Efficiency and Power Characteristics of Switching Regulator Circuits

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Efficiency is often the main objective when using switching regulators. High efficiency means less power drain on the input source (batteries, etc.) and less heat buildup around the regulator, allowing for smaller and lighter power supplies and system enclosures. Practical switching regulator efficiencies typically range from 70%-90%. The highest possible efficiency is desired when battery life is critical or when package size restrictions preclude effective heat removal.

Since switching power supplies are not 100% efficient, the lost power is dissipated as heat. Losses fall into a variety of categories including switching transistors, clamp diodes, filter capacitors, controllers, and inductors. Excessive temperature rise in any of these elements can stress the power supply leading to premature failure. Therefore, proper thermal design is essential for reliable operation of switching supplies.

Calculating the power loss and selecting the proper heat sink or package size for each switching regulator component is not a trivial task. Many variables influence power loss: circuit switching characteristics, DC and AC component characteristics, input and output voltages and load currents. Furthermore, power device contribution to efficiency loss varies with circuit configuration.

To satisfy design requirements various switching regulator approaches are possible. Before selecting the appropriate circuit, a review of the following considerations is useful.

- Input Voltage Range
- Output Voltage
- Maximum Output Current
- Efficiency
- Form Factor
- Maximum and Minimum Temperature Range
- Price

Efficiency varies widely between switching regulator architectures, input voltage and output load conditions.

Therefore, efficiency plots for various input and output conditions accompany each circuit discussed in this application note, allowing comparisons. Squeezing the utmost efficiency out of a switching regulator requires that power components be selected properly. Trade-offs between circuit complexity and efficiency influence final circuit selection.

Basic Step-Up Switching Regulator

A common switching regulator requirement involves converting a lower voltage to a higher voltage. Figure 1 shows a basic step-up switching regulator. The LT1070 is a fully self-contained switching regulator that controls its internal switching transistor current to achieve output voltage regulation.
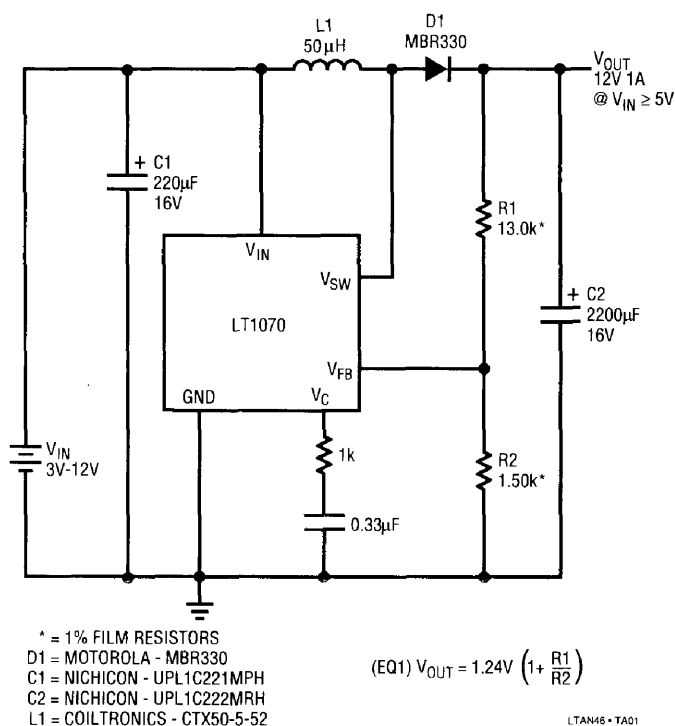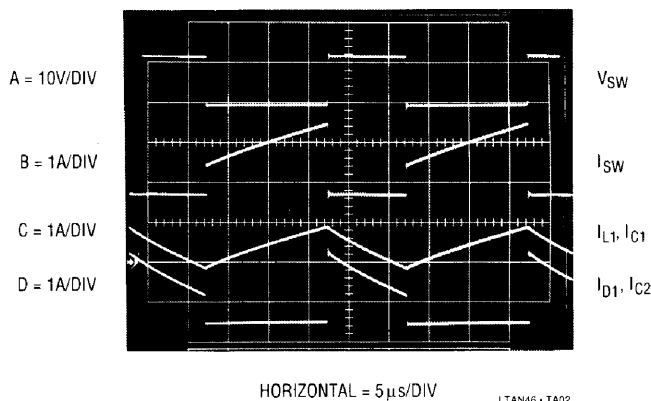
* = 1% FILM RESISTORS
D1 = MOTOROLA - MBR330
C1 = NICHICON - UPL1C221MPH
C2 = NICHICON - UPL1C222MRH
L1 = COILTRONICS - CTX50-5-52

$$(EQ1)\ V_{OUT} = 1.24V \left(1 + \frac{R1}{R2}\right)$$

LTAN46 • TA01

Figure 1. LT1070 Step-Up Switching Regulator

HORIZONTAL = 5μs/DIV

LTAN46 · TA02

Figure 2. LT1070 Step-Up Converter Waveforms[1]

Figure 2 shows regulator operating waveforms. Since the LT1070 has a ground referred power switch, the input voltage is applied across the inductor when it is on. Trace A is the V_{SW} pin voltage and trace B is its current. The inductor current (trace C) rises very slowly as the magnetic field builds up. The rate of change of the current is determined by the voltage applied across the inductor and its inductance. During this interval energy is being stored in the inductor and no power is transferred to the output. When the switch is turned off, energy is no longer transferred to the inductor causing the magnetic field to collapse. The collapsing magnetic field induces a change in voltage across the inductor. The V_{SW} pin voltage rises until clamp diode D1 forward biases. The diode current (trace D) is distributed between the load and the output capacitor. When the diode is reverse biased, the output capacitor provides the load current. The 1% resistors control the

Note 1: Switching Regulator and DC-DC Converter are used interchangeably in this text.
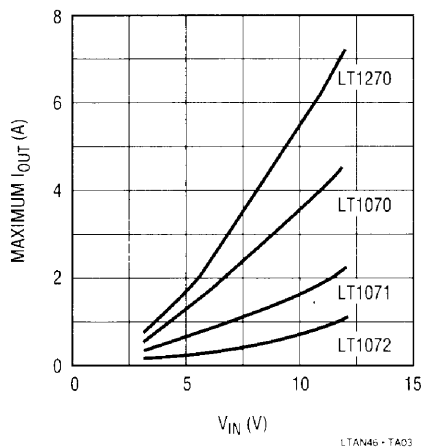

Figure 3. V_{IN} vs I_{OUT} for Boost Converter

regulator's feedback loop, setting output voltage. The RC network at the V_C pin provides loop compensation. Figure 3 shows the maximum output current for various input voltages and power devices.

Trace C is the input capacitor's (C1) current waveform. Capacitor current ramps up when the inductor is storing energy and down when the energy is transferred to the load. The input capacitor provides a low resistance AC return path for the inductor current during switch on and off times. Because of this, both the input capacitor and inductor have the same current waveform.

Trace D is the output capacitor's (C2) current waveform. The output capacitor's waveform is different than the input capacitor's. The output capacitor's current waveform ramps down when charging and goes flat when providing current to the load. The capacitor's ripple current waveform is often misinterpreted. The waveform shows peak to peak current. Capacitor current flows in both directions; into the capacitor when charging and out when discharging. The average capacitor current is zero, but its RMS value is not. Appendix B explains capacitor current waveforms in greater detail.

