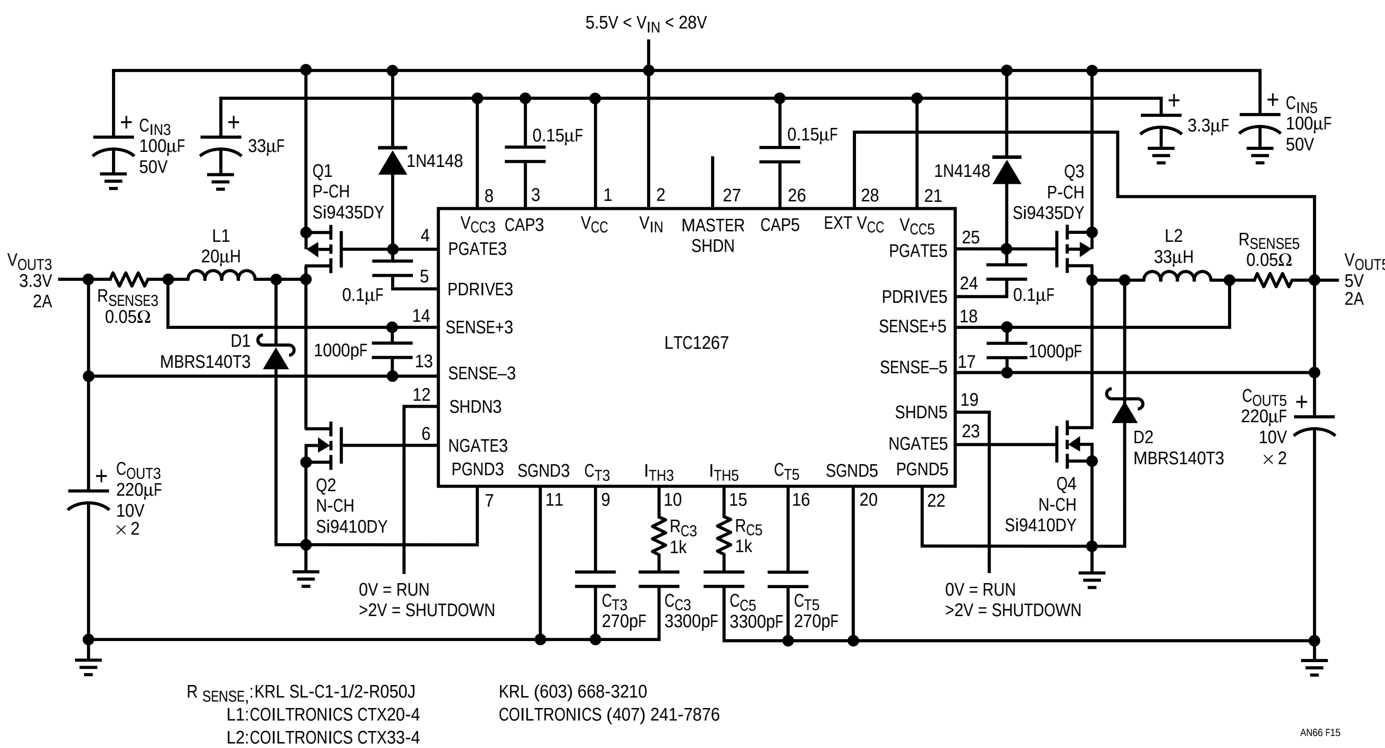


Figure 15. LTC1267 Dual Output 3.3V and 5V High Efficiency Regulator

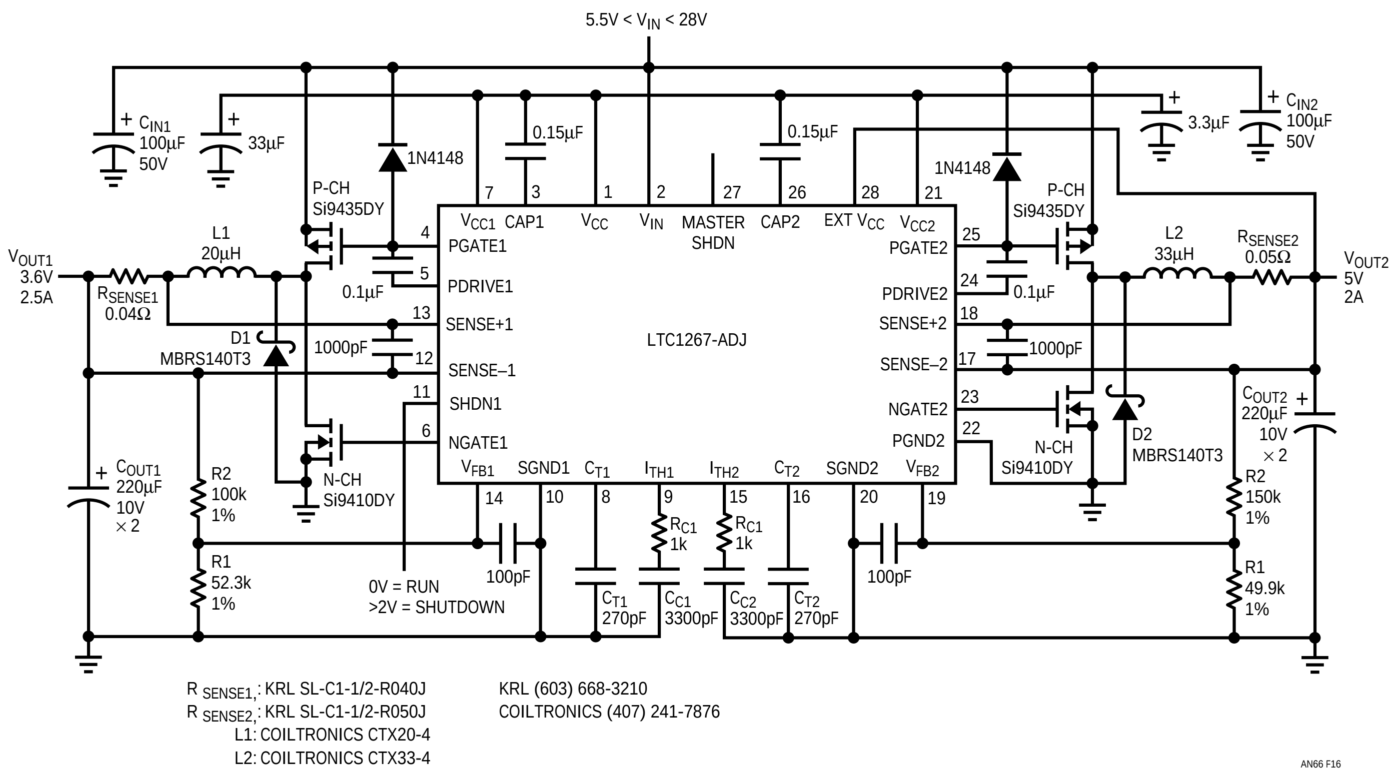


Figure 16. LTC1267 Dual Adjustable High Efficiency Regulator Circuit. Output Voltages Set at 3.6V and 5V

HIGH EFFICIENCY 5V TO 3.3V/1.25A CONVERTER IN 0.6 SQUARE INCHES

by Randy G. Flatness

The next generation of notebook and desktop computers will incorporate a growing number of 3.3V ICs along with 5V devices. As the number of 3.3V devices increases, the current requirements increase. Typically, a high current 5V supply is already available. Thus, the problem is reduced to deriving 3.3V from 5V at high efficiency in a small amount of board space.

High efficiency is mandatory in these applications since converting 5V to 3.3V at 1.25A using a linear regulator would require dissipating over 2W. This is an unnecessary waste of power and board space for heat sinking.

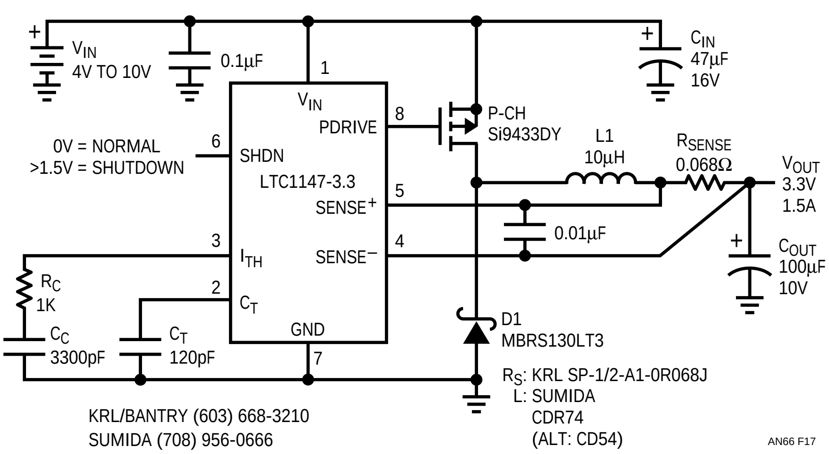


Figure 17. High Efficiency Controller Converts 5V to 3.3V in Minimum Board Area

The LTC1147 SO-8 switching regulator controller accomplishes the 5V to 3.3V conversion with high efficiencies over a wide load current range. The circuit shown in Figure 17 provides 3.3V at efficiencies greater than 90% from 50mA to 1.25A. Using all surface mount components and a low value of inductance (10µH) for L1, the circuit of Figure 17 occupies only 0.6 square inches of PC board area. The efficiency of the circuit in Figure 17 is plotted in Figure 18.

At an output current of 1.25A the efficiency is 90.4%; this means only 0.4W are lost. This lost power is distributed among R_{SENSE}, L1 and the power MOSFETs; thus heat sinking is not required.

The LTC1147 series of controllers use constant off-time current mode architecture to provide clean start-up, accurate current limit and excellent line and load regulation. To maximize the operating efficiency at low output currents, Burst Mode operation is used to reduce switching losses.

The P-channel MOSFET in the circuit of Figure 17 will be on 2/3 of the time with an input voltage of 5V. Hence, this device should be carefully selected to obtain the best performance. This design uses an Si9433DY for optimum efficiency; for lower cost an Si9340DY can be used at a slight reduction in performance.

The circuit in Figure 17 has a no load current of only 160µA. In shutdown, with Pin 6 held high (above 2V), the quiescent current is reduced to less than 20µA with the MOSFET held off. Although the circuit in Figure 17 is specified at a 5V input voltage the circuit will function from 4V to 10V.

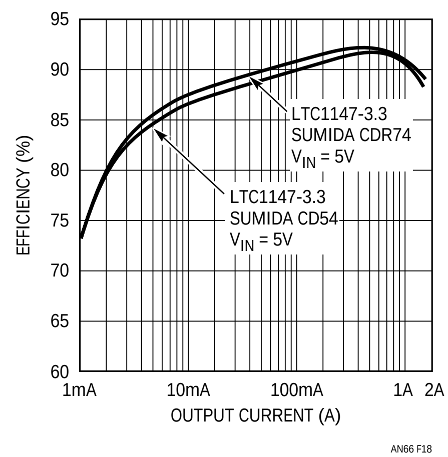


Figure 18. 5V to 3.3V Conversion Efficiency

LT1074/LT1076 ADJUSTABLE 0V TO 5V POWER SUPPLY

by Kevin Vasconcelos

Linear regulator ICs are commonly used in variable power supplies. Common types such as the 317 can be adjusted as low as 1.25V in single-supply applications. At low output voltages power losses in these regulators can be a problem. For example, if an output current of 1.5A is required at 1.25V from an input of 8V, the regulator dissipates more than 10W. Figure 19 shows a DC/DC converter that functionally replaces a linear regulator in this application. The converter not only eliminates power

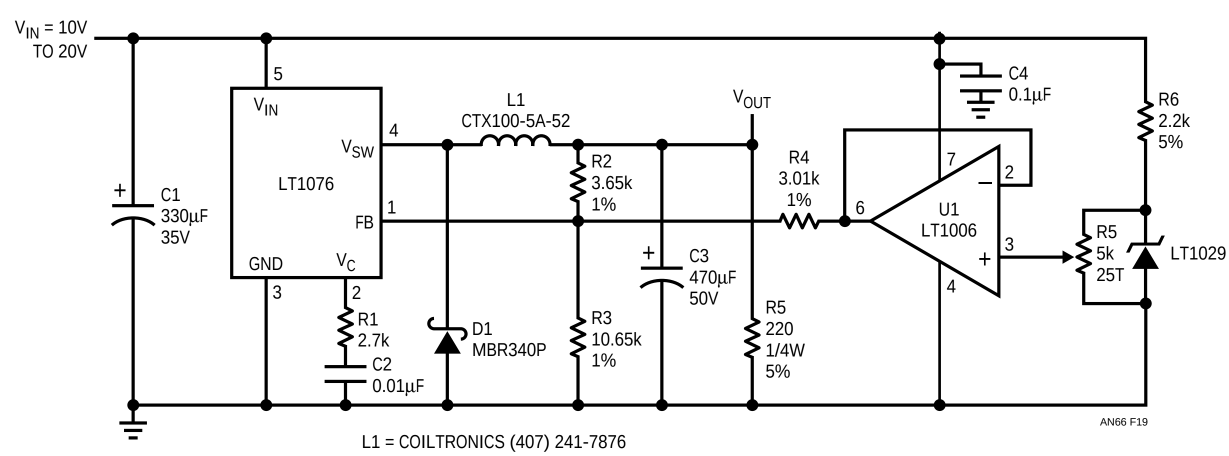


Figure 19. Adjustable LT1074/LT1076 0V to 5V Power Supply

loss as a concern, but can be adjusted for output voltages as low as 25mV while still delivering an output current of 1.5A.

The circuit of Figure 19 employs a basic positive buck topology with one exception: a control voltage is applied through R4 to the feedback summing node at Pin 1 of the LT1076 switching regulator IC, allowing the output to be adjusted from 0V to approximately 6V. This encompasses the 3.3V and 5V logic supply ranges as well as battery pack combinations of one to four D cells.

As R4 is driven from 0V to 5V by the buffer (U1) more or less current is required from R2 to satisfy the loop's desire to hold the feedback summing point at 2.21V. This forces the converter's output to swing over the range of 0V to 6V.

Figure 20 shows a comparison of power losses for a linear regulator and the circuit of Figure 19. The load current is 1.5A in both cases although the LT1076 is capable of 1.75A guaranteed output current in this application and 2A typical. If more current is required the LT1074 can be substituted for the LT1076. This change accommodates outputs up to 5A but at the expense of a heftier diode and coil (D1, L1). An MBR735 and Coiltronics CTX50-2-52 are recommended for 5A service.

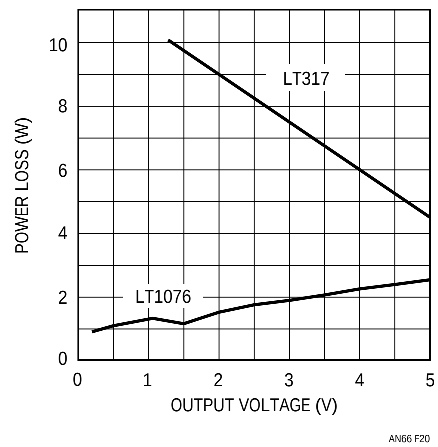


Figure 20. Power Loss Comparison: Linear Regulator vs Figure 19's Power Supply

TRIPLE OUTPUT 3.3V, 5V AND 12V HIGH EFFICIENCY NOTEBOOK POWER SUPPLY

by Randy G. Flatness

LTC1142 Circuit Operation

The application circuit in Figure 22 is configured to provide output voltages of 3.3V, 5V and 12V. The current capability of both the 3.3V and 5V outputs is 2A (2.5A peak). The logic-controlled 12V output can provide 150mA (200mA peak), which is ideal for flash memory applications. The operating efficiency shown in Figure 21 exceeds 90% for both the 3.3V and 5V sections.

The 3.3V section of the circuit in Figure 22 comprises the main switch Q4, synchronous switch Q5, inductor L1 and current shunt R_{SENSE3}. The current sense resistor R_{SENSE} monitors the inductor current and is used to set the output current according to the formula $I_{OUT} = 100mV/R_{SENSE}$. Advantages of current control include excellent line and load transient rejection, inherent short-circuit protection and controlled start-up currents. Peak inductor currents for L1 and T1 of the circuit in Figure 22 are limited to $150mV/R_{SENSE}$ or 3.0A and 3.75A respectively.

When the output current for either regulator section drops below approximately $15mV/R_{SENSE}$, that section automatically enters Burst Mode operation to reduce switching losses. In this mode the LTC1142 holds both MOSFETs off and "sleeps" at 160μA supply current while the output capacitor supports the load. When the output capacitor falls 50mV below its specified voltage (3.3V or 5V) the LTC1142 briefly turns this section back on, or "bursts," to recharge the output capacitor. The timing capacitor pins,

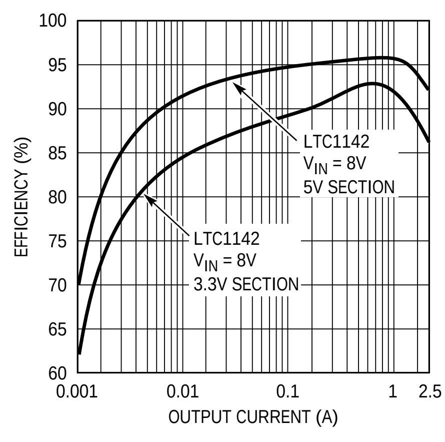


Figure 21. LTC1142 Efficiency

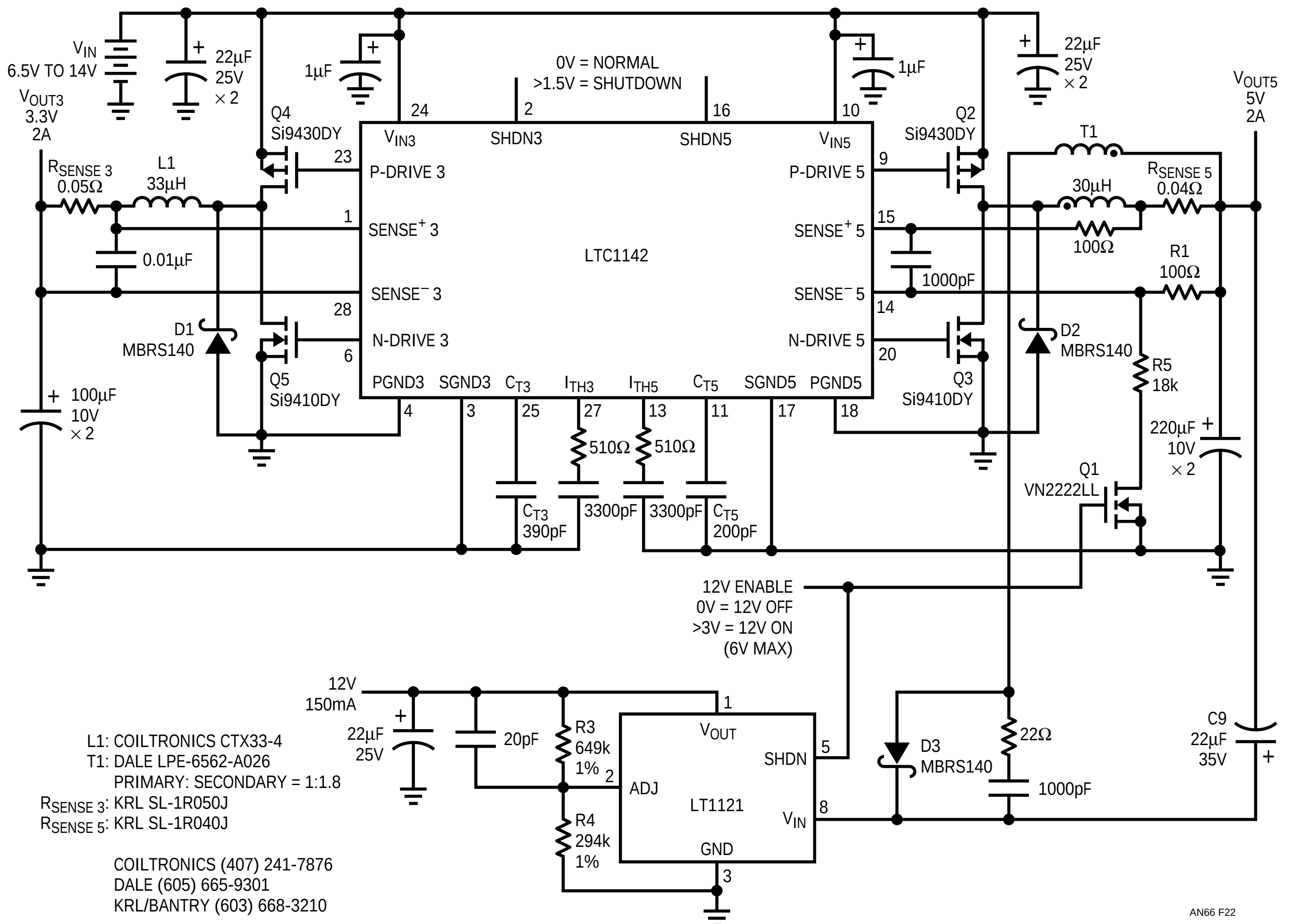


Figure 22. LTC1142 High Efficiency Power Supply Schematic Diagram

which go to 0V during the sleep interval, can be monitored with an oscilloscope to observe burst action. As the load current is decreased the circuit will burst less and less frequently.

The timing capacitors C_{T3} and C_{T5} set the off-time according to the formula $t_{OFF} = 1.3\,(10^4)(C_T)$. The constant off-time architecture maintains a constant ripple current while the operating frequency varies only with input voltage. The 3.3V section has an off-time of approximately 5µs, resulting in a operating frequency of 120kHz with an 8V input. The 5V section has an off-time of 2.6µs and a switching frequency of 140kHz with an 8V input.

Auxiliary 12V Output

The operation of the 5V section is identical to the 3.3V section with inductor L1 replaced by transformer T1. The 12V output is derived from an auxiliary winding on the 5V inductor. The output from this additional winding is rectified by diode D3 and applied to the input of an LT1121 regulator. The output voltage is set by resistors R3 and R4. A turns ratio of 1:1.8 is used for T1 to ensure that the input voltage to the LT1121 is high enough to keep the regulator out of dropout mode while maximizing efficiency.

The LTC1142 synchronous switch removes the normal limitation that power must be drawn from the primary 5V inductor winding in order to extract power from the auxiliary winding. With synchronous switching, the auxiliary 12V output may be loaded without regard to the 5V primary output load, provided that the loop remains in continuous mode operation.

When the 12V output is activated by a TTL high (6V maximum) on the 12V enable line, the 5V section of the LTC1142 is forced into continuous mode. A resistor divider composed of R1, R5 and switch Q1 forces an offset, subtracting from the internal offset at Pin 14. When this external offset cancels the built-in 25mV offset, Burst Mode operation is inhibited.

Auxiliary 12V Output Options

The circuit of Figure 22 can be modified for operation in low-battery count (6-cell) applications. For applications where heavy 12V load currents exist in conjunction with low input voltages (< 6.5V), the auxiliary winding should be derived from the 3.3V instead of the 5V section. As the input voltage falls, the 5V duty cycle increases to the point when there is simply not enough time to transfer energy from the 5V primary winding to the 12V secondary winding. For operation from the 3.3V section, a transformer with a turns ratio of 1:3.25 should be used in place of the 33µH inductor L1. Likewise, a 30µH inductor would replace T1 in the 5V section. With these component changes, the duty cycle of the 3.3V section is more than adequate for full 12V load currents. The minimum input voltage in this case will be determined only by the dropout voltage of the 5V output. The 100% duty cycle inherent in the LTC1142 provides low dropout operation limited only by the load current multiplied by the sum of the resistances of the 5V inductor, Q2 $R_{DS(ON)}$ and current sense resistor R_{SENSE5}.

Extending the Maximum Input Voltage

The circuit in Figure 22 is designed for a 14V maximum input voltage. The operation of the circuit can be extended to over 18V if a few key components are changed. The parts that determine the maximum input voltage of the circuit are the power MOSFETs, the LTC1142 and the input capacitors. With the LTC1142 replaced by an LTC1142HV, an 18V typical (20V maximum) input voltage is allowable. Since the gate drive voltages supplied by the LTC1142 and LTC1142HV are from ground to V_{IN}, the input voltage must not exceed the maximum V_{GS} of the MOSFETs. The MOSFETs specified in Figure 22 have an absolute maximum of 20V, matching that of the LTC1142HV.[1] Finally, the input capacitor's voltage rating will also have to be increased above 12V.

[1]For improved efficiency, CT5 should be charged to 270pF.

THE NEW SO-8 LTC1147 SWITCHING REGULATOR CONTROLLER OFFERS HIGH EFFICIENCY IN A SMALL FOOTPRINT

by Randy Flatness

Introduction

The LTC1147 switching regulator controller is a high efficiency step-down DC/DC converter. It uses the same current mode architecture and Burst Mode operation as the LTC1148/LTC1149 but without the synchronous switch. Ideal for applications requiring up to 1A, the LTC1147 shows 90% efficiencies over two decades of output current.

High Efficiency 5V to 3.3V in a Small Area

The LTC1147 5V to 3.3V converter shown in Figure 23 has 85% efficiency at 1A output with efficiencies greater than 90% for load currents up to 500mA. Using the LTC1147 reduces the power dissipation to less than 500mW. The efficiency plotted as a function of output current is shown in Figure 24.

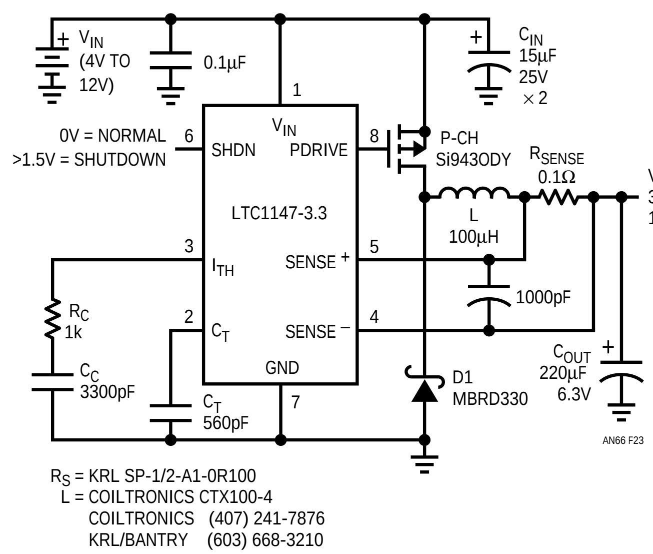


Figure 23. This LTC1147 5V to 3.3V Converter Achieves 92% Efficiency at 300mA Load Current

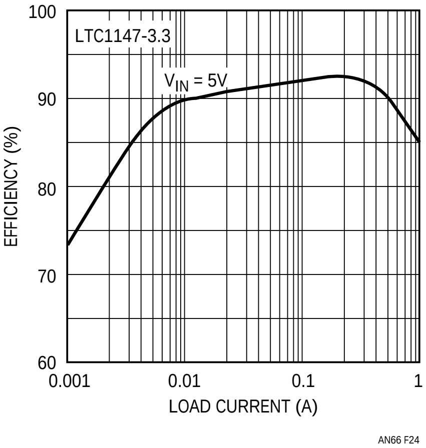


Figure 24. The LTC1147 5V to 3.3V Converter Provides Better Than 90% Efficiency from 20mA to 500mA of Output Current

Giving Up the Synchronous Switch?

The decision whether to use a nonsynchronous LTC1147 design or a fully synchronous LTC1148 design requires a careful analysis of where losses occur. The LTC1147 switching regulator controller uses the same loss reducing techniques as the other members of the LTC1148/LTC1149 family. The nonsynchronous design saves the N-channel MOSFET gate drive current at the expense of increased loss due to the Schottky diode.

Figure 25 shows how the losses in a typical LTC1147 application are apportioned. The gate-charge loss (P-channel MOSFET) is responsible for the majority of the efficiency lost in the midcurrent region. If Burst Mode operation was not employed, the gate charge loss alone would cause the efficiency to drop to unacceptable levels at low output currents. With Burst Mode operation, the DC supply current represents the only loss component that increases almost linearly as output current is reduced. As expected, the I^2R loss and Schottky diode loss dominate at high load currents.

In addition to board space, output current and input voltage are the two primary variables to consider when deciding whether to use the LTC1147. At low input-to-output voltage ratios, the top P-channel switch is on most of the time, leaving the Schottky diode conducting only a small percentage of the total period. Hence, the power lost in the Schottky diode is small at low output currents. This is the ideal application for the LTC1147. As the output current increases the diode loss increases. At high input-to-output voltage ratios, the Schottky diode conducts most of the time. In this situation, any loss in the diode will have a more significant effect on efficiency and an LTC1148 might therefore be chosen.

Figure 26 compares the efficiencies of LTC1147-5 and LTC1148-5 circuits with the same inductor, timing capacitor and P-channel MOSFET. At low input voltages and 1A output current the efficiency of the LTC1147 differs from that of the LTC1148 by less than two percent. At lower

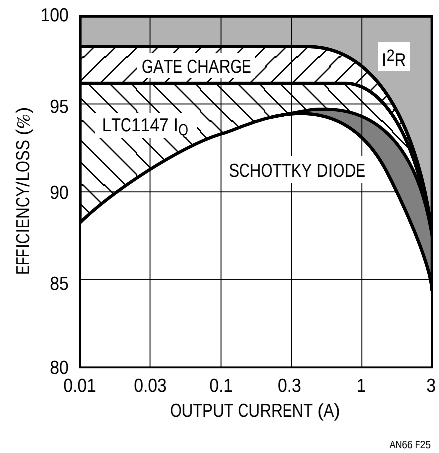


Figure 25. Low Current Efficiency is Enhanced by Burst Mode Operation. Schottky Diode Loss Dominates at High Output Currents

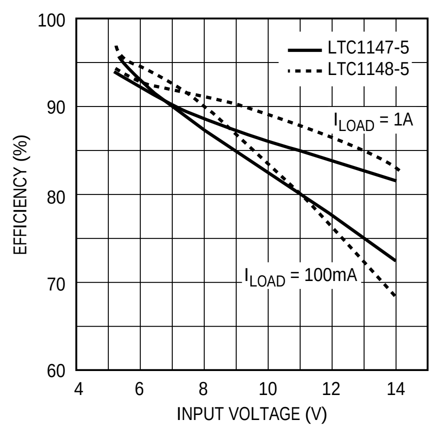


Figure 26. At High Input Voltages Combined with Low Output Currents, the Efficiency of the LTC1147 Exceeds That of the LTC1148

output currents and high input voltages the LTC1147's efficiency can actually exceed that of the LTC1148.

Low Dropout 5V Output Applications

Because the LTC1147 is so well-suited for low input-to-output voltage ratio applications it is an ideal choice for low dropout designs. All members of the LTC1148/LTC1149 family (including the LTC1147) have outstandingly low dropout performance. As the input voltage on the LTC1147 drops, the feedback loop extends the on-time for the P-channel switch (off-time is constant) thereby keeping the inductor ripple current constant. Eventually the on-time extends so far that the P-channel MOSFET is on at DC or at a 100% duty cycle.

With the switch turned on at a 100% duty cycle, the dropout is limited by the load current multiplied by the sum of the resistances of the MOSFET, the current shunt and the inductor. For example, the low dropout 5V regulator shown in Figure 27 has a total resistance of less than 0.2Ω. This gives it a dropout voltage of 200mV at 1A output current. At input voltages below dropout the output voltage follows the input. This is the circuit whose efficiency is plotted in Figure 28.

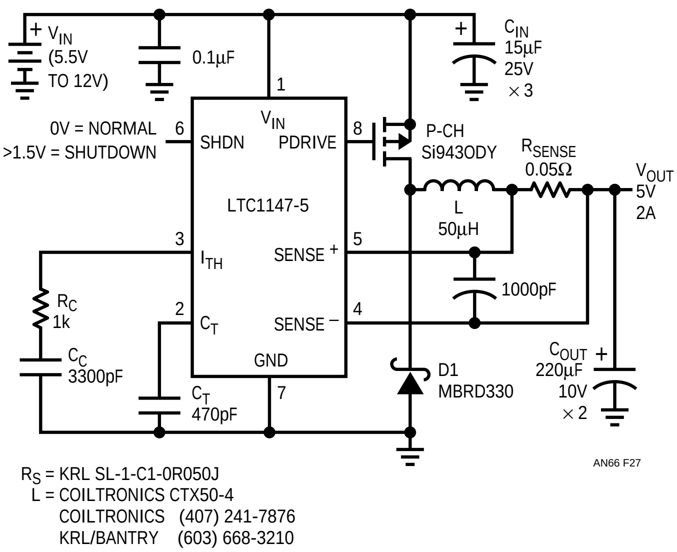


Figure 27. The LTC1147 Architecture Provides Inherent Low Dropout Operation. This LTC1147-5 Circuit Supports a 1A Load with the Input Voltage Only 200mV Above the Output

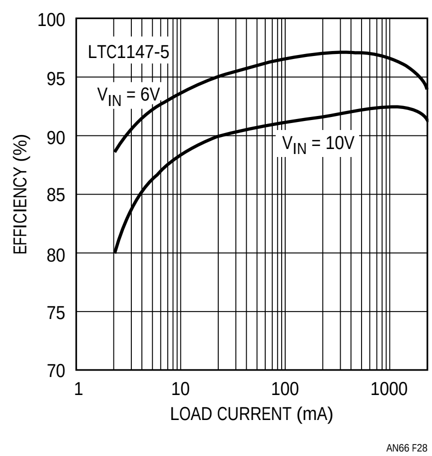


Figure 28. Greater Than 90% Efficiency is Obtained for Load Currents of 20mA to 2A (V_{IN} = 10V)

THE LT1432: 5V REGULATOR ACHIEVES 90% EFFICIENCY

by Carl Nelson

Power supply efficiency has become a highly visible issue in many portable battery-powered applications. Higher efficiency translates directly to longer useful operating time—a potent selling point for products such as notebook computers, cellular phones, data acquisition units, sales terminals and word processors. The "holy grail" of efficiency for 5V outputs is 90%.

For a number of reasons, older designs were limited to efficiencies of 80 to 85%. High quiescent current in the control circuitry limited efficiency at lower output currents. Losses in the power switch, inductor and catch diode all added up to limit efficiency at moderate-to-high output currents. Each of these areas must be addressed in a design that is to have high efficiency over a wide output current range.

Some portable equipment has the additional requirement of high efficiency at extremely light loads (1mA to 5mA). These applications have a sleep mode in which RAM is kept alive to retain information. The instrument may spend days or even weeks in this mode, so battery drain is critical. Ordinary 5V switchers draw quiescent currents of 5mA to 15mA for these light loads. The efficiency of a 12V to 5V converter with 10mA supply current and 1mA load is only 4%. Clearly, some method must be provided to eliminate the quiescent current of the switching regulator control section.

An additional requirement for some systems is full shutdown of the regulator. It would be ideal if a simple logic signal could cause the converter to turn off and draw only a few microamperes of current.

The combination of battery form factors, their discrete voltage steps and the use of higher voltage wall adapters requires a switching regulator that operates with inputs from 6V to 30V. Both of these voltages present problems for a MOS design because of minimum and maximum gate voltage requirements of power MOS switches.

The LT1432 was designed to address all the requirements described above. It is a bipolar control chip that interfaces directly to the LT1070 family of switching regulators and is capable of operating with 6V to 30V inputs. These ICs have a very efficient, quasisaturating NPN switch that mimics the resistive nature of MOS transistors with much smaller die areas. The NPN is a high frequency device with

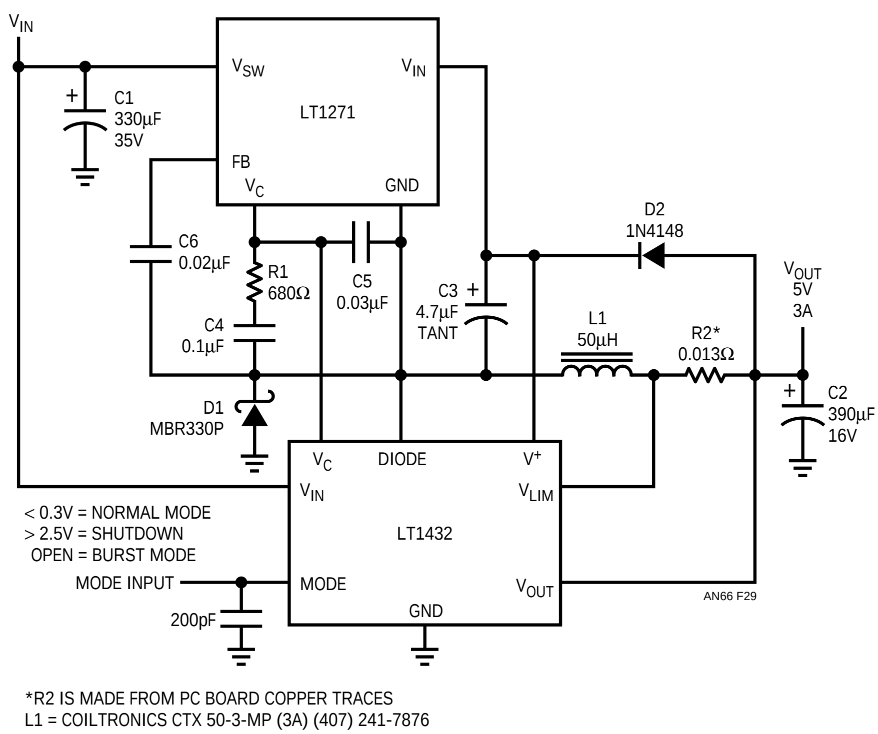


Figure 29. High Efficiency 5V Buck Converter

an equivalent voltage and current overlap time of only 10ns. Drive to the switch is automatically scaled with switch current, so drive losses are also low. Switch and driver losses using an LT1271 with a 12V input and a 5V, 500mA load are only about 2%.

To reduce quiescent current losses, the LT1271 is powered from the 5V output rather than from the input voltage. This is done by pumping the supply capacitor C3 from the output via D2. Quick minded designers will observe that this arrangement does not self-start; accordingly, a parallel path was included inside the LT1432 to provide power to the IC switcher directly from the input during start-up. Equivalent quiescent supply current is reduced to about 3.5mA with this technique.

Catch diode losses cannot be reduced with IC "tricks" unless the diode is replaced with a synchronously driven MOS switch. This is more expensive and still requires the diode to avoid voltage spikes during switch nonoverlap times. The question is, is it worth it?

The following formula was developed to calculate the improvement in efficiency when adding a synchronous switch.

$$\text{Efficiency change} = \frac{(V_{IN} - V_{OUT})(V_f - R_{FET} \bullet I_{OUT})(E)^2}{(V_{IN})(V_{OUT})}$$

With V_{IN} = 10V, V_{OUT} = 5V, V_f (diode forward voltage) = 0.45V, R_{FET} = 0.1Ω and I_{OUT} = 1A the improvement in efficiency is only 2.8%. This does not take into account the losses associated with MOS gate drive, so real improvement would probably be closer to 2%. The availability of low forward voltage Schottky diodes such as the MBR330P makes synchronous switches less attractive than they used to be.

To achieve higher efficiency during sleep, the LT1432 has Burst Mode operation. In this mode the LT1271 is either driven full on, or completely shut down to its micropower state. The LT1432 acts as a comparator with hysteresis instead of a linear amplifier. This mode reduces equivalent input supply current to 1.3mA with a 12V battery. Battery life with NiCd AA cells is over 300 hours with a 1mA 5V load. Burst Mode operation increases output ripple, especially with higher output currents, so maximum load in this mode is 100mA.

The LT1271 normally draws about 50µA to100µA in its shutdown state. A shutdown command to the LT1432 opens all connections to the LT1271 V_{IN} pin so its current drain is eliminated. This leaves only the shutdown current of the LT1432 and the switch leakage of the LT1271, which typically add up to less than 20µA—less than the self-discharge rate of NiCd batteries. For many applications the on/off function is under keystroke control. Digital chips which draw only a few microamps are available for keystroke recognition and power control.

There is no way to design around inductor losses. These losses are minimized by using low loss cores such as molypermalloy or ferrite, and by sizing the core to use wire with sufficient diameter to keep resistive losses low. The 50µH inductor shown has a core loss of 200mW with type-52 powdered iron material and 28mW with molypermalloy. For a 1A load this represents efficiency losses of 4% and 0.56% respectively—a major difference. Ferrite cores would have even lower losses than molypermalloy, but the "moly" has such low losses that ferrites should be chosen for other reasons, such as height, cost, mounting and the like. DC resistance of the inductor shown is 0.02Ω. This represents an efficiency loss of 0.4% at 1A load and 0.8% at 2A. Significant reduction in these resistance losses would require a somewhat larger inductor. The choice is yours.

The LT1432 has a high efficiency current limit with a sense voltage of only 60mV. This has a side benefit in that printed circuit board trace material can be used for the sense resistor. A 3A limit requires a 0.02Ω sense resistor and this is easily made from a small section of serpentine trace. The 60mV sense voltage has a positive temperature coefficient that tracks that of copper so that the current limit is flat with temperature. Foldback current limiting can be easily implemented.

The LT1432 represents a significant improvement in high efficiency 5V supplies that must operate over a wide range of load currents and input voltages. Its efficiency has a very broad peak that exceeds 90%, requiring a new definition of the "holy grail." Logic controlled shutdown, millipower Burst Mode operation and efficient, accurate, current limiting make this regulator extremely attractive for battery-powered applications.

Regulators—Switching (Buck)

Low Power (<1A)

APPLICATIONS FOR THE LTC1265 HIGH EFFICIENCY MONOLITHIC BUCK CONVERTER

by San-Hwa Chee

Efficiency

Figure 30 shows a typical LTC1265-5 application circuit. The efficiency curves for two different input voltages are shown in Figure 31. Note that the efficiency for a 6V input exceeds 90% over a load range from less than 10mA to 850mA. This makes the LTC1265 attractive for all battery operated products and efficiency sensitive applications.

5V to 3.3V Converter

Figure 32 shows the LTC1265 configured for 3.3V output with 1A output current capability. This circuit operates at a frequency of 100kHz. Figure 33 is the efficiency plot of the circuit. At a load current of 100mA the efficiency is at 92%; the efficiency falls to 82% at a 1A output.

2.5mm Typical-Height 5V to 3.3V Regulator

Figure 34 shows the schematic for a very thin 5V to 3.3V converter. For the LTC1265 to be able to source 500mA output current and yet meet the height requirement, a small value inductor must be used. The circuit operates at a high frequency (500kHz typically) increasing the gate charge losses. Figure 35 is the efficiency curve for this application.

Positive-to-Negative Converter

Besides converting from a positive input to positive output, the LTC1265 can be configured to perform a positive-to-negative conversion. Figure 36 shows the schematic for this application.