Figure 4 shows that efficiency for this circuit generally exceeds 75%.
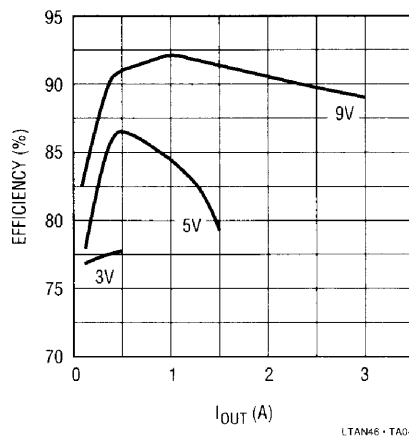
Figure 4. LT1070 Boost Converter Efficiency for Various Input Voltages and Load Currents

Semiconductor Losses

The output diode is often a major source of power loss in switching regulators, especially with output voltages below 10V. Inherent forward conduction losses limit the

LT LINEAR TECHNOLOGY

diode's overall efficiency. For output voltages below 10V, Schottky switching diodes are recommended for their minimum forward voltage drop and fast recovery time characteristics. A Schottky rectifier, with a typical V_f of 0.5V, introduces a loss of 4% for a 12V output and 10% in a 5V output. Its low peak inverse voltage rating limits use to low voltage applications. In higher output voltage applications the silicon diode forward voltage loss is less significant, because the diode's DC loss is a much smaller percent of the total loss.

Another loss element is the LT1070's switching transistor and its control circuitry. The bulk of this power loss is due to the switching transistor's conduction losses. Switch conduction loss includes both actual switch saturation and driver circuitry losses. The internal power switch is a bipolar device with a forced beta of 40. At low input voltage and high switch currents, switch saturation losses dominate with driver losses dominating under converse conditions. Only at light load does the controller's quiescent current significantly affect efficiency.

Capacitor Losses and Considerations

Filter capacitors, if chosen properly, dissipate only a few hundred milliwatts under worse case operating conditions adding very little to overall power loss. The primary concern when choosing a filter capacitor is its RMS ripple current rating. The ripple current rating is selected to limit temperature rise in the capacitor. Capacitors must be kept within manufacturer's ripple current specifications for efficient operation and long lifetime. Power dissipation can be determined by multiplying the capacitor's Effective Series Resistance (ESR) by the square of the RMS current. Physically larger capacitors have higher RMS current ratings than smaller capacitors since they have lower ESR and more surface area to dissipate heat. Another alternative is to parallel several smaller capacitors to achieve low ESR and acceptable component height. Appendix A discusses thermal consideration for aluminum electrolytic capacitors.

RMS current can vary significantly between input and output capacitors. For the boost circuit, the input capacitor RMS current is 400mA RMS, while the output capacitor is 1.3A RMS for a 1A load. Therefore, the output capacitor has to be larger than the input capacitor to handle the
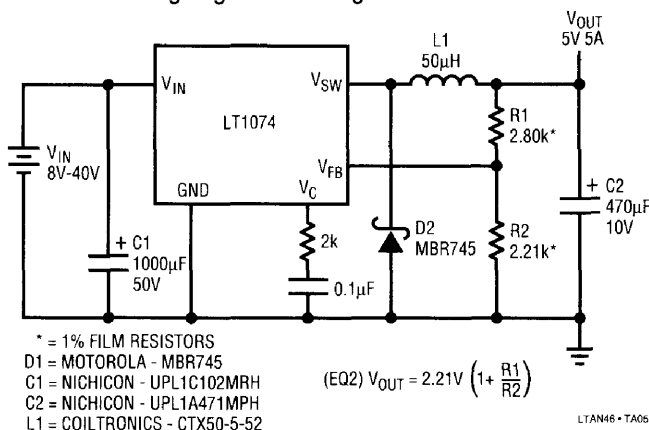
current. Appendix B details RMS current measuring techniques.

Inductor Loss Factors

A properly designed inductor will degrade efficiency by only a small percentage. Inductor losses are broken up into two categories: wire loss (copper loss) and core loss. Trade-offs associated with wire size, inductor value, core volume and core material influence final selection. Power loss in the copper wire only becomes significant at high current levels, since it is proportional to the square of the RMS current. Inductor losses also vary widely for different core materials and input voltages.

Basic Step-Down Switching Regulator

Many regulator requirements involve converting a higher voltage into a lower voltage. Although a linear regulator can do this, it cannot achieve the efficiency of switching regulator-based designs. Figure 5 shows a practical step-down switching regulator using the LT1074.

$$\text{(EQ2)}\ V_{OUT} = 2.21V \left(1 + \frac{R1}{R2}\right)$$

* = 1% FILM RESISTORS
D1 = MOTOROLA - MBR745
C1 = NICHICON - UPL1C102MRH
C2 = NICHICON - UPL1A471MPH
L1 = COILTRONICS - CTX50-5-52

Figure 5. LT1074 Positive Step-Down Switching Regulator

The operating waveforms for this circuit are shown in Figure 6. When the LT1074's high side switch turns on it pulls the V_{SW} pin to within 2.0V of the positive rail. Trace A is the V_{SW} pin voltage and trace B its current. During this period current flows from the input through the LT1074 and the inductor, and into the load. Inductor current is shown in trace C. When the LT1074 power switch turns off, the V_{SW} pin voltage drops until clamp diode D1 forward biases (trace D) providing a path for the inductor to transfer its energy to the output. Figure 7 shows maximum output current for various input conditions.

HORIZONTAL = 2µs/DIV   LTAN46 • TA06
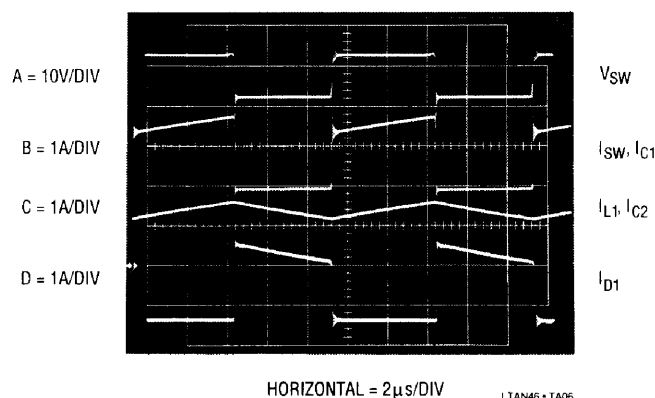
Figure 6. LT1074 Positive Step-Down Converter Waveforms

V_{IN} (V)   LTAN46 • TA07
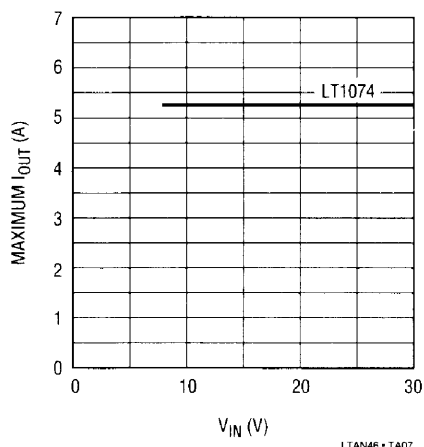
Figure 7. V_{IN} vs I_{OUT} for Positive Buck Converter

The filter capacitors provide a low impedance return path for AC current. The output capacitor's current waveform looks exactly like the inductor's current waveform (trace C), except the capacitor's current has no DC component. The RMS current for the output capacitor is quite low, approximately 150mA RMS with a 1A load. The input capacitor's current waveform is the same as the V_{SW} pin current (trace B). Its ripple current is approximately equal to 1/2 I_{OUT} (500mA) and is noticeably higher compared to the output capacitor current.

Figure 8 shows that efficiency can generally exceed 75%. The output diode and LT1074 switch are the two main loss elements. A Schottky diode is chosen for its low forward voltage drop. It introduces a 5% loss whereas a silicon diode would double this figure. The LT1074 switch has a relatively constant 2.0V loss. At low input voltages efficiency is degraded because this loss makes up a higher

I_{OUT} (A)   LTAN46 • TA08
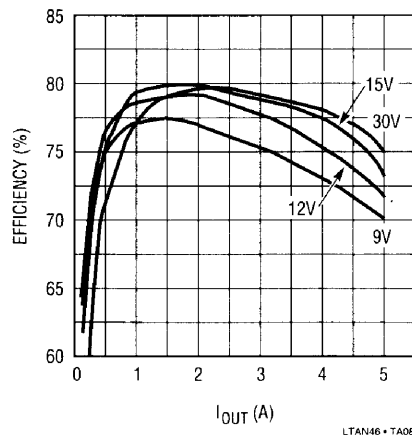
Figure 8. LT1074 Positive Buck Converter Efficiency for Various Input Voltages and Load Currents
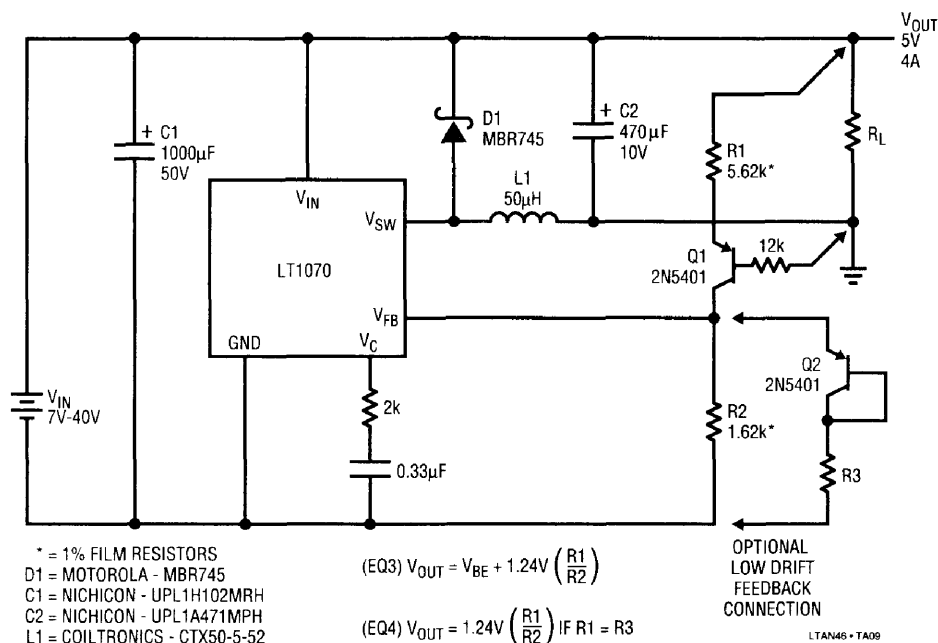
percentage of the available supply. Higher inputs make the fixed loss a smaller percentage, improving efficiency.
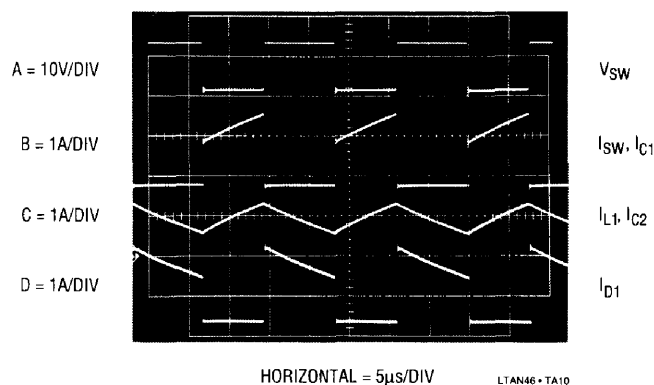
Floating Input Step-Down Switching Regulator

Figure 9 shows a way to obtain significantly higher efficiency at low input voltages. This technique utilizes the low saturation characteristics of the LT1070 power switch to obtain efficiency in excess of 85% for a 9V input. In this circuit the input voltage negative terminal is not connected to the output voltage negative terminal. The input must be allowed to float. A floating input can either be a battery or a galvanically isolated transformer's winding. This circuit is particularly useful in battery application since battery voltages are usually low and maximizing efficiency is often a critical issue.

This circuit operates similarly to Figure 5. The primary difference between this circuit and Figure 5 is the power switch type and location. Here, one side of the inductor is at ground potential and the LT1070's NPN power switch connects the other side to the input supply negative terminal. The LT1074 connects the inductor between the output voltage and the positive supply rail. With the inductor connected to ground instead of the positive rail, a common emitter NPN can be used instead of the much higher loss composite PNP used in the LT1074. This minimizes the power transistor's voltage drop contribution to power loss. Appendix C compares the conduction losses between the two switch types.

Switching waveforms are shown in Figure 10. The current waveforms are identical to those of Figure 6. However, the

* = 1% FILM RESISTORS
D1 = MOTOROLA - MBR745
C1 = NICHICON - UPL1H102MRH
C2 = NICHICON - UPL1A471MPH
L1 = COILTRONICS - CTX50-5-52

$$(EQ3) \ V_{OUT} = V_{BE} + 1.24V \left(\frac{R1}{R2} \right)$$

$$(EQ4) \ V_{OUT} = 1.24V \left(\frac{R1}{R2} \right) \text{ IF R1 = R3}$$

OPTIONAL
LOW DRIFT
FEEDBACK
CONNECTION

LTAN46 • TA09

Figure 9. LT1070 Floating Input Step-Down Switching Regulator

HORIZONTAL = 5μs/DIV   LTAN46 • TA10

**Figure 10. LT1070 Floating Input Step-Down Converter Waveforms**

LTAN46 • TA11
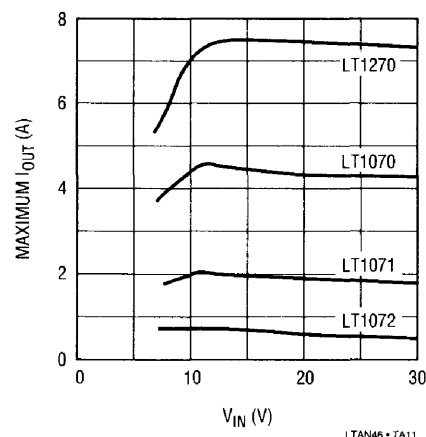
Figure 11. V_{IN} vs I_{OUT} Floating Input Buck Converter

V_{SW} pin voltage transitions are different. For the LT1070 circuit the V_{SW} pin voltage swings from the negative terminal of the input supply to a diode drop above the output voltage. In comparison, the LT1074's V_{SW} pin switches from the positive rail to a diode drop below ground. Figure 11 details the maximum output current for various input voltages and power devices. When the LT1070 operates at a duty cycle greater than 50% its maximum switch current is reduced, which causes the decrease in maximum output current seen at low input voltages.