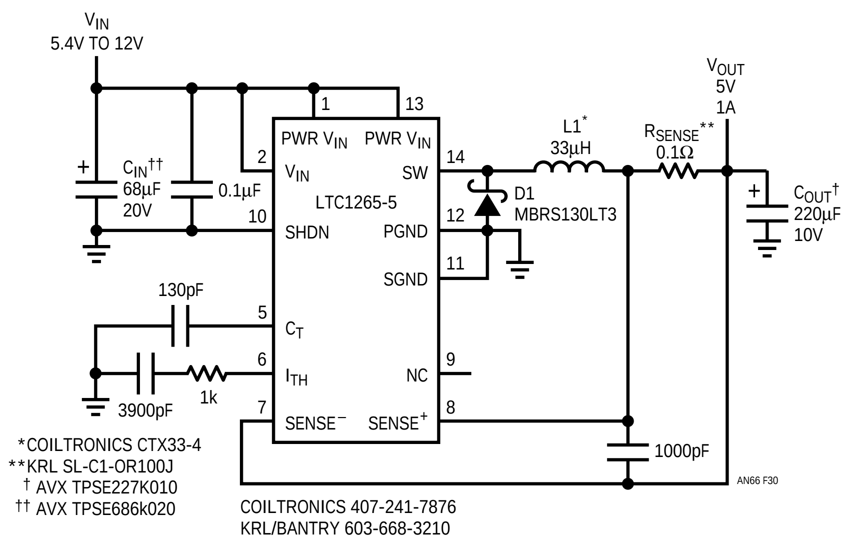


Figure 30. High Efficiency Step-Down Converter

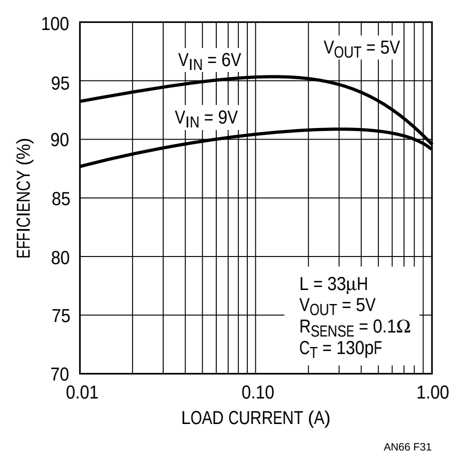
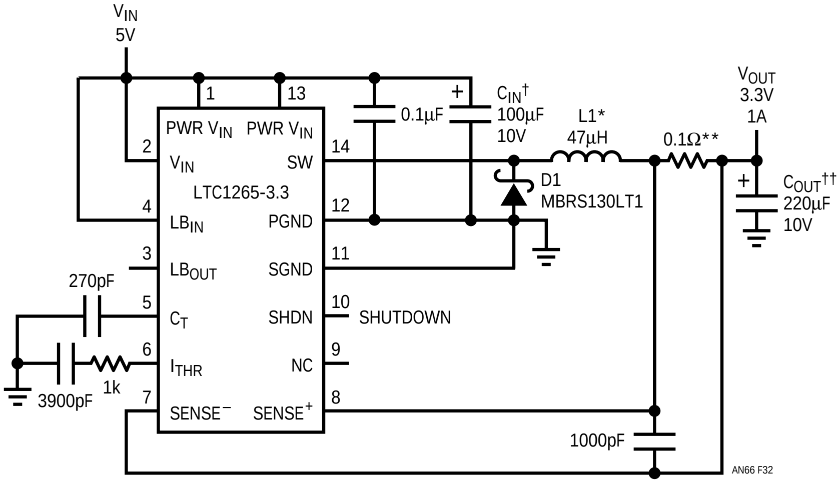


Figure 31. Efficiency vs Load Current



*COILCRAFT D03316-473
**KRL SL-C1-OR100J
† AVX TAJD100K010
†† AVX TAJD226K010

COILCRAFT 708-639-6400
KRL/BANTRY 603-668-3210

Figure 32. High Efficiency 5V to 3.3V Converter

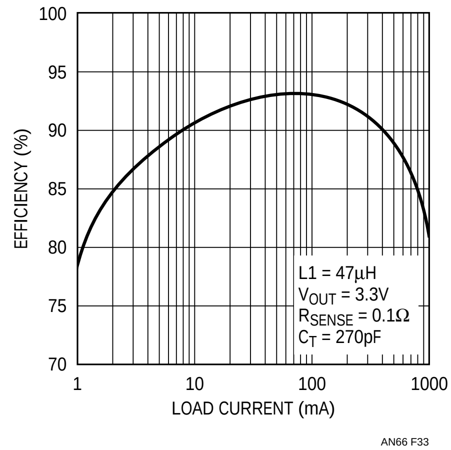
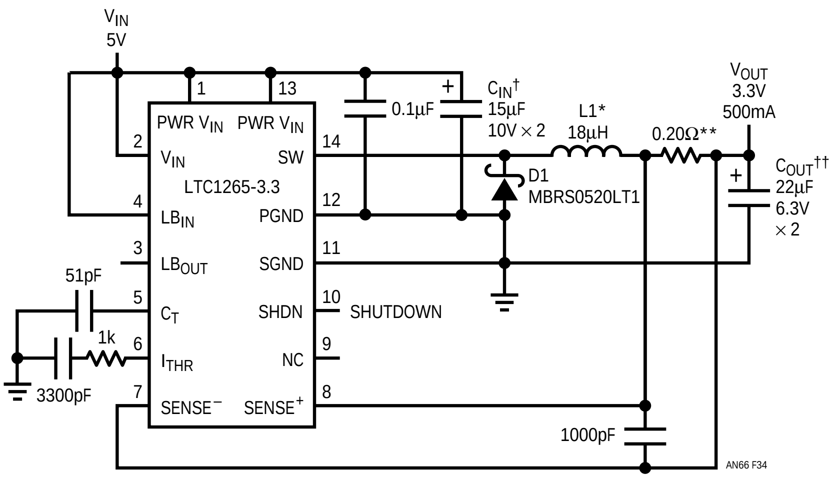


Figure 33. Efficiency vs Load Current



*SUMIDA CLS62-180
**KRL SL-C1-OR200J
† AVX TAJB155K010
†† AVX TAJB225K06

SUMIDA 708-956-0666
KRL/BANTRY 603-668-3210

Figure 34. 2.5mm High 5V to 3.3V Converter (500mA Output Current)

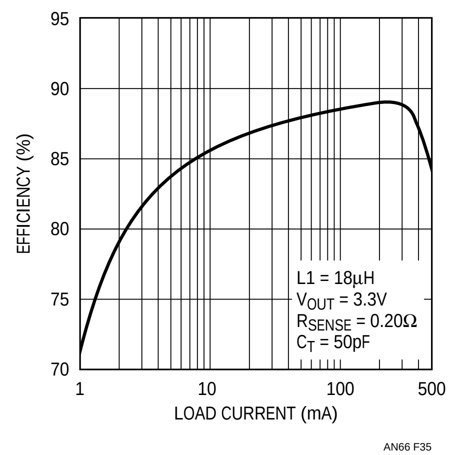
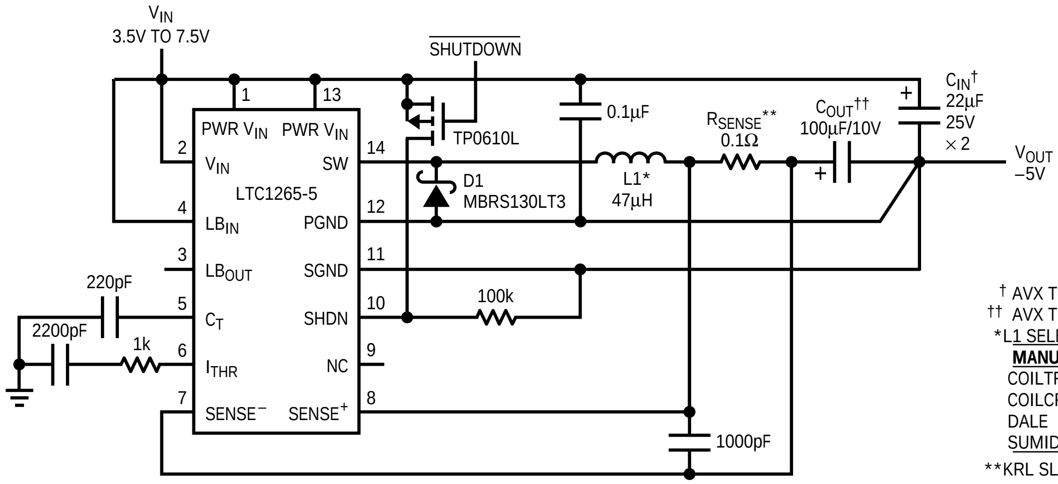


Figure 35. Efficiency vs Load Current



V_{IN} (V)	$I_{OUT(MAX)}$ (mA)
3.5	360
4.0	430
5.0	540
6.0	630
7.0	720
7.5	740

† AVX TPSD226K025
†† AVX TPSD106K010
*L1 SELECTION

MANUFACTURER	PART NO.
COILTRONICS	CTX50-4
COILCRAFT	D03316-473
DALE	LPT4545-500LA
SUMIDA	CD75-470

**KRL SL-C1-OR100J

Figure 36. Positive (3.5 to 7.5V) to Negative (–5V) Converter

Regulators—Switching (Boost)

Medium Power (1A to 4A)

HIGH OUTPUT CURRENT BOOST REGULATOR

by Dimitry Goder

Low voltage switching regulators are often implemented with self-contained power integrated circuits featuring a PWM controller and an onboard power switch. Maximum switch currents of up to 10A are available, providing a convenient means for power conversion over wide input and output voltage ranges. If higher switch currents are required, a controller with an external power MOSFET is a better choice.

Figure 37 shows an LTC1147-based 5V to 12V converter with 3.5A peak output current capability. The LTC1147 is a micropower controller that uses a constant off-time

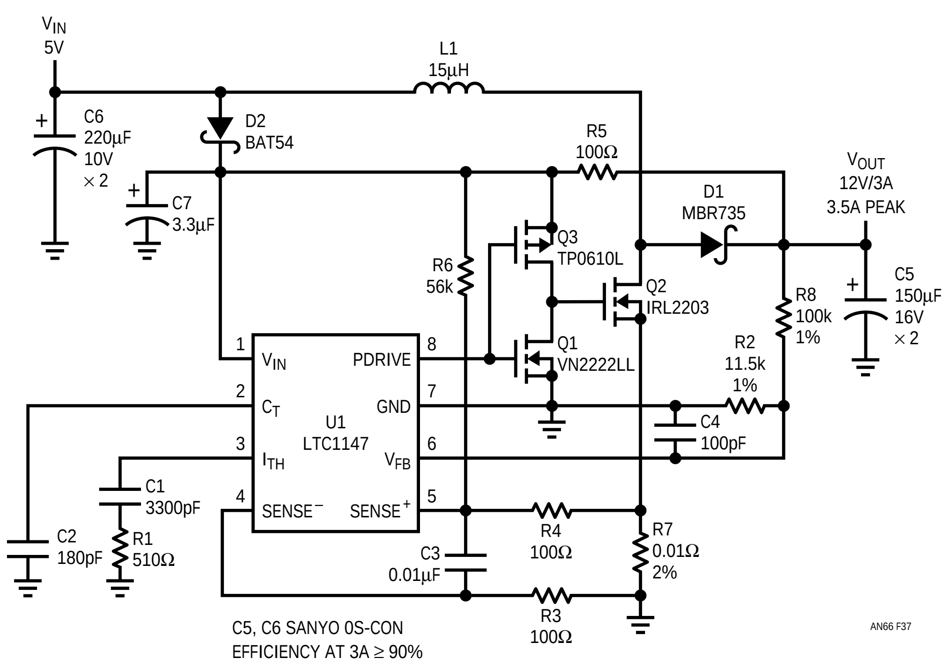


Figure 37. LTC1147-Based 5V to 12V Converter

architecture, eliminating the need for external slope compensation. Current mode control allows fast transient response and cycle-by-cycle current limiting. A maximum voltage of only 150mV across the current-sense resistor R7 optimizes performance for low input voltages.

When Q2 turns on, current starts building up in inductor L1. This provides a ramping voltage across R7. When this voltage reaches a threshold value set internally in the LTC1147, Q2 turns off and the energy stored in L1 is transferred to the output capacitor C5. Timing capacitor C2 sets the operating frequency. The controller is powered from the output through R5 providing 10V of gate drive for Q2. This reduces the MOSFET's ON resistance and allows efficiency to exceed 90% even at full load. The feedback network comprising R2 and R8 sets the output voltage. Current sense resistor R7 sets the maximum output current; it can be changed to meet different circuit requirements.

Regulators—Switching (Boost)

Low Power (<1A)

APPLICATIONS FOR THE LT1372 500kHz SWITCHING REGULATOR
by Bob Essaff

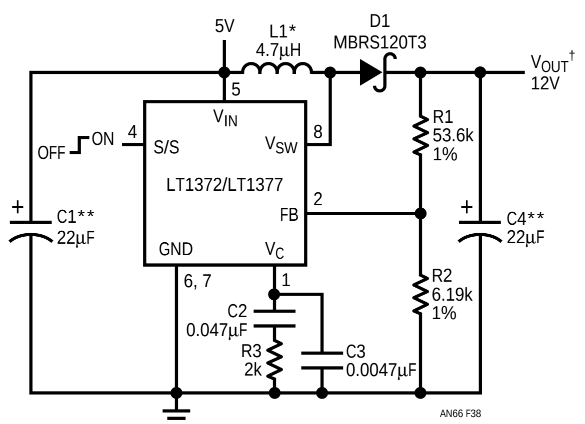
Boost Converter

The boost converter in Figure 38 shows a typical LT1372 application. This circuit converts an input voltage, which can vary from 2.7V to 11V, into a regulated 12V output. Using all surface mount components, the entire boost converter consumes only 0.5 square inches of board space. Figure 39 shows the circuit's efficiency, which can reach 89% on a 5V input.

The reference voltage on the FB pin is trimmed to 1.25V and the output voltage is set by the R1/R2 resistor divider ratio ($V_{OUT} = V_{REF} \bullet (R1/R2 + 1)$). R3 and C2 frequency compensate the circuit.

Positive-to-Negative Flyback with Direct Feedback

A unique feature of the LT1372 is its ability to directly regulate negative output voltages. As shown in the positive-to-negative flyback converter in Figure 40, only two resistors are required to set the output voltage. The reference voltage on the NFB pin is $-2V_{REF}$, making $V_{OUT} = -2V_{REF} \bullet (R2/R3 + 1)$. Efficiency for this circuit reaches 72% on a 5V input.



*COILCRAFT DO1608-472 (4.7µH) OR
COILCRAFT DT3316-103 (10µH) OR
SUMIDA CD43-4R7 (4.7µH) OR
SUMIDA CD73-100KC (10µH) OR
**AVX TPSD226M025R0200

†MAX I_{OUT}

L1	I_{OUT}
4.7µH	0.25A
10µH	0.35A

Figure 38. 5V to 12V Boost Converter

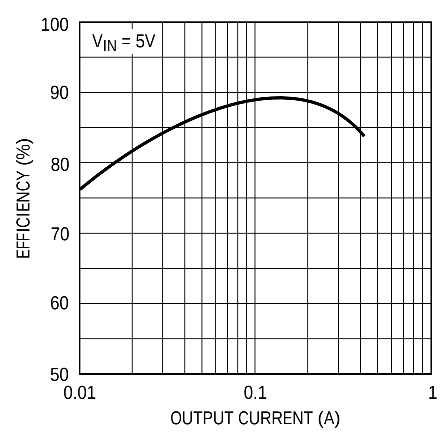


Figure 39. 12V Output Efficiency

Dual Output Flyback with Overvoltage Protection

Multiple-output flyback converters offer an economical means of producing multiple output voltages, but the power supply designer must be aware of cross regulation issues, which can cause electrical overstress on the supply and loads. Figure 41 is a dual-output flyback converter with overvoltage protection. Typically, in multiple-output flyback designs only one output is voltage sensed and regulated. The remaining outputs are "quasi-regulated" by the turns ratios of the transformer secondary. Cross regulation is a function of the transformer used and is a measure of how well the quasi-regulated outputs maintain regulation under varying load conditions. For evenly loaded outputs, as shown in Figure 42, cross regulation can be quite good, but when the loads differ greatly, as in the case of a load disconnect, there may be trouble. Figure 43 shows that when only the 15V output is voltage sensed, the –15V quasi-regulated output exceeds –25V when unloaded. This can cause electrical overstress on the output capacitor, output diode and the load when reconnected. Adding output voltage clamps is one way to fix the problem but the circuit in Figure 41 eliminates this requirement. This circuit senses both the 15V and –15V outputs and prevents either from going beyond its regulating value. Figure 44 shows the unloaded –15V output being held constant. The circuit's efficiency, which can reach 79% on a 5V input, is shown in Figure 45.

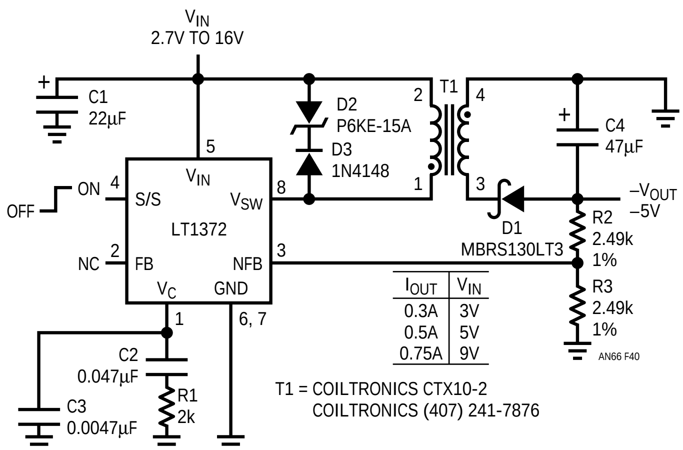


Figure 40. LT1372's Positive-to-Negative Converter with Direct Feedback

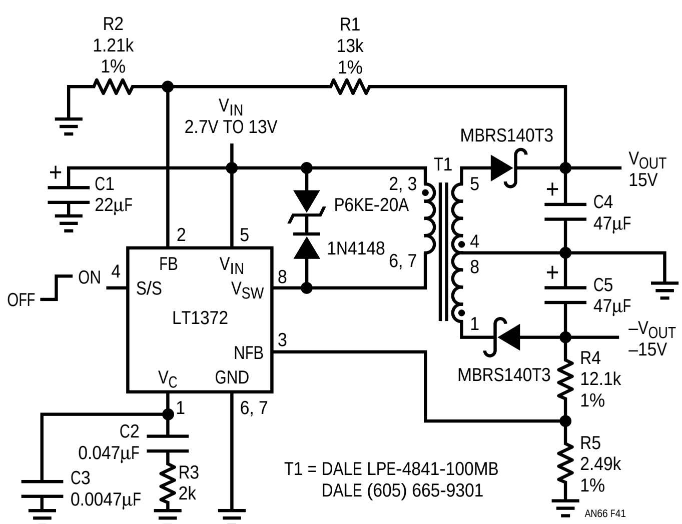


Figure 41. LT1372 Dual Output Flyback Converter with Overvoltage Protection

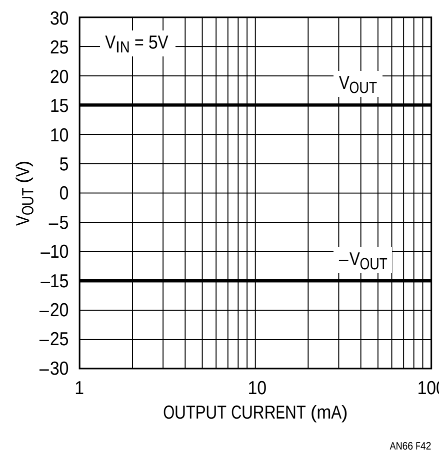


Figure 42. Cross Regulation of Figure 41's Circuit. V_{OUT} and $-V_{OUT}$ Evenly Loaded

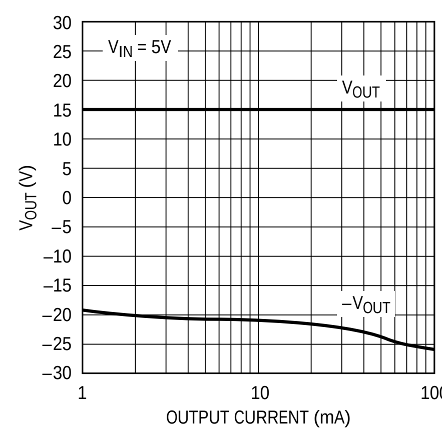


Figure 43. Cross Regulation of Figure 41's Circuit. $-V_{OUT}$ Unloaded; Only V_{OUT} Voltage Sensed

LINEAR TECHNOLOGY

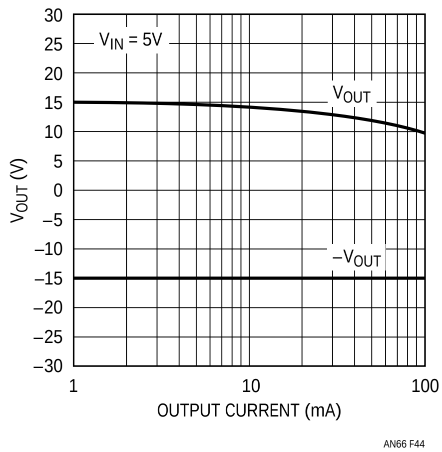


Figure 44. Cross Regulation of Figure 41's Circuit. $-V_{OUT}$ Unloaded; Both $-V_{OUT}$ and V_{OUT} Sensed

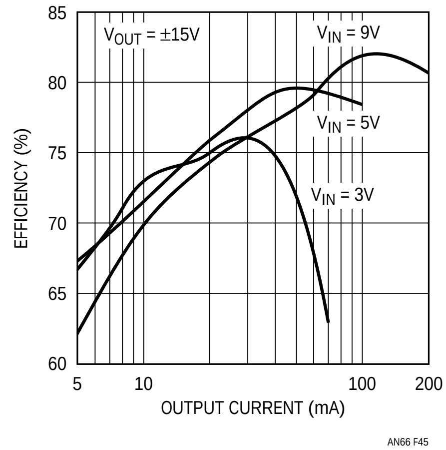


Figure 45. Efficiency of Dual Output Flyback Converter in Figure 41

Regulators—Switching (Buck/Boost)

±5V CONVERTER USES OFF-THE-SHELF SURFACE MOUNT COIL

By Mitchell Lee and Kevin Vasconcelos

Single-output switching regulator circuits can often be adapted to multiple output configurations with a minimum of changes, but these transformations usually call for custom wound inductors. A new series of standard inductors,[1] featuring quadrifilar windings, allows power supply designers to take advantage of these modified circuits but without the risks of a custom magnetics development program.

The circuit shown in Figure 46 fulfills a recent customer requirement for a 9V to 12V input, 5V/800mA and –5V/100mA output converter. It employs a 1:1 overwinding on what is ostensibly a buck converter to provide a –5V output. The optimum solution would be a bifilar wound coil with heavy gauge wire for the main 5V output and smaller wire for the overwinding. To avoid a custom coil design, an off-the-shelf JUMBO-PAC™ quadrifilar wound coil is used. This family of coils is wound with

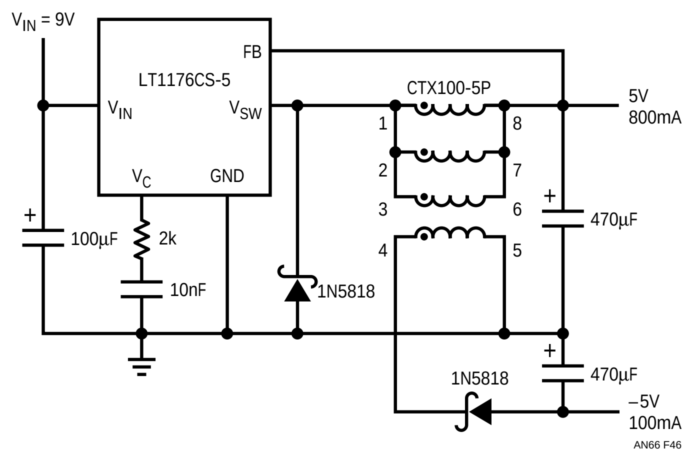


Figure 46. 5V Buck Converter with –5V Overwinding

1:1:1:1 sections. In the application of Figure 46, three sections are paralleled for the main 5V winding and the remaining section is used for the –5V output. This concentrates the copper where it is needed most—on the high current output.

Efficiency with the outputs loaded at 500mA and –50mA is over 80%. Minimum recommended load on the –5V output is 1mA to 2mA, and the –5V load current must always be less than the 5V load current.

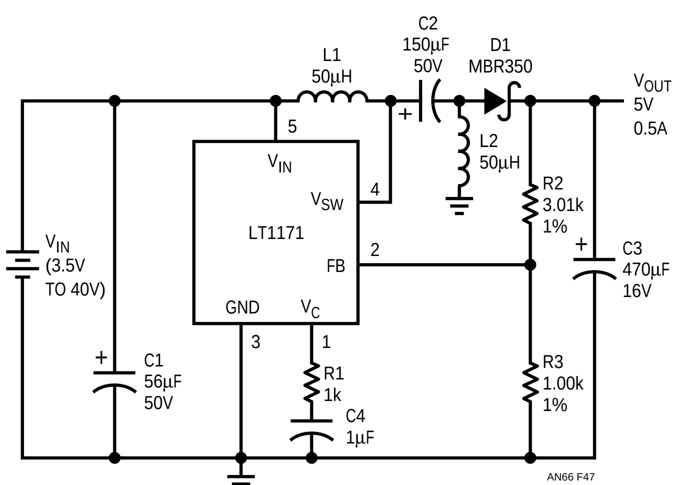
[1] JUMBO-PAC is a trademark of Coiltronics Inc. (407) 241-7876.

SWITCHING REGULATOR PROVIDES CONSTANT 5V OUTPUT FROM 3.5V TO 40V INPUT WITHOUT A TRANSFORMER

by Brian Huffman

A common switching regulator requirement is to produce a constant output voltage from an input voltage that varies above or below the output voltage. This is particularly important for extending battery life in battery-powered applications. Figure 47 shows how an LT1171 switching regulator IC, two inductors and a "flying" capacitor can generate a constant output voltage that is independent of input voltage variations. This is accomplished without the use of a transformer. Inductors are preferred over transformers because they are readily available and more economical.

The circuit in Figure 47 uses the LT1171 to control the output voltage. A fully self-contained switching regulator IC, the LT1171 contains a power switch as well as the control circuitry (pulse-width modulator, oscillator, reference voltage, error amplifier and protection circuitry). The power switch is an NPN transistor in a common-emitter configuration; when the switch turns on, the LT1171's V_{SW} pin is connected to ground. This power switch can handle peak switch currents of up to 2.5A.



C1 = NICHICON (AL) UPL1H560MEH, ESR = 0.250Ω, I_{RMS} = 360mA
C2 = NICHICON (AL) UPL1H151MPH, ESR = 0.100Ω, I_{RMS} = 820mA
C3 = NICHICON (AL) UPL1C471MPH, ESR = 0.090Ω, I_{RMS} = 770mA
L1, L2 = COILTRONICS CTX50-4, DCR = 0.090Ω,
COILTRONICS (407) 241-7876

EQUATION 1: $V_{OUT} = 1.25V (1 + R2/R3)$

Figure 47. LT1171 Provides Constant 5V Output from 3.5V to 40V Input. No Transformer Is Required

Figure 48 shows the operating waveforms for the circuit. In this architecture the capacitor C2 serves as the single energy transfer device between the input voltage and output voltage of the circuit. While the LT1171 power switch is off, diode D1 is forward biased, providing a path for the currents from inductors L1 and L2. Trace A shows inductor L1's current waveform and trace B is L2's current waveform. Observe that the inductor current waveforms occur on top of a DC level. The waveforms are virtually identical because the inductors have identical inductance values and the same voltages are applied across them. The current flowing through inductor L1 is not only delivered to the load but is also used to charge C2. C2 is charged to a potential equal to the input voltage.

When the LT1171 power switch turns on, the V_{SW} pin is pulled to ground and the input voltage is applied across the inductor L1. At the same time, capacitor C2 is connected across inductor L2. Current flows from the input voltage source through inductor L1 and into the LT1171. Trace C shows the voltage at the V_{SW} pin and Trace D is the current flowing through the power switch. The catch diode (D1) is reverse biased and capacitor C2's current also flows through the switch, through ground and into inductor L2. During this interval C2 transfers its stored energy into inductor L2. After the switch turns off the cycle is repeated.

Another advantage of this circuit is that it draws its input current in a triangular waveshape (see Trace A in Figure 48). The current waveshape of the input capacitor is identical to the current waveshape of inductor L1 except that the capacitor's current has no DC component. This type of ripple injects only a modest amount of noise into the input lines because the ripple does not contain any sharp edges.

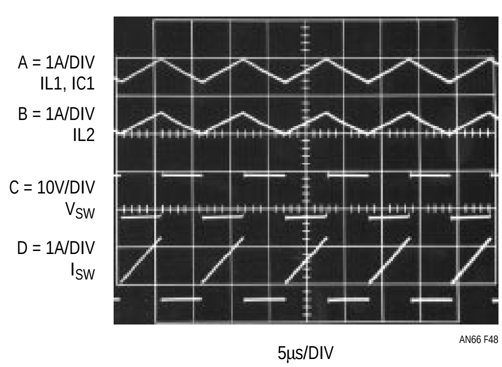


Figure 48. LT1171 Switching Waveforms

Figure 49 shows the efficiency of this circuit for a 0.5A load and maximum output current for various input voltages. The two main loss elements are the output diode (D1) and the LT1171 power switch. A Schottky diode is chosen for its low forward voltage drop; it introduces a 10% loss, which is relatively constant with input voltage variations. At low input voltages the efficiency drops because the LT1171 power switch's saturation voltage becomes a higher percentage of the available input supply.

This circuit can deliver an output current of 0.5A at a 3.5V input voltage. This rises to 1A as input voltage is increased. Above 20V, higher output currents can be achieved by increasing the values of inductors L1 and L2. Larger inductances store more energy, providing additional current to the load. If 0.5A of output current is insufficient, use a higher current part, such as the LT1170.

The output voltage is controlled by the LT1171 internal error amplifier. This error amplifier compares a fraction of the output voltage, via the R1 to R2 divider network shown in Figure 47, with an internal 1.25V reference voltage, and varies the duty cycle until the two values are equal. (The duty cycle is determined by multiplying the switch ON time by the switching frequency.) The RC network (R1 and C4 in Figure 47) connected to the V_C pin provides sufficient compensation to stabilize this control loop. Equation 1 (see Figure 47) can be used to determine the output voltage.

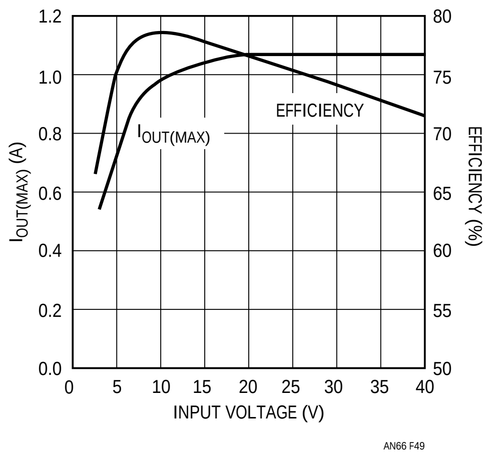


Figure 49. Efficiency and Load Characteristics for Various Input Voltages

SWITCHING REGULATOR PROVIDES ±15V OUTPUT FROM AN 8V TO 40V INPUT WITHOUT A TRANSFORMER

by Brian Huffman

Many systems derive ±15V supplies for analog circuitry from an input voltage that may be above or below the 15V output. The split supply requirement is usually fulfilled by a switcher with a multiple-secondary transformer or by multiple switchers. An alternative approach, shown in Figure 50, uses an LT1074 switching regulator IC, two inductors and a "flying" capacitor to generate a dual-output supply that accepts a wide range of input voltages. This solution is particularly noteworthy because it uses only one switching regulator IC and does not require a transformer. Inductors are preferred over transformers because they are readily available and more economical.

The operating waveforms for the circuit are shown in Figure 51. During the switching cycle, the LT1074's V_{SW} pin swings between the input voltage (V_{IN}) and the negative output voltage ($-V_{OUT}$). (The ability of the LT1074's V_{SW} pin to swing below ground is unusual—most other 5-pin buck switching regulator ICs cannot do this.) Trace A shows the waveform of the V_{SW} pin voltage and Trace B is the current flowing through the power switch.

While the LT1074 power switch is on, current flows from the input voltage source through the switch, through capacitor C2 and inductor L1 (Trace C), and into the load. A portion of the switch current also flows into inductor L2 (Trace D). This current is used to recharge C2 and C4 during the switch OFF time to a potential equal to the positive output voltage (V_{OUT}). The current waveforms for both inductors occur on top of a DC level.

The waveforms are virtually identical because the inductors have identical values and because the same voltage potentials are applied across them during the switching cycles.

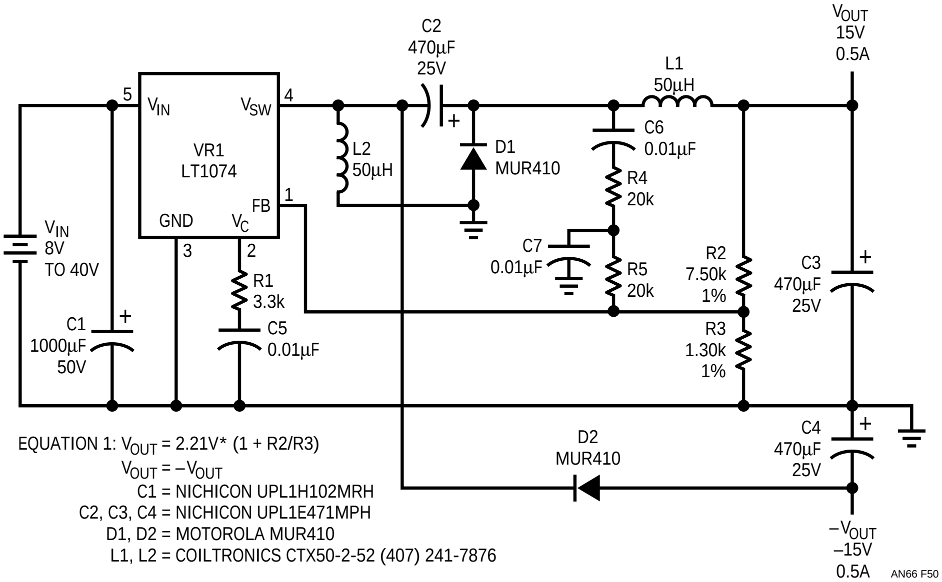


Figure 50. Schematic Diagram for ±15V Version

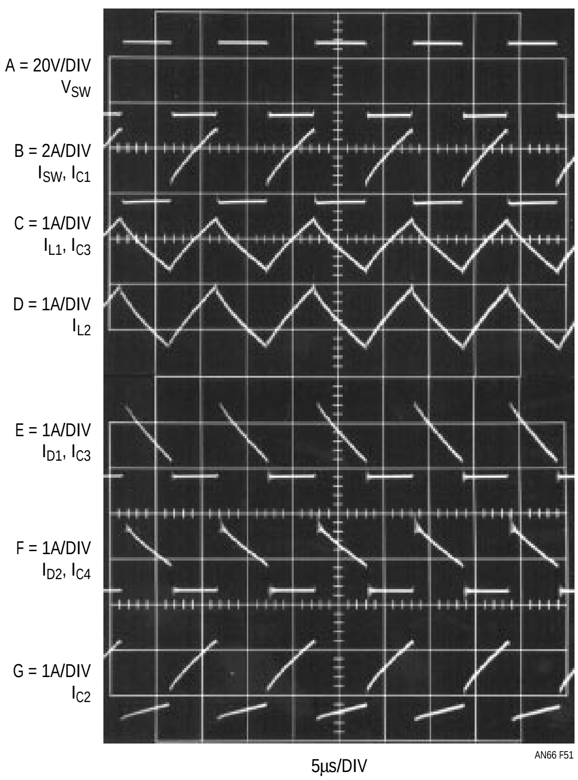


Figure 51. LT1074 Switching Waveforms

When the switch turns off, the current in L1 and L2 begins to ramp downward, causing the voltages across them to reverse polarity and forcing the voltage at the V_{SW} pin below ground. The V_{SW} pin voltage falls until diodes D1 (Trace E) and D2 (Trace F) are forward biased. During this interval the voltage on the V_{SW} pin is equal to a diode drop below the negative output voltage ($–V_{OUT}$). L2's current then circulates between both D1 and D2, charging C2 and C4. The energy stored in L1 is used to replace the energy lost by C2 and C4 during the switch ON time. Trace G is capacitor C2's current waveform. Capacitor C4's current waveform (Trace F) is the same as diode D2's current less the DC component. Assuming that the forward voltage drops of diodes D1 and D2 are equal, the negative output voltage ($–V_{OUT}$) will be equal to the positive output voltage (V_{OUT}). After the switch turns on again the cycle is repeated.

Figure 52 shows the excellent regulation of the negative output voltage for various output currents. The negative

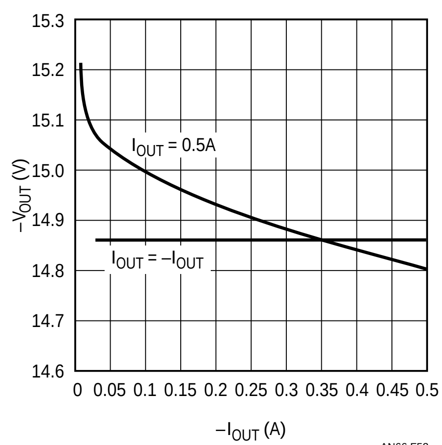


Figure 52. –15V Output Regulation Characteristics

output voltage tracks the positive supply (V_{OUT}) within 200mV for load variations from 50mA to 500mA. Negative output load current should not exceed the positive output load by more than a factor of 4; the imbalance causes loop instabilities. For common load conditions the two output voltages track each other perfectly.

Another advantage of this circuit is that inductor L1 acts as both an energy storage element and as a smoothing filter for the positive output (V_{OUT}). The output ripple voltage has a triangular waveshape whose amplitude is determined by the inductor ripple current (see trace C of Figure 51) and the ESR (effective series resistance) of the output capacitor (C3). This type of ripple is usually small so a post filter is not necessary.

Figure 53 shows the efficiency for a 0.5A common load at various input voltages. The two main loss elements are the output diodes (D1 and D2) and the LT1074 power switch. At low input voltages, the efficiency drops because the switch's saturation voltage becomes a higher percentage of the available input supply.

The output voltage is controlled by the LT1074 internal error amplifier. This error amplifier compares a fraction of the output voltage, via the R2 to R3 divider network shown in Figure 50, with an internal 2.21V reference voltage and then varies the duty cycle until the two values are equal. The RC network (R1 and C5 in Figure 50) connected to the V_C pin along with the R4/R5 and C6/C7 network provides sufficient compensation to stabilize the control loop. Equation 1 can be used to determine the output voltage.

Figure 54 shows the circuit's –5V load regulation characteristics and Figure 55 shows its efficiency.

Refer to the schematic diagram in Figure 56 for modified component values to provide ±5V at 1A.

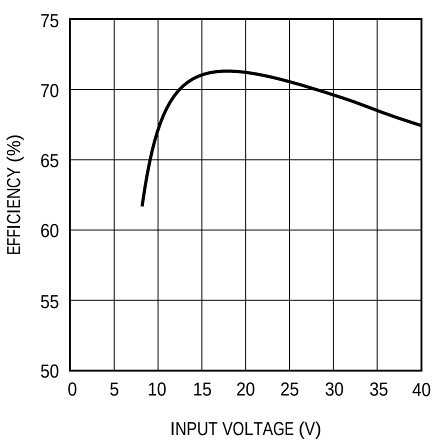


Figure 53. ±15V Efficiency Characteristics with 0.5A Common Load

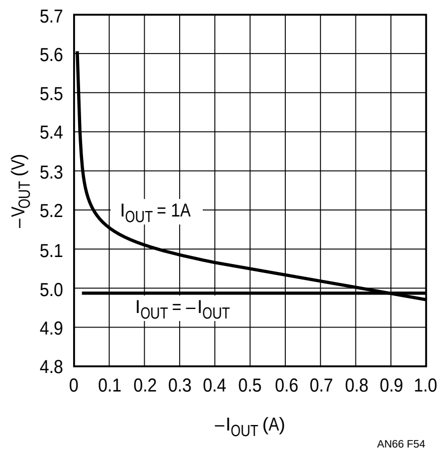


Figure 54. –5V Output Regulation Characteristics

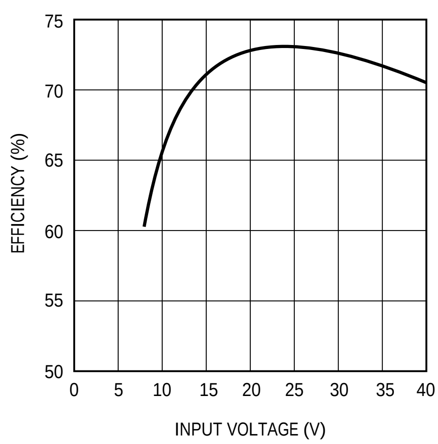


Figure 55. ±5V Efficiency Characteristics with 1A Common Load

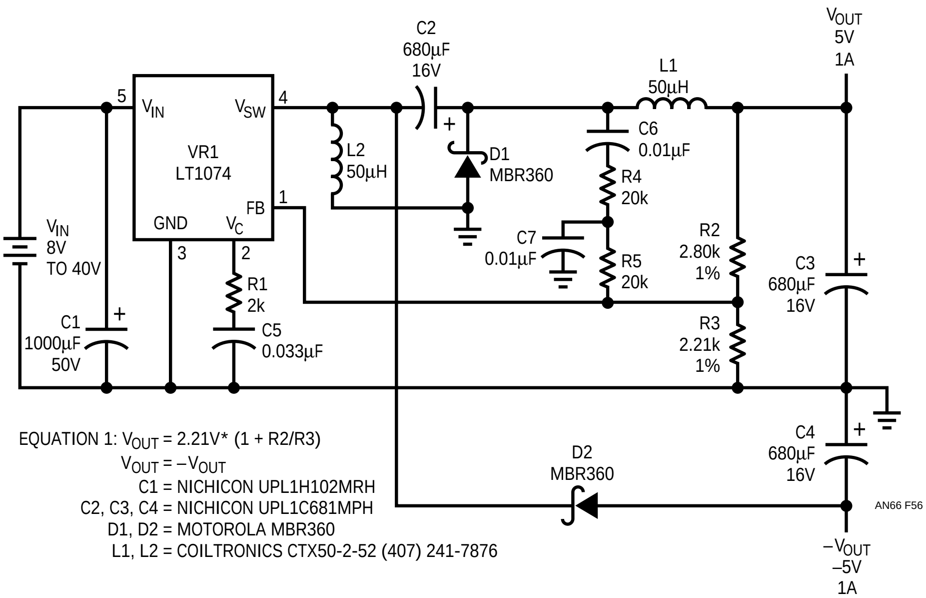


Figure 56. Schematic Diagram for ±5V Version

Regulators—Switching (Inverting)

HIGH EFFICIENCY 12V TO –12V CONVERTER

by Milton Wilcox and Christophe Franklin

It is difficult to obtain high efficiencies from inverting switching regulators because the peak switch and inductor currents must be roughly twice the output current. Furthermore, the switch node must swing twice the input voltage (24V for a 12V inverting converter). The adjustable version of the LTC1159 synchronous stepdown controller is ideally suited for this application, producing a combination of better than 80% efficiency, low quiescent current and 20µA shutdown current.

The 1A circuit shown in Figure 57 exploits the high input-voltage capability of the LTC1159 by connecting the controller ground pins to the –12V output. This allows the simple feedback divider between ground and the output (comprising R1 and R2) to set the regulated voltage, since the internal 1.25V reference rides on the negative output. The inductor connects to ground via the 0.05Ω current-sense resistor.

A unique EXT V_{CC} pin on the LTC1159 allows the MOSFET drivers and control circuitry to be powered from the output of the regulator. In Figure 57 this is accomplished by grounding EXT V_{CC}, placing the entire 12V output voltage across the driver and control circuits (remember the ground pins are at –12V). This is permissible with the LTC1159, which allows a maximum of 13V between the Sense and Ground pins. During start-up or short-circuit conditions, operating power is supplied by an internal 4.5V low dropout linear regulator. This start-up regulator automatically turns off when the output falls below –4.5V.

A cycle of operation begins when Q1 turns on, placing the 12V input across the inductor. This causes the inductor current to ramp to a level set by the error amplifier in the LTC1159. Q1 then turns off and Q2 turns on, causing the current stored in the inductor to flow to the –12V output. At the end of the 5µs off-time (set by capacitor C_T), Q2 turns off and Q1 resumes conduction. With a 12V input the duty cycle is 50%, resulting in a 100kHz operating frequency.

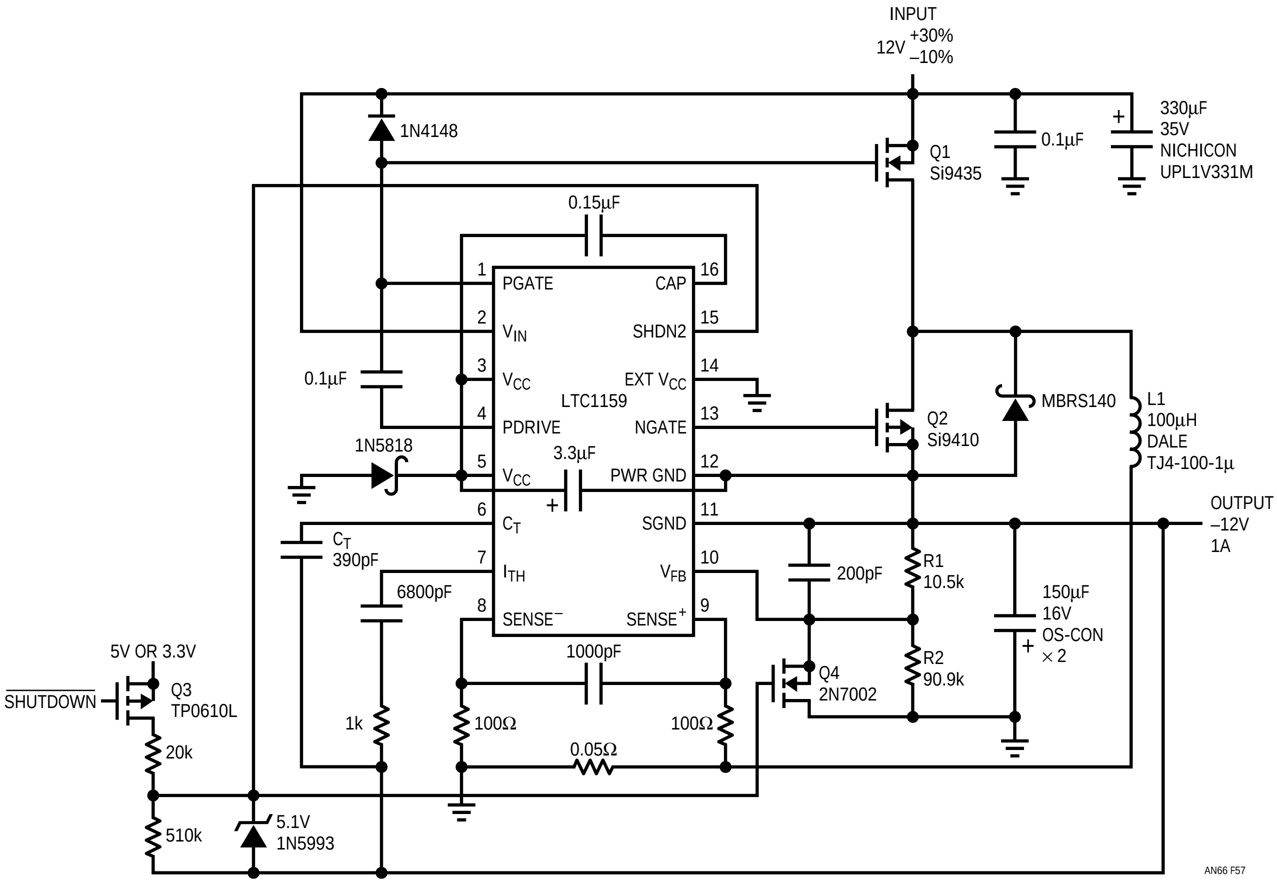


Figure 57. LTC1159 Converts 12V to –12V at 1A

The LTC1159, like other members of the LTC1148 family, automatically switches to Burst Mode operation at low output currents. Figure 57's circuit enters Burst Mode operation below approximately 200mA of load current. This maintains operating efficiencies exceeding 65% over two decades of load current range, as shown in Figure 58. Quiescent current (measured with no load) is 1.8mA.