The feedback senses the output with respect to the GND pin, so a level shift is required from the 5V output. Q1 serves this purpose, introducing only –2mV/°C drift, (see Equation 3). This is normally not objectionable in a logic supply, but can be compensated for with the optional appropriately scaled diode/resistor, (see Equation 4 in Figure 9).

Figure 12 shows this circuit's efficiency characteristics. Efficiency at low input voltages is significantly higher than the previous circuit because of lower power switch losses.
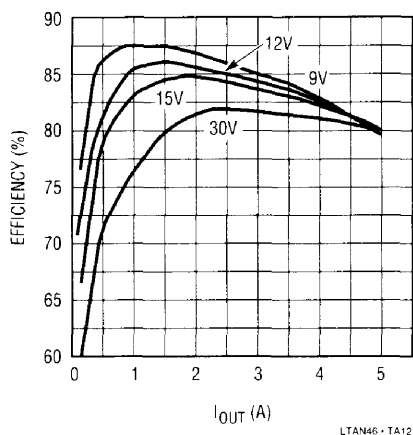
Figure 12. LT1070 Negative Buck Converter Efficiency for Various Input Voltages and Load Currents

At high input voltages and low output current levels the IC's quiescent current reduces the circuit's efficiency. The diode, filter capacitors, and inductor losses are comparable to the previous LT1074 buck circuit.
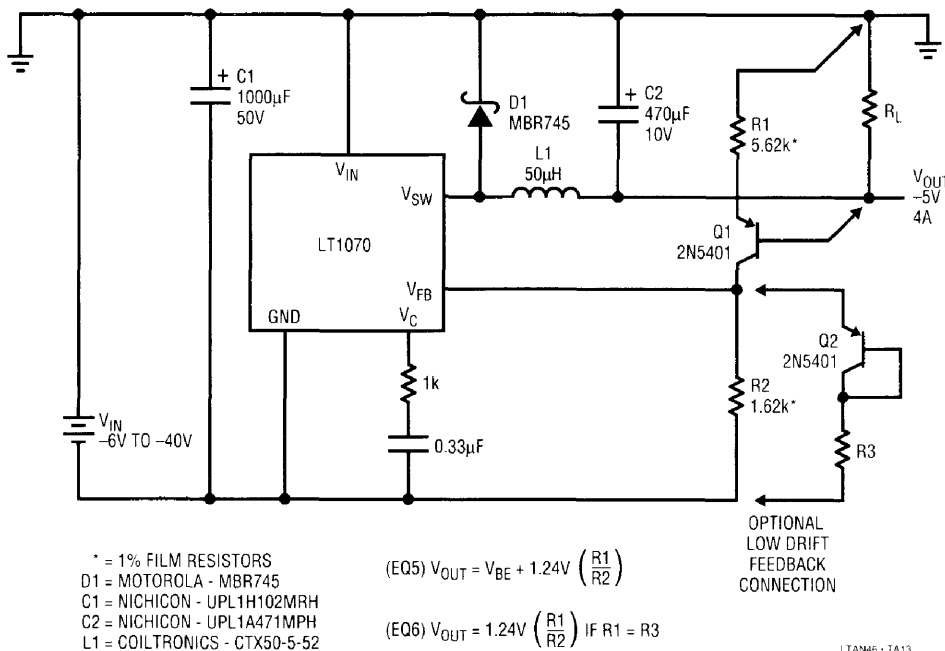
Figure 13 shows this circuit's topology used for negative buck conversion. Its operating waveforms are the same as shown in Figure 10. The common ground connection between the input and output voltage does not change the operating characteristics of the circuit.

High Efficiency Step-Down Switching Regulator

Although more complicated than the previous circuit, the high efficiency circuit in Figure 14 allows a common ground connection between input and output. Here, circuit complexity is traded off for increased efficiency at low input voltages.

The circuit operating characteristics are similar to that of the step-down regulator of Figure 5. In this case, an LT1070 common emitter NPN output switch is used to drive the inductor to the positive rail. Using an NPN switch achieves lower conduction losses than the composite PNP used in the LT1074.

The operating waveforms for this circuit are shown in Figure 15. When the LT1070's switch turns on, it pulls the GND pin to the input voltage. Trace A is the GND pin voltage and trace B is its current. During this period, current flows from the input through the LT1070 and the inductor, and into the load. Inductor current is shown in trace C. When the LT1070 switch turns off, the voltage on the GND pin drops until the clamp diode is forward biased (trace D is clamp diode current), providing a current path for the inductor to transfer its energy to the output. Maximum output current for various input voltages and power devices is shown in Figure 16.

$$(EQ5)\ V_{OUT} = V_{BE} + 1.24V \left(\frac{R1}{R2}\right)$$

$$(EQ6)\ V_{OUT} = 1.24V \left(\frac{R1}{R2}\right)\ IF\ R1 = R3$$

* = 1% FILM RESISTORS
D1 = MOTOROLA - MBR745
C1 = NICHICON - UPL1H102MRH
C2 = NICHICON - UPL1A471MPH
L1 = COILTRONICS - CTX50-5-52

Figure 13. LT1070 Negative Step-Down Switching Regulator

$$(EQ7) \quad V_{OUT} = 2.5V \left(1 + \frac{R1}{R2}\right) + \frac{0.3V \cdot R1}{R3}$$
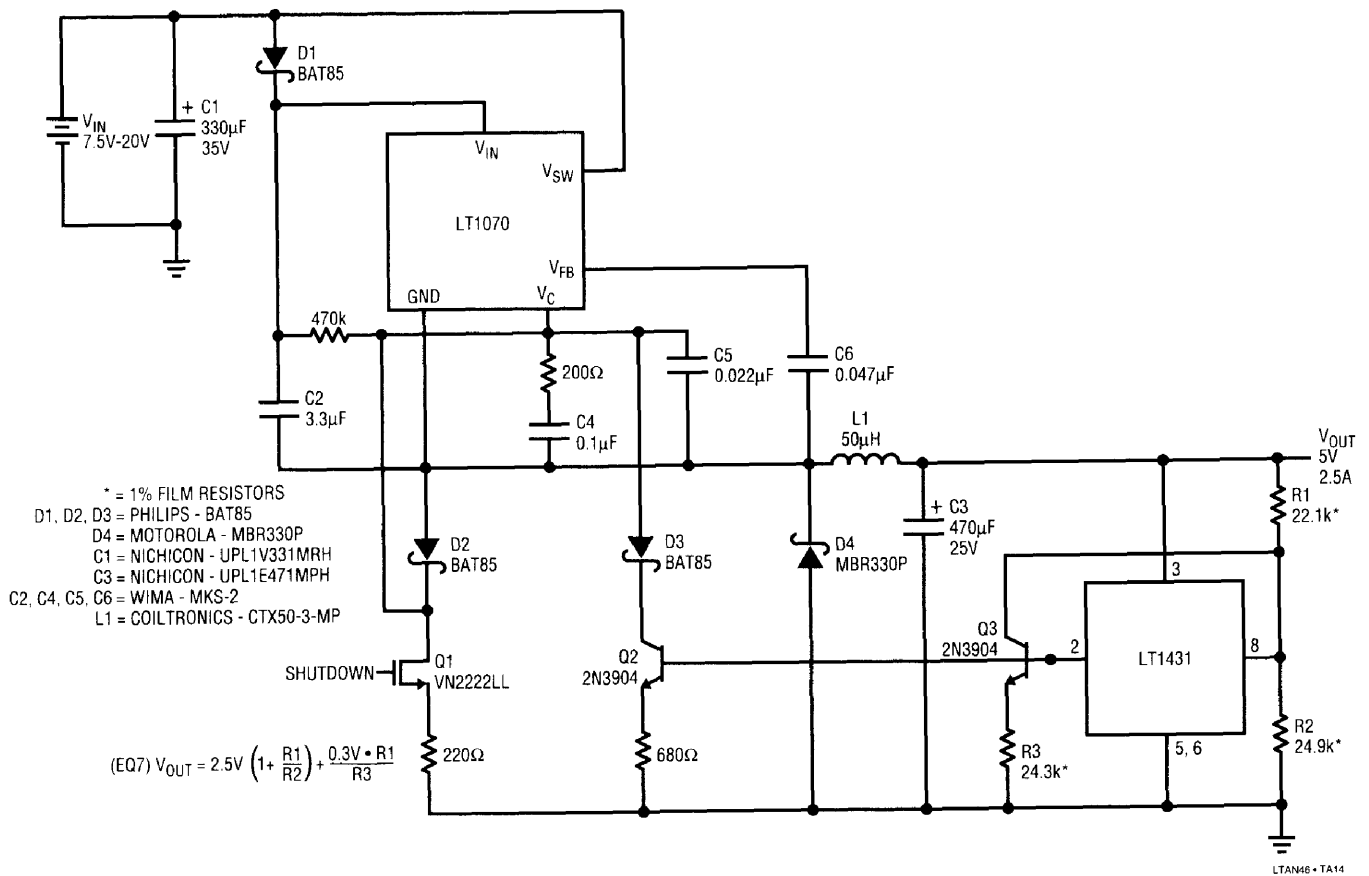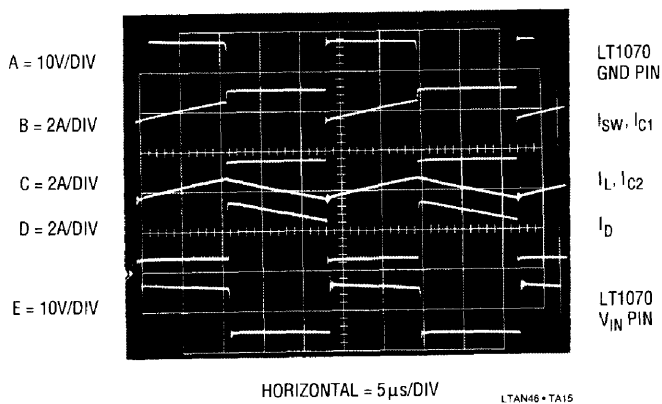
Figure 14. LT1070 High Efficiency Buck Converter

Figure 15. LT1070 High Efficiency Step-Down Converter Waveforms

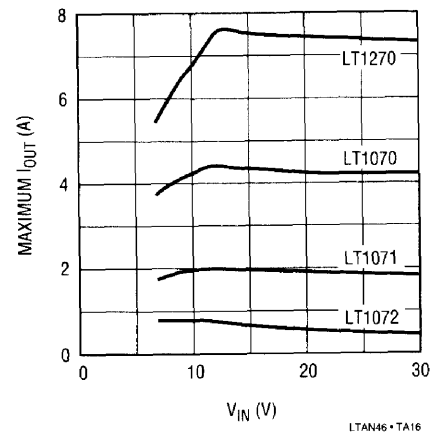**Figure 16. V_{IN} vs I_{OUT} for High Efficiency Buck Converter**

For this circuit to function the V$_{IN}$ pin must be driven above the input voltage (trace E). This is accomplished by bootstrapping C2 off the GND pin of the LT1070. C2 charges up through D1 when the LT1070 switch is off. When the switch turns on, D1 reverse biases allowing the V$_{IN}$ pin to rise above the input voltage. The GND pin is pulled to within a few hundred millivolts of the input voltage. The V$_{IN}$ pin voltage is now twice the input voltage. C2's stored charge provides adequate supply current and base drive for the power switch during this interval.

The output voltage is controlled by the LT1431, an adjustable shunt voltage regulator. The output voltage is set by the ratio of R1 and R2, (see Equation 7 in Figure 14). The LT1431's error amplifier compares the reference pin voltage to its internal 2.5V reference. The LT1431's output drives a shunt transistor, Q2, which absorbs current from the V$_C$ pin of the LT1070, adjusting duty cycle. The V$_C$ pin RC network provides sufficient loop compensation.

It is often desirable to put a switching regulator into "shutdown mode", a condition where the switching regulator is turned off and draws micropower current levels, to maximize battery life. One solution is to place a MOSFET switch in series with the input. This approach requires a large power device and reduces efficiency at high output currents. The LT1070 provides an elegant solution to this problem by integrating a shutdown feature. When the voltage between the V$_C$ and GND pin is pulled below 150mV, the IC shuts down pulling only 150μA. This is implemented by turning on Q1, reducing the circuit's quiescent current from 6mA to 150μA.

When the input voltage is first applied to this circuit, the regulator dumps full short circuit current into the output capacitor, attempting to bring the output up to its regulated value. The output can overshoot its desired value before the control loop is able to idle back the output current. This condition could overdrive Q2, forcing it to pull the voltage between the V$_C$ and GND pin below 150mV, and putting the LT1070 in its shutdown mode. The output voltage would momentarily drop to zero and remain there until the V$_C$ pin rises above 900mV. To prevent this condition from occuring, the V$_C$ pin is clamped by D3 and the 470k resistor is used as a pull up.

Figure 17 shows efficiency approaching 90%. Squeezing the utmost efficiency from this circuit requires care. The
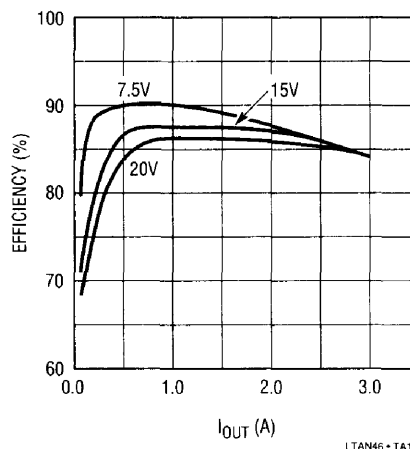
Figure 17. LT1070 High Efficiency Positive Buck Converter Efficiency

power switch and the catch diode conduction losses are the two main loss elements. These devices forward voltage drop must be minimized in order to maximize efficiency. A Schottky diode is used for its minimum forward voltage drop. The inductor selection is not a trivial task since it can add a couple percent loss. Low core loss material such as Molypermalloy, "high flux", "Kool Mu" (Magnetics, Inc.), and ferrite cores should be used. Normal design procedure for wire size may have to be modified to reduce copper loss.
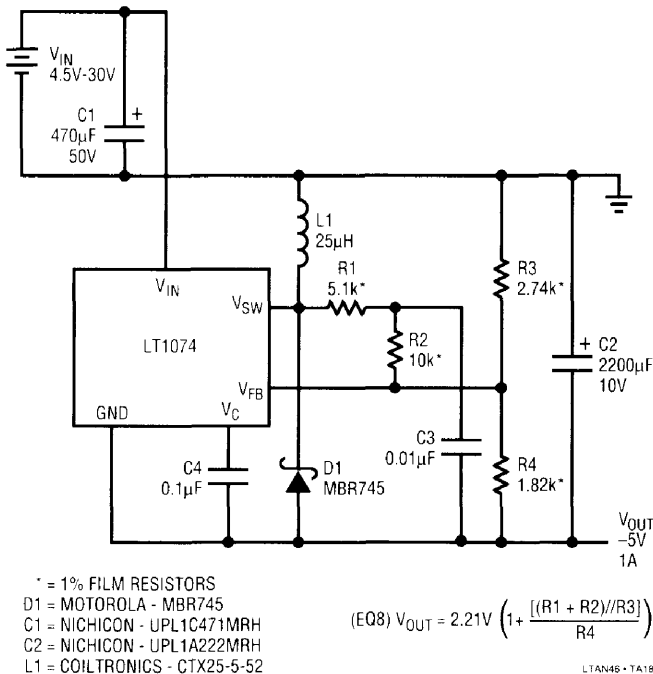
The LT1431 and associated control circuitry can be replaced by an LT1432. Refer to the LT1432 data sheet for further details.