Complete shutdown is achieved by pulling the gate of Q3 low. Q3, which can be interfaced to either 3.3V or 5V logic, creates a 5V shutdown signal referenced to the negative output voltage to activate the LTC1159 Shutdown 2 pin. Additionally, Q4 offsets the V_{FB} pin to ensure that Q1 and Q2 remain off during the entire shutdown sequence. In shutdown conditions, 40µA flows in Q3 and only 20µA is taken from the 12V input.

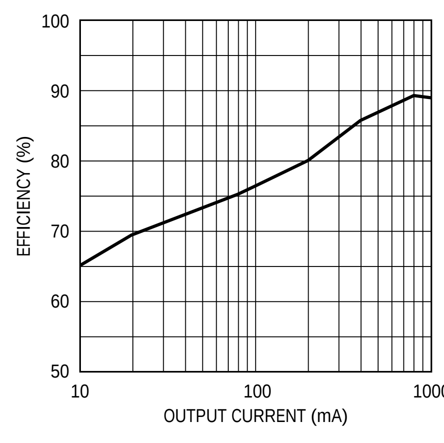


Figure 58. Efficiency Plot of Figure 57's Circuit

REGULATED CHARGE PUMP POWER SUPPLY

by Tommy Wu

The circuit shown in Figure 59 uses an LTC1044A charge pump inverter to convert a 5V input to a –1.7V potential as required for a certain LCD panel. Output regulation is provided by a novel feedback scheme, which uses components Q1, R1 and R2. Without feedback the charge pump would simply develop approximately –5V at its output. With feedback applied, V_{OUT} charges in the negative direction until the emitter of Q1 is biased by the divider comprising R1 and R2. Current flowing in the collector tends to slow the LTC1044A's internal oscillator, reducing the available output current. The output is thereby maintained at a constant voltage.

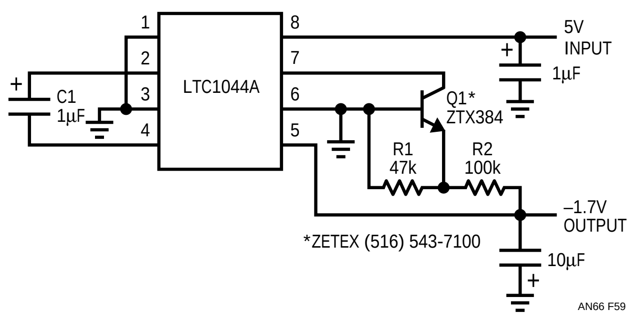


Figure 59. Regulated Charge Pump

In this application less than 5mA output current is required. As a result, charge pump capacitor C1 is reduced to 1µF from the usual 10µF. Curves of output voltage with and without feedback are shown in Figure 60. The equivalent output impedance of the charge pump is reduced from approximately 100Ω to 5Ω.

A variety of output voltages within the limits of the curve in Figure 60 can be set by simply adjusting the V_{BE} multiplier action of Q1, R1 and R2. Tighter regulation or a higher tolerance could be obtained by adding a reference or additional gain, at the expense of increased complexity and cost.

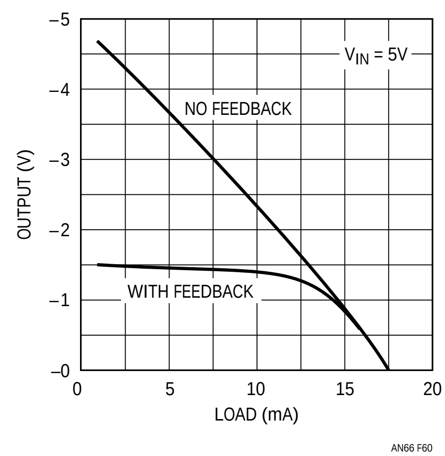


Figure 60. Effect of Feedback on Output Voltage

LTC1174: A HIGH EFFICIENCY BUCK CONVERTER
by San-Hwa Chee and Randy Flatness

The LTC1174 is an 8-pin SO "user-friendly" step-down converter. (A PDIP package is also available.) Only four external components are needed to construct a complete high efficiency converter. With no load it requires only 130µA of quiescent current; this decreases to a mere 1µA upon shutdown. The LTC1174 is protected against output shorts by an internal current limit, which is pin selectable to either 340mA or 600mA. This current limit also sets the inductor's peak current. This allows the user to optimize the converter's efficiency depending upon the output current requirement.

In dropout conditions, the internal 0.9Ω (at a supply voltage of 9V) power P-channel MOSFET switch is turned on continuously (DC), thereby maximizing the life of the battery source. (Who says a switcher has to switch?) In addition to the features already mentioned, the LTC1174 boasts a low-battery detector. All versions function down to an input voltage of 4V and work up to an absolute maximum of 13.5V. For extended input voltage, high voltage parts are also available that can operate up to an absolute maximum of 18.5V.

5V Output Applications

Figure 61 shows a practical LTC1174-5 circuit with a minimum of components. Efficiency curves for this circuit at two different input voltages are shown in Figure 62. Note that the efficiency is 94% at a supply voltage of 6V and load current of 175mA. This makes the LTC1174 attractive to all power sensitive applications and shows clearly why switching regulators are gaining dominance over linear regulators in battery-powered devices.

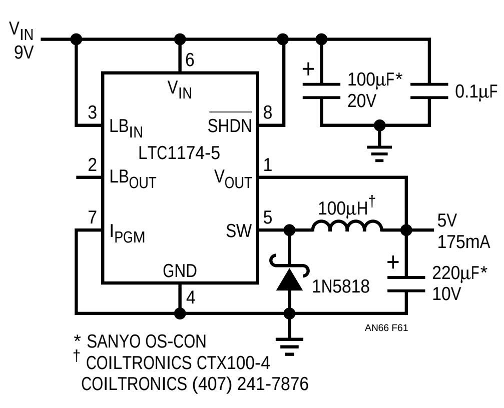


Figure 61. Typical Application for Low Output Currents

If higher output currents are desired Pin 7 (I_{PGM}) can be connected to V_{IN}. Under this condition the maximum load current is increased to 450mA. The resulting circuit and efficiency curves are shown in Figures 63 and 64 respectively.

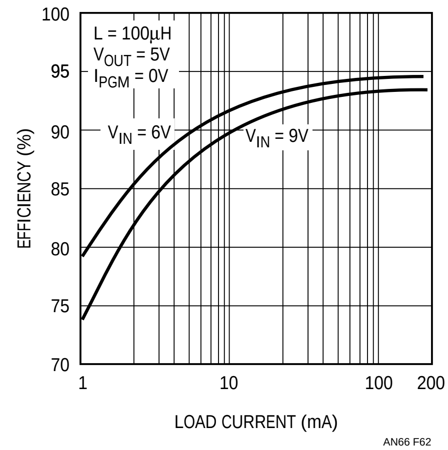


Figure 62. Efficiency vs Load Current

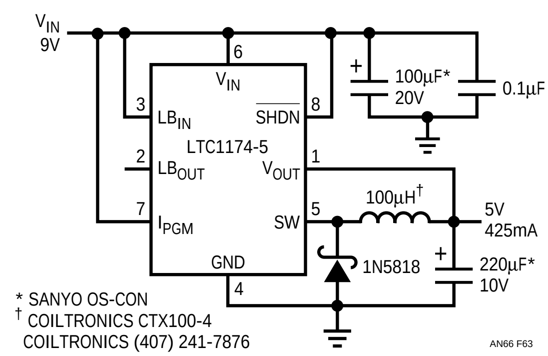


Figure 63. Typical Application for Higher Output Currents

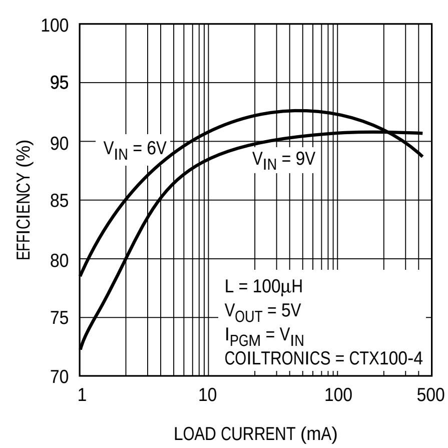


Figure 64. Efficiency vs Load Current

More Applications

Positive-to-Negative Converter

The LTC1174 can easily be set up for a negative output voltage. The LTC1174-5 is ideal for –5V outputs as this configuration requires the fewest components. Figure 65 shows the schematic for this application with low-battery detection capability. The LED will turn on at input voltages below 4.9V. The efficiency of this circuit is 81% at an input voltage of 5V and output current of 150mA.

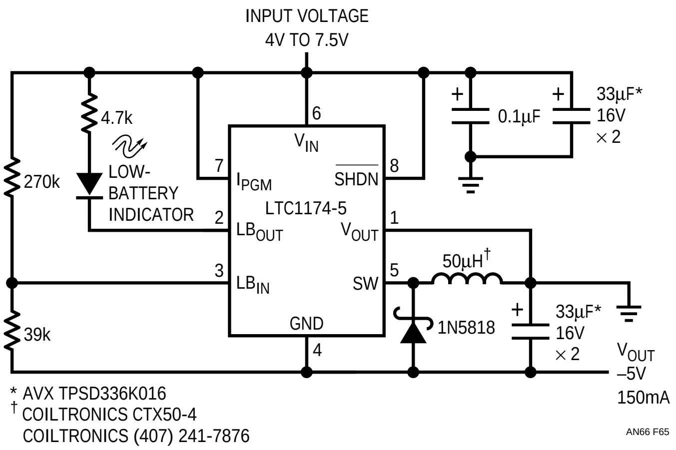


Figure 65. Positive to –5V Converter with Low-Battery Detection

A 5V to 3.3V Converter

The LTC1174-3.3 is ideal for applications that require 3.3V at less than 450mA. A minimum board area surface mount 3.3V regulator is shown in Figure 66. Figure 67 shows that this circuit can achieve efficiency greater than 85% for load currents between 5mA and 450mA.

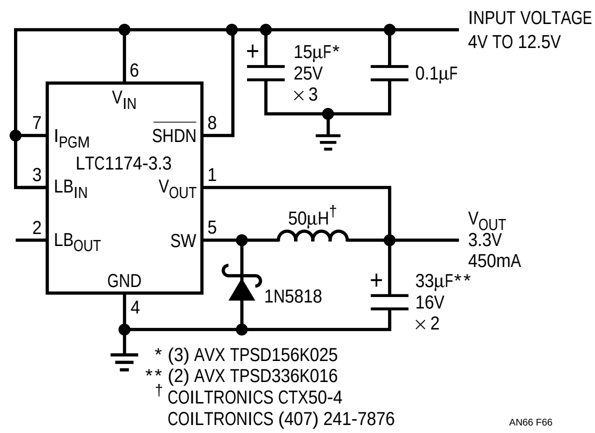


Figure 66. 5V to 3.3V Output Application

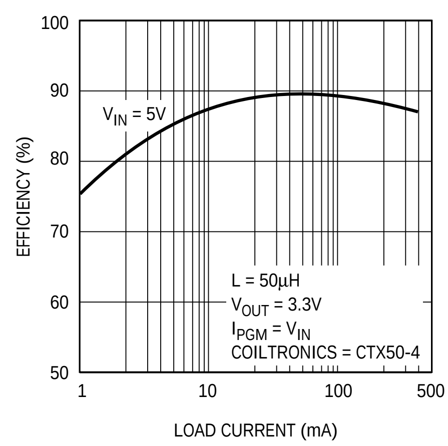


Figure 67. Efficiency vs Load Current

Regulators—Switching (Power Factor Corrected)

THE NEW LT1508/LT1509 COMBINES POWER FACTOR CORRECTION AND A PWM IN A SINGLE PACKAGE

by Kurk Mathews

Typical Application

Figure 68 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 nondissipative snubber, 500VDC switches and automatic core reset guaranteed by the LT1509's 50% maximum duty-cycle limitation. An LT1431 and inexpensive optoisolation are used to close the loop conservatively at 3kHz with excess phase margin (see Figure 69). Figure 70 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 30W, 150W and 300W are shown in Figure 71. The PFC preregulator alone has efficiency numbers of between about 87% and 97% over line and load.

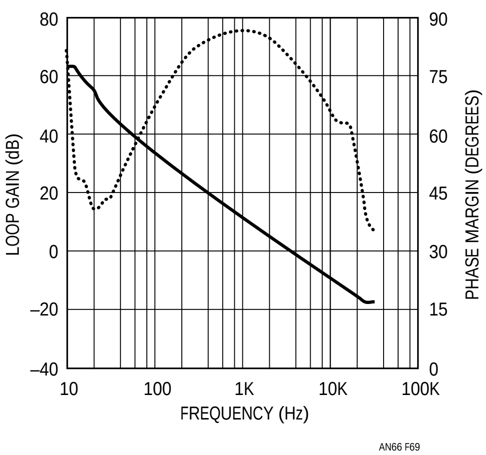


Figure 69. Bode Plot ot the Circuit Shown in Figure 68

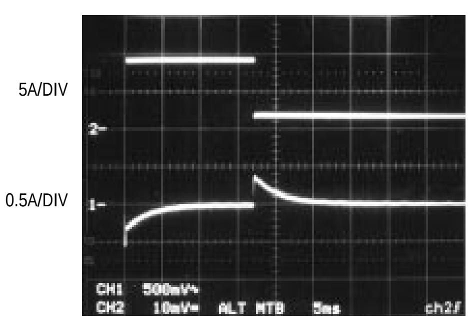


Figure 70. Figure 68's Response to a 2A to ≈10A Load Step

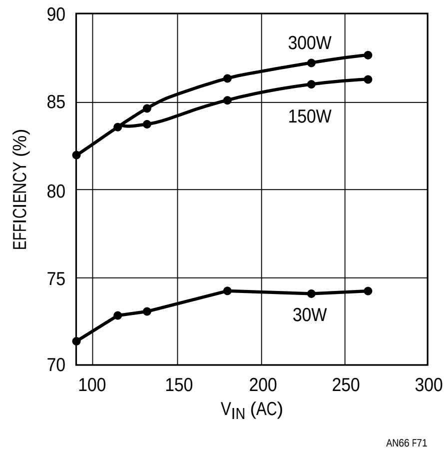


Figure 71. Efficiency Curves for Figure 68's Circuit

Start-up of the circuit begins with the LT1509's V_{CC} bypass capacitors trickle charging through 91kΩ to 16VDC, overcoming the chip's 0.25mA typical start-up current ($V_{CC} \leq$ 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 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 39mA to cover the gate current of the three FETs and high side transformer losses. A 0.15Ω sense resistor senses input current and compares it to a reference current (I_M) created by the outer voltage loop and multiplier. Thus, the input current follows the input line voltage and changes, as necessary, in order to maintain a constant bank voltage. The forward converter sees a voltage input of 382VDC unless the line voltage drops out, in which case the 470µF 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 turn-on spike to the end of the built-in blanking time, necessitating the external blanking transistor.

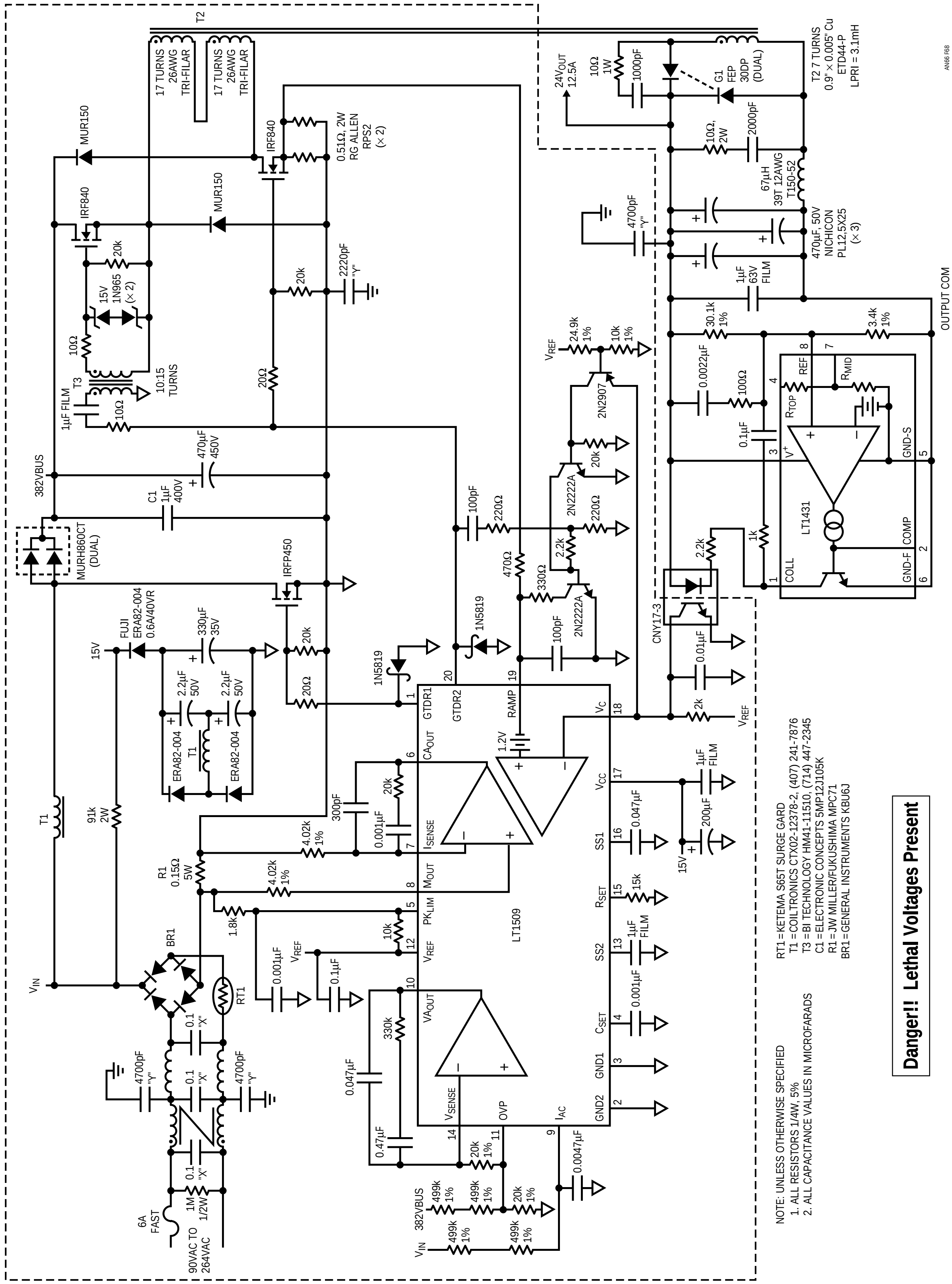


Figure 68. Schematic Diagram of 300W 24VDC Output Power Factor Corrected Universal Input Supply

Regulators—Switching (Discussion)

ADDING FEATURES TO THE BOOST TOPOLOGY

by Dimitry Goder

A boost-topology switching regulator is the simplest solution for converting a 2- or 3-cell input to a 5V output. Unfortunately, boost regulators have some inherent disadvantages, including no short-circuit protection and no shutdown capability. In some battery-operated products, external chargers or adapters can raise the battery voltage to a potential higher than the 5V output. Under this condition a boost converter cannot maintain regulation—the high input voltage feeds through the diode to the output.

The circuit shown in Figure 72 overcomes these problems. An LT1301 is used as a conventional boost converter, preserving simplicity and high efficiency in the boost mode. Transistor Q1 adds short-circuit limiting, true shutdown and regulation when there is a high input voltage.

When the input voltage is lower than 4V and the regulator is enabled, Q1's emitter is driven above its base, saturating the transistor. As a result, the voltages on C1 and C2 are roughly the same and the circuit operates as a conventional boost regulator.

If the input voltage increases above 4V, the internal error amplifier, acting to keep the output at 5V, boosts the voltage on C1 to a level greater than 1V above the input. This voltage controls Q1 to provide the desired output with the transistor operating as a linear pass element. The output does not change abruptly during the switch-over between step-up and step-down modes because it is monitored in both modes by the same error amplifier.

Figure 73 shows efficiency versus input voltage for 5V/100mA output. The break point at 4.25V is evidence of Q1 beginning to operate in a linear mode with an attendant roll-off of efficiency. Below 4.25V the circuit operates as a boost regulator and maintains high efficiency across a broad range of input voltages.

The circuit can be shut down by pulling the LT1301's Shutdown pin high. The LT1301 ceases switching and Q1 automatically turns off, fully disconnecting the output. This stays true over the entire input voltage range.

Q1 also provides overload protection. When the output is shorted the LT1301 operates in a cycle-by-cycle current limit. The short-circuit current depends on the maximum switch current of the LT1301 and on the Q1's gain, typically reaching 200mA. The transistor can withstand overload for several seconds before heating up. For sustained faults the thermal effects on Q1 should be carefully considered.

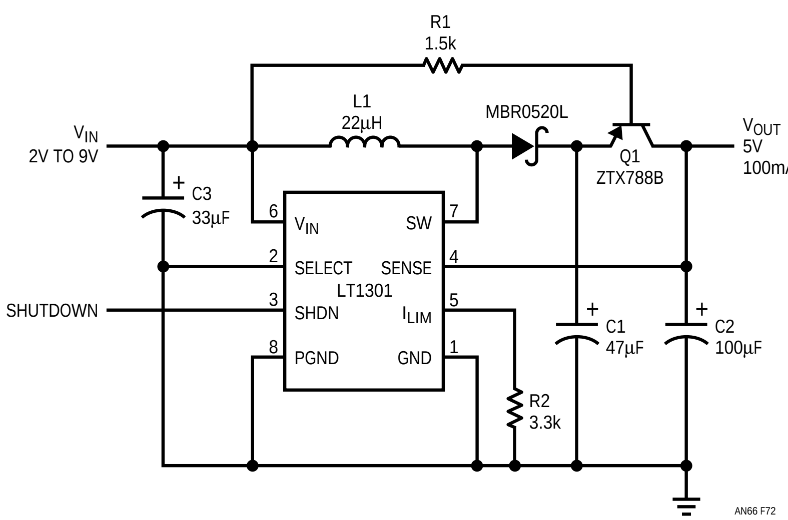


Figure 72. Q1 Adds Short-Circuit Limiting, True Shutdown and Regulation When There Is a High Input Voltage to the LT1301 in Boost Mode

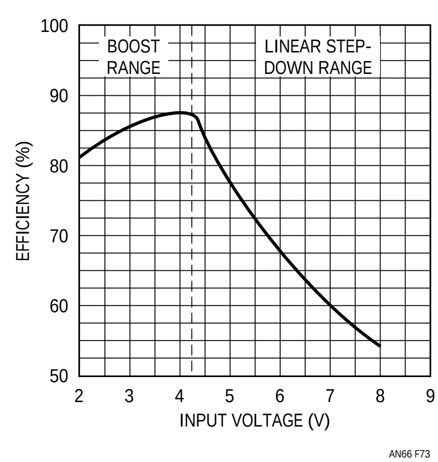


Figure 73. Efficiency vs Input Voltage for 5V/100mA Output

SENSING NEGATIVE OUTPUTS

by Dimitry Goder

Various switching regulator circuits exist to provide positive-to-negative conversion. Unfortunately, most controllers cannot sense the negative output directly; they require a positive feedback signal derived from the negative output. This creates a problem. The circuit presented in Figure 74 provides an easy solution.

The LT1172 is a versatile switching regulator that contains an onboard 100kHz PWM controller and a power switching transistor. Figure 74 shows the LTC1172 configured to provide a negative output using a popular charge pump technique. When the switch turns on, current builds up in the inductor. At the same time the charge on C3 is transferred to output capacitor C4. During the switch off-time, energy stored in the inductor charges capacitor C3. A special DC level-shifting feedback circuit consisting of Q1, Q2, and R1 to R4 senses the negative output.

Under normal conditions Q1's base is biased at a level about 0.6V above ground and the current through resistor R3 is set by the output voltage. If we assume that the base current is negligible, then R3's current also flows through R2, biasing Q2's collector at a positive voltage proportional to the negative output.

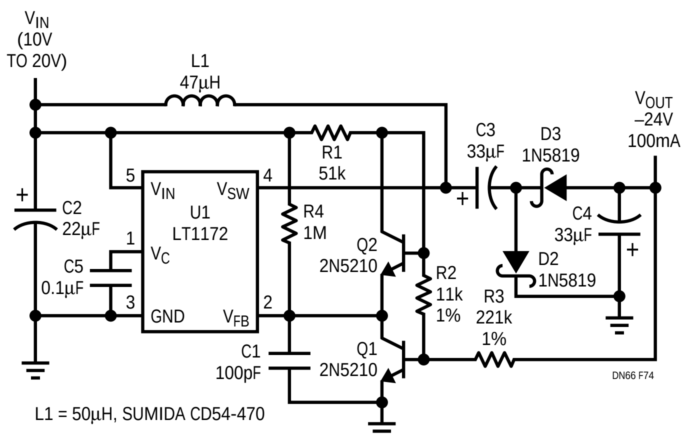


Figure 74. 10V/20V to –24 Converter

Q2 is connected as a diode and is used to compensate for Q1's base-emitter voltage change with temperature and collector current. Both transistors see the same collector current and their base-emitter voltages track quite well. Because the base-emitter voltages cancel, the voltage across R2 also appears on the LT1172's Feedback pin.

The resulting output voltage is given by the following formula:

$$V_{OUT} = V_{FB}\frac{R3}{R2} - V_{BE}$$

where V_{FB} is the LT1172 internal 1.244V reference and V_{BE} is Q1's base/emitter voltage (≈ 0.6V). The V_{BE} term in the equation denotes a minor output voltage dependency on input voltage and temperature. However, the variation due to this factor is usually well below 1%.

Essentially, Q1 holds its collector voltage constant by changing its collector current and will function properly as long as some collector current exists. This puts the following limitation on R1: at minimum input voltage the current through R1 must exceed the current through R2. This is reflected by the following inequality:

$$R1 < R2\frac{V_{IN(MIN)} - V_{FB} - V_{BE}}{V_{FB}}$$

If the input voltage drops below the specified limit (e.g., under a slow start-up condition) and Q1 turns off, R4 provides the LT1172 Feedback pin with a positive bias and the output voltage decreases. Without R4 the Feedback pin would not get an adequate positive signal, forcing the LT1172 to provide excessive output voltage and resulting in possible circuit damage.

The feedback configuration described above is simple yet very versatile. Only resistor value changes are required for the circuit to accommodate a variety of input and output voltages. Exactly the same feedback technique can be used with flyback, "Cuk" or inverting topologies, or whenever it is necessary to sense a negative output.

Regulators—Switching (Micropower)

3-CELL TO 3.3V BUCK/BOOST CONVERTER

by Dimitry Goder

Obtaining 3.3V from three 1.2V (nominal) cells is not a straightforward task. Since battery voltage can be either below or above the output, common step-up or step-down converters are inadequate. Alternatives include using more complex switching topologies, such as SEPIC, or a switching boost regulator plus a series linear pass element. Figure 75 presents an elegant implementation of the latter approach.

The LT1303 is a Burst Mode switching regulator that contains control circuitry, an onboard power transistor and a gain block. When the input voltage is below the output, U1 starts switching and boosts the voltage across C2 and C3 to 3.3V. The gain block turns on Q1 because the feedback network R3 to R5 biases the low-battery comparator input (LBI) 20mV below the reference. In this mode the circuit operates as a conventional boost converter, sensing output voltage at the FB pin.

When the input voltage increases, it eventually reaches a point where the regulator ceases switching and the input voltage is passed unchanged to capacitor C2. The output voltage rises until the LBI input reaches the reference voltage of 1.25V, at which point Q1 starts operating as a series pass element. In these conditions the circuit functions as a linear regulator with the attending efficiency roll-off at higher input voltages.

For input voltages derived from three NiCd or NiMH cells, the circuit described provides excellent efficiency and the longest battery life. At 3.6V, where the battery spends most of its life, efficiency exceeds 91%, leaving all alternative topologies far behind.

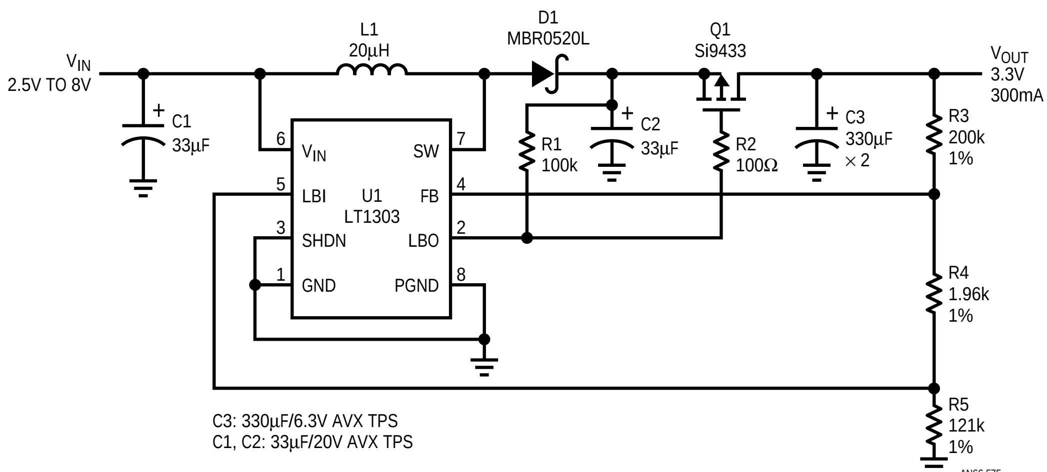


Figure 75. 3-Cell to 3.3V Buck/Boost Converter

LT1111 ISOLATED 5V SWITCHING POWER SUPPLY

by Kevin R. Hoskins

Circuit Description

Many applications require isolated power supplies. Examples include remote sensing, measurement of signals riding on high voltages, remote battery-powered equipment, elimination of ground loops and data acquisition systems where noise elimination is vital. In each situation the isolated circuitry needs a floating power source. In some cases batteries or an AC line transformer can be used for power. Alternately, the DC/DC converter shown here creates an accurately regulated, isolated output from a 5V source. Moreover, it eliminates the optoisolator feedback arrangements normally associated with fully isolated converters.

Figure 76 shows a switching power supply that generates an isolated and accurately regulated 5V at 100mA output. The circuit consists of an LT1111, configured as a flyback converter, followed by an LT1121 low dropout, micropower linear regulator. An LTC1145 (winner of EDN's IC Innovation of the Year Award) provides micropower isolated feedback.

The LT1111 is a micropower device that operates on only 400µA (max). This micropower operation is important for energy-conscious applications. It works well with surface mount inductors such as the Coiltronics OCTA-PAC shown in the schematic. Although the LT1111's internal power switch handles up to 1A, a 100Ω resistor (R1) limits the peak switch current to approximately 650mA. This maximizes converter efficiency. One side benefit of limiting the peak switch current is that the circuit becomes insensitive to inductance. The circuit operates satisfactorily with an inductance in the range of 20µH to 50µH.

It is important that capacitor C2 have low effective series resistance (ESR) and inductance (ESL) to minimize output ripple voltage. Although aluminum capacitors are abundant and inexpensive, they will perform poorly in this switcher application because of their relatively high ESR and ESL. Two good choices that meet C2's low ESR and ESL requirements are the AVX TPS and Sanyo OS-CON™ capacitor series.

Circuit Operation

The LT1111 is configured to operate as a flyback converter. The voltage on the transformer's secondary is rectified by D2, filtered by C2 and applied to the LT1121's input. As the LT1121's input voltage continues to rise, its output will regulate at 5V. The LT1121's input voltage continues increasing until the differential between input and output equals approximately 600mV. At this point Q1 begins conducting, turning on the LTC1145 isolator. The output of the LTC1145 goes high, turning off the converter. The feedback from the LTC1145 gates the LT1111's oscillator, controlling the energy transmitted to the transformer's secondary and the LT1121's input voltage. The oscillator is gated on for longer periods as the LT1121's load current increases. Q1's gain and the feedback through the LTC1145 force the converter loop to maintain the LT1121 just above dropout, resulting in the best efficiency. The LT1121 provides current limiting as well as a tightly regulated low noise output.

OS-CON is a trademark of SANYO Electric Co., LTD.

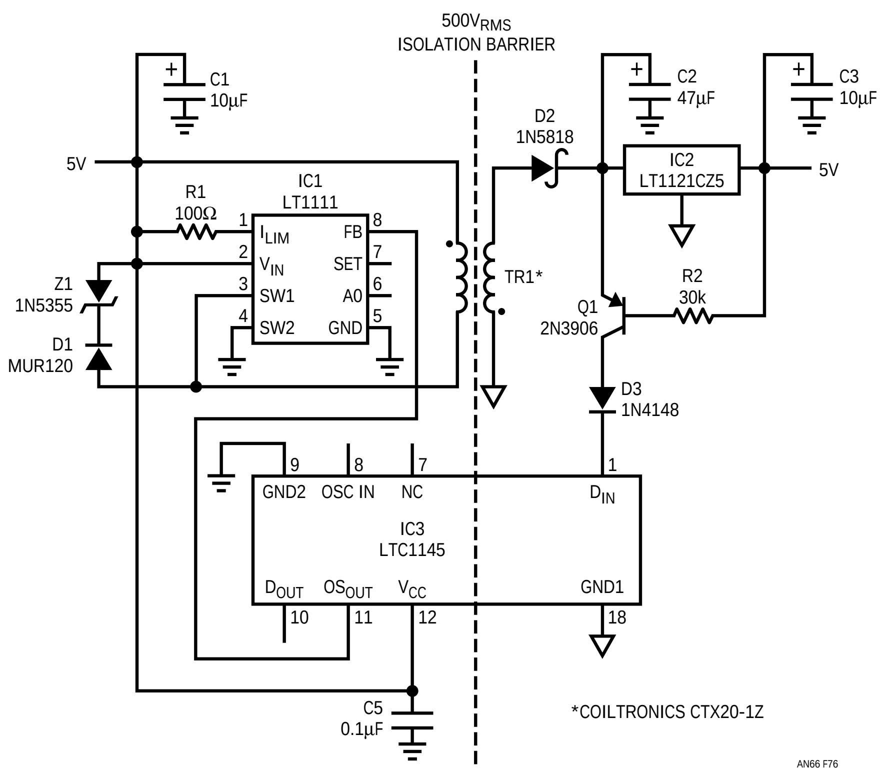


Figure 76. Circuit Generates Isolated, Regulated 5V at 100mA

LOW NOISE PORTABLE COMMUNICATIONS DC/DC CONVERTER
by Mitchell Lee

Portable communications products pack plenty of parts into close proximity. Digital clock noise must be eliminated not only from the audio sections but also from the antenna, which, by the very nature of the product, is located only inches from active circuitry. If a switching regulator is used in the power supply, it becomes another potential source of noise. The LTC1174 stepdown converter is designed specifically to eliminate noise at audio frequencies while maintaining high efficiency at low output currents.

Figure 77 shows an all surface mount solution for a 5V, 120mA output derived from five to seven NiCd or NiMH cells. Small input and output capacitors are used to conserve space without sacrificing reliability. In applications where it is desired, a shutdown feature is available; otherwise, short this pin to V_{IN}.

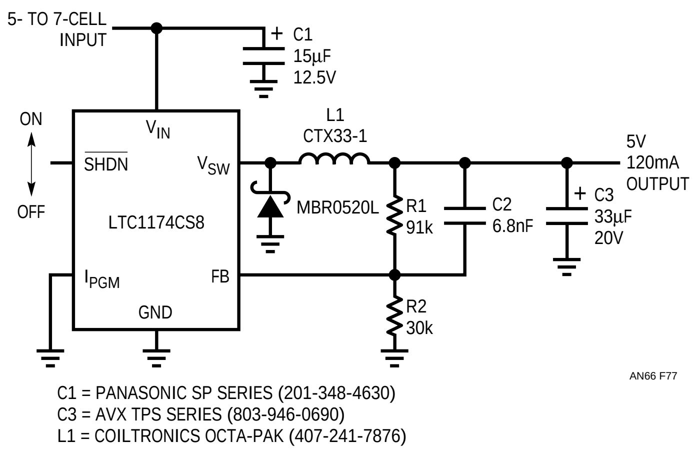


Figure 77. Low Noise, High Efficiency Step-Down Regulator for Personal Communications Devices

The LTC1174's internal switch, which is connected between V_{IN} and V_{SW}, is current controlled at a peak threshold of approximately 340mA. This low peak threshold is one of the key features that allows the LTC1174 to minimize system noise compared to other chips that carry significantly higher peak currents, easing shielding and filtering requirements and decreasing component stress.

To conserve power and maintain high efficiency at light loads, the LTC1174 uses Burst Mode operation. Unfortunately, this control scheme can also generate audio frequency noise at both light and heavy loads. In addition to electrical noise, acoustical noise can emanate from capacitors and coils under these conditions. A feedforward capacitor (C2) shifts the noise spectrum up out of the audio band, eliminating these problems. C2 also reduces peak-to-peak output ripple, which measures approximately 30mV over the entire load range.

The interactions of load current, efficiency and operating frequency are shown in Figure 78. High efficiency is maintained at even low current levels, dropping below 70% at around 800µA. No-load supply current is less than 200µA, dropping to approximately 1µA in shutdown mode. The operating frequency rises above the telephony bandwidth of 3kHz at a load of 1.2mA. Most products draw such low load currents only in standby mode with the audio circuits squelched, when noise is not an issue.

The frequency curve depicted in Figure 78 was measured with a spectrum analyzer, not a counter. This ensures that the lowest frequency noise peak is observed, rather than a faster switching frequency component. Any tendency to generate subharmonic noise is quickly exposed using this measurement method.

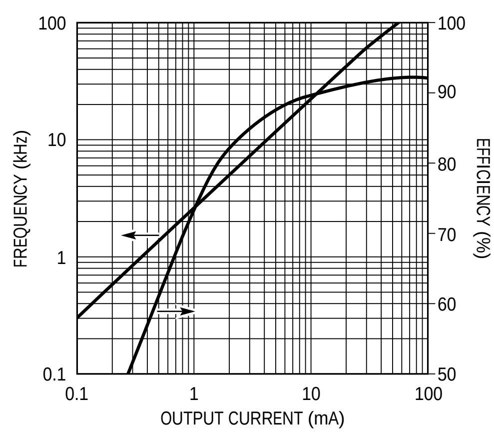


Figure 78. Parameter Interaction Chart for Figure 77's Circuit

APPLICATIONS FOR THE LT1302 MICROPOWER DC/DC CONVERTER

by Steve Pietkiewicz

2- or 3-Cell to 5V Converter

Figure 79 shows a 2- or 3-cell to 5V DC/DC converter that can deliver up to 600mA from a 2-cell input (2V minimum) or up to 900mA from a 3-cell input (2.7V minimum). R1 and R2 set the output voltage at 5V. The 200pF capacitor from FB to ground aids stability; without it the FB pin can act as an antenna and pick up dV/dt from the switch node, causing some instability in switch current levels at heavy loads. L1's inductance value is not critical; a minimum of 10µH is suggested in 2-cell applications (although this guideline is ignored in the 2-cell to 12V circuit shown later). Lower values typically have less DC resistance and can handle higher current. Transient response is better with low inductance but more output current can be had with higher values. Peak current in Burst Mode operation increases as inductance decreases, due to the finite response time of the current sensing comparator in the LT1302. The Coilcraft DO3316 series inductors have been found to be excellent in terms of performance, size and cost but their open construction results in some magnetic flux spray; try Coiltronics' OCTA-PAC series if EMI is a problem. Transient response with a load step of 25mA to 525mA is detailed in Figure 80. There is no overshoot upon load removal because switching stops entirely when output voltage rises above the comparator threshold. Undershoot at load step is less than 5%. The circuit's efficiency at various input voltages is shown in Figure 81.

Although efficiency graphs present useful information, a more "real world" measure of converter performance comes from battery lifetime chart recordings. Many systems require high power for a short time, for example to spin up a hard disk or transmit a packet of data. Figures 83, 84 and 85 present battery life data with a load profile of 50mA for 9 seconds and 550mA for 1 second, as detailed in Figure 82. At the chart speeds used, individual 10 second events are not discernable and the battery voltage appears as a very thick line. Figure 83 shows operating life using a 2-cell alkaline (Eveready E91) battery. Battery voltage (pen B) drops 400mV as the output load changes

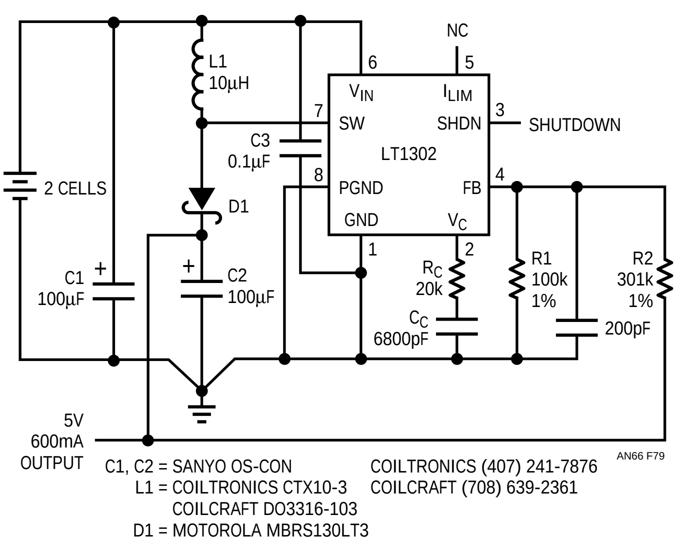


Figure 79. 2- or 3-Cell to 5V Converter Delivers 600mA, 1A From 3.3V Supply

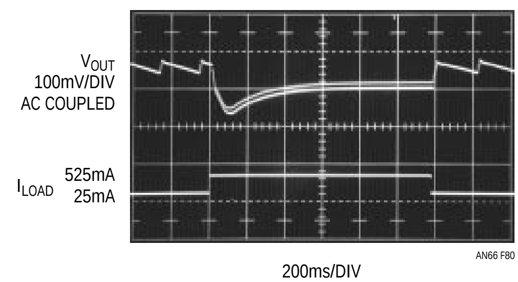


Figure 80. Transient Response of DC/DC Converter with 2.5V Input. Load Step is 25mA to 525mA

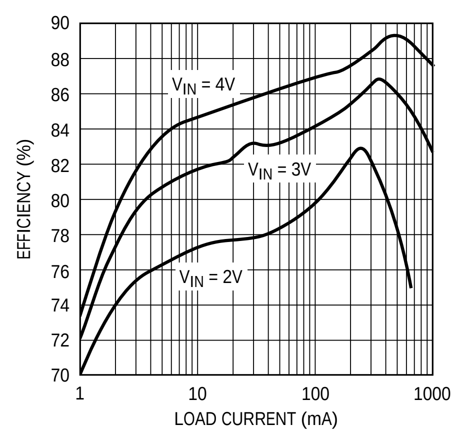


Figure 81. Efficiency of Figure 79's Circuit

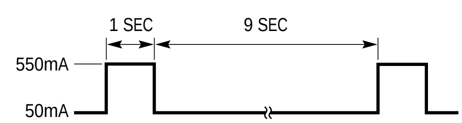


Figure 82. Load Profile for Battery Life Curves in Figures 83, 84 and 85

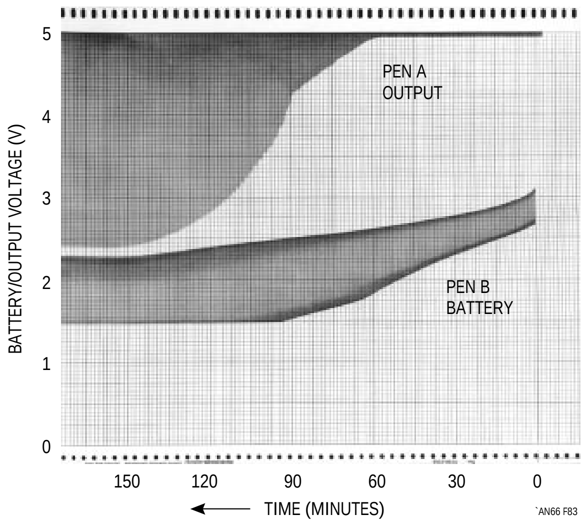


Figure 83. 2-Cell Alkaline Battery to 5V Converter with Load Profile of Figure 82 Gives 63 Minutes Operating Life. Battery Life Decreases When 550mA Load is Applied; Impedance is 330mΩ When Fresh. Output Voltage Drops at 550mA Load After 63 Minutes But Converter Can Still Deliver 50mA

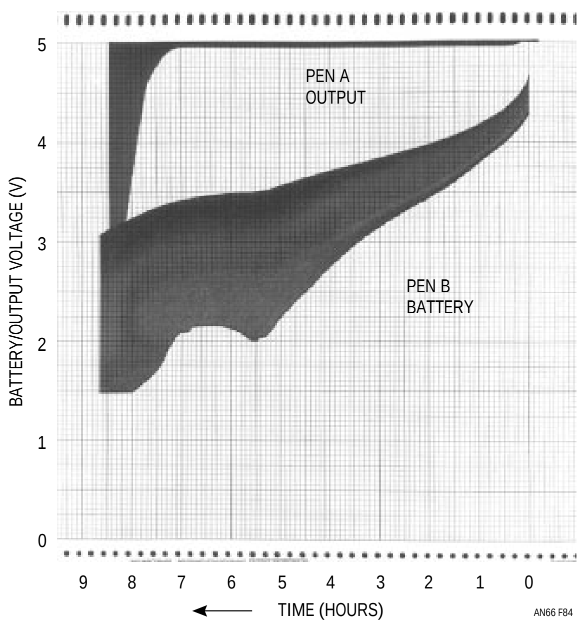


Figure 84. 3-Cell Alkaline Battery to 5V Converter with Pulsed Load Has 7.3 Hours Operating Life

from 50mA to 550mA. Battery impedance (330mΩ when fresh) can be derived from this data. After 63 minutes the battery voltage drops substantially below 2V when the output load is 550mA, causing the output voltage (pen A) to drop. The output returns to 5V when the load drops to 50mA. The LT1302's undervoltage lockout prevents the battery voltage from falling below 1.5V until the battery is completely discharged (not shown on the chart).

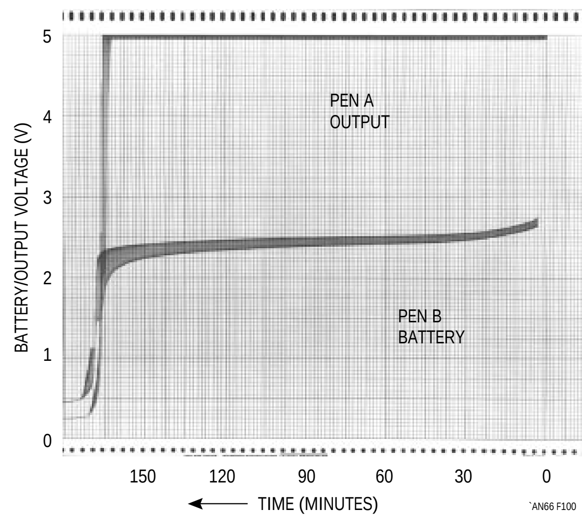


Figure 85. 2-Cell NiCd Battery to 5V Converter Shows Dramatically Lower ESR of NiCds Compared to Alkalines. Battery Impedance Is 80mΩ. Although the 600mA Hour NiCd Has 1/4 the Energy of 2.4A/Hr Alkalines with 50mA/550mA Loads NiCds Outlast Alkalines by a Factor of 2.8. Low Cell Impedance is Maintained Until the Battery Is Completely Discharged

A 3-cell alkaline battery has a significantly longer life, as shown in Figure 84. Note that the time scale here is one hour per inch. Usable life is about 7.3 hours, a sevenfold improvement over the 2-cell battery. Again, battery impedance causes the battery voltage (pen B) to drop as the load changes from 50mA to 550mA. The increasing change between the loaded and unloaded battery voltage over time is due to both increased current demand on the battery as its voltage decreases and increasing battery impedance as it is discharged.

Replacing the 2-cell alkaline with a 2-cell NiCd (AA Gates Millennium) battery results in a surprise shown in Figure 85. Although these AA NiCd cells have one-fourth the energy of AA alkaline cells, operating life is 2.8 times greater with the 50mA/550mA load profile. Dramatically lower battery impedance (80mΩ for the NiCd versus 330mΩ for the alkaline) is the cause. Battery voltage (pen B) drops just 100mV as the output load changes from 50mA to 550mA, compared to 400mV for alkalines. Additionally, impedance stays relatively constant over the life of the battery. This comparison clearly illustrates the limitations of alkaline cells in high power applications.

2-Cell to 12V Converter

Portable systems with PCMCIA interfaces often require 12V at currents of up to 120mA. Figure 86's circuit can

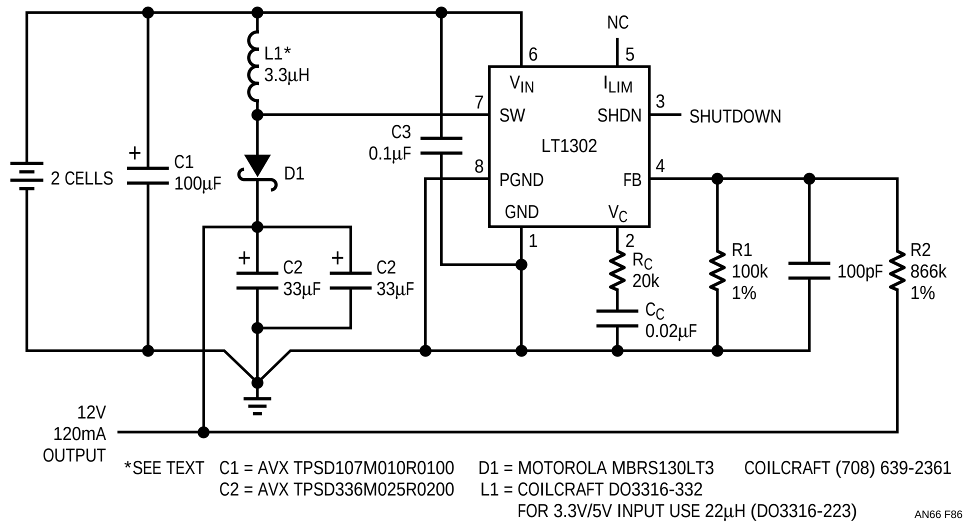


Figure 86. 2-Cell to 12V DC/DC Converter Delivers 120mA. Changing L1's Value Allows Operation from 3.3V/5V Supply

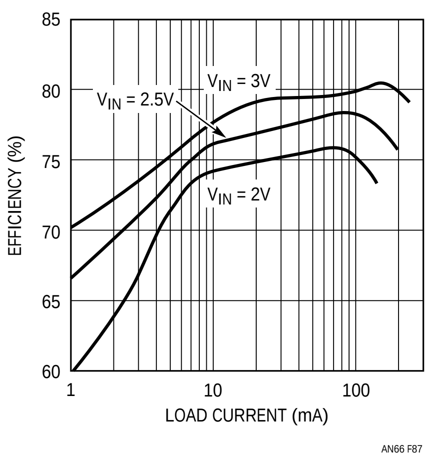


Figure 87. 2-Cell to 12V Converter Efficiency

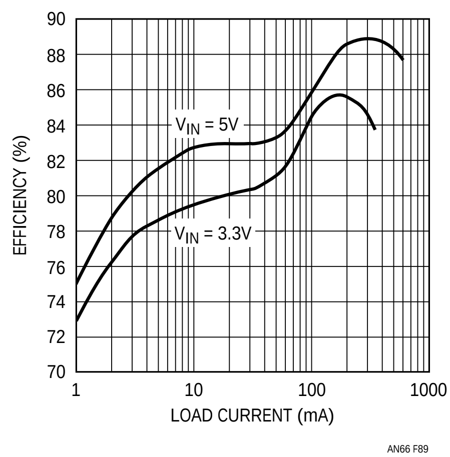


Figure 89. 3.3V/5V to 12V Converter Efficiency

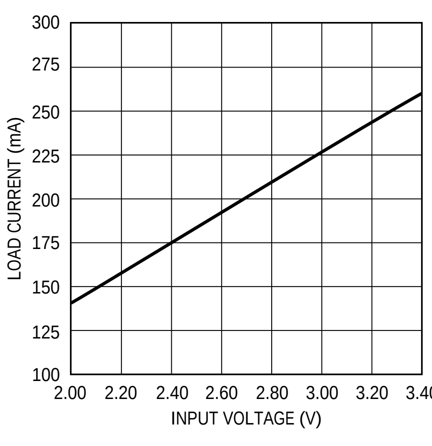


Figure 88. Maximum Load Current of 2-Cell to 12V Converter vs Input

generate 12V at over 120mA from a 2-cell battery. Operating the converter in continuous mode requires a higher duty cycle than the LT1302 provides, so a very low inductance (3.3µH) must be used in order to provide enough output current in discontinuous mode. Efficiency for this circuit is in the 70% to 80% range, as Figure 87's graph shows. Battery life at this power level would be short with a continuous load but the most common application for this voltage/current level, flash memory programming, has a rather low duty factor. Maximum output current versus input voltage is shown in Figure 88. To operate this circuit from a 3-cell battery change L1's value to 6.8µH. This will result in lower peak currents, improving efficiency substantially.

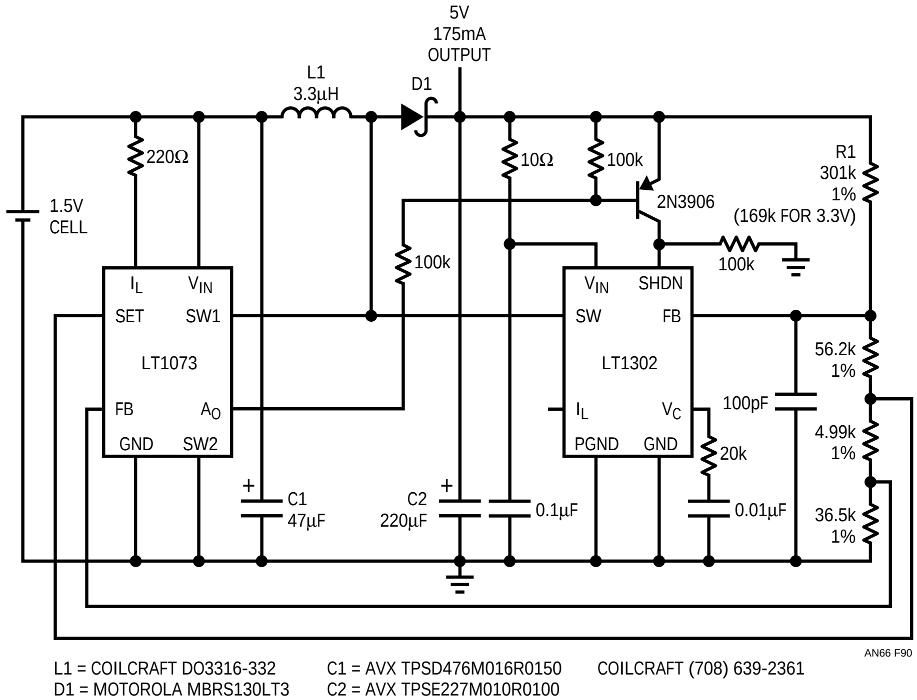


Figure 90. Single Cell to 5V Converter Delivers 150mA. Changing R1 to 169k Provides 3.3V at 250mA

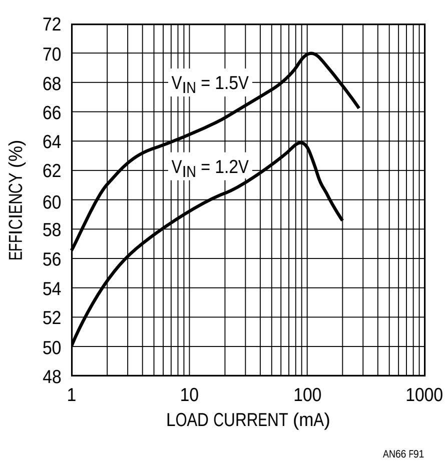


Figure 91. Single Cell to 5V Converter Efficiency

Changing L1's value to 22µH allows the circuit to operate from a 3.3V or 5V supply. Up to 350mA can be generated from 3.3V; 600mA can be delivered from a 5V input. Efficiency, pictured in Figure 89, exceeds 80% over much of the load range and peaks at 89% with a 5V input.

Single Cell to 5V/150mA Converter

Stand alone single-cell converters can typically provide no more than 40mA to 50mA at 5V from a single cell. When more power is required, the LT1302 can be used in conjunction with a single-cell device.[1] Figure 90's circuit operates from a single cell and delivers 5V at 150mA output. Although the LT1302 requires a minimum V_{IN} of 2V, single-cell operation can be achieved by powering the LT1302 from the 5V output. At start-up V_{OUT} is equal to the cell voltage minus a diode drop. The LT1073 initially puts the LT1302 in its shutdown state. The LT1073 switches L1, causing L1's current to alternately build up and dump into C2. When V_{OUT} reaches approximately 2V the LT1073's Set pin goes above 212mV causing A_O to go low. This pulls the LT1302's SHDN pin low, enabling it. The output, now booted by the much higher power LT1302, quickly reaches 2.4V. When the LT1073's FB pin reaches 212mV its switching action stops. The brief period when the LT1073 and LT1302 are switching simultaneously has no detrimental effect. When the output reaches 5V the LT1073 has ceased switching. Circuit efficiency is in the 60 to 70% range as shown in Figure 91.

3-Cell to 3.3V/12V Buck/Boost Converter

Obtaining 3.3V from three cells is not a straightforward task; a fresh battery measures over 4.5V and a fully depleted one 2.7V. Since battery voltage can be both

[1] Williams, Jim. "200mA Output, 1.5 to 5V Converter." *Linear Technology* III:1 (February, 1993) p. 17.

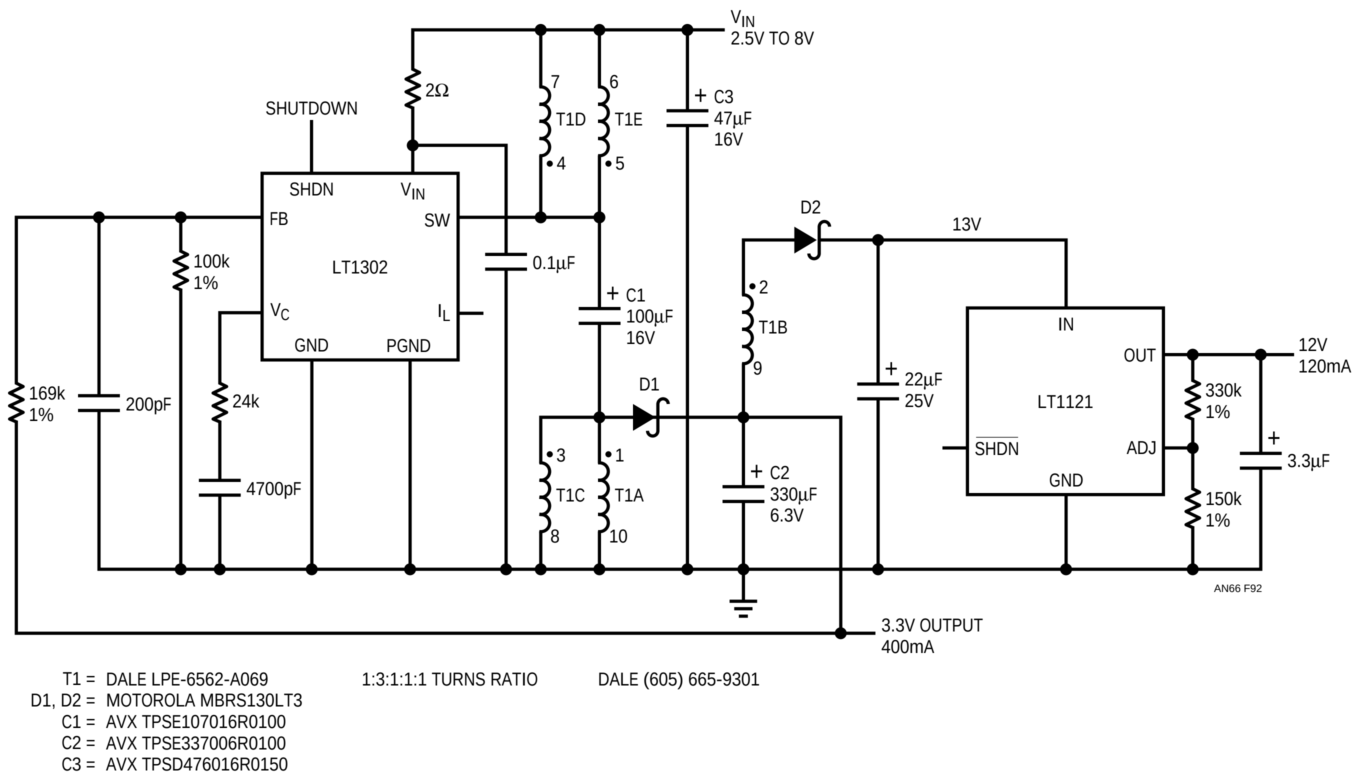


Figure 92. 3-Cell to 3.3V Buck/Boost Converter with Auxiliary 12V Regulated Output

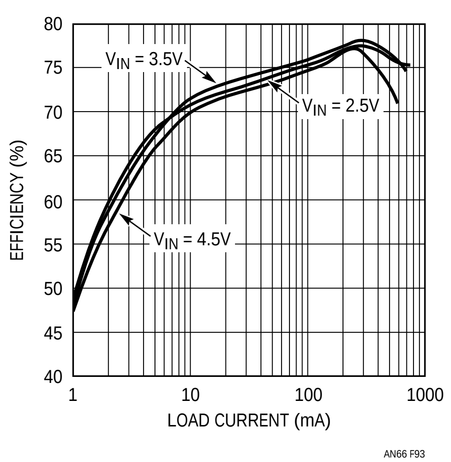


Figure 93. 3.3V Buck/Boost Converter Efficiency

above and below the output, common step-up (boost) or step-down (buck) converters are inadequate. Figure 92's circuit provides an efficient solution to the problem using just one magnetics component and also provides an auxiliary 12V output. When the LT1302's switch is on its SW pin goes low, causing current buildup in T1D and T1E (windings are paralleled to achieve lower DC resistance). D1's anode goes to $-V_{IN}$ because of the phasing of T1C/T1A relative to T1D/T1E. C1 is charged to V_{IN}. When the switch opens, SW flies high to a voltage of $V_{IN} + V_{OUT} + V_{DIODE}$. Energy is transferred to the output by magnetic coupling from T1D/T1E to T1C/T1A and by current flowing through C1. During this flyback phase, T1A/T1C has 3.3V plus a diode drop across the windings. T1B, which has a 3:1 turns ratio, has approximately 10V to 11V impressed upon it. T1B "stands" on the 3.3V output, resulting in about 13V to14V at the input of the LT1121 linear regulator, which then precisely regulates the 12V output. Since this output is not directly regulated by the LT1302, it cannot be loaded without having at least a small load on the directly regulated 3.3V output. The LT1121 can be turned off by pulling its SHDN pin low, isolating the load from the output. Figure 93 shows the circuit's efficiency for various input voltages.

Construction Hints

The high speed, high current switching associated with the LT1302 mandates careful attention to layout. Follow the suggested component placement in Figure 94 for proper operation. High current functions are separated by the package from sensitive control functions. Feedback

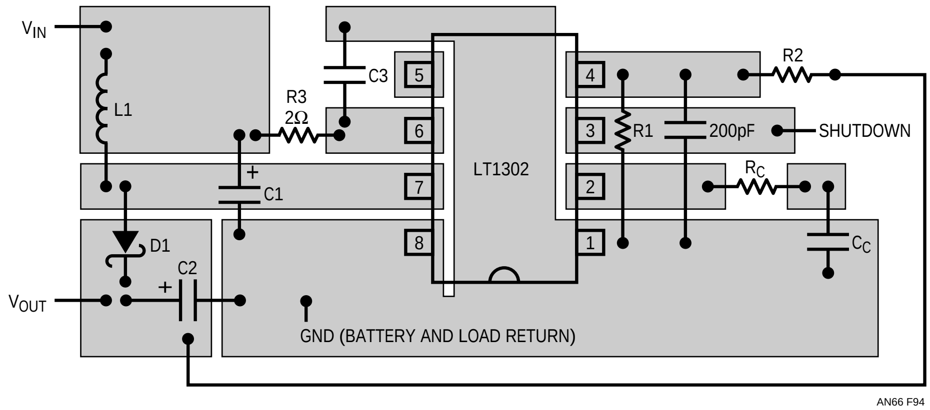


Figure 94. Suggested Component Placement for LT1302

resistors R1 and R2 should be close to the Feedback pin (Pin 4). Noise can easily be coupled into this pin if care is not taken. If the LT1302 is operated off a 3-cell or higher input, R3 (2Ω) in series with V_{IN} is recommended. This isolates the device from noise spikes on the input voltage. Do not install R3 if the device must operate from a 2V input, as input current will cause the LT1302's input voltage to go below 2V. The 0.1µF ceramic bypass capacitor C3 (use X7R not Z5U) should be mounted as close as possible to the package. Grounding should be segregated as illustrated. C3's ground trace should not carry switch current. Run a separate ground trace up under the package as shown. The battery and load return should go to the power side of the ground copper. Adherence to these rules will result in working converters with optimum performance.

CLOCK-SYNCHRONIZED SWITCHING REGULATOR HAS COHERENT NOISE

by Jim Williams, Sean Gold and Steve Pietkiewicz

Gated oscillator type switching regulators permit high efficiency over extended ranges of output current. These regulators achieve this desirable characteristic by using a gated oscillator architecture instead of a clocked pulse-width modulator. This eliminates the "housekeeping" currents associated with the continuous operation of fixed-frequency designs. Gated oscillator regulators simply self-clock at whatever frequency is required to maintain the output voltage. Typically, loop oscillation frequency ranges from a few Hertz into the kiloHertz region depending upon the load.

This asynchronous variable frequency operation seldom creates problems; some systems, however, are sensitive to this characteristic. The circuit in Figure 95 slightly modifies a gated-oscillator-type switching regulator by synchronizing its loop oscillation frequency to the system's clock. In this fashion the oscillation frequency and its attendant switching noise, although variable, are made coherent with system operation.

Circuit operation is best understood by temporarily ignoring the flip-flop and assuming that the LT1107 regulator's A_{OUT} and FB pins are connected. When the output voltage decays, the Set pin drops below V_{REF}, causing A_{OUT} to fall. This causes the internal comparator to switch high, biasing the oscillator and output transistor into conduction. L1 receives pulsed drive and its flyback events are deposited into the 100µF capacitor via the diode, restoring output voltage. This overdrives the Set pin, causing the IC to switch OFF until another cycle is required.

The frequency of this oscillatory cycle is load dependent and variable. If a flip-flop is interposed in the A_{OUT}/FB pin path as shown, the frequency is synchronized to the system clock. When the output decays far enough (trace A, Figure 96) the A_{OUT} pin (trace B) goes low. At the next clock pulse (trace C) the flip-flop Q2 output (trace D) sets low, biasing the comparator-oscillator. This turns on the power switch (V_{SW} pin is trace E), which pulses L1. L1

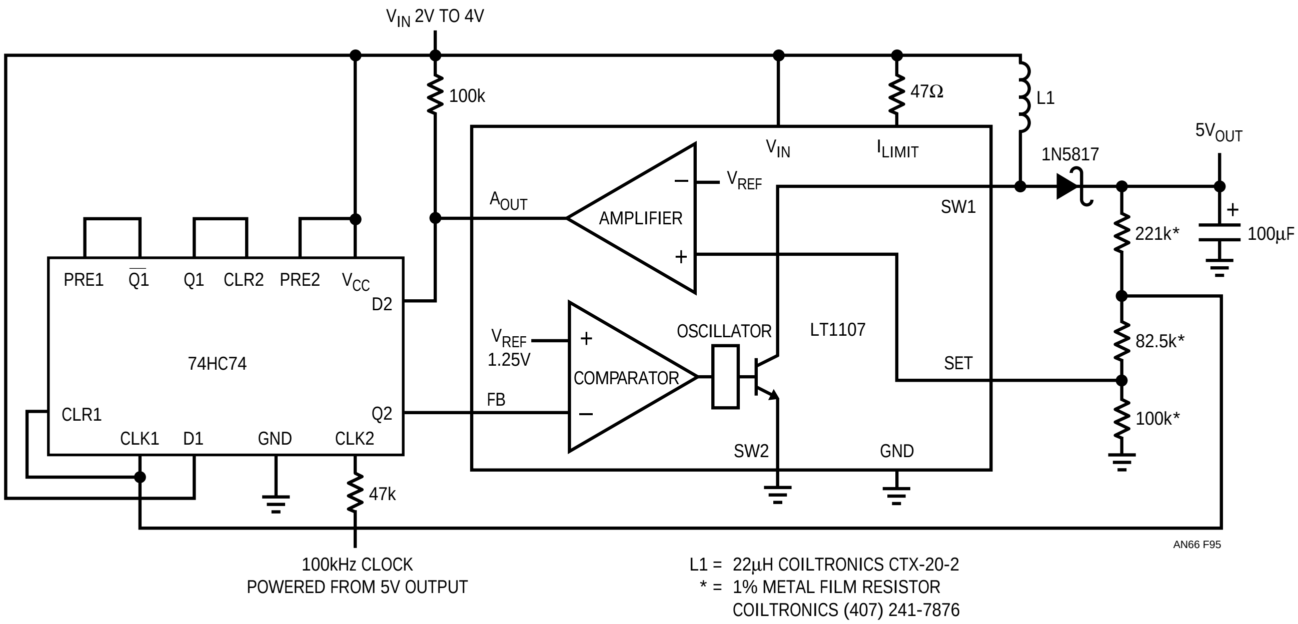


Figure 95. A Synchronizing Flip-Flop Forces Switching Regulator Noise To Be Coherent with the Clock

responds in flyback fashion, depositing its energy into the output capacitor to maintain output voltage. This operation is similar to the previously described case except that the sequence is forced to synchronize with the system clock by the flip-flop's action. Although the resulting loop's oscillation frequency is variable, it and all its attendant switching noise are synchronous and coherent with the system clock.

Because of its sampled nature, this clocked loop may not start. To ensure start-up, the flip-flop's remaining section is connected as a buffer. The CLR1/CLK1 line monitors output voltage via the resistor string. If the circuit does not start, Q1 goes high, CLR2 sets and loop operation commences. Although the circuit shown is a step-up type, any switching regulator configuration can use this technique.

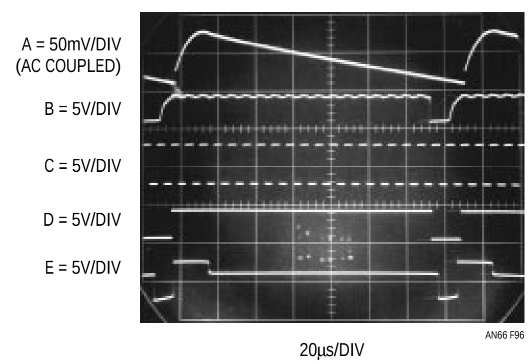


Figure 96. Waveforms for the Clock Synchronized Switching Regulator. The Regulator Switches (Trace E) Only on Clock Transitions (Trace C) Resulting in Clock Coherent Output Noise (Trace A)

BATTERY-POWERED CIRCUITS USING THE LT1300 AND LT1301

by Steve Pietkiewicz

5V from 2 Cells

Figure 97's circuit provides 5V from a 2-cell input. Shutdown is effected by taking the Shutdown pin high. V_{IN} current drops to 10µA in this condition. This simple boost topology does not provide output isolation, and in shutdown the load is still connected to the battery via L1 and D1. Figure 98 shows the efficiency of the circuit with a range of input voltages, including a fresh battery (3V) and an "almost-dead" battery (2V). At load currents below a few milliamperes, the 120µA quiescent current of the device becomes significant, causing the fall off in efficiency detailed in the figure. At load currents in the 20mA to 200mA range, efficiency flattens out in the 80% to 88% range, depending on the input. Figure 99 details circuit operation. V_{OUT} is shown in trace A. The burst repetition pattern is clearly shown as V_{OUT} decays, then steps back up due to switching action. Trace B shows the voltage at the switch node. The damped high frequency waveform at the end of each burst is due to the inductor "ringing off," forming an LC tank with the switch and diode capacitance. It is not harmful and contains far less energy than the high speed edge that occurs when the switch turns off. Switch current is shown in trace C. The current comparator inside the LT1300 controls peak switch current, turning off the switch when the current reaches approximately 1A.

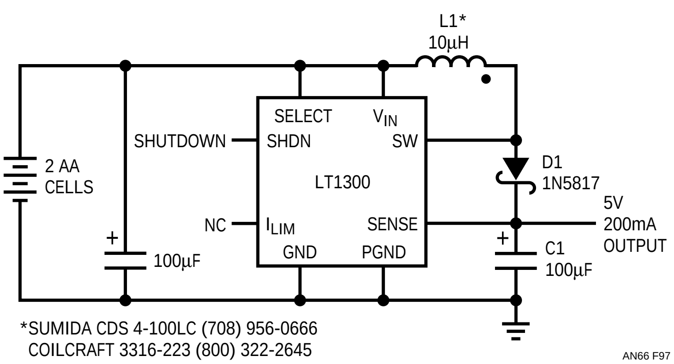


Figure 97. 2-Cell to 5V DC/DC Converter Delivers >200mA with a 2V Input

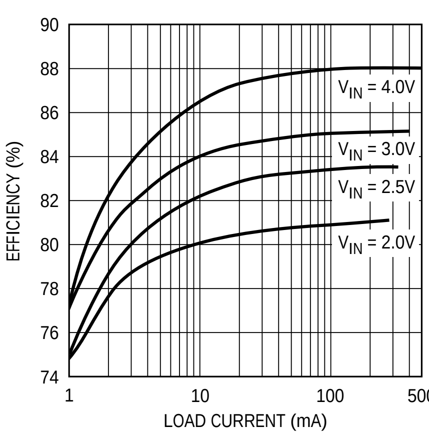


Figure 98. Efficiency of Figure 112's Circuit

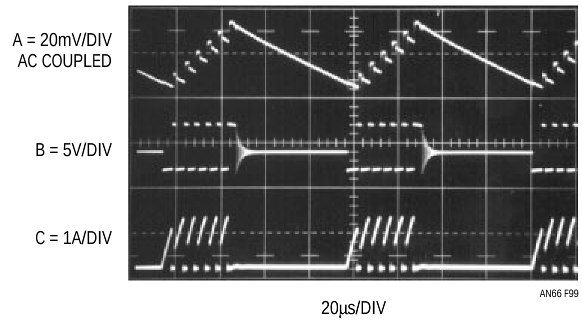


Figure 99. Burst Mode Operation In Action

Although efficiency curves present useful information, a more important measure of battery-powered DC/DC converter performance is operating life. Figures 100 and 101 detail battery life tests with Figure 97's circuit at load currents of 100mA and 200mA, respectively. Operating-life curves are shown using both Eveready E91 alkaline cells and new L91 "Hi-Energy" lithium cells. These lithium cells, new to the market, are specifically designed for high drain applications. The performance advantage of lithium is about 2:1 at 100mA load current (Figure 100), increasing to 2.5:1 at 200mA load (Figure 101). Alkaline cells perform poorly at high drain rates because their internal impedance ranges from 0.2Ω to 0.5Ω, causing a large voltage drop within the cell. The alkaline cells feel quite warm at 200mA load current, the result of I^2R losses inside the cells.

The reduced power circuit shown in Figure 102 can generate 5V at currents up to 50mA. Here the I_{LIM} pin is grounded, reducing peak switch current to 400mA. Lower profile components can be used in this circuit. The capacitors are C-case size solid tantalum and inductor L1 is the tallest component at 3.2mm. The reduced peak current also extends battery life, since the I^2R loss due to internal battery impedance is reduced. Figure 103 details efficiency versus load current for several input voltages and

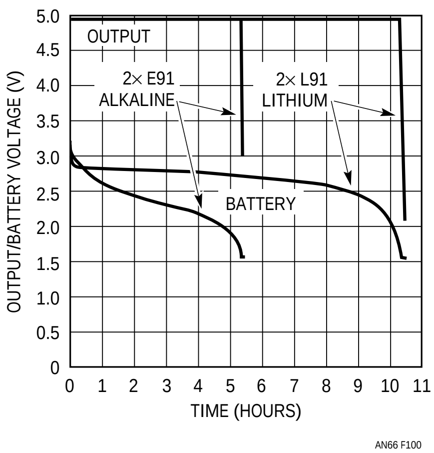


Figure 100. Two Eveready L91 Lithium AA Cells Provide Approximately Twice the Life of E91 Alkaline Cells at a 100mA Load Current

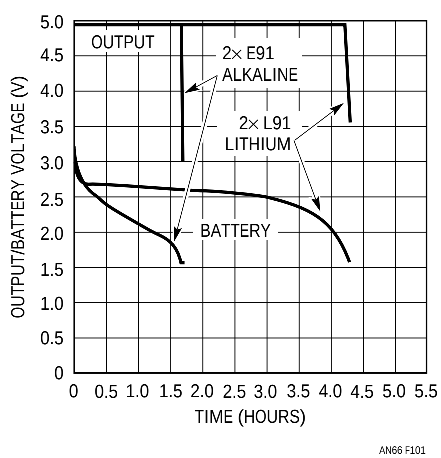


Figure 101. Doubling Load Current to 200mA Causes E91 Alkaline Battery Life to Drop by 2/3; L91 Lithium Battery Shows 2.5:1 Difference in Operating Life

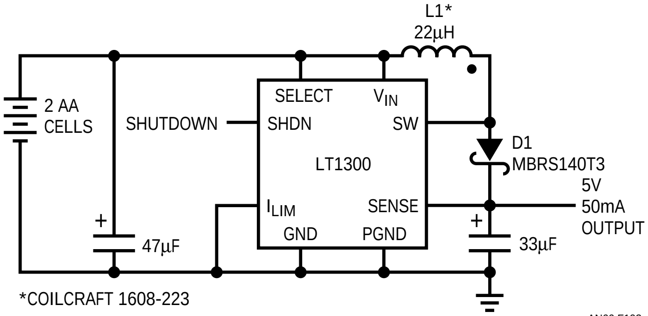


Figure 102. Lower Power Applications Can Use Smaller Components. L1 Is Tallest Component at 3.1mm

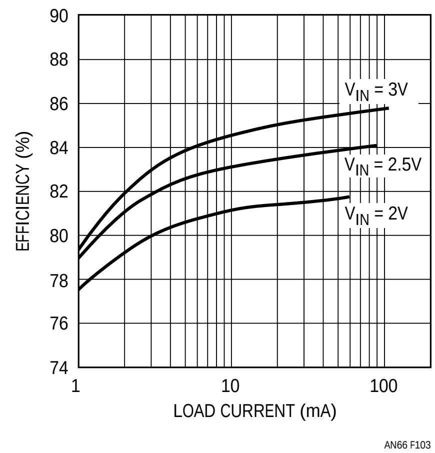


Figure 103. Efficiency of Figure 102's Circuit

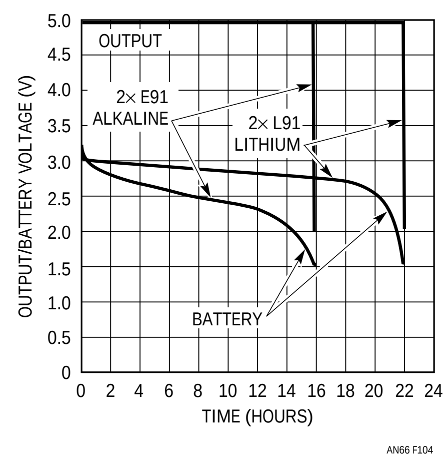


Figure 104. 50mA Load and Reduced Switch Current Are Kind to E91 AA Alkaline Battery; the Advantages of L91 Lithium Are Not as Evident

Figure 104 shows battery life at a 50mA load. Note that the L91 lithium battery lasts only about 40% longer than the alkaline. The higher cost of the lithium cells makes the alkaline cells more cost effective in this application. A pair of Eveready AAA alkaline cells (type E92) lasts 96.6 hours with 5mA load, very close to the rated capacity of the battery.

A 4-Cell Application

A 4-cell pack is a convenient, popular battery size. Alkaline cells are sold in 4-packs at retail stores and 4 cells usually provide sufficient energy to keep battery replacement frequency reasonable. Generating 5V from 4 cells, however, is a bit tricky. A fresh 4-cell pack has a terminal voltage of 6.4V, but at the end of its life the pack's terminal voltage is around 3.2V; hence, the DC/DC converter must step the voltage either up or down depending on the state of the batteries. A flyback topology with a costly custom designed transformer could be employed but Figure 105's circuit gets around these problems by using a flying capacitor scheme along with a second inductor. The circuit also isolates the input from the output, allowing the output to go to 0V during shutdown. The circuit can be divided conceptually into boost and buck sections. L1 and the LT1300 switch comprise the boost or step-up section and L2, D1 and C3 comprise the buck or step-down section. C2 is charged to V_{IN} and acts as a level shift between the two sections. The switch node toggles between ground and $V_{IN} + V_{OUT}$, and the L2/C2 diode node toggles between $-V_{IN}$ and $V_{OUT} + V_D$. Figure 106 shows efficiency versus load current for the circuit. All four energy storage elements must handle power, which accounts for the lower efficiency of this circuit compared to the simpler boost circuit in Figure 97. Efficiency is directly related to the ESR and DCR of the capacitors and inductors used. Better capacitors cost more money. Better inductors do not necessarily cost more but they do take up more space. Worst-case RMS current through C2 occurs at minimum input voltage and measures 0.4A at full load with a 3V input. C2's specified maximum RMS current must be greater than this worst-case current. The Sanyo capacitors noted specify a maximum ESR of 0.045Ω with a maximum ripple current rating of 2.1A. The Gowanda inductors specify a maximum DCR of 0.058Ω.

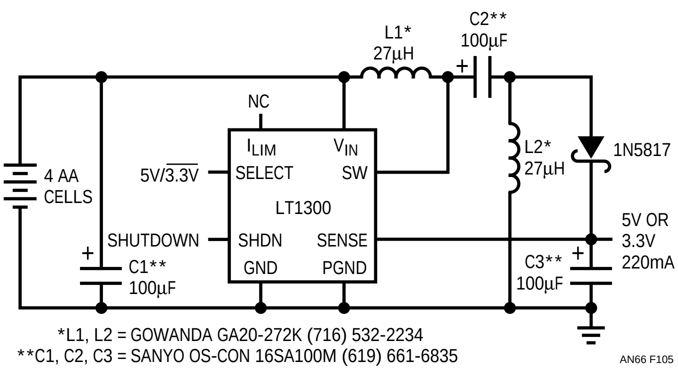


Figure 105. 4-Cell to 3.3V or 5V Converter Output Goes to Zero When in Shutdown. Inductors May Have, but Do Not Require Coupling; a Transformer or Two Separate Units Can Be Used

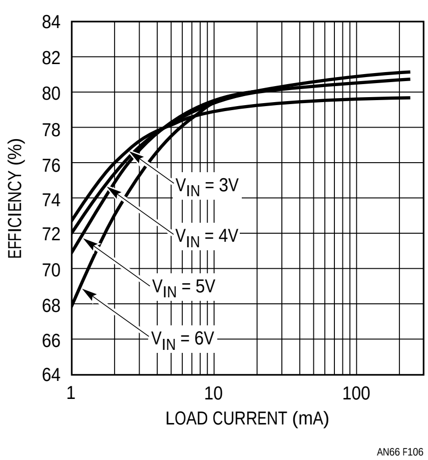


Figure 106. Efficiency of Up/Down Converter in Figure 105

LT1301 Outputs: 5V or 12V

The LT1301 is identical to the LT1300 in every way except output voltage. The LT1301 can be set to a 5V or 12V output via its Select pin. Figure 107 shows a simple 3.3V or 5V to 12V step-up converter. It can generate 120mA at 12V from either 3.3V or 5V inputs, enabling the circuit to provide VPP on a PCMCIA card socket. Figure 108 shows the circuit's efficiency. Switch voltage drop is a smaller percentage of input voltage at 5V than at 3.3V, resulting in the higher efficiency at 5V input.

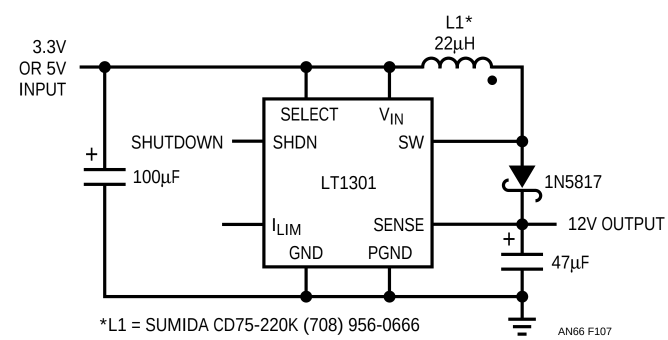


Figure 107. LT1301 Delivers 12V from 3.3V or 5V Input

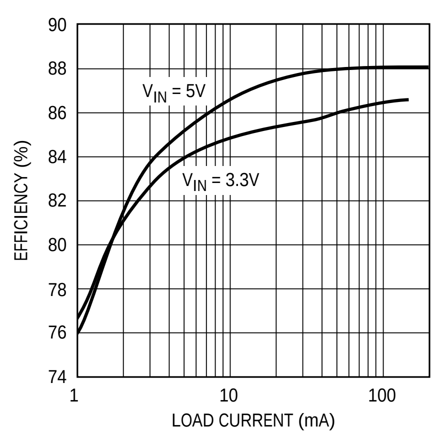


Figure 108. Efficiency of Figure 122's Circuit

BATTERY-POWERED CIRCUITS USING THE LT1304 MICROPOWER DC/DC CONVERTER WITH LOW-BATTERY DETECTOR

by Steve Pietkiewicz

A 2-Cell to 5V Converter

A compact 2-cell to 5V converter can be constructed using the circuit in Figure 109. 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.

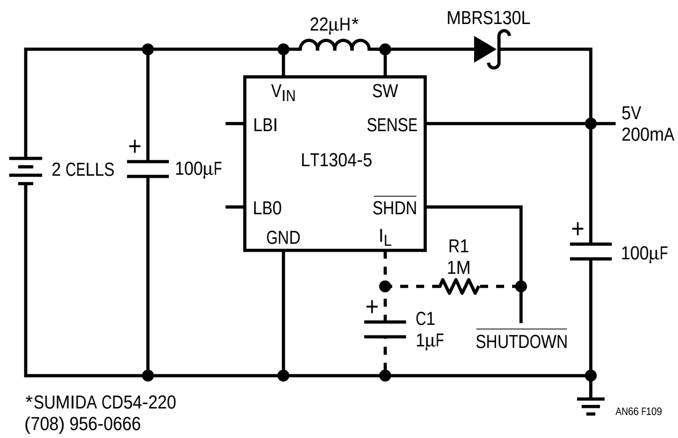


Figure 109. 2-Cell to 5V/200mA Boost Converter Takes Four External Parts. Components with Dashed Lines Are for Soft Start (Optional)

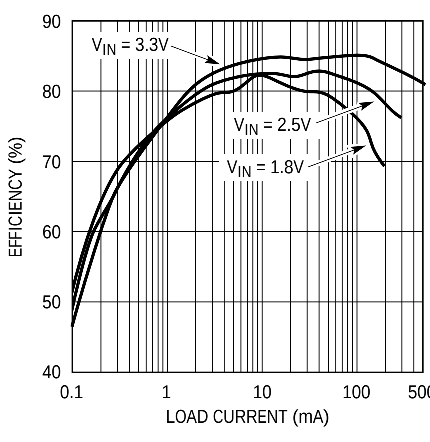


Figure 110. 2-Cell to 5V Converter Efficiency

Efficiency is detailed in Figure 110. Micropower Burst Mode 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 111. 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 112 details start-up behavior without soft start circuitry (R1 and C1 in Figure 109). 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 109's circuit limits inrush current at start-up, providing for a smoother turn-on as indicated in Figure 113.