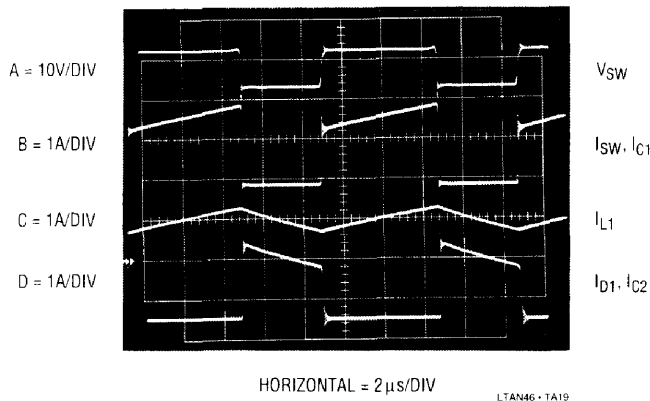## Positive to Negative Buck Boost Switching Regulator

A frequent switching regulator application is to produce a negative output from a positive supply, usually 5V. One approach is to use a transformer in a flyback topology, but transformers are not off-the-shelf components. They are expensive, especially in low volumes, and can have long delivery times. Alternately, a negative output is easily obtained with a simple inductor. Inductors are more desirable than transformers in many converter designs because they are readily available and economical. The negative output requirement is easily fulfilled by the circuit shown in Figure 18.

The operating waveforms are shown in Figure 19. When the LT1074 switch turns on, current flows from the input through the LT1074 and into the inductor. Trace A is the switch pin voltage and trace B is its current. During this

* = 1% FILM RESISTORS
D1 = MOTOROLA - MBR745
C1 = NICHICON - UPL1C471MRH
C2 = NICHICON - UPL1A222MRH
L1 = COILTRONICS - CTX25-5-52

$$\text{(EQ8)} \quad V_{OUT} = 2.21V \left(1 + \frac{[(R1 + R2)//R3]}{R4}\right)$$

Figure 18. LT1074 Positive to Negative Switch

HORIZONTAL = 2μs/DIV

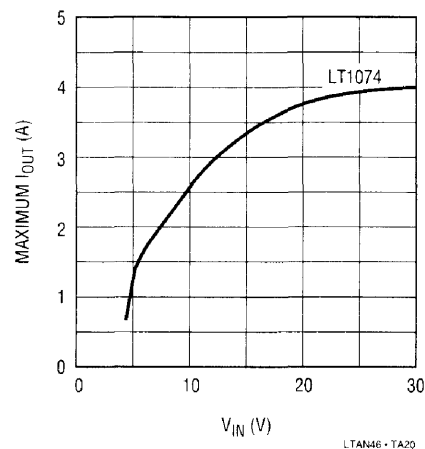**Figure 19. LT1074 Positive to Negative Converter Waveforms**

period current ramps up in the inductor (trace C) as energy is stored in the core. When the LT1074 power switch turns off, the voltage on the V_{SW} pin drops until clamp diode D1 forward biases (trace D). This provides a current path for the inductor to transfer its energy to the output. Figure 20 shows this circuit can provide a 1A output for a 4.5V input.

In this architecture the LT1074's GND pin is tied to the negative output rather than to ground. This technique eliminates a level shifting op amp in the feedback path. Feedback is sensed from circuit ground, and the regulator forces its feedback pin to 2.21V above its GND pin. The

output voltage can be varied by changing the R1-R2-R3-R4 divider ratio, (see Equation 10 in Figure 26). The LT1074 controls duty, cycle to achieve output voltage regulation.

Positive to negative converters have a "right half plane zero" in the transfer function, which makes them particularly hard to frequency stabilize, especially with low input voltage. R1, R2 and C3 form an AC feedforward path needed for loop compensation at low input voltages. They can be omitted for $V_{IN} > 10V$, or $V_{IN}/V_{OUT} > 2$. This network along with C4 provides stable loop frequency compensation.

Efficiency generally exceeds 60% as shown in Figure 21. Efficiency is degraded at low input voltages where the LT1074 power switch is responsible for the majority of the efficiency loss. Its 2.0V switch voltage drop cuts the

Figure 20. V_{IN} vs I_{OUT} for Positive to Negative Converter
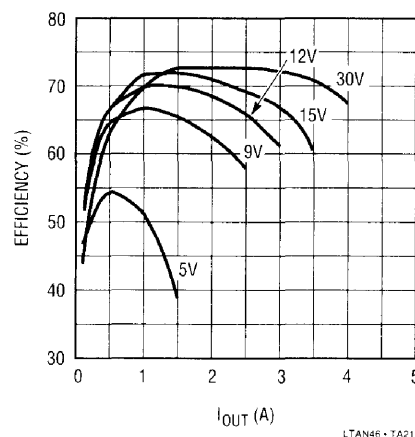
Figure 21. LT1074 Positive to Negative Converter Efficiency for Various Input Voltages and Load Currents

efficiency by 28% at a 5V input; another 10% is contributed by the output diode. Higher input voltages make the fixed LT1074 voltage drop a smaller percentage, improving efficiency.

High Efficiency Positive to Negative Switching Regulator

The previous circuit has excellent efficiency performance above a 12V input. However, at low input voltages the efficiency falls off dramatically. The LT1074's internal power switch voltage drop is the major contributor to the degradation of efficiency. Figure 22 shows an alternative approach to generate a negative output from a positive input. Here, the LT1070 low loss switching transistor achieves remarkable efficiency levels, even with a 5V input.

This circuit is reminiscent of the high efficiency buck converter of Figure 14. The control circuitry and the manner in which the LT1070 power switch is driven are

identical. However, the power components route the current through the same course as the previous positive to negative design; therefore, their switching characteristics are the same.

The switching waveforms for this circuit are shown in Figure 23. Trace A is the LT1070's GND pin voltage and trace B is its current. Current flows through the LT1070 and the inductor when the power switch is on. When the switch turns off, the voltage on the LT1070's GND pin drops until diode D4 forward biases. Inductor current (trace C) then flows through D4 (trace D).