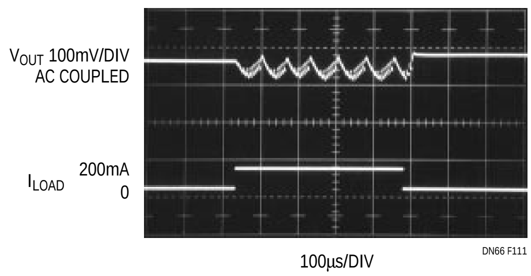


Figure 111. Boost Converter Load Transient Response with V_{IN} = 2.2V

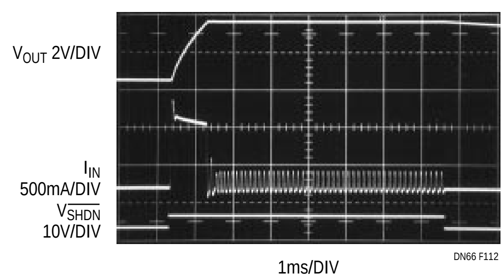


Figure 112. Start-Up Response. Input Current Rises Quickly to 1A. V_{OUT} Reaches 5V in Approximately 1ms. Output Drives 20mA Load

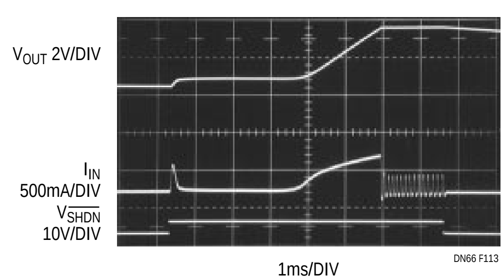


Figure 113. Start-Up Response with 1µF/1MΩ Components in Figure 109 added. Input Current Is More Controlled. V_{OUT} Reaches 5V in 6ms. Output Drives 20mA Load

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 114 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.

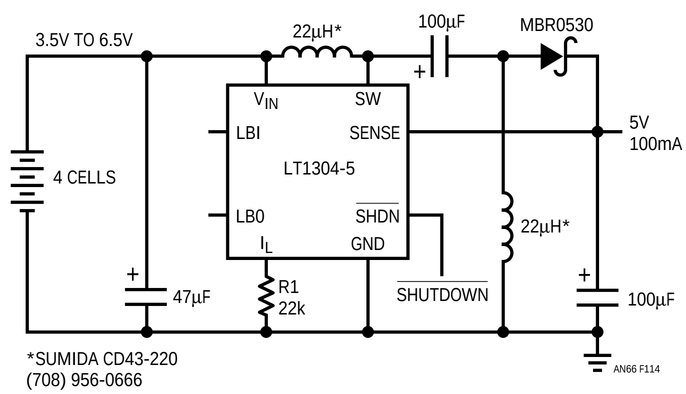


Figure 114. 4-Cell to 5V Step-Up/Step-Down Converter, Also Known as SEPIC (Single-Ended Primary Inductance Converter). Low Profile Components Are Used Throughout

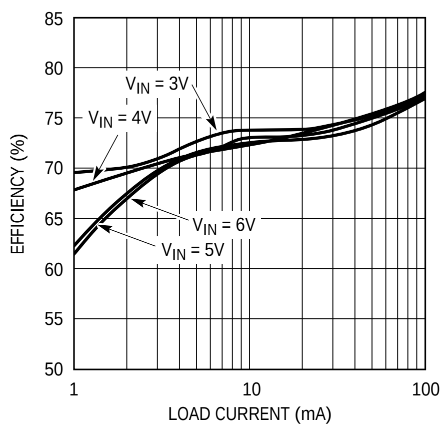


Figure 115. Efficiency Plot of SEPIC Converter Shown in Figure 114

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 116 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 $\overline{\text{SHDN}}$ 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

Super Burst is a trademark of Linear Technology Corporation.

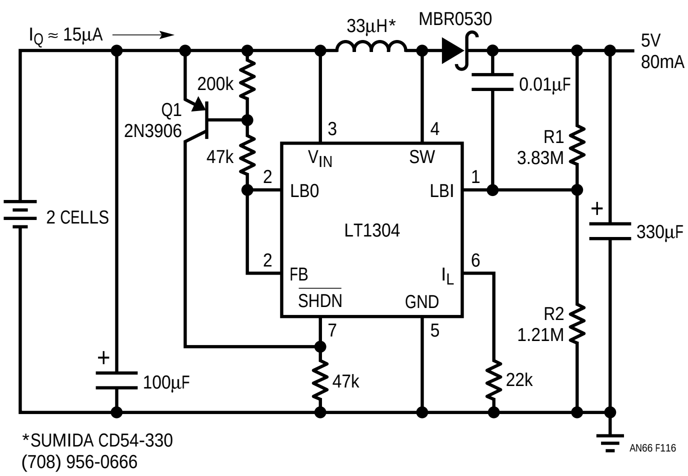


Figure 116. Super Burst Mode Operation 2-Cell to 5V DC/DC Converter Draws Only 15µA Unloaded. 2 AA Alkaline Cells Will Last for Years

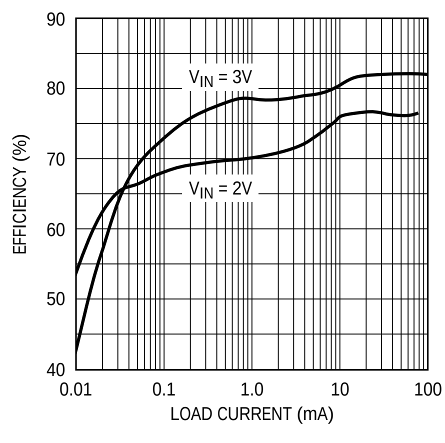


Figure 117. Super Burst Mode Operation DC/DC Converter Efficiency

117. The converter is approximately 70% efficient at a 100μA load, 20 points higher than the circuit of Figure 109. Even at a 10μA load, efficiency is in the 40% to 50% range, equivalent to 100μW to 120μW total power drain from the battery. In contrast, Figure 109's circuit consumes approximately 300μW to 400μW unloaded.

An output capacitor charging cycle or "burst" is shown in Figure 118, 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 119 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 116's circuit.

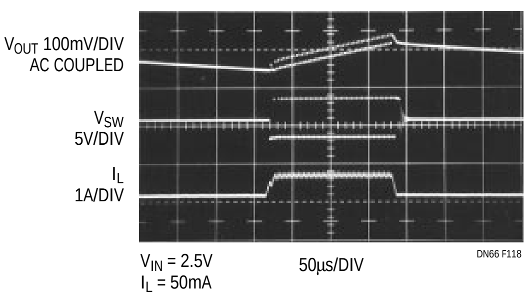


Figure 118. Super Burst Mode Operation in Action

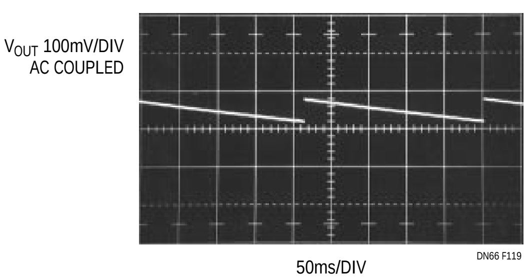


Figure 119. Super Burst Mode Operation Circuit with 100μA Load, Burst Occurs Approximately Once Every 240ms

Layout

The LT1304 switch turns on and off very quickly. For best performance we suggest the component placement in Figure 120. 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 121[1]. 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 122. Transition time is similar to that of Figure 121. Adherence to the layout suggestions will result in working DC/DC converters with a minimum of trouble.

[1]Instrumentation for oscillographs of Figures 121 and 122 include Tektronix P6032 active probe, Type 1S1 sampling unit and type 547 mainframe.

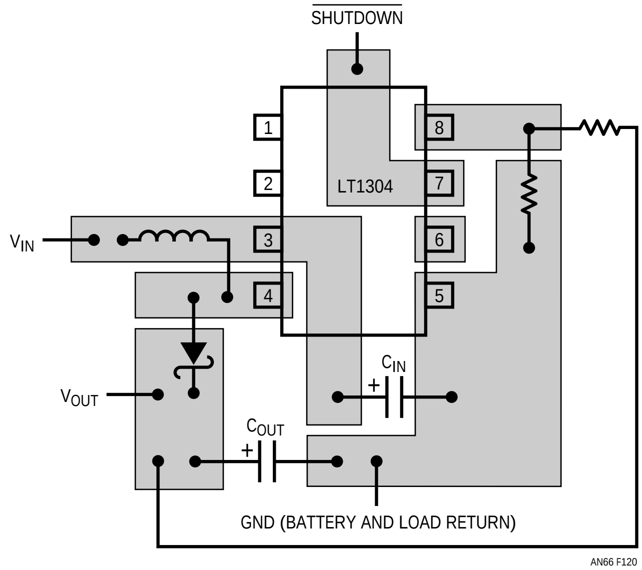


Figure 120. Suggested Layout for Best Performance. Input Capacitor Placement as Shown Is Highly Recommended. Switch Trace (Pin 4) Copper Area is Minimized

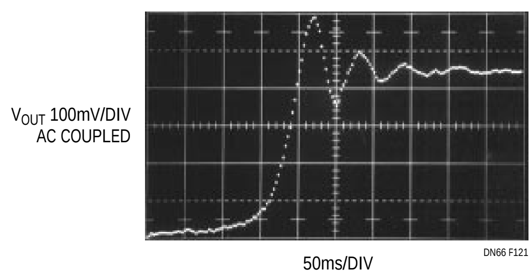


Figure 121. LT1304 Switch Rise Time Is in the 5ns Range. These Types of Edges Emphasize the Need for Proper PC Board Layout

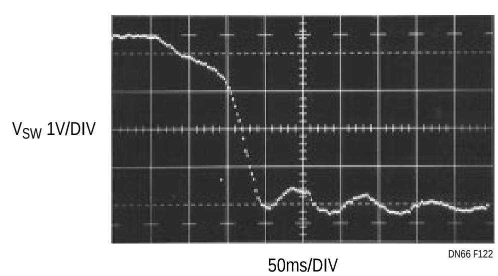


Figure 122. Switch Fall Time. Lower Slope in Second and Third Graticules Shows Effect of Lead and Bond Wire Inductance

AUTOMATIC LOAD SENSING SAVES POWER IN HIGH VOLTAGE CONVERTER

by Mitchell Lee

There are a surprising number of high output voltage applications for LTC's micropower DC/DC converter family. These applications include electroluminescent panels, specialized sensing tubes and xenon strobes. One of the key features of the micropower converters is low quiescent current. Since the quiescent current is far less than the self-discharge rate of common alkaline cells, the traditional ON/OFF switch can be eliminated in cases where the load is intermittent or where the load is shut down under digital control.

The maximum switch voltage for many micropower devices is 50V. For higher outputs the circuit shown in Figure 123 is often recommended. It combines a boost regulator and a charge pump tripler to produce an output voltage of up to 150V. The output is sensed through a divider network, which consumes a constant current of about 12µA. This doesn't seem like much, but reflected back to the 3V battery it amounts to over 3mA. Together with the LT1107's 320µA quiescent current the battery current is 3.5mA under no load. In standby applications this is unacceptably high, even for two D cells.

A circuit consisting of transistors Q1 and Q2 was added to reduce the standby current to an acceptable level. When a load of more than 50µA is present, Q1 turns on, Q2 turns off and the 9.1M resistor (R4) serves as a feedback path. R2, R3 and R4 regulate the output at 128V.

If the load current drops below 50µA, Q1 turns off and Q2 turns on, shorting out R4. With R4 out of the way, R2 and R3 regulate the output to approximately 15V. The measured input current under this condition is only 350µA, just slightly higher than the chip's no-load quiescent current. When the load returns, Q1 senses the excess current and the output automatically rises to its nominal value of 128V.

This automatic feedback switching scheme improves the battery current by a factor of ten and eliminates the need for a mechanical ON/OFF switch in applications where the load is under digital control.

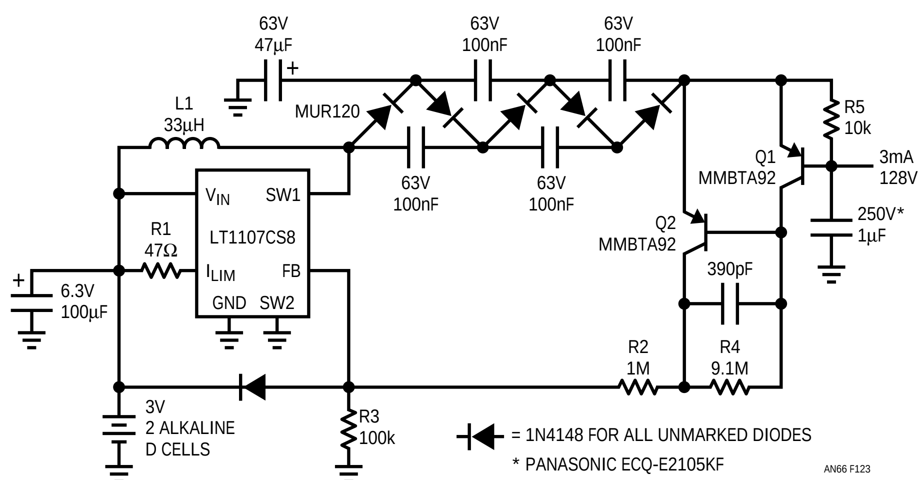


Figure 123. Automatic Shutdown Reduces Battery Current to 350µA

Regulators—Switching (Micropower)

Backlight

HIGH EFFICIENCY EL DRIVER CIRCUIT

by Dave Bell

Electroluminescent (EL) lamps are gaining popularity as sources of LCD backlight illumination, especially in small, handheld products. Compared with other backlighting technologies, EL is attractive because the lamp is thin, lightweight, rugged and can be illuminated with little power.

EL lamps are capacitive in nature, typically exhibiting around 3000pF/in^2, and require a low frequency (50Hz to 1kHz) 120V_{RMS} AC drive voltage. Heretofore, this has usually been generated by a low frequency blocking oscillator using a large transformer.

Figure 124 depicts a high efficiency EL driver that can drive a relatively large (12 in^2) EL lamp using a small high frequency transformer. The circuit is self-oscillating and delivers a regulated triangle wave to the attached lamp. Very high conversion efficiency may be obtained using this circuit, even matching state-of-the-art CCFL backlights at modest brightness levels (10 to 20 foot-lamberts).

Since an EL lamp is basically a lossy capacitor, the majority of the energy delivered to the lamp during the charge half-cycle is stored as electrostatic energy (1/2CV^2). Overall conversion efficiency can be improved by almost 2:1 if this stored energy is returned to the battery during the discharge half-cycle. The circuit of Figure 124 operates as a flyback converter during the charge half-cycle, taking energy from the battery and charging the EL capacitance. During the discharge half-cycle the flyback converter operates in the reverse direction, taking energy back out of the EL lamp and returning it to the battery. Nearly 50% of the energy taken during the charge half-cycle is returned during the discharge half-cycle; hence the 2:1 efficiency improvement.

During the charge half-cycle, the LT1303 operates as a flyback converter at approximately 150kHz, ramping the current in T1's 10µH primary inductance to approximately 1A on each switching pulse. When the LT1303's internal power switch turns off, the flyback energy stored in T1 is delivered to the EL lamp through D3 and C5. Successive high frequency flyback cycles progressively charge the EL capacitance until 300V is reached on the "+" side of C5. At this point the feedback voltage present at the LT1303's LBI input reaches 1.25V, causing the internal comparator to change state.

When the LT1303's internal comparator changes state, the open-collector driver at the LBO output is released. This places the circuit into discharge mode and reverses the operation of the flyback energy transfer. Q3 turns on and removes the gate drive from Q2A, thereby disabling switching action on the primary of T1. Flip-flop U2A is also clocked, resulting in a high level on the $\overline{Q}$ output; this positive feedback action keeps LBI above 1.25V. Even though Q2A is turned off the LT1303's SW pin still switches into pull-up resistor R4. The resulting pulses at the SW pin are used to clock U2B and to drive a "poor man's" current mode flyback converter on the secondary of T1.

Every clock pulse to flip-flop U2B turns on Q2B and draws current from the EL lamp through C5, T1, D2 and Q4. (Q4 must be a 600V rated MOSFET to withstand the high peak voltages present on its drain during normal operation.) Current ramps up through T1's 2.25mH secondary inductance until the voltage across current sense resistor R12 reaches approximately 0.6V. At this point Q5 turns on, providing a direct clear to U2B and thereby terminating the pulse. Energy taken from the EL lamp and stored in T1's inductance is then transferred back to the battery through D1 and T1's primary winding. This cycle repeats at approximately 150kHz until the voltage on C5 ratchets down to approximately 0V. Once C5 is fully discharged, the preset input on U2A will be pulled low, forcing the voltage on the LT1303's LBI input to ground and initiating another charge half-cycle.

This circuit produces a triangle voltage waveform with a constant peak-to-peak voltage of 300V, but the frequency of the triangle wave depends on the capacitance of the attached EL lamp. A 12 in^2 lamp has approximately 36nF of capacitance, which results in a triangle wave frequency of approximately 400Hz. This produces approximately 17FL of light output from a state-of-the-art EL lamp. Because of the "constant power" nature of the charging flyback converter, light output remains relatively constant with changes in the battery voltage. In addition, since EL lamp capacitance decreases with age, the circuit tends to minimize brightness reduction with lamp aging. C5, R9, and R10 maintain a zero average voltage across the EL lamp terminals—an essential factor for reliable lamp operation.

Two options exist for EL lamps with different characteristics. Larger lamps can be supported by specifying an LT1305 instead of the LT1303 shown in Figure 124. The LT1305 will terminate switch cycles at 2A instead of 1A, thereby delivering four times as much energy (energy stored in T1 is defined by $1/2LI^2$). The value of R12 must also be reduced to 7.5Ω to increase the discharge flyback current by the same ratio. For smaller lamps or for brightness adjustment, the circuit may be "throttled" by connecting the LT1303/LT1305's FB pin to a small current-sense resistor in the lower leg of the EL lamp.

Not only does the depicted circuit operate very efficiently, it takes output fault conditions in stride. The circuit, with C5 rated at 300V, tolerates indefinite short-circuit and open-circuit conditions across its EL lamp output pins.

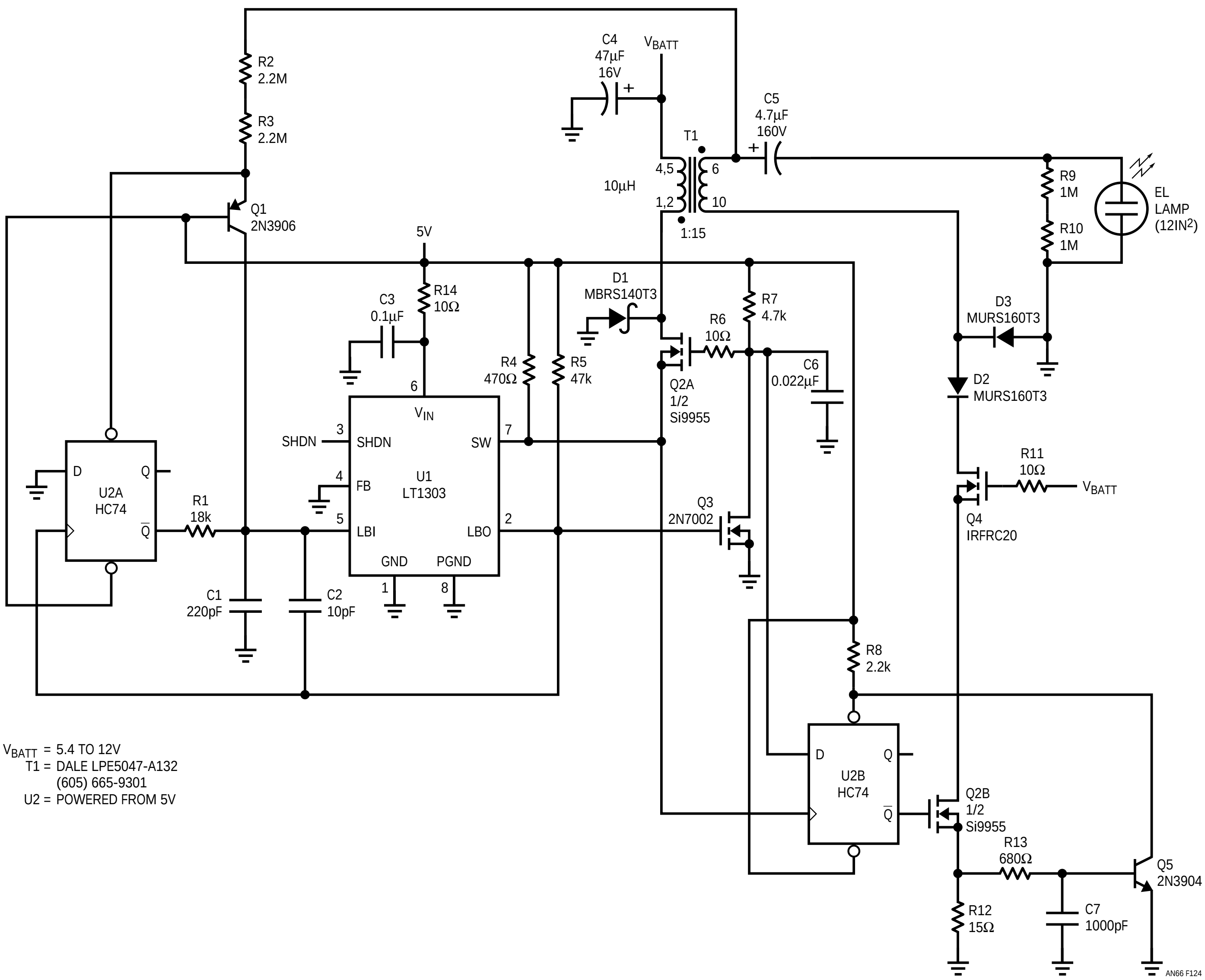


Figure 124. High Efficiency EL Driver Circuit

A LOW POWER, LOW VOLTAGE CCFL POWER SUPPLY

by Steve Pietkiewicz

Most recently published CCFL driver circuits require an input supply of 7V to 20V and are optimized for bulb currents of 5mA or more. This precludes lower power operation from 2- or 3-cell batteries often used in PDAs, palmtop computers and the like. A CCFL power supply that operates from 2V to 6V is shown in Figure 125. This circuit can drive a small (75mm) CCFL over a 100µA to 2mA range.

The circuit uses an LT1301 micropower DC/DC converter IC in conjunction with a current driven Royer class converter comprising T1, Q1 and Q2. When power is applied along with intensity adjust voltage V_A, the LT1301's I_{LIM} pin is driven slightly positive, causing maximum switching current to flow through the IC's internal Switch pin (SW). L1 conducts current, which flows from T1's center tap, through the transistors, into L1. L1's current is deposited in switched fashion to ground by the regulator's action.

The Royer converter oscillates at a frequency set primarily by T1's characteristics (including its load) and the 0.068µF capacitor. L1 sets the magnitude of the Q1-to-Q2 tail current, and hence, T1's drive level. The 1N5817 diode maintains L1's current flow when the LT1301's switch is off. The 0.068µF capacitor combines with L1's characteristics to produce sine wave voltage drive at the Q1 and Q2 collectors. T1 furnishes voltage step-up and about $1400V_{P-P}$ appears at its secondary. Alternating current flows through the 22pF capacitor into the lamp. On positive half cycles the lamp's current is steered to ground via D1. On negative half cycles the lamp's current flows through Q3's collector and is filtered by C1. The LT1301's I_{LIM} pin acts as a zero summing point with about 25µA bias current flowing out of the pin into C1. The LT1301 regulates L1's current to maintain equality of Q3's average collector current, representing one-half the lamp current, and R1's current, represented by V_A/R1. When V_A is set to zero the I_{LIM} pin's bias current forces about 100µA bulb current.

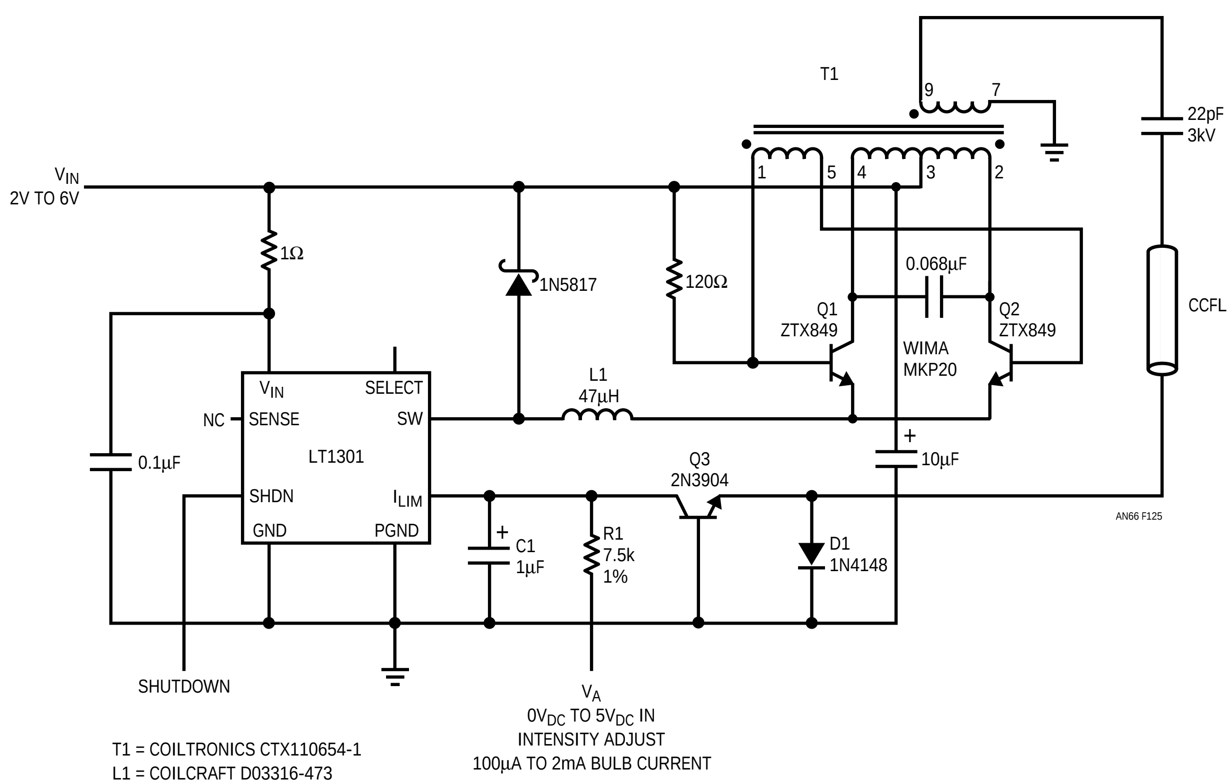


Figure 125. CCFL Power Supply

ALL SURFACE MOUNT EL PANEL DRIVER OPERATES FROM 1.8V TO 8V INPUT

by Steve Pietkiewicz

Electroluminescent (EL) panels offer a viable alternative to LED, incandescent or CCFL backlighting systems in many portable devices. EL panels are thin, rugged, lightweight and consume little power. They require no diffuser and emit an aesthetically pleasing blue-green light. EL panels, being capacitive in nature, typically exhibit about 3000pF per square inch of panel area and require low frequency (50Hz to 1kHz) $120V_{RMS}$ AC drive. This has traditionally been generated using a low frequency blocking oscillator with a transformer. Although this technique is efficient, transformer size renders the circuit unusable in many applications due to space constraints. Moreover, low frequency transformers are not readily available in surface mount form, complicating assembly.

Figure 126's circuit solves these problems by using an LT1303 micropower switching regulator IC along with a small surface mount transformer in a flyback topology. The 400Hz drive signal is supplied externally. When the drive signal is low, T1 charges the panel until the voltage at point A reaches 240VDC. C1 removes the DC component from the panel drive, resulting in 120VDC at the panel. When the input drive signal goes high the LT1303's FB pin is also pulled high, idling the IC and turning on Q1. Q1's collector pulls point A to ground and the panel to –120VDC. C2 can be added to limit voltage if the panel is disconnected or open. R3 provides intensity control by varying output voltage. Intensity can also be modulated by varying the drive signal's frequency.

Flyback transformer T1 (Dale LPE5047-A132) has a 10µH primary inductance and a 1:15 turns ratio. It measures 12mm by 13.3mm and is 6.3mm high. The 1:15 turns ratio generates high voltage at the output without exceeding the allowable voltage on the LT1303's Switch pin. Schottky diode D1 is required to prevent ringing at the SW pin from forward biasing the IC's substrate diode. Because of T1's low leakage inductance the flyback spike does not exceed 22V. No snubber network is required since the LT1303 SW pin can safely tolerate 25V. R1 and C3 provide decoupling for the IC's V_{IN} pin. Feedback resistor R2 is made from three 3.3M units in series instead of a single 10M resistor. This lessens the possibility of output voltage reduction due to PC board leakage shunting the resistor. Shutdown is accomplished by bringing the IC's SHDN pin high. For minimum current drain in shutdown the 400Hz drive signal should be low.

Figure 127 details relevant circuit waveforms with a 22nF load (about 7 inches of panel) and a 5V input. Trace A is the panel voltage. Trace B shows Switch pin action. The circuit's input current is pictured in trace C and trace D is the 400Hz input signal. The circuit's efficiency measures about 77%. With a 5V input the circuit can deliver $100V_{RMS}$ at 400Hz into a 44nF load. More voltage can be obtained at lower drive frequencies.

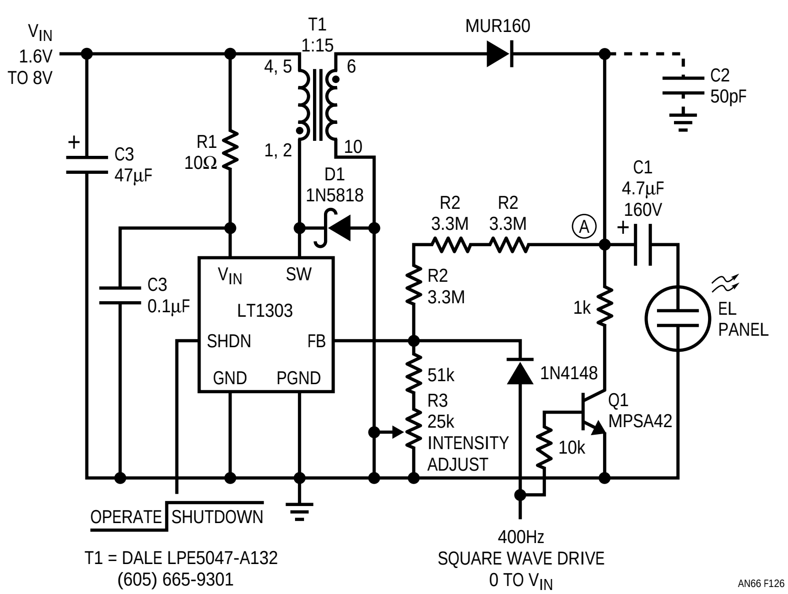


Figure 126. LT1303 Circuit Drives EL Panel

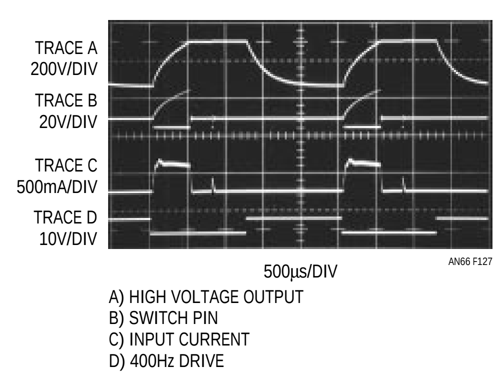


Figure 127. Oscillograph of Relevant Circuit Waveforms

A DUAL OUTPUT LCD BIAS VOLTAGE GENERATOR

by Jon A. Dutra

With the many different kinds of LCD displays available, systems manufacturers often want the option of deciding the polarity of the LCD bias voltage at the time of manufacturing.

The circuit in Figure 128 uses the LT1107 micropower DC/DC converter with a single inductor. The LT1107 features an I_{LIM} pin that enables direct control of maximum inductor current. This allows the use of a smaller inductor without the risk of saturation. The LT1111 could also be used with a resulting reduction in output power.

Circuit Operation

The circuit is basically an AC-coupled boost topology. The feedback signal is derived separately from the outputs, so loading of the outputs does not affect loop compensation. Since there is no direct feedback from the outputs, load regulation performance is reduced. With 28V out, from 10% to 100% load (4mA to 40mA), the output voltage sags by about 0.65V. From 1mA to 40mA load, the output voltage sags by about 1.4V. This is acceptable for most displays.

Output noise is reduced by using the auxiliary gain block (AGB) in the feedback path. This added gain effectively reduces the hysteresis of the comparator and tends to randomize output noise. With low ESR capacitors for C2 and C4, output noise is below 30mV over the output load range. Output power increases with $V_{BATTERY}$, from about 1.4W out with 5V in to about 2W out with 8V or more. Efficiency is 80% over a broad output power range.

If only a positive or negative output voltage is required, the two diodes and two capacitors associated with the unused output can be eliminated. The 100kΩ load is required on each output to load a parasitic voltage doubler created by the capacitance of diodes D2 and D4. Without this minimum load, the output voltage can go up to almost 50% above the nominal value.

Component Selection

The voltage at the Switch pin SW1 swings from 0V to V_{OUT} plus two diode drops. This voltage is AC coupled to the positive output through C1 and D1 and to the negative output through C3 and D3. The full output current flows through C1 and C3. Most tantalum capacitors are not rated for current flow and their use can result in field failures.

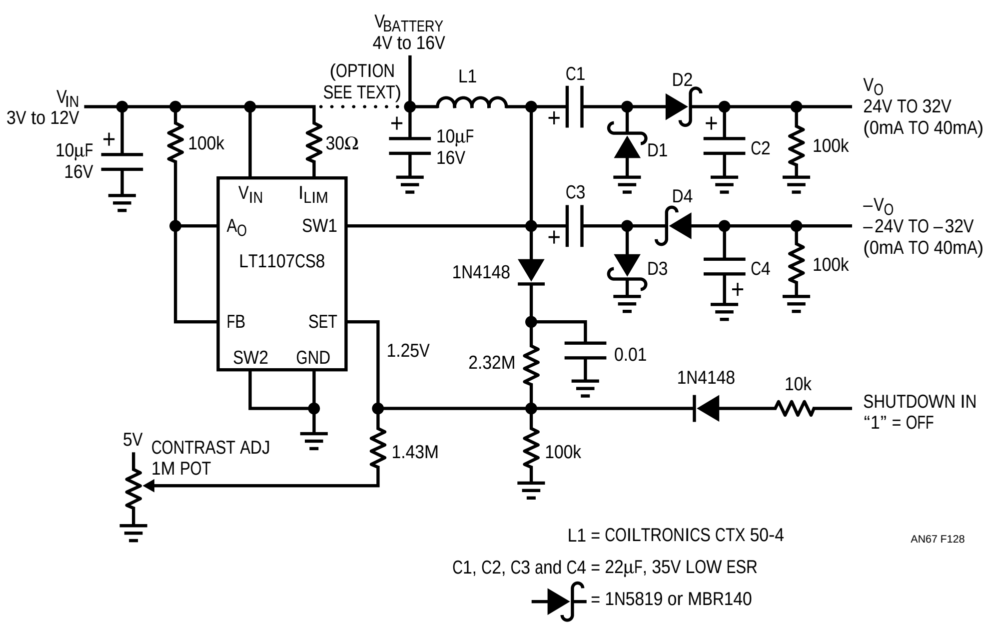


Figure 128. LT1107 Dual Output LCD Bias Generator Schematic Diagram

Use a rated tantalum or a rated electrolytic for longer system life. At lower output currents or higher frequencies, using monolithic ceramics is also feasible.

One could replace the 1N5819 Schottky diodes with 1N4148 types for lower cost, with a reduction in efficiency and load regulation characteristics.

Shutdown

The circuit can be shut down in several ways. The easiest is to pull the Set pin above 1.25V; however, this consumes 300µA in shutdown conditions. A lower power method is to turn off V_{IN} to the LT1107 by means of a high side switch or by simply disabling a logic supply. This drops quiescent current from the $V_{BATTERY}$ input below 10µA. In both cases V_{OUT} drops to 0V. In the event that $+V_{OUT}$ does not need to drop to zero, C1 and D1 can be eliminated.

Output Voltage Adjustment

The output voltage can be adjusted from any voltage above $V_{BATTERY}$ up to 46V with proper passive components. Output voltage can be controlled by the user with DAC, PWM or potentiometer control. By summing currents into the feedback node, the output voltage can be adjusted downward.

LCD BIAS SUPPLY

by Steve Pietkiewicz

An LCD requires a bias supply for contrast control. The supply's variable negative output permits adjustment of display contrast. Relatively little power is involved, easing RF radiation and efficiency requirements. An LCD bias generator is shown in Figure 129. In this circuit, U1 is an LT1173 micropower DC/DC converter. The 3V input is converted to 24V by U1's switch, L2, D1 and C1. The Switch pin (SW1) also drives a charge pump composed of C2, C3, D2 and D3 to generate –24V. Line regulation is less than 0.2% from 3.3V to 2V inputs. Although load regulation suffers somewhat because the –24V output is not directly regulated, it measures 2% for loads from 1mA to 7mA. The circuit will deliver 7mA from a 2V input at 75% efficiency.

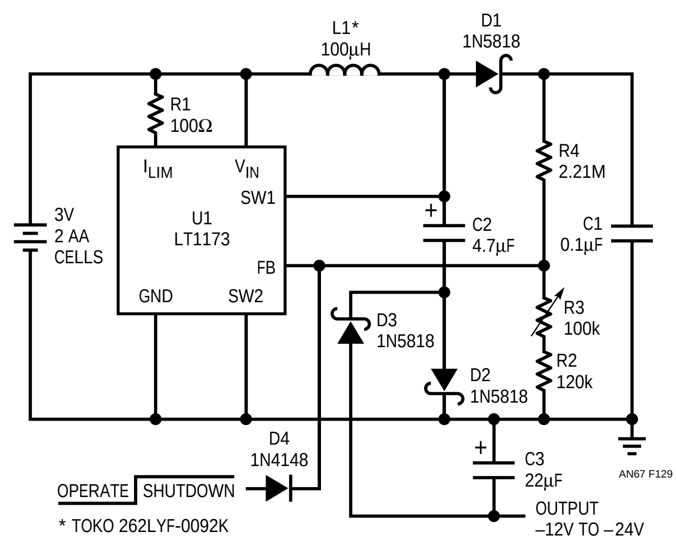


Figure 129. DC/DC Converter Generates LCD Bias

Regulators—Switching (Micropower)

VPP Generator

LTC1262 GENERATES 12V FOR PROGRAMMING FLASH MEMORIES WITHOUT INDUCTORS

by Anthony Ng and Robert Reay

Flash memories require a 5V V_{CC} supply and an additional 12V supply for write or erase cycles. The 12V supply can be a system supply or be generated from the 5V supply using a DC/DC converter circuit. Single supply flash memories (i.e., those with the 12V converter built-in) are available, but these memories have lower capacities and slower write/erase performance and therefore are not as popular as memories without a built-in 12V supply. Flash memories require that the 12V supply be regulated to within 5% and not exceed the permitted maximum voltage (14V for Intel ETOX™ memories). The LTC1262 offers a simple and cost effective 12V programming supply to meet these requirements.

Figure 130 shows a typical application circuit. The only external components needed are four surface mount capacitors. The LTC1262 uses a triple charge pump technique to convert 5V to 12V. It operates from 4.75V to 5.5V and delivers 30mA while regulating the 12V output to within 5%. The TTL-compatible SHDN pin can be driven directly by a microprocessor. When the SHDN pin is taken high (or floated) the LTC1262 enters shutdown mode. In this state the supply current of the LTC1262 is reduced to 0.5µA typical and the 12V output drops to V_{CC}. When SHDN is taken low, the LTC1262 leaves shutdown mode and the output rises to 12V without any potentially harmful overshoot (see Figure 131).

The LTC1262 is available in both 8-pin PDIP and narrow SO packages. With small surface mount capacitors, the complete 12V supply takes up very little space on a printed circuit board. In power sensitive applications, such as PCMCIA flash cards for portable PCs, the LTC1262 shutdown current is low enough to preclude the need for external switching devices when the system is inactive.

ETOX is a trademark of Intel Corporation.

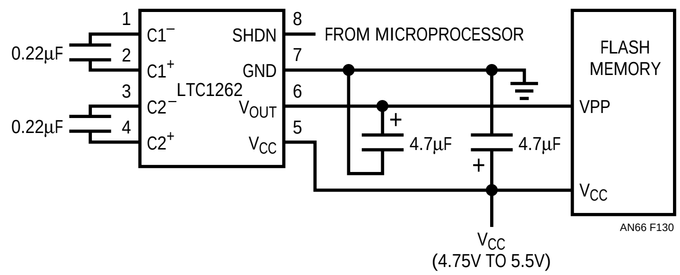


Figure 130. Typical LTC1262 Application Circuit

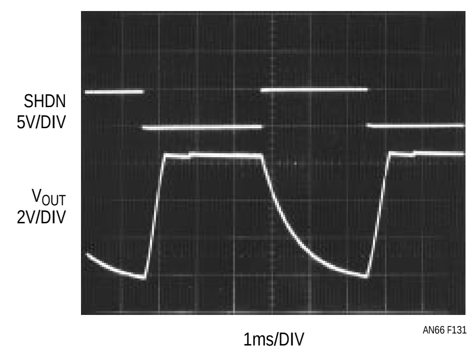


Figure 131. LTC1262 Taken In and Out of Shutdown

FLASH MEMORY VPP GENERATOR SHUTS DOWN WITH 0V OUTPUT

by Sean Gold

Nonvolatile "flash" memories require a well controlled 12V bias (VPP) for programming. The tolerance on VPP is ±5% for 12V memories. Excursions in VPP above 14V or below –0.3V are destructive. VPP is often generated with a boost regulator whose output follows the input supply when shut down. It is sometimes desirable to force VPP to 0V when the memory is not in use or is in read-only mode.

The circuit in Figure 132 generates a smoothly rising 12V, 60mA supply that drops to 0V under logic control. Figure 133 illustrates the programming cycle. Shortly after driving the SHDN pin high, the LT1109-12 switching regulator drives L1, producing high voltage pulses at the device's Switch pin. The 1N5818 Schottky diode rectifies these pulses and charges a reservoir capacitor C2. Q1 functions as a low on-resistance pass element. The 1N4148 diode clamps Q1 for reverse voltage protection. The circuit does not overshoot or display unruly dynamics because the regulator gets its DC feedback directly from the output at Q1's collector. Minor slew aberrations are due to Q1's switching characteristics.

Even with the additional losses introduced by Q1, efficiency is 83% with a 60mA load. Line and load regulation are both less than 1%. Output ripple is about 100mV under light loads. Quiescent current drops to 400µA when shut down. All components shown in Figure 132 are available in surface mount packages, making the circuit well suited for flash memory cards and other applications where minimizing PC board space is critical.

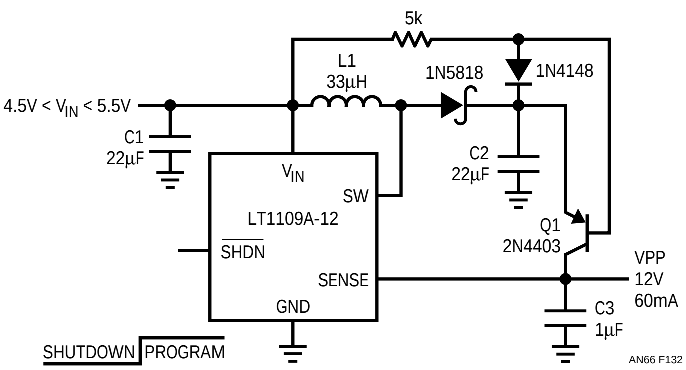


Figure 132. Boost Mode Switching Regulator with Low R_{ON} Pass Transistor for Flash Memory Programming

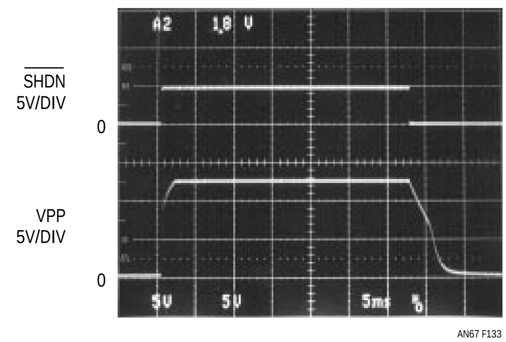


Figure 133. Input and Output Waveforms for the Flash Memory Programming Circuit

Regulators—Linear

LOW NOISE WIRELESS COMMUNICATIONS POWER SUPPLY

by Mitchell Lee and Kevin Vasconcelos

Shown in Figure 134 is a 5V power supply we designed for a synthesizer oscillator. The original design used a 3-terminal regulator but the regulator's voltage noise contributed to excessive phase noise in the oscillator, leading us to this solution. Of prime importance is noise over the 10Hz to 10kHz band. Careful measurements show a 40dB improvement over standard 3-terminal regulators.

The regulator is built around a 5V buried-Zener reference. It is the buried Zener's inherently low noise that makes the finished supply so quiet. Measured over a 10Hz to 10kHz band the 5V output contains just $7\mu V_{RMS}$ noise at full load. The 10Hz to 10kHz noise can be further reduced to $2.5\mu V_{RMS}$ by adding a 100µH, 1000µF output filter. The noise characteristics of the reference are tested and guaranteed to a maximum of 11µV over the band of interest.

An external boost transistor, the ZBD949, provides gain to meet a 200mA output current requirement. Current limiting is achieved by ballasting the pass transistor and clamping base drive. Although our application only requires 200mA, it is possible to extend the output current to at least 1A by selecting an appropriate ballast resistor and addressing attendant thermal considerations in the pass transistor.

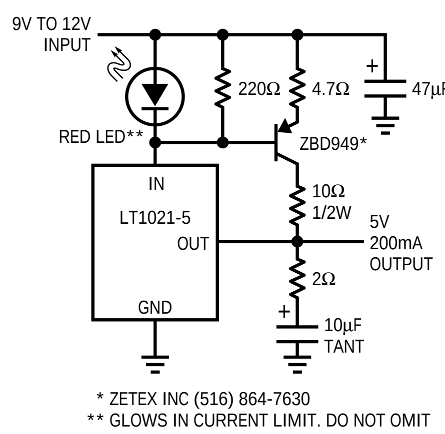


Figure 134. Ultralow Noise 5V, 200mA Supply. Output Noise Is $7\mu V_{RMS}$ Over a 10Hz to 10kHz Bandwidth. Reference Noise Is Guaranteed Less Than $11\mu V_{RMS}$. Standard 3-Terminal Regulators Have 100 Times the Noise and No Guarantees

AN LT1123 ULTRALOW DROPOUT 5V REGULATOR

by Jim Williams and Dennis O'Neill

Switching regulator post regulation, battery-powered apparatus and other applications often require low V_{IN}/V_{OUT} or low dropout linear regulators. For post regulators this is needed for high efficiency. In battery circuits lifetime is significantly affected by regulator dropout. The LT1123, a new low cost reference/control IC, is designed specifically for cost effective duty in such applications. Used in conjunction with a discrete PNP power transistor, the 3-lead TO-92 unit allows very high performance positive regulator designs. The IC contains a 5V bandgap reference, error amplifier, NPN Darlington driver and circuitry for current and thermal limiting.

A low dropout example is the simple 5V circuit of Figure 135 using the LT1123 and an MJE1123 PNP transistor. In operation, the LT1123 sinks Q1 base current through the Drive pin to servo control the FB (feedback) pin to 5V. R1 provides pull-up current to turn Q1 off and R2 is a drive limiter. The 10µF output capacitor (C_{OUT}) provides frequency compensation. The LT1123 is designed to tolerate a wide range of capacitor ESR so that low cost aluminum electrolytics can be be used for C_{OUT}. If the circuit is located more than six inches from the input source, the optional 10µF input capacitor (C_{IN}) should be added.

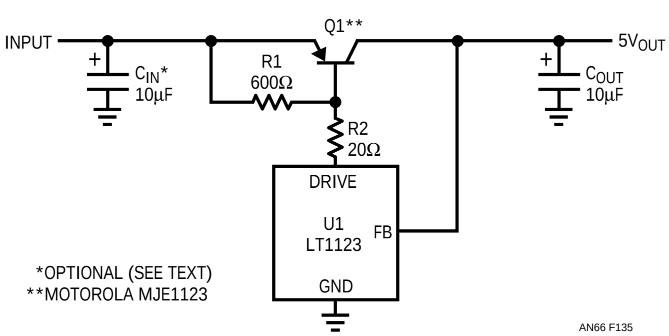


Figure 135. The LT1123 5V Regulator Features Ultralow Dropout

Normally, such configurations require external protection circuitry. Here, the MJE1123 has been cooperatively designed by Motorola and LTC for use with the LT1123. The device is specified for saturation voltage for currents up to 4A with base drive equal to the minimum LT1123 drive current specification. In addition, the MJE1123 is specified for min/max beta at high current. Because of this factor and the defined LT1123 drive, simple current limiting is practical. Excessive output current causes the LT1123 to drive Q1 hard until the LT1123 current limits. Maximum circuit output current is then a product of the LT1123 current and the beta of Q1. The foldback characteristic of the LT1123's drive current combined with the MJE1123 beta and safe area characteristics provide reliable short-circuit limiting. Thermal limiting can also be accomplished by mounting the active devices with good thermal coupling.

Performance of the circuit is notable as it has lower dropout than any monolithic regulator. Line and load regulation are typically within 5mV and initial accuracy is typically inside 1%. Additionally, the regulator is fully short-circuit protected with a no load quiescent current of 1.3mA.

Figure 136 shows typical circuit dropout characteristics in comparison with other IC regulators. Even at 5A the LT1123/MJE1123 circuit dropout is less than 0.5V, decreasing to only 50mV at 1A. Totally monolithic regulators cannot approach these figures, primarily because their power transistors do not offer the MJE1123 combination of high beta and excellent saturation. For example, dropout is ten times lower than in 138 types and significantly better than all the other IC types. Because of Q1's high beta, base drive loss is only 1% to 2% of output current, even at high output currents. This maintains high efficiency under the low V_{IN}/V_{OUT} conditions the circuit will typically see. As an exercise, the MJE1123 was replaced with a 2N4276 germanium device. This provided even lower dropout performance but limiting couldn't be production guaranteed without screening.

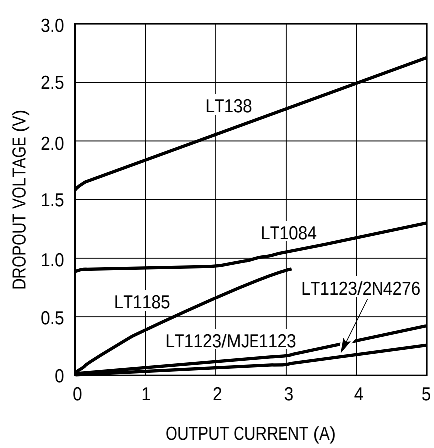


Figure 136. LT1123 Regulator Dropout Voltage vs Output Current

Regulators—Linear (Microprocessor Power)

LT1580 LOW DROPOUT REGULATOR USES NEW APPROACH TO ACHIEVE HIGH PERFORMANCE

by Craig Varga

Enter the LT1580

The 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.

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 1/100 of the output load current or about 70mA for a 7A load.

Circuit Examples

Figure 137 shows a circuit designed to deliver 2.5V from a 3.3V source with 5V available for the control voltage. Figure 138 shows the response to a load step of 200mA to 4.0A. The circuit is configured with a 0.33µF Adjust pin bypass capacitor. The performance without this capacitor is shown in Figure 139. This difference in performance is the reason for providing the Adjust pin on the fixed voltage devices. A substantial savings in expensive output decoupling capacitance may be realized by adding a small "1206-case" ceramic capacitor at this pin.

Figure 140 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.

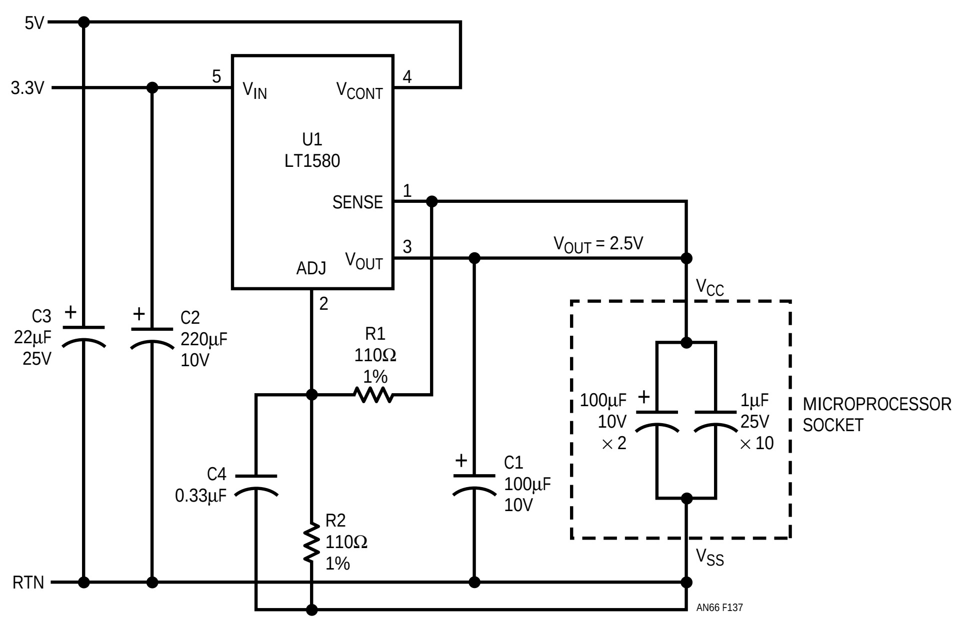


Figure 137. LT1580 Delivers 2.5V from 3.3V at up to 6A

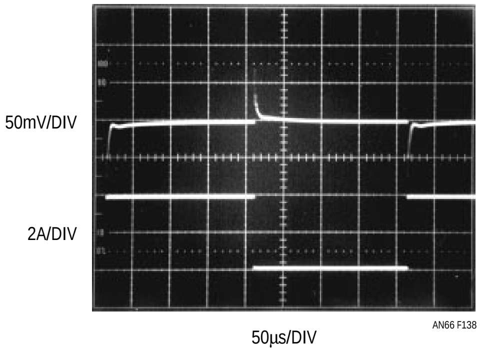


Figure 138. Transient Response of Figure 137's Circuit with Adjust Pin Bypass Capacitor. Load Step Is from 200mA to 4A

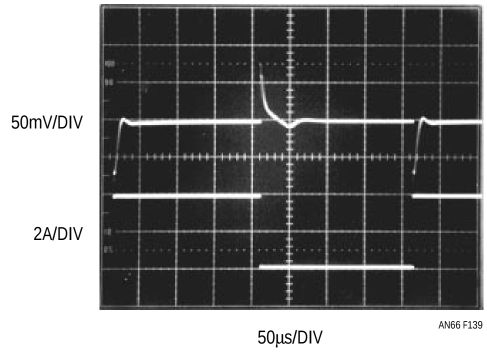


Figure 139. Transient Response Without Adjust Pin Bypass Capacitor. Otherwise, Conditions Are the Same as in Figure 138

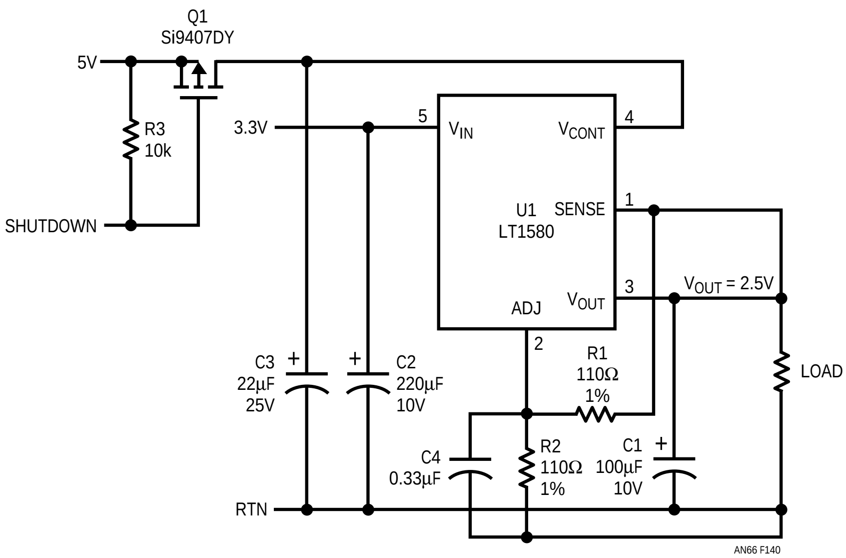


Figure 140. Small FET Adds Shutdown Capability to LT1580 Circuit

LT1585: NEW LINEAR REGULATOR SOLVES LOAD TRANSIENTS

by Craig Varga

The latest hot new microprocessors have added a significant complication to the design of the power supplies that feed them. These devices have the ability to switch from consuming very little power to requiring several amps in tens of nanoseconds. To add a further complication, they are extremely intolerant of supply voltage variations. Gone are the days of the popcorn 3-terminal regulator and the 0.1µF decoupling capacitor. The LT1585 is the first low dropout regulator specifically designed for tight output voltage tolerance (optimized for the latest generation processors) and fast transient response.

Figure 141 shows the kind of response that can and must be achieved if these microprocessors are to operate reliably. Figure 142 details the first several microseconds of

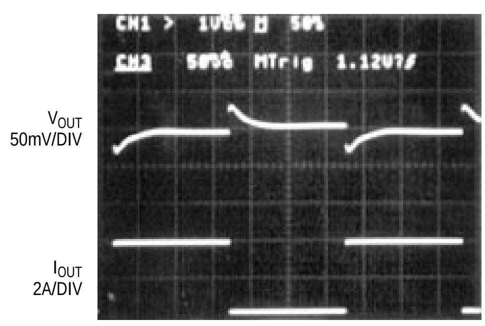


Figure 141. Transient Response of 200mA to 4A Load Step

the transient in Figure 141. The load change in this case is 3.8A in about 20ns. Two parasitic elements dominate the transient performance of the system. Both are controlled by the type, quantity and location of the decoupling capacitors in the system.

Anatomy of a Load Transient

The instantaneous droop at the leading edge of the transient is the result of the sum of the effects of the equivalent series resistance (ESR) and the equivalent series inductance (ESL) from the output capacitor(s) terminal(s) to the load connection. Note that these contributions also include the lead trace parasitics from the capacitor(s) to the load.

The resistive component is simply $\Delta I \bullet ESR$. The droop to point A, 23.6mV, is the ESR contribution. Calculating ESR:

$$23.6mV/3.8A = 0.0062\Omega$$

The effects of inductance are predicted by the formula $V = LdI/dt$. The voltage from point A to the bottom of the trough is the inductive contribution (13.4mV). ESL is calculated to be 0.07nH. After the load current stops rising the inductive effects end, bringing the voltage to point B. At this point the curve settles into a gentle droop. The droop rate is $dV/dt = I/C$. There is about 1300µF of useful capacitance on the board in this case (see Figure 143). As the regulator output current starts to approach the new load current, the droop rate lessens until the regulator supplies the full load current. This is the inflection point in the curve. Since the regulator now measures the output voltage as being too low, it overshoots the load current and recharges the output capacitors to the correct voltage.

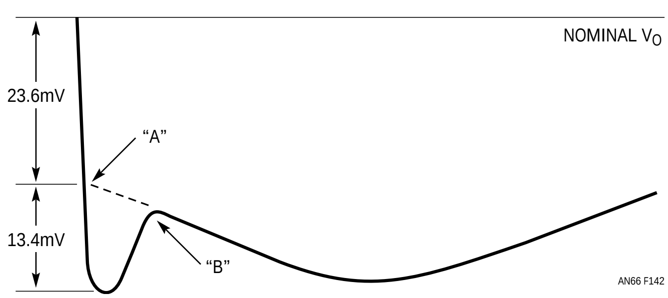


Figure 142. Detailed Sketch of First Few Microseconds of Transients

Faster Regulator Means Fewer Capacitors, Less Board Space

The regulator has one major effect on the system's transient behavior. The faster the regulator, the less bulk capacitance is needed to keep the droop from becoming excessive. It is here that the advantage of the LT1585 shows up. The response time of the LT1585 is about one-half that of the last generation 3-terminal regulators.

The response in the first several hundred nanoseconds is controlled by the careful placement of bypass capacitors. Figure 143 is a schematic diagram of the circuit but the layout is critical so consult the LTC factory for circuit and layout information.

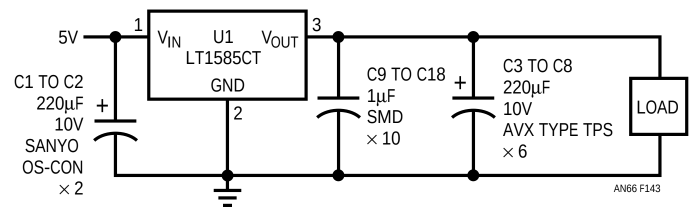


Figure 143. Schematic Diagram: LT1585 Responding to Fast Transients

Battery Chargers

CHARGING NiMH/NiCd OR Li-Ion WITH THE LT1510

by Chia Wei Liao

Charging NiMH or NiCd Batteries

The circuit in Figure 144 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:

$$\frac{2.465}{R1||R2}\ (2000)\ (\text{when Q1 ON})$$

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

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 microcontroller is used to control the charging, connect the DAC current-sink output to the PROG pin.

Charging Li-Ion Batteries

The circuit in Figure 145 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 adapter is unplugged.

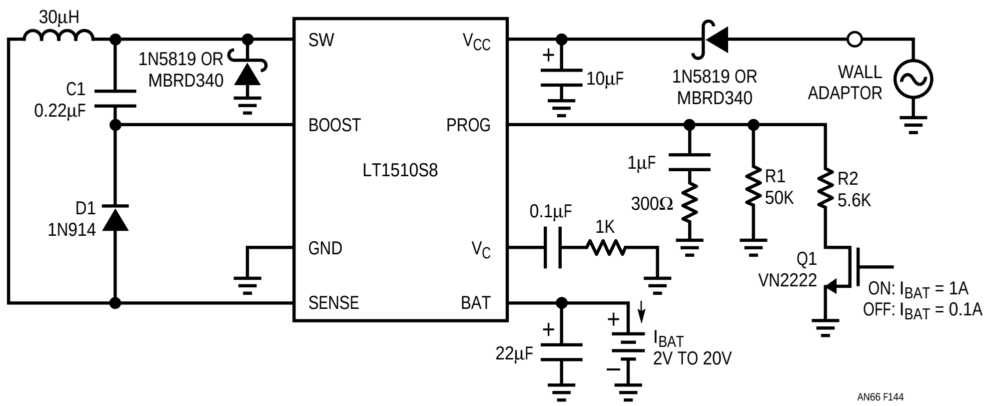


Figure 144. Charging NiCd or NiMH Batteries

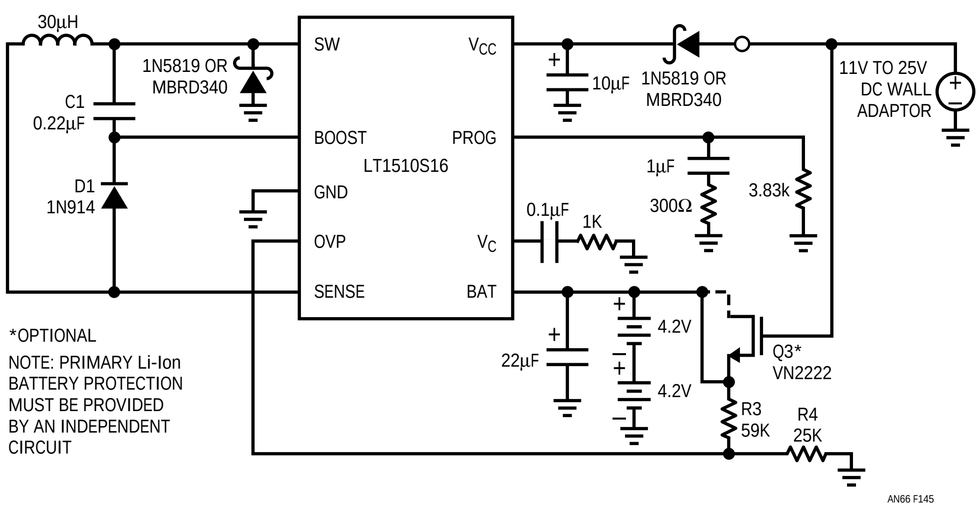


Figure 145. Charging Li-Ion Batteries

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 redistributing 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.

LITHIUM-ION BATTERY CHARGER

by Dimitry Goder

Lithium-ion (Li-Ion) rechargeable batteries are quickly gaining popularity in a variety of applications. The main reasons for the success of Li-Ion cells are higher power density and higher terminal voltage compared to other currently available battery technologies. The basic charging principle for a Li-Ion battery is quite simple: apply a constant voltage source with a built-in current limit. A depleted battery is charged with a constant current until it reaches a specific voltage (usually 4.2V per cell), then it floats at this voltage for an indefinite period. The main difficulty with charging Li-Ion cells is that the floating voltage accuracy needs to be around 1%, with 5% current-limit accuracy. These two targets are fairly difficult to achieve. Figure 146 shows the schematic of a full solution for a Li-Ion charger.

The battery charger is built around the LTC1147, a high efficiency step-down regulator controller. The IC's constant off-time architecture and current mode control ensure circuit simplicity and fast transient response. At the beginning of the ON cycle, P-channel MOSFET Q1 turns on and the current ramps up in the inductor. An internal current comparator senses the voltage proportional to the inductor current across sense resistor R13. When this voltage reaches a preset value, the LTC1147 turns Q1 off for a fixed period of time set by C1. After the off-time the cycle repeats.

To provide an accurate current limit, U3A and Q2 are used to sense the charging current separately from the LTC1147. U3A forces the voltage across R11 to match the average drop across the current sense resistor R13. This voltage sets Q2's drain current, which flows unchanged to the source. As a result the same voltage appears across R9, which is now referenced to ground. Since C5 provides high frequency filtering, U3A shifts the average value of the output current. N-channel MOSFET Q2 ensures correct circuit operation even under short-circuit conditions by allowing current sensing at potentials close to ground.

U3B monitors voltage across R9 and acts to keep it constant by comparing it to the reference voltage. Diode D3 is connected in series with U3B's output, allowing the circuit to operate as a current limiter. The current feedback circuit is not active if the output current limit has not been reached.

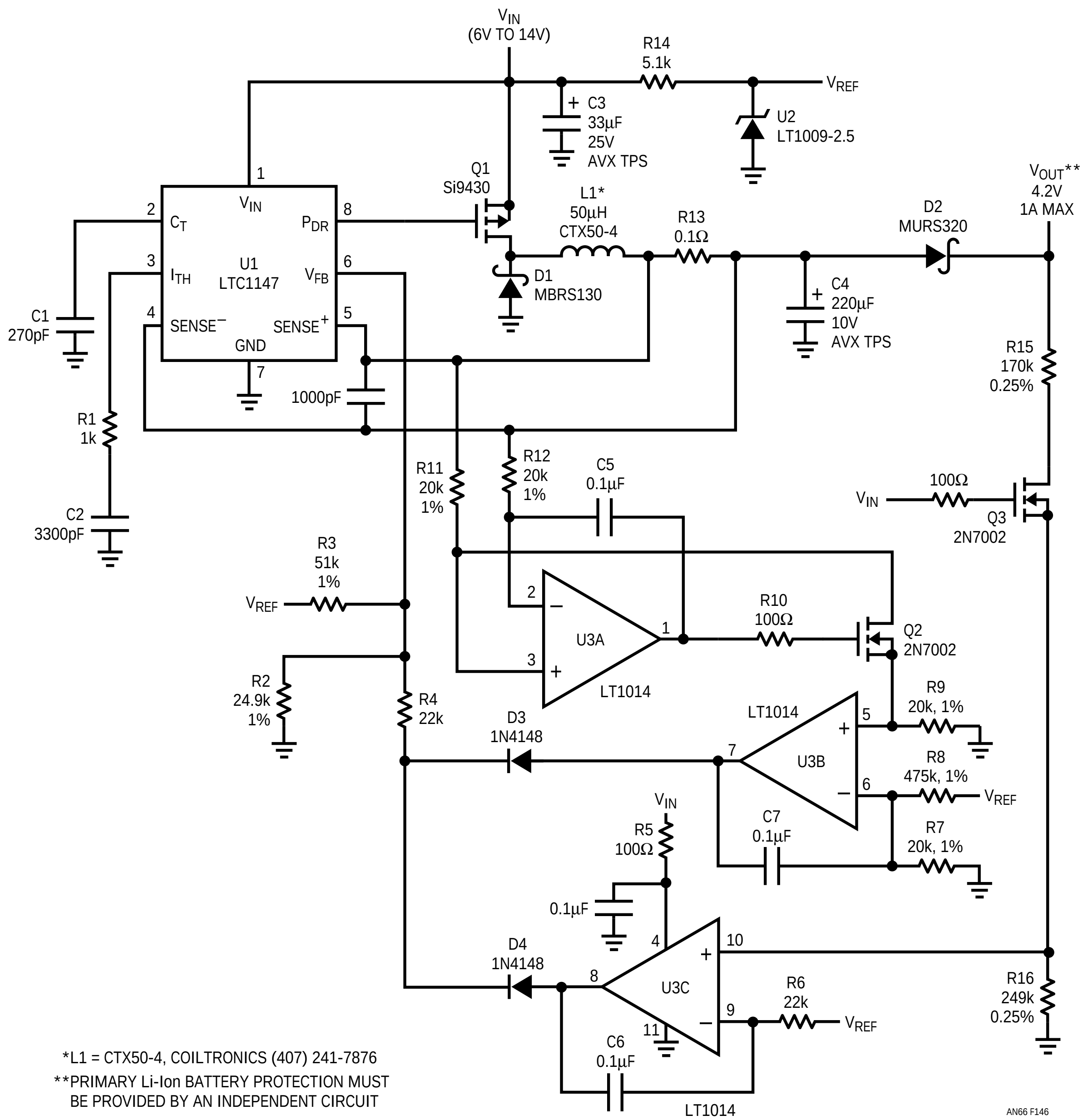


Figure 146. Li-Ion Battery Charger Schematic

U3C provides the voltage feedback by comparing the output voltage to the reference. The feedback resistor ratio [R16/(R15 + R16)] sets the output at exactly 4.2V. U3C has a diode (D4) connected in series with its output. This diode ensures that the voltage and current feedback circuits do not operate at the same time. The reference voltage is supplied by the LT1009, with a guaranteed initial tolerance of 0.2%. Together with the 0.25% feedback resistors, the circuit provides less than 1% output voltage error over temperature.

When the input voltage is not present Q3 is automatically turned off and the feedback resistors do not discharge the battery. Diode D2 is connected in series with the output, preventing the battery from supplying reverse current to the charger.

SIMPLE BATTERY CHARGER RUNS AT 1MHz

by Mitchell Lee

Fast switching regulators have reduced coil sizes to the point that they are no longer the largest components on the board. A case in point is the LT1377, which can operate at 1MHz with inductances under 10µH.

The circuit shown in Figure 147 was designed for a customer who wanted to charge a 4-cell NiCd pack from a 5V logic supply. (This circuit will work equally well with a 3.3V input.) Clearly the circuit needs an output voltage greater than 5V, which is handled easily by the LT1377 boost regulator. The output current is limited to approximately 50mA by a V_{BE} current sensor (Q1/R1) controlling the Feedback pin (2) of the LT1377. This current is perfect for slow charging or trickle charging AA NiCd batteries.

Battery chargers are commonly subject to a number of fault conditions, which must be addressed in the design phase. First, what happens when the battery is disconnected? In a boost regulator the output voltage will increase without bound and blow up either the output capacitor or switch. Some voltage limiting is necessary, and D2 serves this purpose. If the voltage on C3 rises to 11.25V, D2 takes over the control loop at the Feedback pin.

Another potential calamity is an output short circuit; a related fault results from connecting a battery pack containing one or more shorted cells, such that the terminal voltage is less than about 4V. Under either of these circumstances, unlimited current flows from the 5V input supply, through D1 and Q1's base-emitter junction, frying at least Q1.

Q2 has been added to allow full current control even when the output voltage is less than the input voltage. In normal operation, where the output is boosted higher than 5V, Q2 is fully on. Its gate is held at 1.25V (Pin 2 feedback voltage) and its source is greater than 5V; hence it has no choice but to be fully enhanced. Q2 becomes more functional when the output voltage drops to around 4V. First of all, at 4V input the switching regulator stops switching because more than 50mA current flows and the feedback pin is pulled up above 1.25V—Q1 makes sure of that. But as Q1's collector continues to rise, Q2 is gradually cut off, at least to the extent necessary to starve the drain current back to about 50mA. This action works right down to V_{OUT} = 0. In a short circuit Q2 dissipates about 200mW, not too much for a surface mount MOSFET.

This circuit is useful for four to six cells and the output current can be modified somewhat by changing sense resistor R1. A reasonable range is from very low currents (1mA or less) up to 100mA. The current will diminish as Q1's V_{BE} drops about 0.3%/°C with temperature.

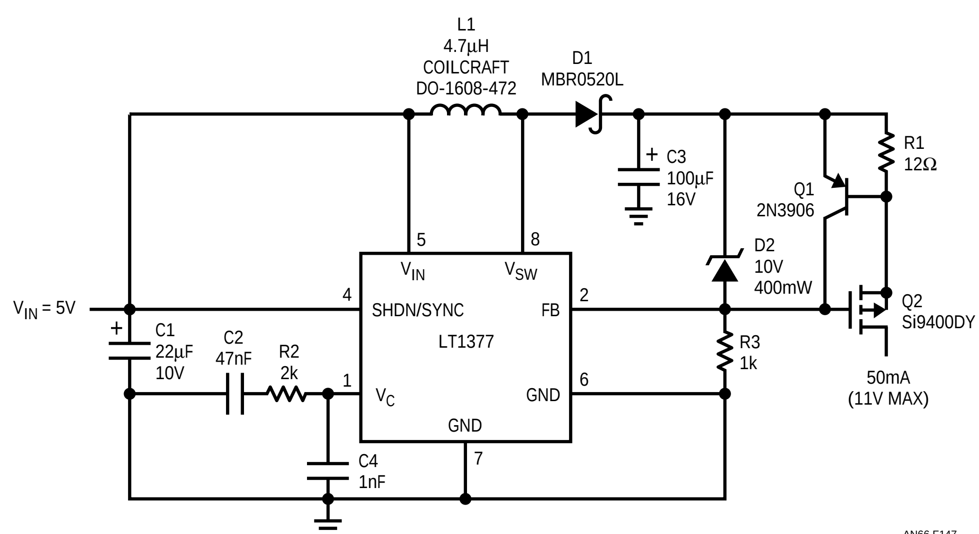


Figure 147. Battery Charger Schematic Diagram

A PERFECTLY TEMPERATURE-COMPENSATED BATTERY CHARGER

by Mitchell Lee and Kevin Vasconcelos

Battery charging circuits are usually greeted with a yawn, but this lead-acid charger offers a combination of features that sets it apart from all others. It incorporates a low dropout regulator, temperature compensation, dual-rate charging, true negative ground and consumes zero standby current.

The LT1083 family of linear regulators exhibits dropout characteristics of less than 1.5V as compared to 2.5V in standard regulators. A smaller regulator drop allows for lower input voltages and less power dissipation in the regulator. In this application the regulator is used to control charging voltage and limit maximum charging current.

The temperature compensation employed in this circuit, unlike diode-based straight-line approximations, follows the true curvature of a lead-acid cell. This prevents over or undercharging of the battery during periods of extended low or high ambient temperatures. Temperature compensation is conveniently provided by a Tempsistor® as shown in Figure 148. The Tempsistor is used to generate a temperature-dependent current, which, in turn, adjusts the charger's output voltage to match that of the battery. The match is within 100mV for a 12V battery over a range of –10°C to 60°C. The best place for the Tempsistor is directly under the battery with the battery resting on a pad of styrofoam.

Tempsistor is a registered trademark of Thermodisc Inc.

Q1 provides a low voltage disconnect function that reduces the charger standby current to zero. When the input voltage (from a rectified transformer) is available, Q1 is biased ON and Q2 is turned ON. Q2 connects the various current paths on the output of the regulator to ground, activating the charging circuitry. If the input voltage is removed, Q1 and Q2 turn off, and all current paths from the battery to ground (except for the load, of course) are interrupted. This prevents unnecessary battery drain when the charging source is not available.

A dual-rate charging characteristic is achieved by means of a current-sense resistor (R_S) and a sense comparator (LT1012). If the battery charge current exceeds the float-current threshold of $10mV/R_S$, the comparator pulls the gate of Q3 low, increasing the output voltage by 600mV. This sets the charging voltage to 14.4V at 25°C. After the battery reaches full charge the current will fall below the $10mV/R_S$ threshold and the LT1012 will short out R7, reducing the output by 600mV to a float level of 13.8V.

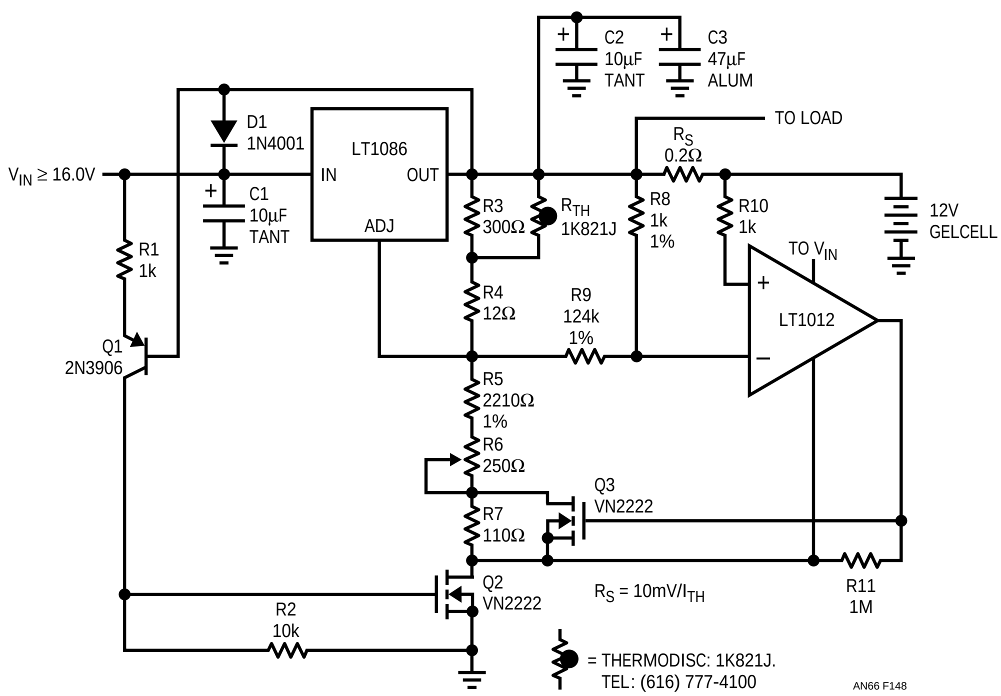


Figure 148. Battery Charger Follows Temperature Coefficient of a Lead-Acid Cell Very Accurately

Both the float and charging voltages can be trimmed by R6; R7 sets the 600mV difference between them.

With the charging source connected, the sense resistor R_S measures only battery current. This eliminates the tendency found in some schemes for the charger to trip on load current.

Table 1 simplifies the selection of an appropriate regulator for batteries of up to 48 Ampere-hours (Ah). The selection is based on providing a minimum available charge current of at least C/4, where C represents the battery's Ampere-hour capacity. The next larger regulator may be required in applications where sustained load currents of greater than C/10 are expected.

If you want to set the trip current to an exact figure, the current shunt R_S can be calculated as $R_S = 10mV/I_{TH}$. For a threshold of C/100 this reduces to $R_S = 1/C$.

Table 1. The Regulator Should Be Chosen to Provide at Least C/4 Charging Current

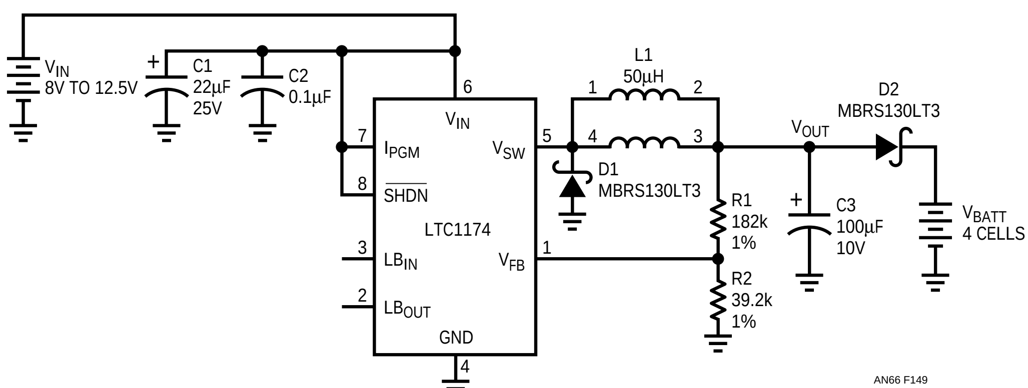
BATTERY CAPACITY	DEVICE	MAXIMUM CHARGING CURRENT	FLOAT CURRENT THRESHOLD	SENSE RESISTOR (SHUNT)
≤3Ah	LT1117	0.8A	20mA	0.5Ω
3Ah to 6Ah	LT1086	1.5A	50mA	0.2Ω
6Ah to 12Ah	LT1085	3.2A	100mA	0.1Ω
12Ah to 24Ah	LT1084	5.5A	200mA	0.05Ω
24Ah to 48Ah	LT1083	8.0A	400mA	0.025Ω

A SIMPLE 300mA NiCd BATTERY CHARGER

by Randy G. Flatness

Low current battery charger circuits are required in handheld products such as palmtop, pen-based and fingertip computers. The charging circuitry for these applications must use surface mount components and consume minimal board space. The circuit shown in Figure 149 meets both of these requirements.

The circuit shown in Figure 149 uses an LTC1174 to control the charging circuit. A fully self-contained switching regulator IC, the LTC1174 contains both a power switch and the control circuitry (constant off-time controller, reference voltage, error amplifier and protection circuitry). The internal power switch is a P-channel MOSFET transistor in a common-source configuration; consequently when the switch turns on, the LTC1174's V_{SW} pin is



C1 = AVX (TA) TPSD226M025R0200 ESR = 0.200 I_{RMS} = 0.775A
C3 = AVX (TA) TPSD107M010R0100 ESR = 0.100 I_{RMS} = 1.095A
D1, D2 = MOTOROLA SCHOTTKY VBR = 30V
L1 = COILTRONICS CTX50-2P DCR = 0.212 IDC = 0.729A TYPE 52 CORE
COILTRONICS (407) 241-7876

$V_{OUT} = 1.25V \cdot (1 + R1/R2) = 7.0V$

$$\text{FAST CHARGE} \approx 0.6A - \frac{(V_{BATT} + 0.6V) \cdot 4\mu s}{2 \cdot L} \quad \text{(EQ.1)}$$

Figure 149. 4-Cell, 300mA LTC1174 Battery Charger Implemented with All Surface Mount Components

connected to the input voltage. This power switch handles peak currents of 600mA. The LTC1174's architecture allows it to achieve 100% duty cycle, forcing the internal P-channel MOSFET on 100% of the time.

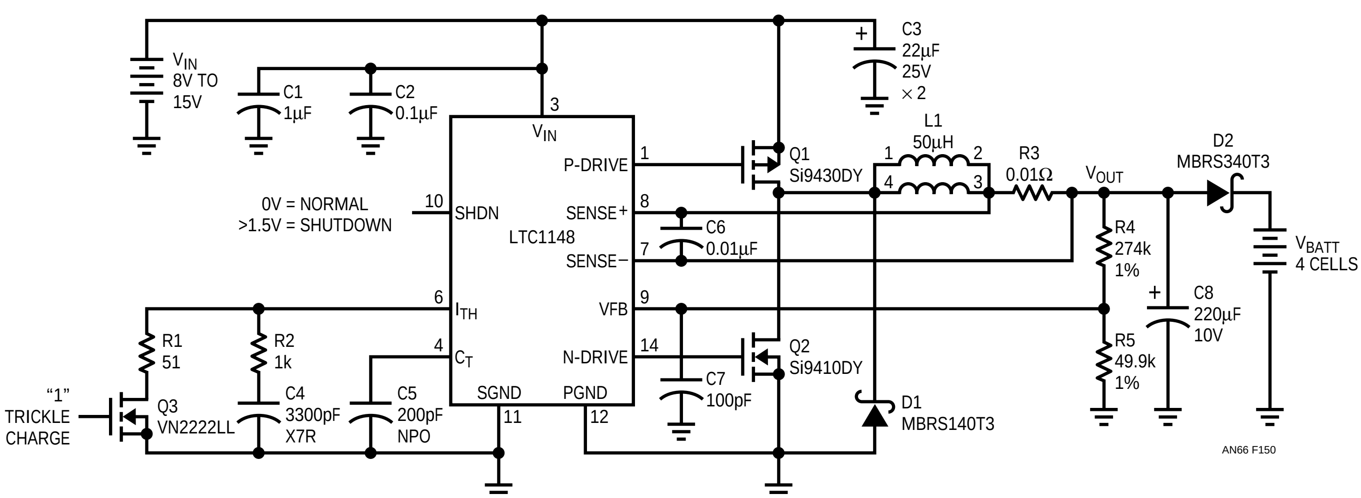
When the batteries are being charged, the resistor divider network (R1 and R2) forces the LTC1174's Feedback pin (V_{FB}) below 1.25V, causing the LTC1174 to operate at the maximum output current. An internal 0.1Ω resistor senses this current and sets it at approximately 300mA, according to equation 1 (shown on the schematic). When the batteries are disconnected, the error amplifier drives the Feedback pin to 1.25V, limiting the output voltage to 7.0V. Diode D2 prevents the batteries from discharging through the divider network when the charger is shut down. In shutdown mode less than 10µA of supply current is drawn from the input supply.

HIGH EFFICIENCY (>90%) NiCd BATTERY CHARGER CIRCUIT PROGRAMMABLE FOR 1.3A FAST CHARGE OR 100mA TRICKLE CHARGE

by Brian Huffman

Battery charger circuits are of universal interest to laptop, notebook and palmtop computer manufacturers. High efficiency is desirable in these applications to minimize the power dissipated in the surface mount components. The circuit shown in Figure 150 is designed to charge four NiCd cells at a 1.3A fast charge or a 100mA trickle charge with efficiency exceeding 90%. This circuit can be modified easily to handle up to eight NiCd cells.

The circuit uses an LTC1148 in a step-down configuration to control the charge rate. The LTC1148 is a synchronous switching regulator controller that drives external, complementary power MOSFETs using a constant off-time current mode architecture. When the LTC1148's P-drive output pulls the gate of Q1 low, the P-channel MOSFET turns on and connects one side of the inductor to the input voltage. During this period, current flows from the input through Q1,



C1 = (TA)
C3 = AVX (TA) TPSD226K025R0200 ESR = 0.200 IRMS = 0.775A
C8 = AVX (TA) TPSE227M010R0100 ESR = 0.100 IRMS = 1.149A
Q1 = SILICONIX PMOS BVDSS = 20V RDS_{ON} = 0.125 C_{RSS} = 400pF Q_G = 25nC θJA = 50°C/W
Q2 = SILICONIX NMOS BVDSS = 30V RDS_{ON} = 0.050 C_{RSS} = 160pF Q_G = 50nC θJA = 50°C/W
D1, D2 = MOTOROLA SCHOTTKY VBR =40V
R3 = KRL SP-1/2-A1-0R100J Pd = 0.75V
L1 = COILTRONICS CTX50-4 DCR = 0.175 IDC = 1.350A KOOL Mµ CORE

ALL OTHER CAPACITORS ARE CERAMIC

COILTRONICS (407) 241-7876
KRL (809) 668-3210

V_{OUT} = 1.25V • (1 + R4/R5) = 8.1V

FAST CHARGE = 130mV/R3 = 1.3A (EQ. 1)
TRICKLE CHARGE = 100mA (SEE FIGURE 2)

Figure 150. 4-Cell, 1.3A Battery Charger Implemented in Surface Mount Technology

through the inductor and into the battery. When the LTC1148 P-drive pin goes high, Q1 is turned off and the voltage on the drain of Q1 drops until the clamp diode is forward biased. The diode conducts for a very short period of time, until the LTC1148 internal circuitry senses that the P-channel is fully off, preventing the simultaneous conduction of Q1 and Q2. Then the N-drive output goes high, turning on Q2, which shorts out D1. Now the inductor current flows through the N-channel MOSFET instead of through the diode, increasing efficiency. This type of switching architecture is known as synchronous rectification.

During the fast-charge interval, the resistor divider network (R4 and R5) forces the LTC1148's Feedback pin (V_{FB}) below 1.25V, causing the LTC1148 to operate at the maximum output current. R3, a 0.1Ω resistor, senses the current and sets it at approximately 1.3A according to equation 1 in Figure 150. When the batteries are disconnected, the error amplifier forces the Feedback pin to 1.25V, limiting the output voltage to 8.1V. Diode D2 prevents the batteries from discharging through the divider network when the charger is shut down. In shutdown mode the circuit draws less than 20µA from the input supply.

The dual rate charging is controlled by Q3, which can be toggled between fast charge and trickle charge. The trickle charge rate is set by resistor R1. Figure 151 is a graph showing the value of R1 for a given trickle charge output current. The trickle charge current can also be varied by using an op amp to force the Threshold pin voltage within its 0V to 2V range. Figure 152 shows the output current as a function of Threshold pin voltage.