Figure 25 shows this circuit's efficiency approaches 70% for a 5V input. The higher efficiency at low input voltage results from the LT1070's extremely low switch conduction loss. The effect of switch loss can be seen by comparing the efficiency characteristics of Figure 21 and Figure 25. The LT1074-based circuit's efficiency drops dramatically at inputs below 9V.
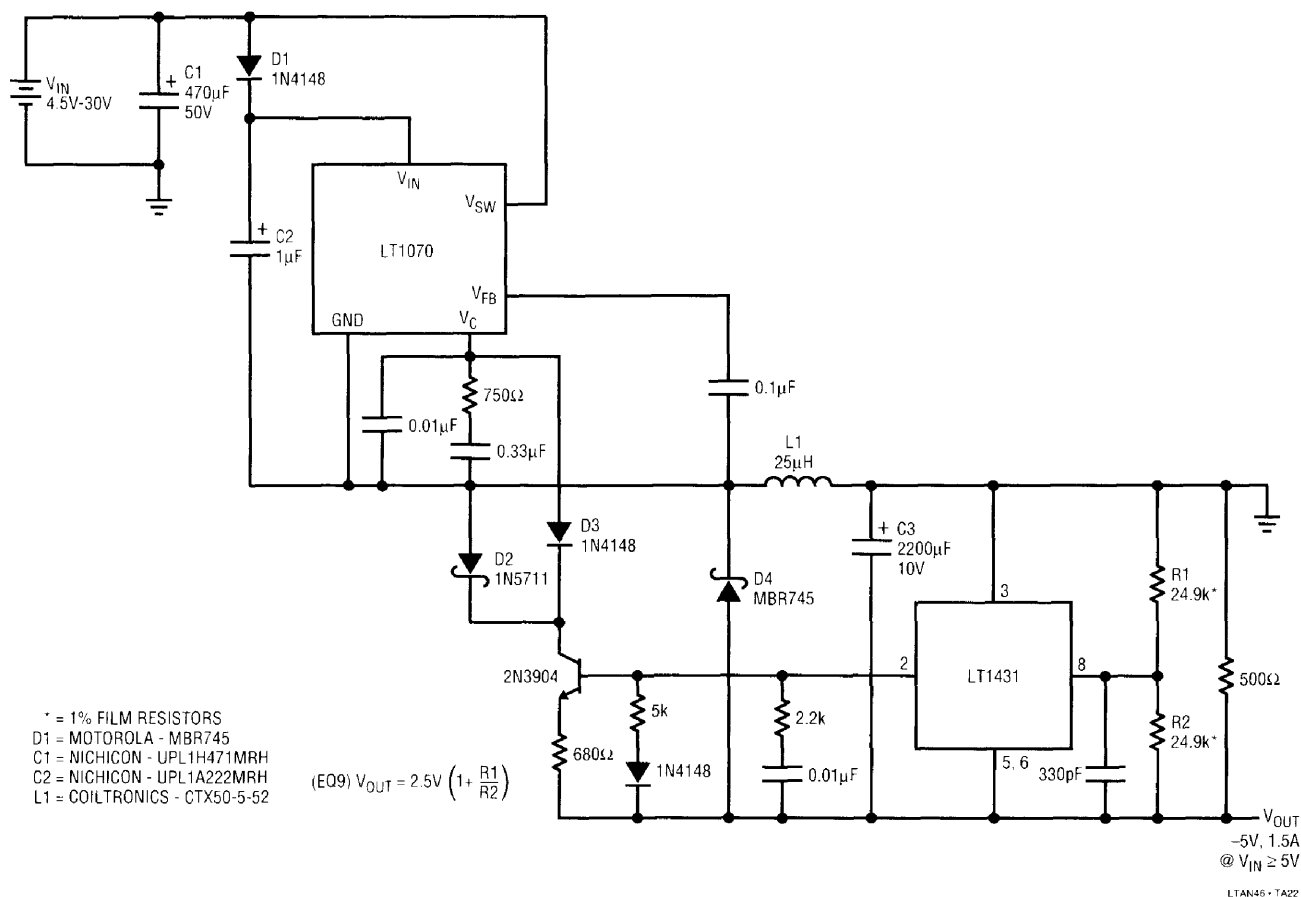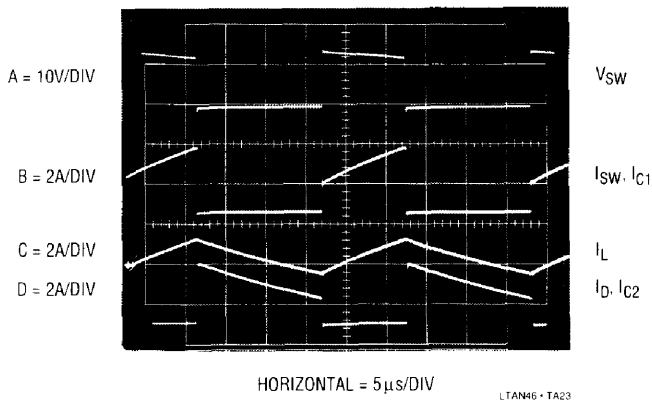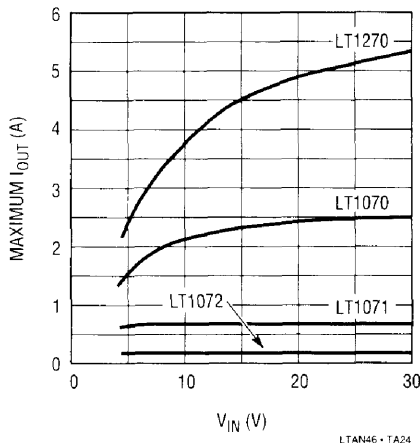

$$(EQ9)\ V_{OUT} = 2.5V \left(1 + \frac{R1}{R2}\right)$$

Figure 22. LT1070 High Efficiency Positive to Negative Switching Regulator

A = 10V/DIV   V_SW

B = 2A/DIV   I_SW, I_C1

C = 2A/DIV   I_L

D = 2A/DIV   I_D, I_C2

HORIZONTAL = 5μS/DIV

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Figure 23. LT1070 High Efficiency Positive to Negative Converter Waveforms


Figure 24. V$_{IN}$ vs I$_{OUT}$ for High Efficiency Positive to Negative Converter

LT1070 Negative to Positive Switching Regulator

This converter topology can maintain a constant output voltage whether the absolute value of the negative input voltage is greater or less than the positive output voltage. This is extremely desirable in battery operated electronic equipment. This flexibility reduces the number of battery cells required and provides a constant output voltage over the battery's complete operating range, maximizing battery life. The circuit is shown in Figure 26. This technique can be used only if the input and output voltage negative terminals are not connected to each other.

This circuit operates similar to the boost regulator of Figure 1, but is intended for buck-boost conversion. Here, the positive terminal of the battery is connected to ground
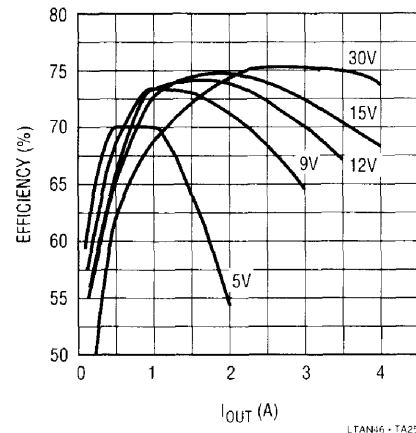
Figure 25. LT1070 High Efficiency Positive to Negative Converter Efficiency for Various Input Voltages and Load Current

and the LT1070's GND pin is connected to the negative terminal. The feedback pin senses with respect to GND pin, so Q1 provides a level shift from the 5V output.