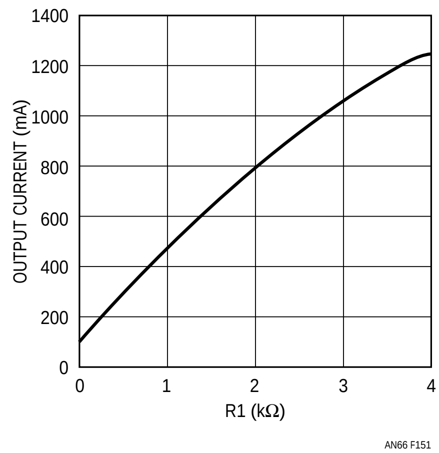


Figure 151. LTC1148 Output Current Voice Trickle Charge Set Resistance (R1)

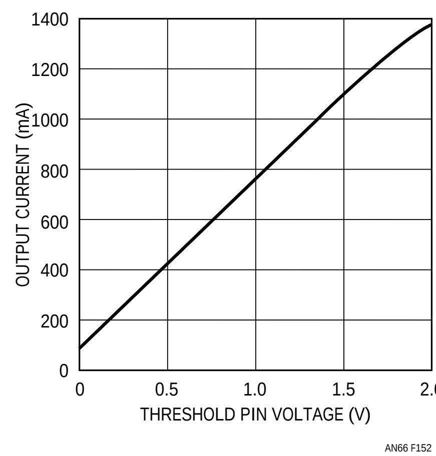


Figure 152. LTC1148 Output Current vs Forced Threshold Pin Voltage

Power Management

LT1366 RAIL-TO-RAIL AMPLIFIER CONTROLS TOPSIDE CURRENT

by William Jett and Sean Gold

Topside Current Source

The circuit shown in Figure 153 takes advantage of the LT1366's rail-to-rail input range to form a wide-compliance current source. The LT1366 adjusts Q1's gate voltage to force the voltage across the sense resistor (R_{SENSE}) to equal the voltage from the supply to the potentiometer's wiper. A rail-to-rail op amp is needed because the voltage across the sense resistor must drop to zero when the divided reference voltage is set to zero. Q2 acts as a constant current sink to minimize error in the reference voltage when the supply voltage varies.

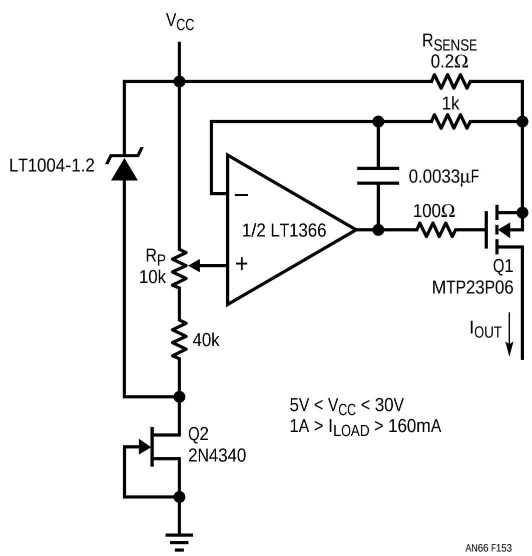


Figure 153. Topside Current Source

The circuit can operate over a wide supply range (5V < V_{CC} < 30V). At low input voltage, circuit operation is limited by the MOSFET's gate-drive requirements. At high input voltage, circuit operation is limited by the LT1366's absolute maximum ratings and the output power requirements.

In this example the circuit delivers 1A at 200mV of sense voltage. With a 5V input supply the power dissipation is 5W. For operation at 70°C ambient temperature, the MOSFET's heat sink must have a thermal resistance of:

$$\theta_{HS} = \theta_{JA\ SYSTEM} - \theta_{JC\ FET} = 55°C/5W - 1.25°C/W = 9.75°C/W$$

This is easily achievable with a small heat sink. When input voltages are greater than 5V the use of a larger heat sink or derating of the output current is necessary.

The circuit's supply regulation is about 0.03%/V. The output impedance is equal to the MOSFET's output impedance multiplied by the op amp's open-loop gain. Degradations in current source compliance occur when the voltage across the MOSFET's on-resistance and the sense resistor drops below the voltage required to maintain the desired output current. This condition occurs when $V_{CC} - V_{OUT} < I_{LOAD}(R_{SENSE} + R_{ON})$.

High Side Current Sense Amplifier

In power control it is sometimes necessary to sense load current at low loss near the input supply. The current sense amplifier shown in Figure 154 amplifies the voltage across a small value sense resistor by the ratio of the current source resistors (R2/R1). The LT1366 forces the low power MOSFET's gate voltage such that the sense voltage appears across a current source resistor R1. The resulting current in Q1's drain is converted to a ground referred voltage at R2. ($V_O = I_{IN} R_S$ [R2/R1])

The circuit takes advantage of the LT1366's ability to sense signals up to the supply rail, which permits the use of small value, low loss sense resistors. The LT1366 and the gain setting resistors are also biased at low current to reduce losses in the current sense.

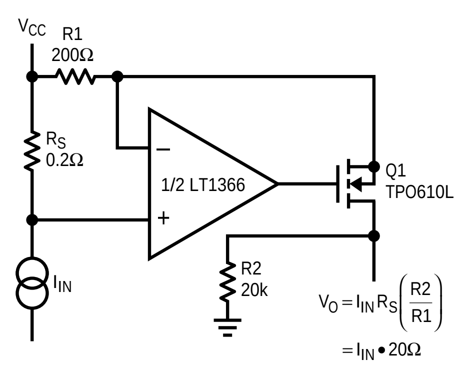


Figure 154. High Side Current Sense Amplifier

AN ISOLATED HIGH SIDE DRIVER

by James Herr

Introduction

The LTC1146 low power digital isolator draws only 70μA of supply current with V_{IN} = 5V. Its low supply current feature is well suited for battery-powered systems that require isolation, such as an isolated high side driver. The LTC1146A is rated at $2500V_{RMS}$ and is UL approved. The LTC1146 is intended for less stringent applications and is rated at 500VDC.

Theory of Operation

Optoisolators available today require supply currents in the milliampere range even for low speed operation (less than 20kHz). This high supply current is another drain on the battery. Figure 155 shows the alternative of using an LTC1146A to drive an external power MOSFET (IRF840) at speeds to 20kHz with V^+ = 300V.

The Input pin of LTC1146A must be driven with a signal that swings at least 3V (referred to GND1, which is a floating ground). The O_S pin outputs a square wave corresponding to the input signal but with a time delay. The amplitude of the output square wave is equal to the potential at the V_{CC} pin. The TL4426 is a high speed MOSFET driver used here to supply gate drive current to the power MOSFET. The power supply to the LTC1146A and the TL4426 is bootstrapped from a 13V supply referred to system ground. C1 supplies the current to both the LTC1146A and the TL4426 when the power MOSFET is being turned on. Its value should be increased when the input signal's ON time increases. D3 prevents the output from swinging negative due to stray inductance. If the output goes below ground, the gate-to-source voltage of the IRF840 rises. This high potential could damage the power MOSFET. The output slew rate should be limited to 1000V/μs to prevent glitches on the O_S output of the LTC1146A.

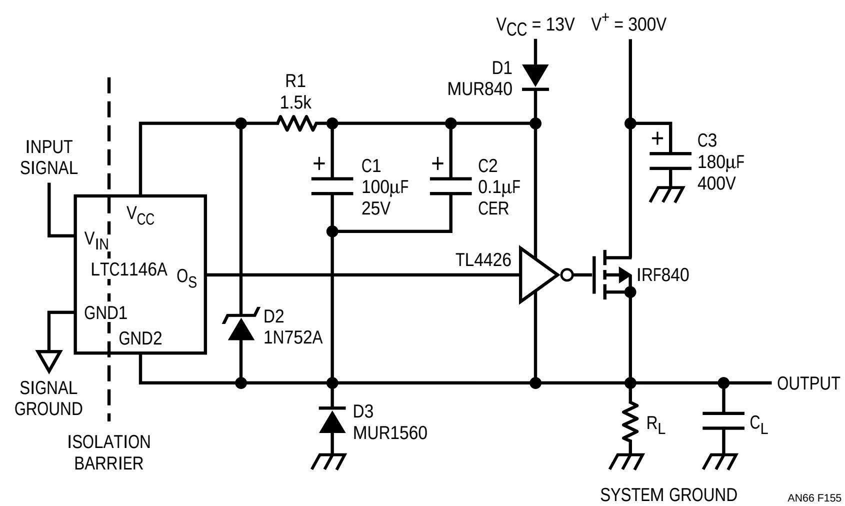


Figure 155. Isolated High Side Driver Schematic Diagram

LTC1163: 2-CELL POWER MANAGEMENT

by Tim Skovmand

The LTC1163 1.8V to 6V high side MOSFET driver allows inexpensive N-channel switches to be used to efficiently manage power in 2-cell systems such as palmtop computers, portable medical equipment, cellular telephones and personal organizers.

Any supply voltage above 3V, such as 3.3V, 5V or 12V, can be generated by step-up converters powered from a 2-cell supply. Step-up regulators are typically configured as shown in Figure 156. An inductor is connected directly to the 2-cell battery pack and switched by a large (1A) switch. The inductor current is then passed through a low drop Schottky rectifier to charge the output capacitor to a voltage higher than the input voltage. Unfortunately, when the regulator is shut down, the inductor and diode remain connected and the load may leak significant current in standby.

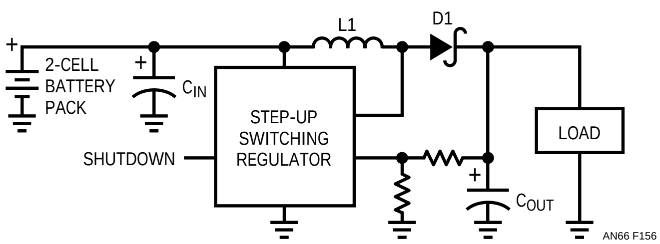


Figure 156. Typical Step-Up Converter Topology

One possible solution to this problem is to add a low $R_{DS(ON)}$ MOSFET switch between the battery pack and the input of the regulator to completely disconnect it and the load from the battery pack. MOSFET switches, however, cannot operate directly from 2-cell battery supplies because the gate voltage is limited to 3V with fresh cells and 1.8V when the cells are fully discharged.

The LTC1163 solves this problem by generating gate drive voltages that fully enhance high side N-channel switches when powered from a 2-cell battery pack, as shown in Figure 157. The standby current with all three drivers switched off is typically 0.01µA. The quiescent current rises to 85µA per channel with the input turned on and the charge pump producing 10V (above ground) from a 3V supply. The surface mount MOSFET switches shown are guaranteed to be less than 0.1Ω with V_{GS} = 5V and less than 0.12Ω with V_{GS} = 4V and therefore have extremely low voltage drops.

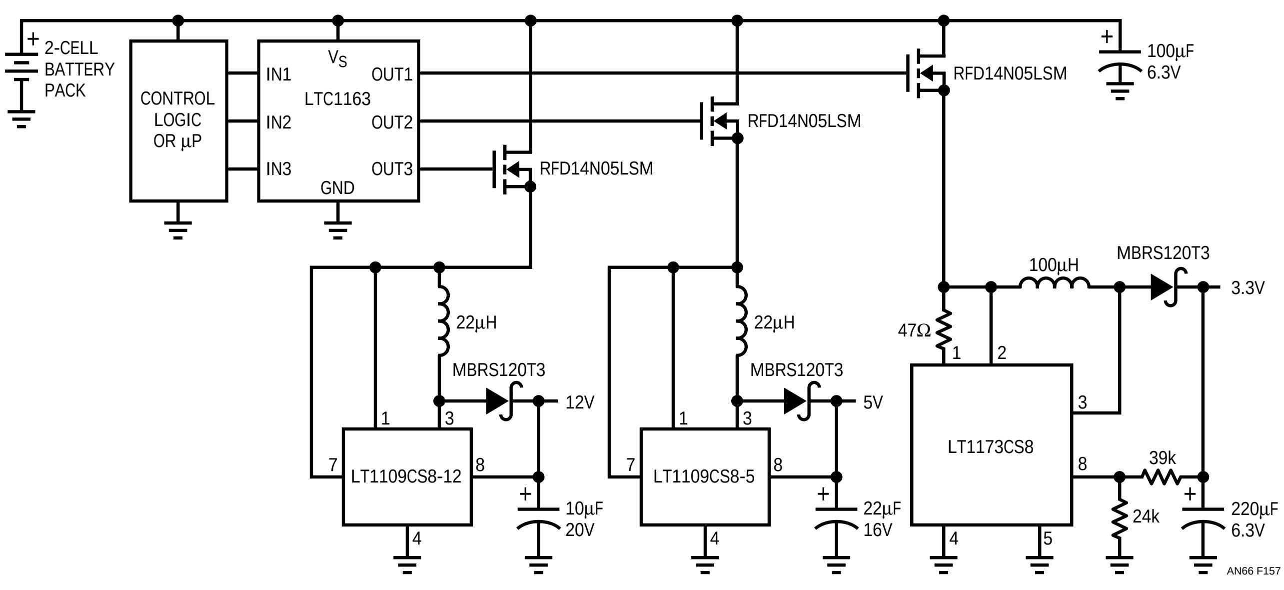


Figure 157. Complete 2-Cell to 3.3V, 5V and 12V Power Management System

LTC1157 SWITCH FOR 3.3V PC CARD POWER

by Tim Skovmand

Computers designed to accept PC cards—plug-in modules specified by the Personal Computer Memory Card International Association (PCMCIA)—have special hardware features to accommodate these pocket-sized cards. PCMCIA-compliant cards require power management electronics that conform to the height restrictions of the three standard configurations: 3.3mm, 5mm and 10.5mm. These height limitations dramatically reduce the available options for power management on the card itself. For example, high efficiency switching regulators to convert the incoming 5V down to 3.3V for the on-card 3.3V logic require relatively large magnetics and filter capacitors, which are not always available in packaging thin enough to meet the tight height requirements.

One possible approach to the problem of supplying power to a 3.3V PC card is to switch the input supply voltage from 5V to 3.3V after the card has been inserted and the attribute ROM has informed the computer of the card's voltage and current requirements. The switching regulator, housed in the computer, switches the power supplied to the connector from 5V to 3.3V.

A window comparator and ultralow drop switch on the PC card, Q1 in Figure 158, closes after the supply voltage drops from 5V to 3.3V, ensuring that the sensitive 3.3V logic on the card is never powered by more than 3.6V or less than 2.4V. A second switch, Q2, is provided on the card to interrupt power to 3.3V loads that can be idled when not in use.

The built-in charge pumps in the LTC1157 drive the gates of the low $R_{DS(ON)}$ N-channel MOSFETs to 8.7V when powered from a 3.3V supply. The LT1017 and the LTC1157 are both micropower and are supplied by a filter (R5 and C2) that holds the supply pins high long enough to ensure that the MOSFET gates are fully discharged immediately after the card is disconnected from the power supply. A large value bleed resistor, R6, further ensures that the high impedance gate of Q1 is not inadvertently charged up when the card is removed or when it is stored.

All of the components shown in Figure 158 are available in thin, surface mount packaging and occupy a very small amount of surface area. Further, the power dissipation is extremely low because the LTC1157 and LT1017 are micropower and the MOSFET switches are very low $R_{DS(ON)}$.

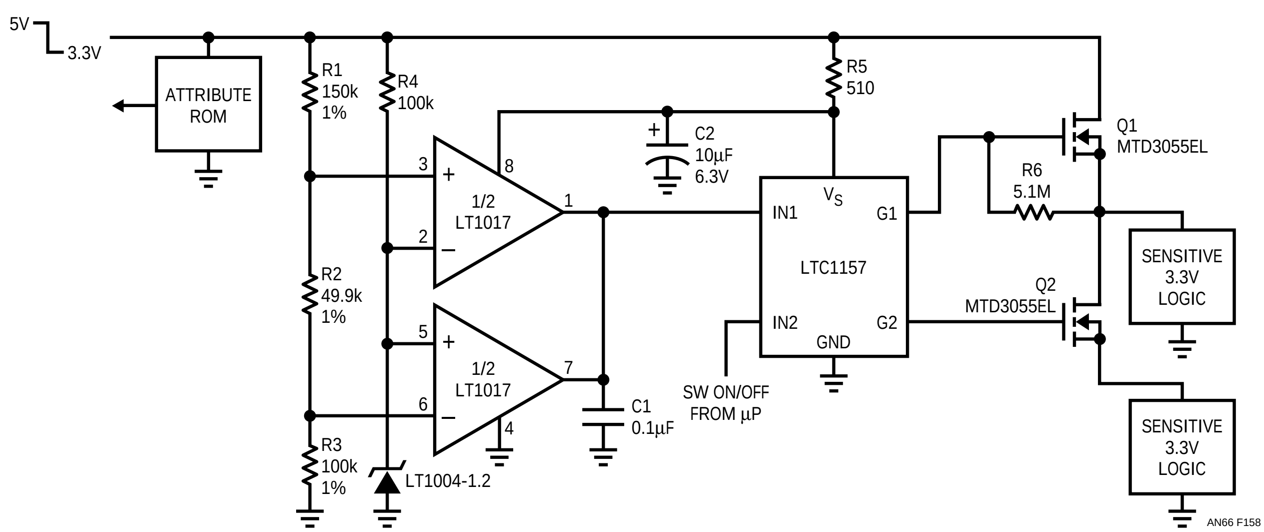


Figure 158. 3.3V PCMCIA Card Power Switching

THE LTC1157 DUAL 3.3V MICROPOWER MOSFET DRIVER

by Tim Skovmand

The LTC1157 dual micropower MOSFET driver makes it possible to switch either supply- or ground-referenced loads through a low $R_{DS(ON)}$ N-channel switch. The LTC1157's internal charge pump boosts the gate drive voltage 5.4V above the positive rail (8.7V above ground), fully enhancing a logic level, N-channel MOSFET for 3.3V high side switching applications.

LTC1157 Switches Two 3.3V Loads

Figure 159 illustrates how two surface mount MOSFETs and the LTC1157 (also available in SO-8 packaging) can be used to switch two 3.3V loads. The gate rise and fall times are typically in the tens of microseconds, but can be slowed by adding two resistors and a capacitor as shown on the second channel. Slower rise and fall times are sometimes required to reduce the start-up current demands of large supply capacitors which might otherwise glitch the main supply.

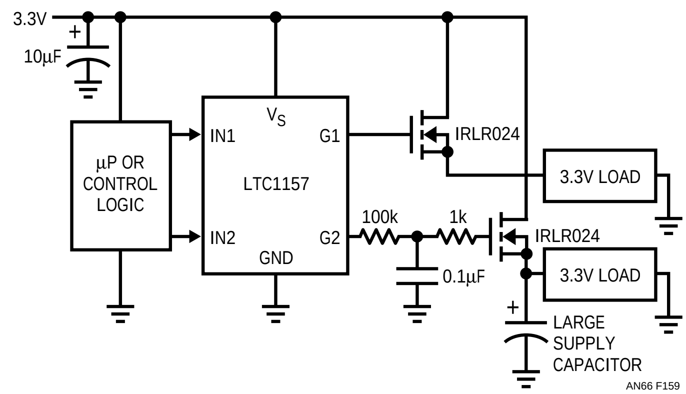


Figure 159. LTC1157 Used to Switch Two 3.3V Loads

THE LTC1155 DOES LAPTOP COMPUTER POWER BUS SWITCHING, SCSI TERMINATION POWER OR 5V/3A EXTREMELY LOW DROPOUT REGULATOR

by Tim Skovmand

The LTC1155 is a new micropower MOSFET driver specifically designed for low voltage, high efficiency switching applications such as those found in laptop or notebook computers. Three applications for this versatile part are detailed here.

Figure 160 is a schematic diagram that demonstrates the use of the LTC1155 for switching the power buses in a laptop computer system. The disk drive, display, printer and the microprocessor system itself are selectively engaged via high side switching with minimum loss and are shut down completely when not in use.

The quiescent current of the LTC1155 is designed to be extremely low in both the OFF and ON states, so that efficiency is preserved even when driving loads that require very little current to operate in standby, but require much larger peak currents when in operation. This combination of a low $R_{DS(ON)}$ MOSFET and an efficient driver delivers the maximum energy to the load.

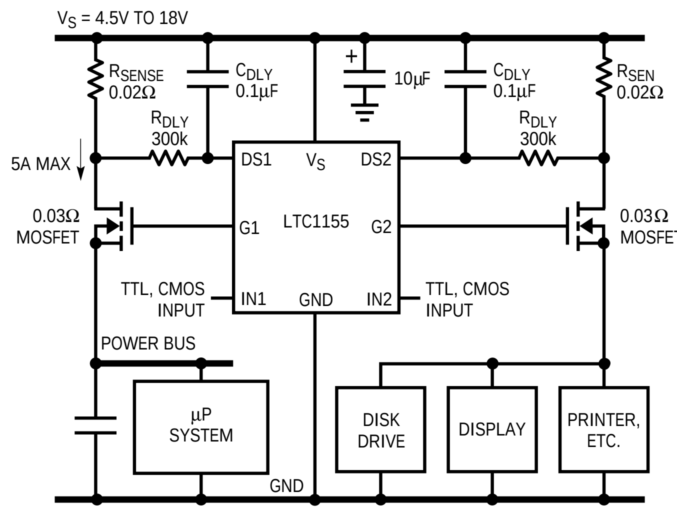


Figure 160. Laptop Computer Power Bus Switching

Protected SCSI Termination Power

The circuit shown in Figure 161 demonstrates how the LTC1155 provides protected power to SCSI terminators. The LTC1155 is initially triggered by the free-running 1Hz oscillator (it could also be triggered by a pulse from the microprocessor) and latches ON via the positive feedback provided by R_{FB}. The power MOSFET gate is driven to 12V and the MOSFET is fully enhanced.

The delay afforded by the two delay components, R_{DLY} and C_{DLY}, ensures that the protection circuit is not triggered by a high inrush-current load. If, however, the source of the MOSFET is shorted to ground or if the output of LT1117 is shorted, the delay will be exceeded and the MOSFET will be held OFF until the pulse from the free-running oscillator resets the input again. The drain sense resistor, R_{SENSE}, is selected to trip the LTC1155 protection circuitry when the MOSFET current exceeds 1A. This current limit protects both the LT1117 and any peripheral system powered by the SCSI termination power line.

The delay time afforded by R_{DLY} and C_{DLY} is chosen to be considerably smaller than the reset time period (>100:1), so that very little power is dissipated while the short-circuit condition persists, i.e., the LTC1155 will deliver small pulses of current during every reset time period until the short-circuit condition is removed.

The LTC1155 and the LT1117, as well as the power MOSFET shown, are available in surface mount packaging and therefore consume very little board space.

Extremely Low Voltage Drop Regulator

An extremely low voltage drop regulator can be built around the LTC1155 and a low resistance power MOSFET as shown in Figure 162. The LTC1155 charge pump boosts the gate voltage above the supply rail and continuously charges a 0.1µF reservoir capacitor. The LT1431 works against this capacitor and the 100k series resistor to control the MOSFET gate voltage and maintain a constant 5V at the output.

The regulator is switched ON and OFF by the control logic or the microprocessor to conserve power in the standby mode. The LTC1155 standby current drops to about 10µA when the input is switched OFF. The total ON current, including the LT1431 is less than 1mA.

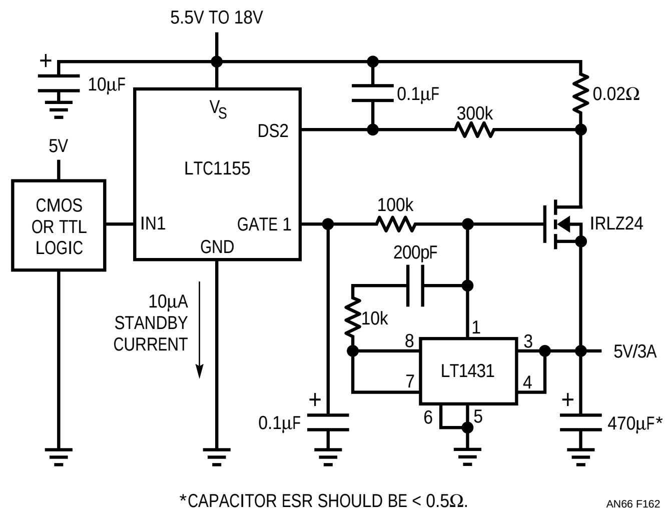


Figure 162. 5V/3A Extremely Low Voltage Drop Regulator

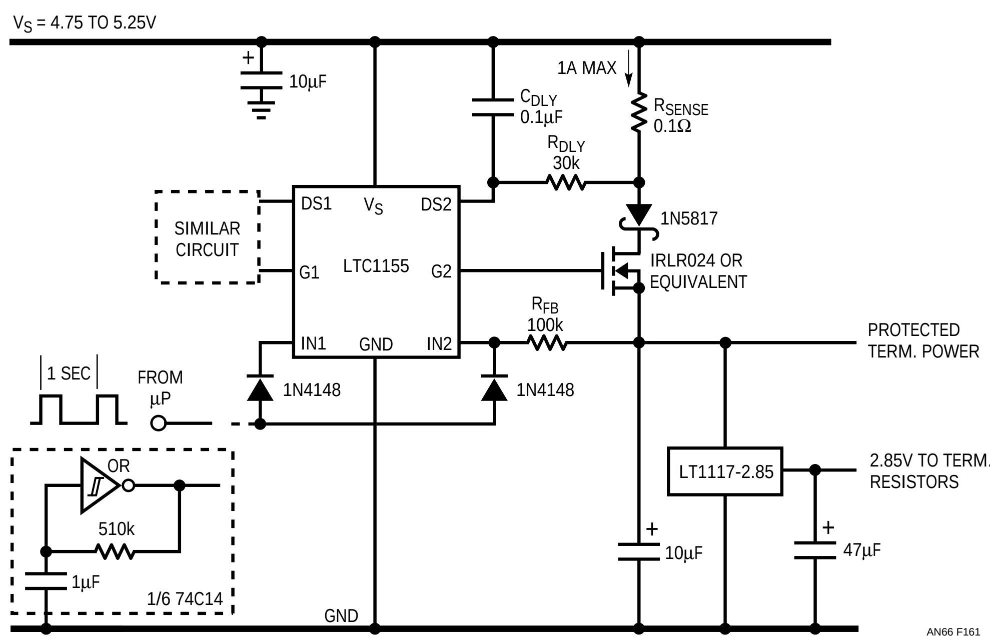


Figure 161. SCSI Termination Power with Short-Circuit Protection

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 163 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 163 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 163 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.

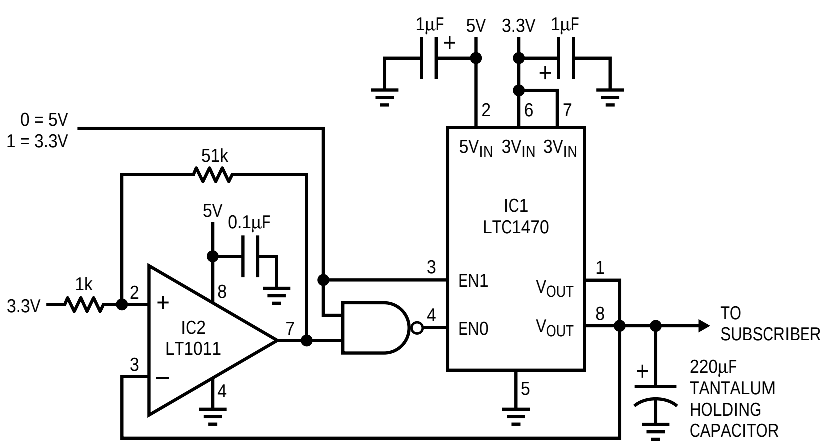


Figure 163. Schematic Diagram of 3.3V and 5V Switchover Circuit

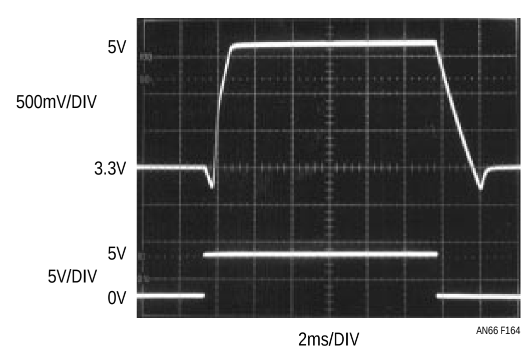


Figure 164. Oscillograph of the Switchover Waveform Showing Smooth Transitions

A FULLY ISOLATED QUAD 4A HIGH SIDE SWITCH

by Milton Wilcox

High side switching in hostile environments often requires isolation to protect the controlling logic from transients on the "dirty" power ground. The circuit shown in Figure 165 drives and protects four low $R_{DS(ON)}$ power MOSFET switches over a wide operating supply range. The LT1161 drivers are protected from transients of up to 60V on the supply pins and 75V on the gate pins. Fault indication is provided by an inexpensive logic gate.

Each of the four LT1161 switch channels has a completely self-contained charge pump, which drives the gate of the N-channel MOSFET switch 12V above the supply rail when the corresponding Input pin is taken high. The specified MOSFET device types have a maximum $R_{DS(ON)}$ of 0.028Ω, resulting in a total switch drop (including sense resistor) of only 0.15V at 4A output current.

The LT1161 independently protects and restarts each MOSFET. It senses drain current via the voltage drop across a current shunt R_S. When the current in one switch exceeds approximately 6A (62mV/0.01Ω) the switch is turned off without affecting the other switches. The switch remains off for 50ms (set by external timing capacitor C_T), after which the LT1161 automatically attempts to restart it. If the fault is still present this cycle repeats until the fault is removed, thus protecting the MOSFET. Current shunts are readily available in both through-hole and surface mount case styles. AN53 has additional information on shunts. Connect the LT1161 V^+ pins directly to the top of the current shunts (see LT1161 data sheet).

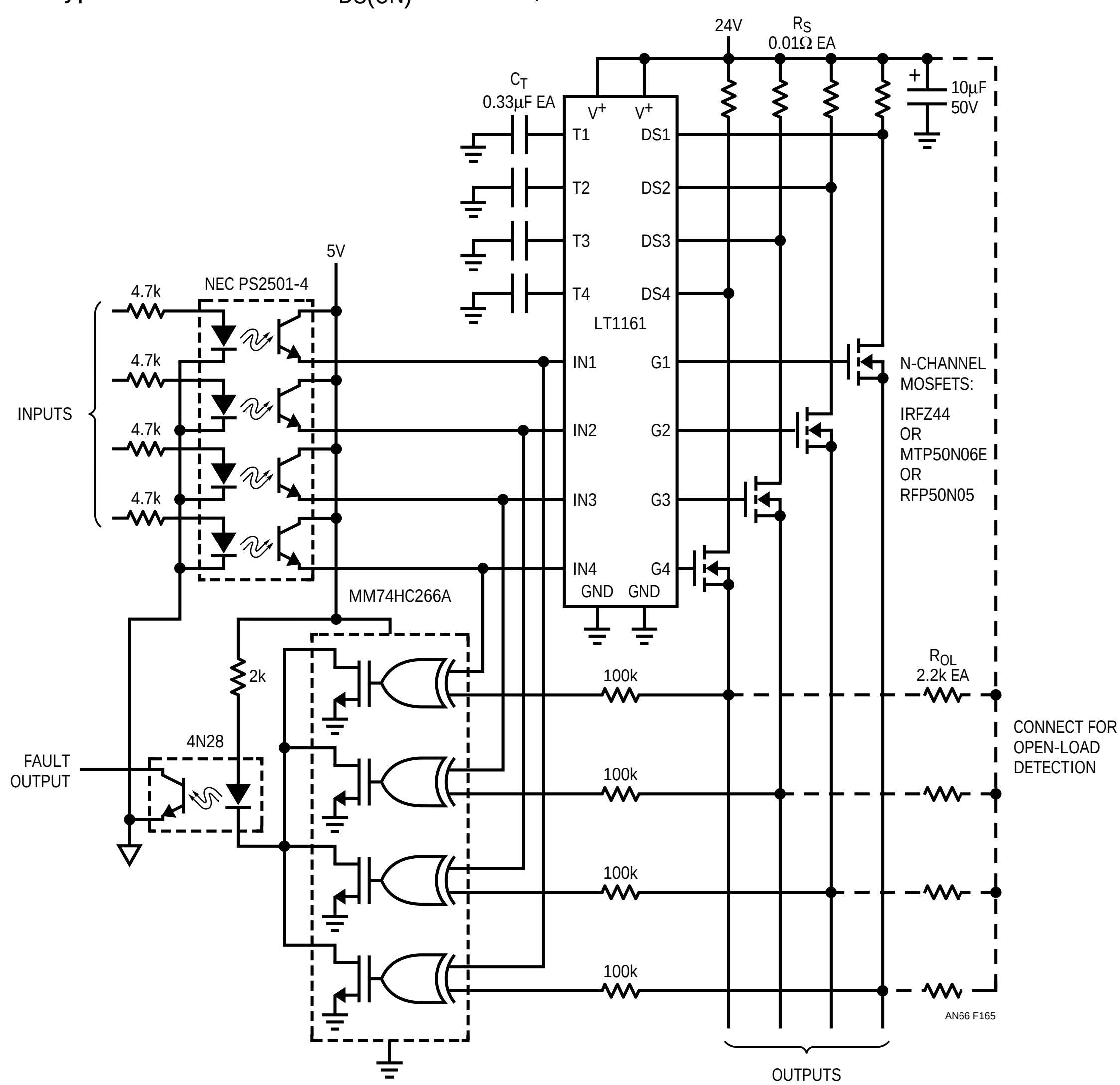


Figure 165. Protected Quad High Side Switch Has Isolated Inputs and Fault Output

The highest MOSFET dissipation occurs with a "soft short" (one in which the current is above the normal operating level but still below the current limit threshold). This can cause dissipation in Figure 165's circuit to rise, in the worst-case to 2W, requiring modest heatsinking. When an output is directly shorted to ground the average dissipation is very low because the MOSFET conducts only during brief restart attempts.

Fault indication is provided by a low cost exclusive NOR gate. In normal operation a low on the LT1161 input forces a low on the output and a high forces a high. If an input is high and the corresponding output is low (i.e., short circuited), the output of the exclusive NOR gate activates the isolated fault output. Similarly, by adding resistor R_{OL} the low input/high output state can be used to diagnose an open load condition. Adjusting the value of R_{OL} sets the output current at which the load is considered to be open. For example, in Figure 165 with V_{SUPPLY} = 24V, a fault would be indicated if the load could not sink 10mA.

Figure 165's circuit is ideal for driving resistive or inductive loads such as solenoids. However, the circuit can be tailored for capacitive or high inrush loads as well. Consult the LT1161 data sheet for information on programming current limit, delay time and automatic restart period to handle other loads. The LT1161 is available in both PDIP and surface mount packaging.

THE LTC1153 ELECTRONIC CIRCUIT BREAKER

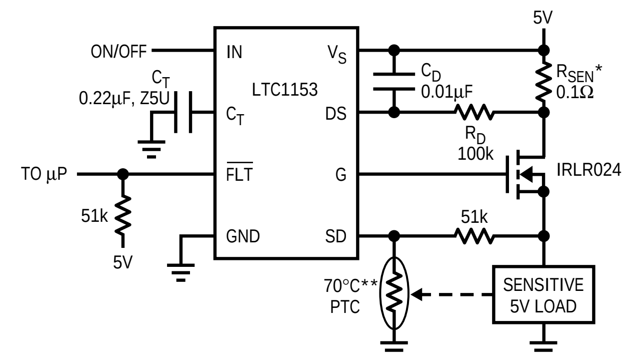
by Tim Skovmand

The LTC1153 electronic circuit breaker is designed to work with a low cost, N-channel power MOSFET to interrupt power to a sensitive electronic load in the event of an overcurrent condition. The breaker is tripped by an overcurrent condition and remains tripped for a period of time programmed by an external timing capacitor, C_T. The switch is then automatically reset and the load momentarily retried. If the load current is still too high, the switch is shut down again. This cycle continues until the overcurrent condition is removed, thereby protecting the sensitive load and the power MOSFET.

5V/1A Circuit Breaker with Thermal Shutdown

The trip current, trip-delay time and autoreset period are programmable over a wide range to accommodate a variety of load impedances. Figure 166 demonstrates how the LTC1153 is used in a typical circuit breaker application. The DC trip current is set by a small valued resistor, R_{SEN}, in series with the drain lead, which drops 100mV when the current limit is reached. In the circuit of Figure 166, the DC trip current is set at 1A (R_{SEN} = 0.1Ω).

The trip-delay time is set by the two delay components, R_D and C_D, which establish an RC time constant in series with the drain sense resistor, producing a trip delay that is shorter for increasing breaker current (similar to that of a mechanical circuit breaker). Figure 167 is a graph of the trip-delay time versus the circuit breaker current for a 1ms



* ALL COMPONENTS SHOWN ARE SURFACE MOUNT.
** IMS026 INTERNATIONAL MANUFACTURING SERVICE, INC. (401) 683-9700
RL2006-100-70-30-PT1 KEYSTONE CARBON COMPANY (814) 781-1591

AN67 F166

Figure 166. LTC1153 5V/1A Circuit Breaker with Thermal Shutdown

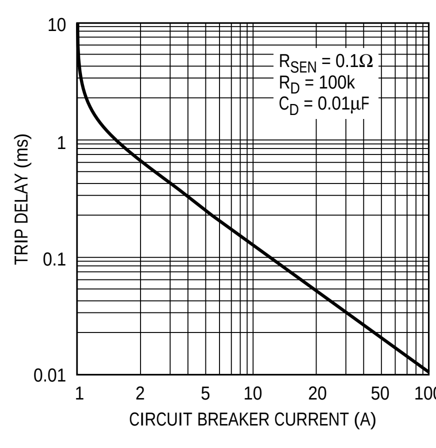


Figure 167. Trip Delay Time vs Circuit Breaker Current (1ms RC Time Constant for the Circuit of Figure 166)

RC time constant. Note that the trip time is 0.63ms at 2A, but falls to 55µs at 20A. This characteristic ensures that the load and the MOSFET switch are protected against a wide range of overload conditions.

The autoreset time is typically set in the range of 10s of milliseconds to a few seconds by selecting the timing capacitor, C_T. The autoreset period for the circuit in Figure 190 is 200ms, i.e., the circuit breaker is automatically reset (retried) every 200ms until the overload condition is removed.

An open-drain fault output is provided to warn the host microprocessor whenever the circuit breaker has been tripped. The microprocessor can either wait for the autoreset function to reset the load, or shut the switch OFF after a fixed number of retries.

The shutdown input interfaces directly with a PTC thermistor to sense overtemperature conditions and trip the circuit breaker whenever the load temperature or the MOSFET switch temperature exceeds a safe level. The thermistor shown in Figure 166 trips the circuit breaker when the load temperature exceeds approximately 70°C.

LTC1153: DC Motor Protector

A 5V DC motor can be powered and protected using the circuit shown in Figure 168. The DC current delivered to the motor is limited to 5A and a rather long trip delay is used to ensure that the motor starts properly. The motor temperature is also continuously monitored and the breaker is tripped if the motor temperature exceeds 70°C. The fault output of the LTC1153 informs the host microprocessor whenever the breaker is tripped. The microprocessor can disable the motor if a set number of faults occur or it can initiate a retry after a much longer period of time has elapsed. A rectifier diode across the motor returns the motor current to ground and restricts the output of the switch to less than 1V below ground.

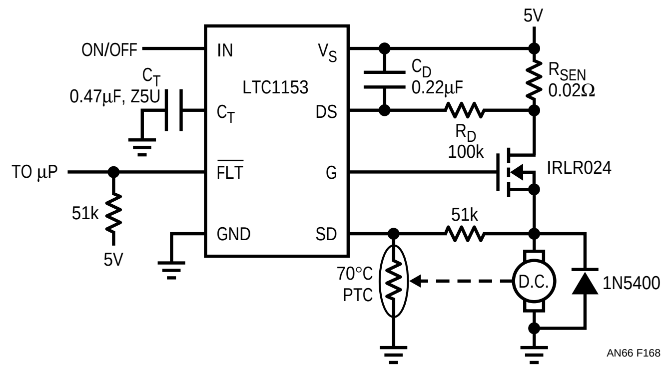


Figure 168. DC Motor Driver with Overcurrent and Overtemperature Protection

LTC1477: 0.07Ω 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 overvoltage or overcurrent 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 169 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."

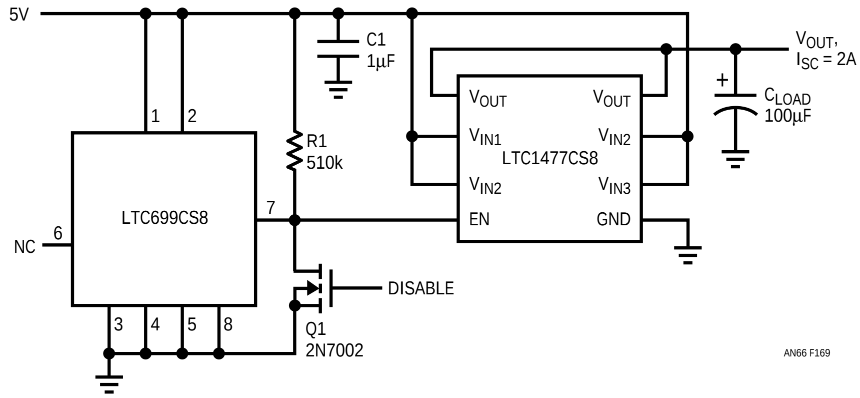


Figure 169. "Hot Swap" Circuit Featuring LTC1477 and LTC699

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.

Miscellaneous

PROTECTED BIAS FOR GaAs POWER AMPLIFERS

by Mitchell Lee

Portable communications devices such as cellular telephones and answer-back pagers rely on small GaAsFET-based 0.1W to 1.0W RF amplifiers as the transmitter output stage. The main power device requires a negative gate bias supply, which is not readily available in a battery-operated product. The circuit shown in Figure 170 not only develops a regulated negative gate bias, it also switches the positive supply, protects against the loss of gate bias, limits power dissipation in the amplifier under high standing-wave ratio (SWR) conditions and protects against amplifier failures that might otherwise short circuit the battery pack.

Negative bias is supplied by an LTC1044 charge pump inverter and the amplifier's positive supply is switched by an LTC1153 electronic circuit breaker. An open-collector switch can be used to turn the LTC1044 inverter off by grounding the OSC pin (Pin 7). When off the LTC1044 draws only 2µA.

The negative output from the LTC1044 is sensed by a 2.5V reference diode (IC2) and Q2. With no negative bias available, Q2 is off and Q3 turns on, pulling the LTC1153's control input low. This shuts off the GaAs amplifier. Total standby power, including the LTC1044, is approximately 25µA.

If the LTC1044's OSC pin (Pin 7) is released, a negative output nearly equal in magnitude to the battery input voltage appears at V_{OUT} (Pin 5). The negative bias is regulated by R1, IC2 and Q2's base-emitter junction. Q2 saturates, shutting Q3 off and thereby turning the LTC1153 on.

The LTC1153 charges the N-channel MOSFET (Q4) gate to 10V above the battery potential, switching Q4 fully on. Power is thus applied to the GaAs amplifier.

The nominal negative bias is –3.2V, comfortably assuring the –2.5V minimum specified for the amplifier. Total

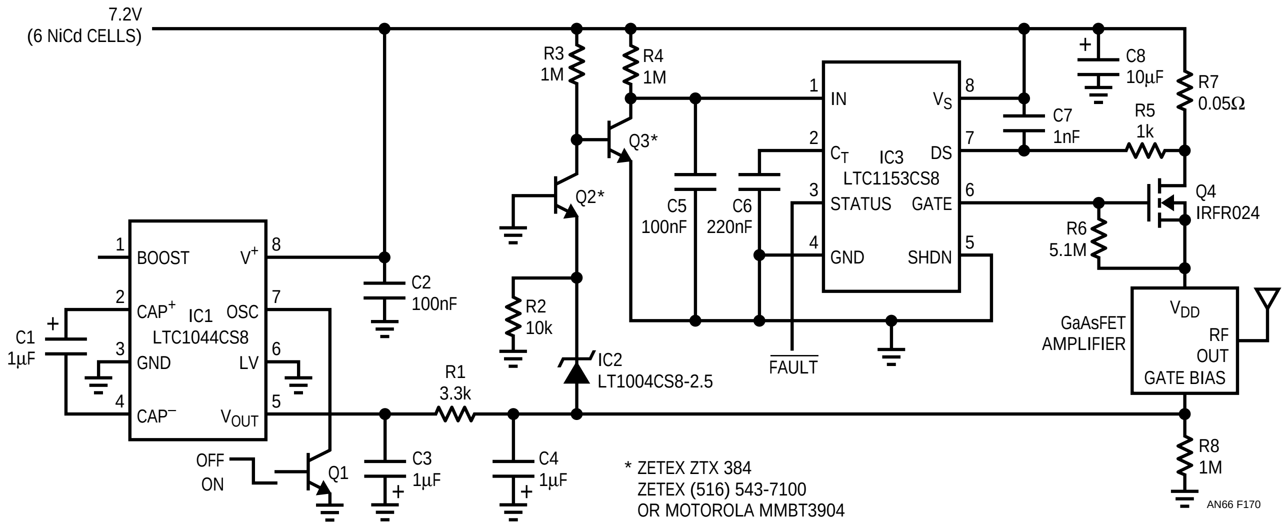


Figure 170. Schematic Diagram

quiescent current, exclusive of the GaAs amplifier drain supply, is approximately 1.5mA in the ON state.

Short circuits or overcurrent conditions in the GaAs amplifier can damage the circuit board, the batteries or both. The LTC1153 senses the amplifier's supply current and turns Q4 off if it is over 2A. After a timeout period set by C6 (200ms) the LTC1153 tries again, turning Q4 on. If the amplifier's supply current is still too high the LTC1153 trips off again. This cycle continues until the fault condition is cleared. Under fault conditions the LTC1153's Status pin (Pin 3) is low. As soon as the fault is cleared the LTC1153 resets and normal operation is restored.

LT1158 H-BRIDGE USES GROUND REFERENCED CURRENT SENSING FOR SYSTEM PROTECTION

by Peter Schwartz

The LT1158 half-bridge motor driver incorporates a number of powerful protection features. Some of these, such as its adaptive gate drive, are dedicated in function. Others are open to a variety of uses depending upon application requirements. The circuit shown in Figure 171 takes advantage of the wide common mode input range of the LT1158's fault comparator to perform ground referenced current sensing in an H-bridge motor driver. By using ground referenced sensing, protection can easily be provided against overloaded, stalled or shorted motors. For overloads and stalls the circuit becomes a constant current chopper, regulating the motor's armature current to a preset maximum value. For shorted loads the circuit protects itself by operating at a very low duty cycle until the short is cleared.

Setting Up for Ground Referenced Sensing

The circuit of Figure 171 is essentially a straightforward LT1158 H-bridge of the "sign/magnitude" type. (See the LT1158 data sheet for a description of component functions.) In many LT1158 applications, a current sense resistor is placed in each upper MOSFET source lead. This circuit, however, senses the IR drop across one resistor (R1) common to the sources of both lower MOSFETs. In Figure 171, U1's $\overline{\text{FAULT}}$ output activates the constant current protection mode (for overloads and stalls) and U2's $\overline{\text{FAULT}}$ output indicates a shorted load. Hence, given a maximum continuous motor current of 15A, R1's value is easily determined: V_{SENSE^+} of U1 must exceed V_{SENSE^-} by the LT1158's internal threshold of 110mV in order for $\overline{\text{FAULT}}$ to go low. 15A • R1 = 0.110V, so R1 = (0.110V/15A) at 0.0075Ω. The $\overline{\text{FAULT}}$ pin of U2 should go low when I_{R1} is 24A, so a 1.6:1 voltage divider is added at U2's Sense$^+$ input. R2, R3, C1 and C2 filter any switching spikes that appear across R1.

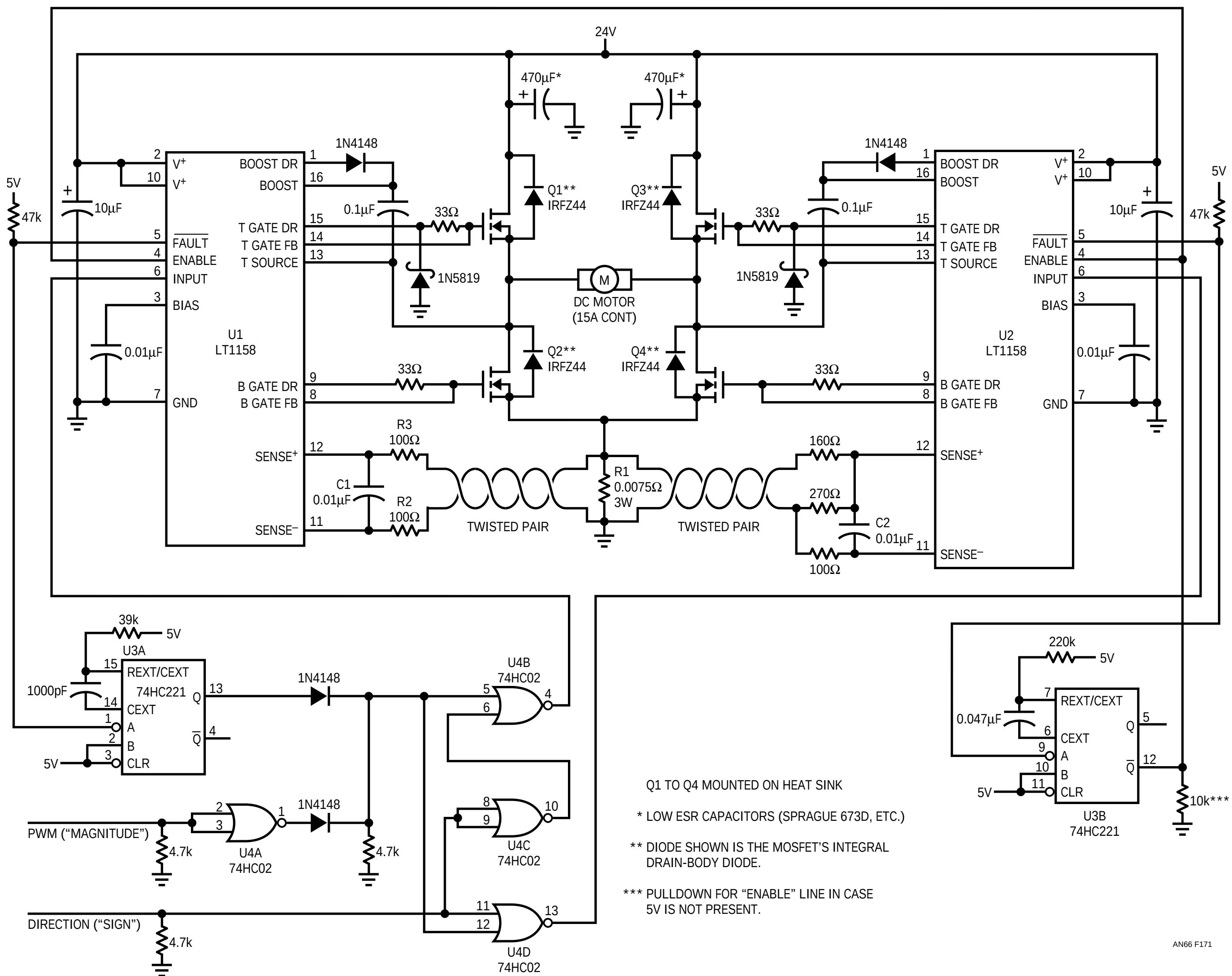


Figure 171. H-Bridge Motor Driver with Ground Referenced Current Sensing

Closing the Loop on Overloads

If the motor is overloaded or stalled, its back EMF will drop, causing the armature current to increase at a rate determined primarily by the motor’s inductance. Without protection this current could rise to a value limited only by supply voltage and circuit resistance. The necessary protection is provided via the feedback loop formed by U1’s FAULT output, U3A, U4B and U4D. When I_{R1} exceeds 15A, the FAULT pin of U1 conducts, triggering the 40µs monostable U3A. The Q output of U3A in turn forces the outputs of U4B and U4D to a logic low state, turning off Q1 or Q3, and turning on both Q2 and Q4. For the time during which U3A’s Q output is high, the motor current decays through the path formed by the motor’s resistance, plus the on-resistance of Q2 and Q4 in series. In this application, turning both lower MOSFETs on is preferable to forcing all four MOSFETs off, as it provides a low resistance recirculation path for the motor current. This reduces motor and supply ripple currents, as well as MOSFET dissipation. At the end of U3A’s 40ms timeout the H-bridge turns on again. If the overload still exists, the current quickly builds up to the U1 FAULT trip point again and the 40ms timeout repeats. This feedback loop holds the motor current approximately constant at 15A for any combination of supply voltage and duty cycle that would otherwise cause an excess current condition. When the motor’s current draw falls below 15A, the circuit resumes normal operation.

Opening the Loop on Shorts

In the event of a short across the motor terminals the current through the H-bridge rises faster than the U1/U3A loop can regulate it. This could easily exceed the safe operating area limits of the MOSFETs. The solution is simple: when the fault comparator of U2 detects that $I_{R1} \geq$ 24A, monostable U3B is triggered. The $\overline{Q}$ output of U3B will then hold the enable line of the two LT1158s low for 10ms, resulting in a rapid shutdown and a very low duty cycle. After the 10ms shutdown interval, U3B's $\overline{Q}$ output will return high and the bridge will be reenabled. If the motor remains shorted, U3B is triggered again, causing another 10ms shutdown. When the short is cleared, circuit operation returns to that described above.

A Final Note

As a class, sign/magnitude H-bridge systems are susceptible to MOSFET and/or motor damage if the motor velocity is accelerated rapidly, or the state of the DIRECTION line is switched while the motor is rotating. This is especially true if the motor/load system has high inertia. The circuit of Figure 171 is designed to provide protection under these conditions: the motor may be commanded to accelerate and to change direction with no precautions. For the case of deceleration, however, it's generally best to use a controlled velocity profile. If a specific application requires the ability to operate with no restrictions upon the rate of change of duty cycle, there are straightforward modifications to Figure 171 that allow this. Please contact the factory for more information.

LT1158 ALLOWS EASY 10A LOCKED ANTIPHASE MOTOR CONTROL

by Milton Wilcox

Allowing synchronous control of two N-channel power MOSFETs operating from 5V to 30V, the LT1158 half-bridge driver effectively deals with the many problems and pitfalls encountered in the design of high efficiency motor control and switching regulator circuits.

Figures 172a and 172b illustrate a locked antiphase motor drive in which the motor stops if either side is shorted to ground (since a 50% input duty cycle is used to stop the motor in locked antiphase operation, the motor would

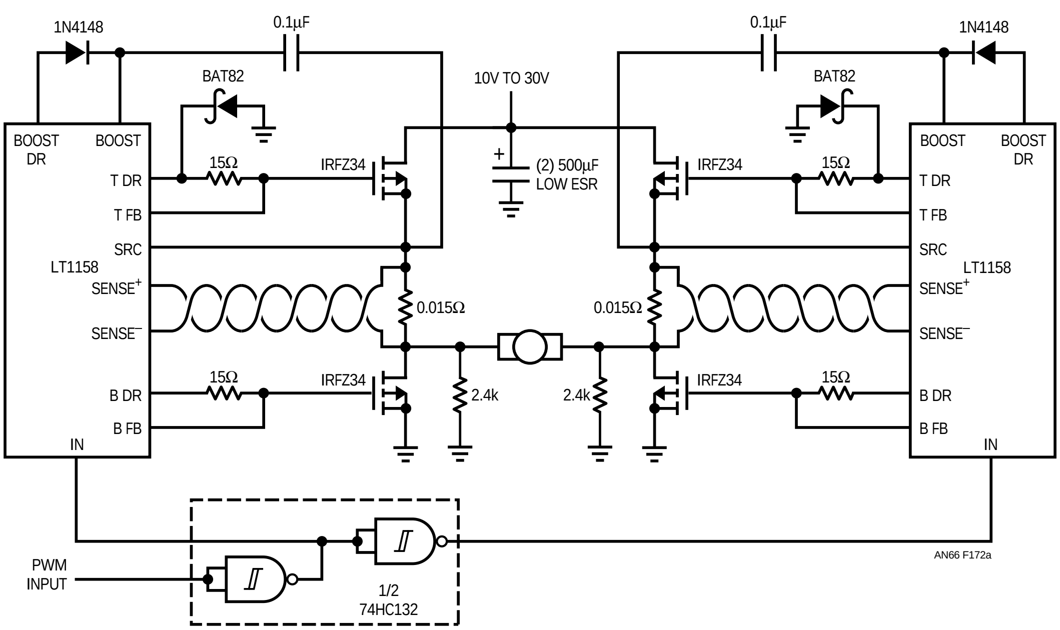


Figure 172a. 10A Locked Antiphase Full-Bridge Circuit Operates Over Wide Supply Range

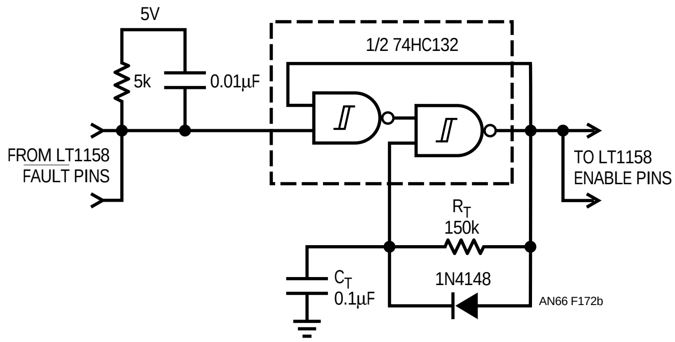


Figure 172b. Protection Logic Stops Motor if Either Side Is Shorted to Ground

normally accelerate to half speed with one side shorted). When a fault is detected by either LT1158, the Figure 172b latch is set, disabling both LT1158s. The circuit periodically tries restarting the motor at a time interval determined by R_T and C_T. If the short still exists, the disabled state is resumed within 20µs, far too short a time to move the motor.

The LT1158 can be used with virtually any N-channel power MOSFET, including 5-lead current sensing MOSFETs. This configuration offers the benefit of no-loss current sensing, since a current shunt is no longer needed in the source. In addition, R_{SENSE} increases by a factor of 1000 or more: from milliohms to ohms. The LT1158 can also be used with logic level MOSFETs for operation as low as 4.5V if a Schottky boost diode is used and connected directly to the supply.

The LT1158 N-channel power MOSFET driver anticipates all of the major pitfalls associated with the design of high efficiency bridge circuits. The designed-in ruggedness and numerous protection features make the LT1158 the best solution for 5V to 30V medium-to-high current synchronous switching applications.

ALL SURFACE MOUNT PROGRAMMABLE 0V, 3.3V, 5V AND 12V VPP GENERATOR FOR PCMCIA

by Jon A. Dutra

Generating the VPP voltage for a PCMCIA port in laptop computers has become more complicated with PCMCIA standard 2.0. The VPP line must come up to 5V initially until the card "tuple" tells the card its type and VPP voltage. For example, a 3.3V SRAM card must have VPP adjusted to 3.3V. If it is a flash memory card, 12V must be supplied during programming. During card insertion, 0V is desirable to unconditionally prevent latch-up. Shutdown supply current must be as low as possible and the supply must not overshoot. This design idea presents a circuit (Figure 173) that meets these specifications. The same topology could be useful for generating other programmable supplies.

The circuit uses the LT1107 micropower DC/DC converter with a single surface mount transformer. The LT1107 features an I_{LIM} pin that enables direct control of maximum inductor current. This allows use of a smaller transformer without risk of saturation. The LT1111 could also be used with a reduction in output power.

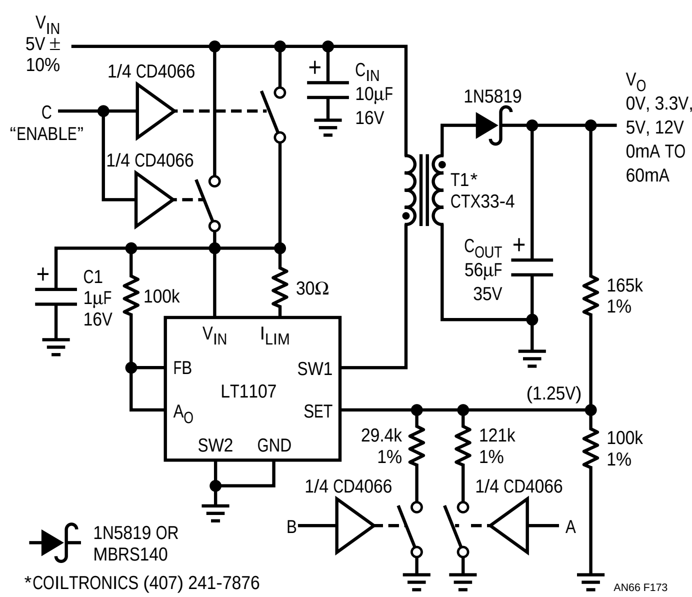


Figure 173. Schematic Diagram for VPP Generator

Circuit Operation

The circuit is basically a gated-oscillator flyback topology. The SET pin of the LT1107 is held at 1.25V by negative feedback. Summing currents into the SET pin to zero for the three different output states yields three equations with three unknown resistor values. The resistor values are easily solved for using Mathametica, MathCad or classical techniques. Table 1 shows the output voltage truth table.

Table 1

INPUTS			OUTPUTS	
A	B	ENABLE	V_O	NOTE
X	X	0	0V	Off
1	1	1	12V	12V
1	0	1	5V	5V
0	1	1	10.3V	Not Used
0	0	1	3.3V	3.3V

Output noise is reduced by using the auxiliary gain block (AGB) in the feedback path. This added gain effectively reduces the hysteresis of the comparator and tends to randomize output noise. With a low ESR capacitor for C1, output noise is below 30mV over the output load range.

Output power increases with $V_{BATTERY}$ from about 1.4W out with 5V in to about 2W out with 8V or more. Efficiency is 62% to 76% over a broad output power range. No minimum load is required.

Component Selection

Substantial current flows through C_{IN} and C_{OUT}. Most tantalum capacitors are not rated for current flow and can result in field failures. Using a rated tantalum or rated electrolytic will result in longer system life.

Shutdown

The circuit is shut down by using two sections of the CD4066 in parallel as a high side switch. Alternatively, simply disabling the logic supply to the V_{IN} and I_{LIM} nodes of the LT1107 will shut it down. This drops quiescent current from the $V_{BATTERY}$ input below 2µA. When the device is shut down V_{OUT} drops to 0V.

A TACHLESS MOTOR SPEED REGULATOR

by Mitchell Lee

A common requirement in many motor applications is a means of maintaining constant speed with variable loading or variable supply voltage. Speed control is easily implemented using tachometer feedback, but the cost of a tach may be prohibitive in many situations and adds mechanical complications to the product. A lower cost solution with no moving parts is presented here.

Motor speed changes under conditions of varying loads because of the effects of series loss terms in the motor. The effects of the predominant contributors to loss, copper and brush/commutator resistance (collectively known as R_M), are best understood by considering the circuit model for a motor (see Figure 174). A motor's back EMF (V_M) is proportional to speed (n) and the motor current (I_M) is proportional to the load torque (T). The following equation predicts the speed of the motor for any given condition of loading:

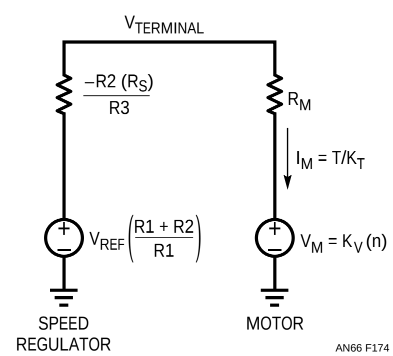


Figure 174. On the Right is Shown an Equivalent Circuit for a Motor. On the Left is the Model for a Circuit Which Will Stablize the Motor's Speed Against Changes in Supply Voltage and Loading

$$n = \frac{V_{TERMINAL}}{K_V} - T\left(\frac{R_M}{(K_T)(K_V)}\right) \qquad (1)$$

where K_V and K_T are constants of proportionality for rotational velocity and torque. For a fixed terminal voltage, the speed of the motor must decrease as increasing load torque is applied to the shaft. For a fixed load, the speed of the motor will also change if the supply (terminal) voltage is changed.

A voltage regulator fixes the problem of a varying terminal voltage, but the only way to eliminate torque from Equation (1) is by reducing R_M to zero. Physically this is impractical, but an electrical solution exists.

If a motor is driven from a regulated source whose output impedance is opposite in sign and equal in magnitude to R_M (see Figure 174), the result is a motor that runs at a constant speed—regardless of loading and power source variations. Figure 175 shows a circuit that does it all. The LT1170 is configured as a buck/boost converter, which can take a wide ranging 3V to 20V input source and produce a regulated output of, say, 6V. The circuit shown can deliver 1A at 6V with a 5V input, adequate for many small permanent-magnet DC motors.

To cancel the effects of the motor resistance, a negative output impedance is introduced with an op amp and a current sense resistor (R_S). As the motor current increases, the LT1006 responds by increasing the motor terminal voltage by an amount equal to $(I_M)(R_M)$. Depending on the value of R3, the speed can be made to increase, decrease or stay the same under load. If R3 is just right, the motor speed will remain constant until the LT1170 reaches full power and the circuit runs out of steam.

Many small motors in the 1W to 10W class are not well characterized. In order to choose proper component values for a given motor, figures for R_M and V_M are necessary. Fortunately, these are easily measured using a DVM and a motor characterization test stand. If you don't have a motor characterization test stand, it is also possible to use a lathe or drill press to do the job.

Chuck up the candidate motor's shaft in a variable speed drill press or lathe, which is set to run at the same speed you're intending to operate the motor. Clamp down the motor frame so it won't spin. Turn on the big machine, and measure the open-circuit motor terminals with a DVM. This is the motor voltage, V_M, as shown in Figure 174. Switch the meter to measure the motor's short circuit current, I_{SC}. Motor resistance $R_M = V_M/I_{SC}$. With these figures the other component values can be calculated:

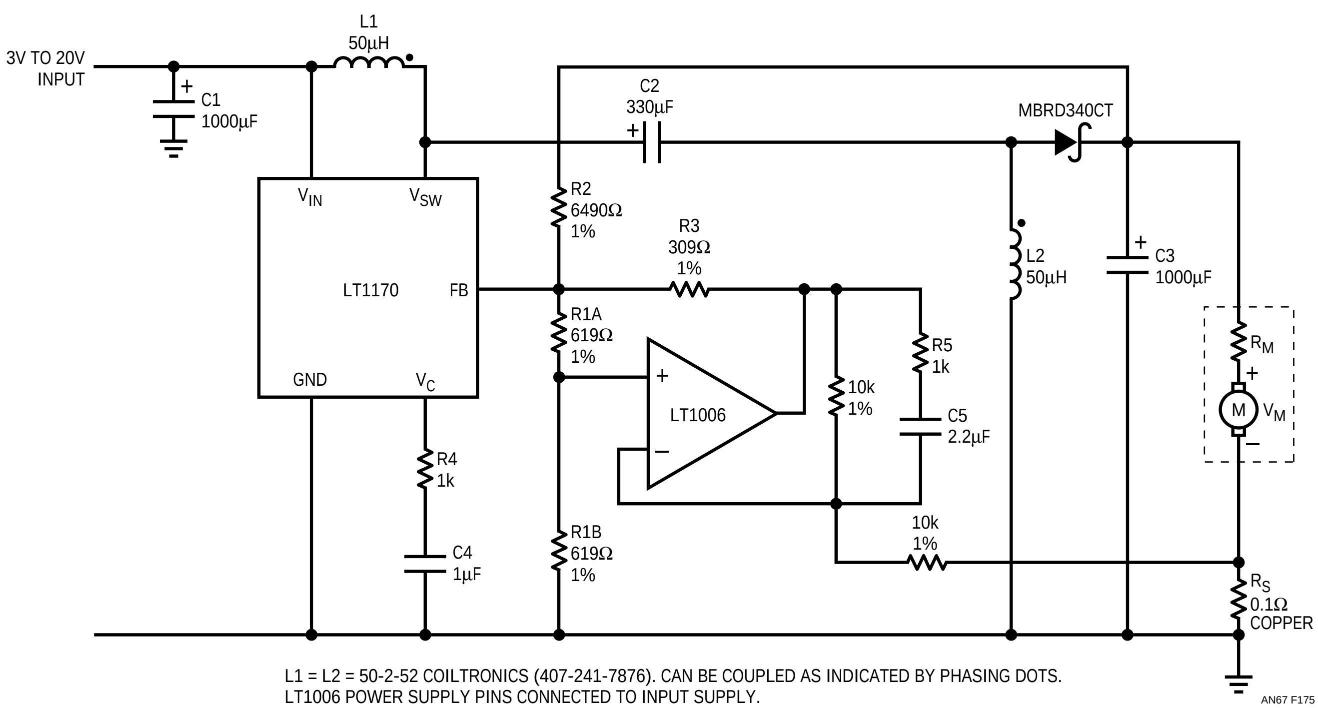


Figure 175. Tachless Motor Speed Regulator

I_{MAX} = motor current at full load

V_{REF} = 1.244V

R1 = series combination of 619Ω + 619Ω = 1238Ω

$R_S \le 1/I_{MAX}$ (drops less than 1V at maximum load)

$$R2 = (V_M \bullet R1/V_{REF}) - R1 \quad (2)$$

$$R3 = (R2 \bullet R_S)/(R_M + R_S) \quad (3)$$

The component values shown in Figure 175 are for a small motor with the following characteristics measured at 360RPM: V_M = 7.8V, I_{SC} = 3.7A, R_M = 2.1Ω, $I_{MAX} \approx$ 1A.

R_S, a copper resistor, is either located close to or wound around the motor to assist in tracking changes in armature resistance with temperature. Copper has a strong, 3930ppm/°C temperature coefficient, matching the TC of the motor winding.

Setup Procedure

Initial tests should be performed with a potentiometer in place of, and twice the value of, R3. R5 and C5 should be disconnected; remove all loading from the motor. Check the motor's unloaded speed, and adjust R2 if necessary to set it precisely.

With the motor driving a nominal load, decrease R3 until the motor commences "hunting." R3 will be near the nominal calculated value. This threshold is very close to optimum motor resistance cancellation. R5 and C5 offer a convenient means of compensating for frictional and inertial effects in the mechanical system, eliminating instabilities. System stability should be evaluated under a variety of loading conditions. The effect of R5 is to reduce the negative output impedance of the circuit at high frequencies. Systems with a net positive impedance are inherently stable.

When the system stability is satisfactory, a final adjustment of R3 can be made to achieve the desired speed regulation under conditions of varying loads. These final values can be used in production. Note that R2 defines the regulated speed value and may be production trimmed in precision applications.

LT1161: … AND BACK AND STOP AND FORWARD AND REST — ALL WITH NO WORRIES AT ALL

by Peter Schwartz and Milt Wilcox

Many applications of DC motors require not only the ability to turn the motor on and off, but also to control its direction of rotation. When directional control is involved, the need for rapid deceleration (electronic braking) can also be assumed. A microcontroller interface (logic-level control) is a necessity in modern systems, as is protection of both the motor controller and the motor itself. With the advent of high power, logic-level N-channel MOSFETs, it is a straightforward matter to build the lower half of an H-bridge suitable for the versatile control of DC motor loads. Equivalent performance P-channel MOSFETs, however, are still expensive devices of limited availability, even without logic-level capability. Therefore, motor control circuits commonly use N-channel devices for the upper half of the H-bridge as well. The trick is to do this without requiring an additional power supply to provide bias for the upper MOSFET gates, while ensuring the necessary system protections.

A Complete, Six-Part Plan

The circuit shown in Figure 176 is a complete H-bridge motor driver, with six distinct modes of operation:

- ❏ Motor Forward Rotation — In this mode, Q1 and Q4 are on, and Q2 and Q3 are off.
- ❏ Motor Reverse Rotation — In this mode, Q2 and Q3 are on, and Q1 and Q4 are off.
- ❏ Motor Stop — Here, a rapid stop is performed by using "plugging braking," wherein the motor acts as a generator to dissipate mechanical energy as heat in the braking circuit's resistance.
- ❏ Motor Idle — All four MOSFETs are turned off. The motor is, in effect, disconnected from the H-bridge driver.
- ❏ Load Protect — If the motor is overloaded or stalled for an excessive period, the on-chip fault detection and protection circuitry of the LT1161 will shut the motor off for programmed interval, then turn it back on.

- ❏ Short-Circuit Protect—If a source-to-ground short is detected on either Q1 or Q2, the on-chip fault detection and protection circuitry of the LT1161 will shut off the MOSFET at risk for the programmed interval and then attempt to turn the circuit back on.

Figure 176 shows a straightforward H-bridge using four N-channel MOSFETs (Q1 to Q4). The lower MOSFETs (Q3 and Q4) are logic-level devices to allow direct drive from 5V logic. The upper MOSFETs (Q1 and Q2) are driven via level translation circuitry integral to the LT1161. Input 1 of the LT1161 controls a charge pump in the IC, whose output is developed on Gate 1. Similarly, Input 2 controls a charge pump whose output is available on Gate 2. The Gate outputs have voltage swings from 0V to (V_{CC} + 12V), which is more than sufficient to enhance a standard threshold N-channel MOSFET, such as the IRFZ34. D3 is added to Q1 as a gate-source protection diode to prevent excessive voltage from appearing across the gate-source terminals of Q1. This could otherwise happen under certain conditions of "motor-idle" operation. D4 serves the same function for Q2.

The Logic Behind It All

The logic of the circuit is straightforward and could be replaced by a microcontroller in many applications. CMOS inverters U1 and U2 drive the lower MOSFETs directly from a 5V supply, with the RCD networks on their inputs providing the necessary timing to prevent shoot-through currents in the MOSFET switches. Inverter U3 and NOR gate U5 work together to turn Gate 1 and hence Q1 on when point A is at a logical high. This also ensures that C3 is charged to a logical high to take U2's output low and turn Q3 off. Under these conditions, with point B low (or left floating), U1 will turn Q4 on and U6 will hold Gate 2 and hence Q2 off. If point A is now immediately taken low (or left floating), and point B is taken high, the symmetry of the

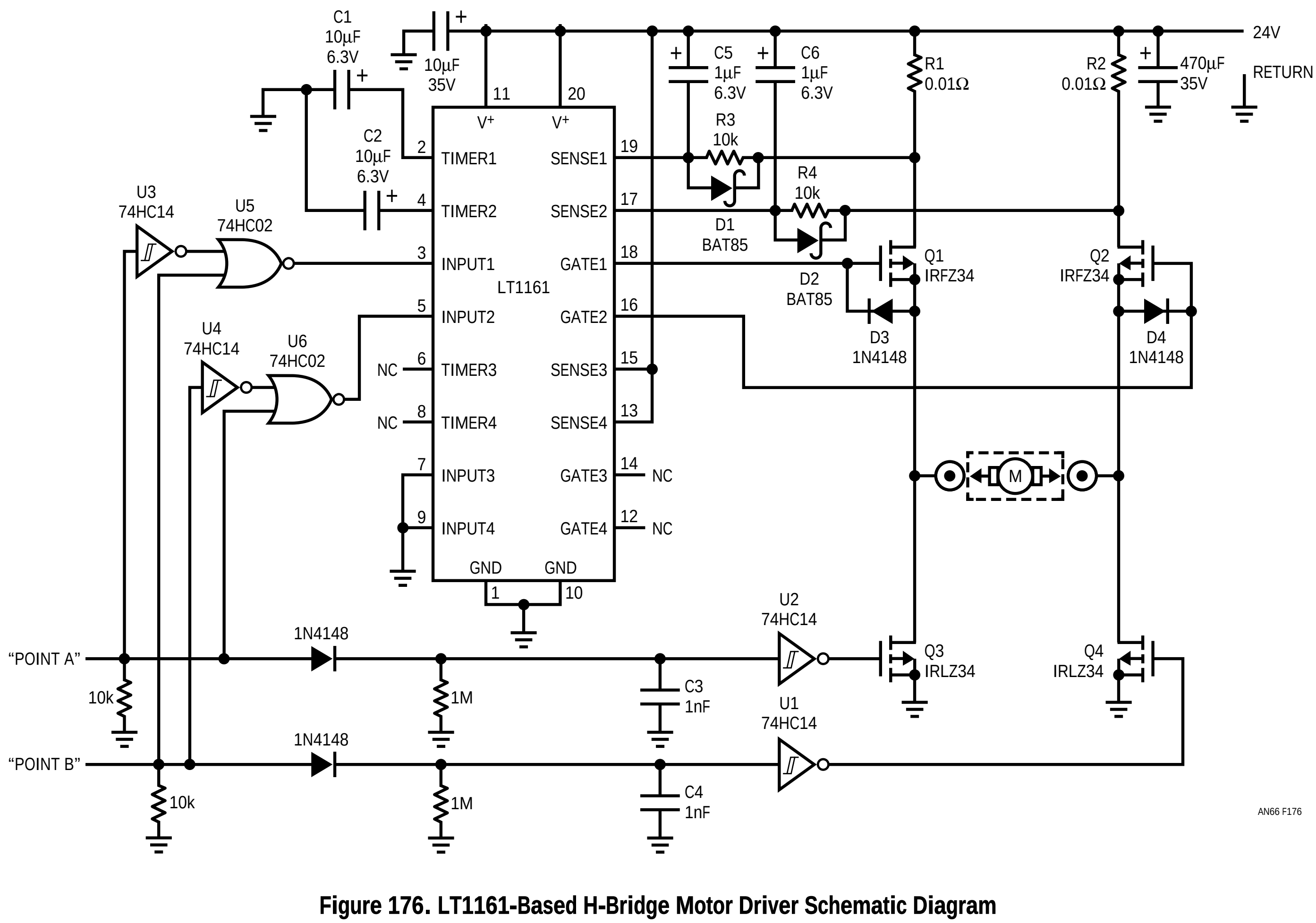


Figure 176. LT1161-Based H-Bridge Motor Driver Schematic Diagram

logic will reverse these conditions—but only after C3 has discharged to the point where the output of U2 can go high to turn Q3 on. This is the shoot-through prevention mentioned previously.

There are two exceptions to the symmetry of the logic: if both point A and point B are low, both upper MOSFETs are turned off while both lower MOSFETs are turned on. Under these conditions, the kinetic energy stored in the motor and its load is used to drive the motor as a generator. This produces a current through the motor winding, Q3 and Q4. In this "plugging braking" mode, the motor's energy is largely dissipated as I^2R losses and a rapid stop occurs. If point A and point B are both high, all four MOSFETs will be turned off and the motor is essentially disconnected from the electrical circuit. Although primarily included as a cross-conduction interlock in the event that both inputs should ever be high at the same time (things do happen on the test bench), this can also be useful in situations where it is desirable that the motor coast down from a higher velocity to a lower one.

Just a Few Grams… But Lots of Protection

In addition to its level translation and charge pump features, the LT1161 also provides comprehensive protection features via its Sense 1 and Sense 2 pins. Each Sense pin is the (–) input to an on-chip comparator, with the (+) input to that driver's comparator fixed at a level 65mV (nominal, 50mV minimum) below the LT1161's V^+ input. If a Sense pin goes more than 65mV below V^+, several things happen: the corresponding Gate output is rapidly pulled to ground, the capacitor on the Timer pin is dumped to ground and the charge pump is shut off. The charge pump will remain shut off, and the Gate pin will remain clamped to ground until the Timer capacitor has charged back up to 3V from an on-chip 14µA current source. When the capacitor reaches this 3V threshold, the internal charge pump starts up again and the clamp from the Gate pin to ground is removed. The net effect of this is that, if one of the Sense pins is pulled 65mV below V^+, the MOSFET turns off for a period that is set by the value of the capacitor connected to the Timer pin. At the end of this programmed interval the circuit will automatically restart. If the fault has been cleared, the protection circuitry then becomes transparent to the system. This shutdown/retry cycle will repeat until the fault is cleared.

The fault scenarios for which protection is required are, as mentioned above, an overloaded or stalled motor or a source-to-ground short on Q1 or Q2. In each case such a fault will cause excessive current to flow through the affected upper MOSFET; this current is readily transformed into a voltage by a current shunt resistor. Allowing for a 5A motor current under load, this yields a resistor value of [5A/50mV (min)] = 0.01Ω for R1 and R2. To allow for inrush current when the motor starts up or changes direction, delay networks (R3/C5 and R4/C6) have been added to each half of the H-bridge. At a 20A startup current, the values shown give a 3ms delay. The value of the capacitor can be changed to affect longer or shorter delays as needed (the resistor value should not be raised above 10k). A short-to-ground fault, however, requires a shutdown in microseconds, not milliseconds. This is accomplished by adding two BAT85 signal level Schottky diodes (D1 and D2) in parallel with the 10k delay resistors. At a fault current of approximately 45A, which is easily attained in the short-circuit case, V_{SHUNT} = 0.45V. At this voltage the appropriate diode conducts to temporarily bypass the delay resistor, allowing the LT1161 to turn off the imperiled MOSFET within 20µs (typical). In each case, the retry interval is programmed by C1 and C2; the 10µF shown gives a time-out of about 1.8 seconds.

The LT1161 is a quad driver IC, capable of providing drive and protection for two additional MOSFETs beyond those shown in Figure 176.

SIMPLE THERMAL ANALYSIS — A REAL COOL SUBJECT FOR LTC REGULATORS

by Alan Rich

As the temperatures go up… so go the problems with voltage regulators.

Introduction

Linear Technology Corporation applications engineers get lots of calls saying, "that $X%#@& voltage regulator is so hot I can't touch it!" The purpose of the article is to show you, the design engineer, how to perform simple thermal calculations to determine regulator temperature and select the proper package style and/or heat sink. In addition, it will show an alternate method of specifying thermal parameters on LTC voltage regulators.

Definition of Terms

Power dissipation is the parameter that causes a regulator to heat up; the unit for power is watts. Power is the product of the voltage across a linear regulator times the load current (see Figure 177).

Thermal resistance is a measure of the flow of heat from one surface to another surface; the unit of thermal resistance is °C/watt. Common terms for thermal resistance that show up on most LTC data sheets are:

θ_{JC}—thermal resistance from the junction of the die to the case of the package

θ_{JA}—thermal resistance from the junction of the die to the ambient temperature

Some typical LTC regulators and their thermal characteristics are shown in Table 1.

Table 1. θ_{JC} and θ_{JA} for Three LTC Regulators

DEVICE	θ_{JC} (°C/W)	θ_{JA} (°C/W)
LT1005CT	5.0	—
LT1083MK	1.6	—
LT1129CT	5.0	50

There are several other common thermal resistance terms:

θ_{CS} — thermal resistance from the case of the package to the heat sink

θ_{SA} — thermal resistance from a heat sink surface to the ambient temperature

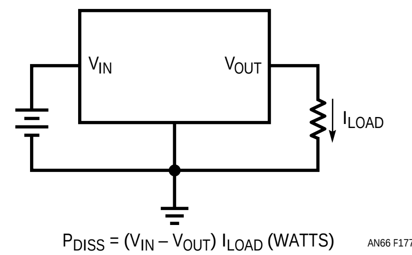


Figure 177. Typical Linear Regulator Circuit

The last two terms are determined by how a regulator is mounted to the heat sink and by the properties of the heat sink. Heat sinks are used to decrease the thermal resistance and therefore lower the temperature rise of the regulator.

Temperature is a term with which we are all very familiar. All thermal calculations will use the Centigrade scale or °C.

T_J — temperature of the junction of the regulator die
T_C — temperature of the case of the regulator
T_A — ambient temperature

The maximum operating junction temperature, $T_{J\,MAX}$ for LTC regulators is shown on the device data sheet.

What is Thermal Analysis?

The goal of any thermal analysis is to determine the regulator junction temperature, T_J, to ensure that this temperature is less than either the regulator rating or a design specification. In the simplest case, temperature rise is calculated by multiplying the power times the total of all thermal resistance:

$$T_R = P(\theta_{TOTAL})$$

θ_{TOTAL} includes the thermal resistance junction-to-case (θ_{JC}), thermal resistance case-to-heat sink (θ_{CS}), and thermal resistance heat sink-to-ambient (θ_{SA}).

T_R represents the temperature rise above the ambient temperature; therefore, to determine the actual junction temperature of the regulator, the ambient temperature must be added to T_R:

Regulator junction temperature =
Ambient Temperature + T_R

For example, consider a circuit using an LT1129CT operating in a 50°C enclosure with an input voltage of 8VDC, an output voltage of 5VDC and a load current of 1A[1].

[1]The LT1129CT is guaranteed for 700mA, but could be selected to output 1A.

The power dissipated by the LT1129CT is:

$P = (V_{IN} – V_{OUT})(I_{LOAD}) = (8V – 5V)(1a) = 3W$

The first question is, does this circuit need a heat sink?

Since we have assumed no heat sink on the LT1129CT for the purpose of this calculation, we must use thermal resistance from junction to ambient, θ_{JA} = 50°C/W.

$T_J = P(\theta_{JA}) + T_A = 3W(50°C/W) + 50°C$

$= 150°C + 50°C = 200°C$

The junction temperature, T_J, that we just calculated is greater than the LT1129CT's maximum junction temperature specification of 125°C; therefore this circuit must use a heat sink.

Now the task at hand is to calculate the correct heat sink to use. The selected heat sink must hold the junction temperature at less than 125°C for the LT1129CT.

$T_J = P(\theta_{TOTAL}) + T_A$
$125°C = 3W(\theta_{TOTAL}) + 50°C$
θ_{TOTAL} = 25°C/W and
$\theta_{TOTAL} = \theta_{JC} + \theta_{CS} + \theta_{SA}$

For this configuration:

θ_{JC} = 5°C/W (LT1129CT data sheet)
θ_{CS} = 0.2°C/W (typical for heat sink mounting)
θ_{SA} = heat sink specification

Plugging in these numbers:

$25°C/W = 5°C/W + 0.2°C/W + \theta_{SA}$
$\theta_{SA} = 19.8°C/W$

Therefore, the heat sink selected must have a thermal resistance of less than 19.8°C/W to hold the LT1129CT junction temperature at less than 125°C. Obviously, the lower the heat sink thermal resistance, the lower the LT1129CT junction temperature. A lower junction temperature will increase reliability.

Now, let's consider a circuit using an LT1129CT operating in a 50°C enclosure with an input voltage of only 6VDC, an output voltage of 5VDC, and a load current of 1A.

The power dissipated by the LT1129CT is:

$P = (V_{IN} – V_{OUT})(I_{LOAD}) = (6V – 5V)1A = 1W$

Does this circuit need a heat sink? Again, for the purposes of the calculation, we must use thermal resistance from junction to ambient, θ_{JA}= 50°C/W for the LT1129CT.

$T_J = P(\theta_{JA}) + T_A = 1W(50°C/W) + 50°C$

$= 50°C + 50°C = 100°C$

The junction temperature T_J that we just calculated is now less than the LT1129CT's maximum junction temperature specification of 125°C. Therefore this circuit does not need a heat sink. This illustrates the advantage of a low dropout regulator like the LT1129CT.

An Alternative Method for Specifying Thermal Parameters

Linear Technology Corp. has introduced an alternative method to specify and calculate thermal parameters of voltage regulators. Previous regulators, with a single thermal resistance junction-to-case (θ_{JC}), used an average of temperature rise of the control and power sections. This could easily allow excessive junction temperature under certain conditions of ambient temperature and heat sink thermal resistance.

Several LTC voltage regulators include thermal resistance and maximum junction temperature specifications for both the control and power sections, as shown in

Table 2. Two Examples Showing Thermal Resistance of Control and Power Sections of LTC Regulators

	CONTROL		POWER	
DEVICE	θ_{JC}	T_{JMAX}	θ_{JC}	T_{JMAX}
LT1083MK	0.6°C/W	150°C	1.6°C/W	200°C
LT1085CT	0.7°C/W	125°C	3.0°C/W	150°C

As an example, let's calculate the junction temperature for the same application shown before, using an LT1085CT instead of the LT1129CT. Once again, we are operating in a 50°C enclosure; the input voltage is 8VDC, the output voltage is 5VDC and the load current is 1A.

The power dissipated by the LT1085CT is the same as before, 3W. We will assume we have selected a heat sink with a thermal resistance, θ_{SA} of 10°C/W. First calculate the control section of the LT1085CT:

θ_{JC} = 0.7°C/W (LT1085CT data sheet)
θ_{CA} = 0.2°C/W (typical)
θ_{SA} = 10°C/W

$\theta_{TOTAL} = \theta_{JC} + \theta_{CA} + \theta_{SA}$ = 0.7°C/W + 0.2°C/W + 10°C/W = 10.9°C/W

To determine the control section junction temperature:

$T_J = P(\theta_{TOTAL}) + T_A$ = 3W(10.9°C/W) + 50°C = 82.7°C ($T_{J\,MAX}$ = 125°C)

To calculate the power section of the LT1085CT:

θ_{JC} = 3°C/W (LT1085CT data sheet)
$\theta_{TOTAL} = \theta_{JC} + \theta_{CA} + \theta_{SA}$ = 3°C/W + 0.2°C/W + 10°C/W = 13.2°C/W

To determine the power section junction temperature:

$T_J = P(\theta_{TOTAL}) + T_A$ = 3W(13.2°C/W) + 50°C = 89.6°C ($T_{J\,MAX}$ = 150°C)

In both cases, the junction temperature is below the maximum rating for the respective section; this ensures reliable operation.

Conclusion

This article is an introduction to thermal analysis for voltage regulators; however, the techniques also apply to other devices, including operational amplifiers, voltage references, resistors, and the like. For the more advanced student of thermal analysis, it can be shown that there is a direct analogy between electronic circuit analysis and thermal analysis, as shown in Table 3.

Table 3. analogy Between Thermal Analysis and Electronic Circuit Analysis

THERMAL WORLD	ELECTRICAL WORLD
Power	Current
Temperature Differences	Voltage
Thermal Resistance	Resistance

All standard electronic network analysis techniques (Kirchhoff's laws, Ohm's law) and computer circuit analysis programs (SPICE) can be applied to complex thermal systems.

ALPHABETICAL INDEX (BY MAJOR CATEGORIES)

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