Figure 27 shows the circuit operating waveforms. They resemble those of the boost regulator (Figure 2). The primary difference between the two circuits is that the inductor current does not flow through input capacitor C1 during the switch off time. This is noticeable in the input capacitor's ripple current waveform (trace B). The ripple current is significantly higher, compared to the boost circuit, since the current is being pulled from the input capacitor in pulses. This increased ripple current necessitates a larger input capacitor. Maximum output current for various input voltages and power devices is shown in Figure 28.

Figure 29 shows circuit efficiency in excess of 75%. Once again the main contributors to power loss are the switching diode and LT1070's internal power switch.

Flyback Converter

Many applications require multiple regulated output potentials. A popular output combination is 5V and ±12V as implemented in the circuit shown in Figure 30.

This circuit is a basic flyback regulator. The transformer transfers energy from the 12V input to the 5V and ±12V outputs. Figure 31 shows the operating waveforms for this circuit. Trace A is the voltage at the V$_{SW}$ pin and trace B is its current. During the V$_{SW}$ on-time, the V$_{SW}$ pin is pulled

$* = 1\%$ FILM RESISTORS
D1 = MOTOROLA - MBR745
C1 = NICHICON - UPL1H102MRH
C2 = NICHICON - UPL1A472MRH
L1 = COILTRONICS-CTX50-5-52

$$(EQ10)\ V_{OUT} = V_{BE} + 1.24V \left(\frac{R1}{R2}\right)$$

$$(EQ11)\ V_{OUT} = 1.24 \left(\frac{R1}{R2}\right) \text{ IF R1 = R3}$$
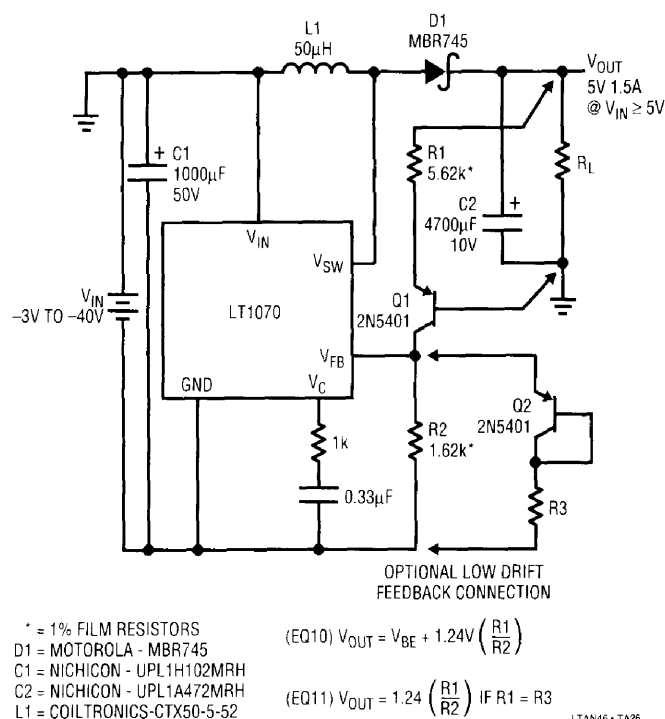
Figure 26. LT1070 Negative to Positive Switching Regulator

HORIZONTAL = 5µs/DIV
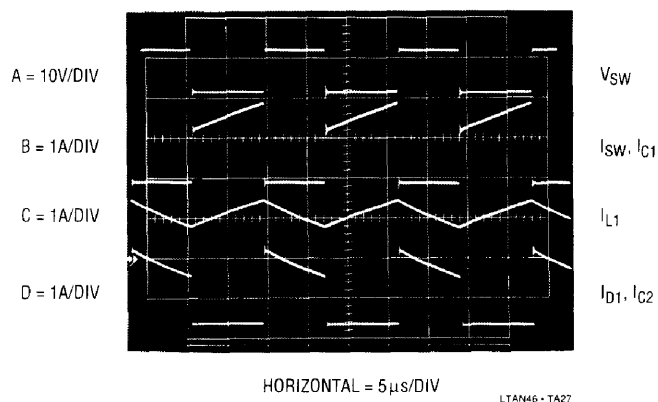
Figure 27. LT1070 Negative to Positive Converter Waveforms

to ground forcing the input voltage across the primary winding. After the initial jump, the primary current (trace C) rises slowly as the magnetic field builds up. The magnetic field in the core induces a voltage on the secondary windings, which is proportional to the input voltage times the turns ratio. However, no power is transferred to the outputs because the rectifier diodes are reverse biased. The energy is stored in the transformer's magnetic field. When the switch is turned off, energy is no longer
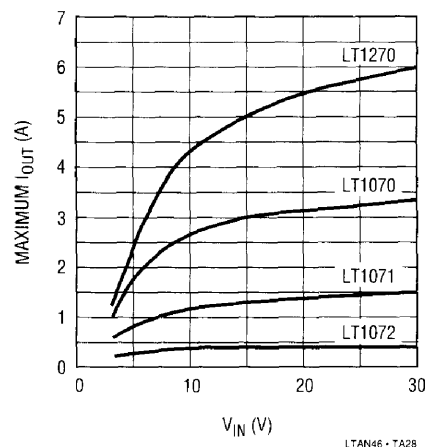
Figure 28. V_{IN} vs I_{OUT} for Negative to Positive Converter
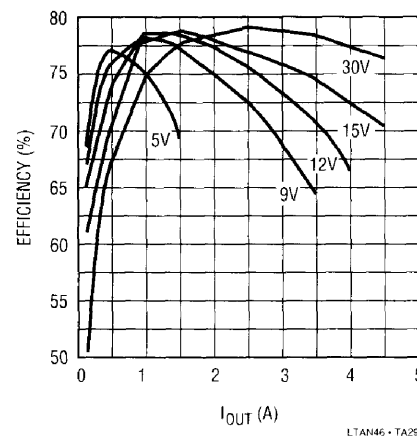
Figure 29. LT1070 Negative to Positive Converter Efficiency for Various Input Voltages and Load Currents

transferred to the transformer causing the magnetic field to collapse. The collapsing magnetic field induces a change in voltage across the transformer's windings. During this transition the V_{SW} pin's voltage flies to a potential above the input voltage, the secondary forward biases the rectifier diodes, and the transformer's energy is transferred to the outputs. Trace D is the voltage seen on the 5V secondary and trace E is its current.

This is not an ideal transformer, so not all the energy is coupled into its secondary windings. The energy left in the primary winding causes the overvoltage spike seen on the V_{SW} pin (trace F). This phenomenon is modeled by an inductor term placed in series with the primary winding. When the switch is turned off, current continues to flow in the primary winding, causing D1 and D2 to conduct (trace G). This diode network clamps the flyback voltage spike,

* = 1% FILM RESISTORS
** = CURRENT LIMIT = (15k/R_LIM) A
L1 = COILTRONICS - CTX02-11808-1
L2 = COILTRONICS - CTX5-5-FR
D1 = MOTOROLA - MBR745
D2 = MOTOROLA - MUR110
D3 = MOTOROLA - P6KE36A
C1 = NICHICON - UPL1E471MRH6
C2, C3 = NICHICON - UPL1A222MRH6
C4 = NICHICON - UPL1A221MPH
C5, C6 = NICHICON - UPL1E471MPH

(EQ12) $V_{OUT} = 1.24V \left(1 + \dfrac{R1}{R2}\right)$

(EQ13) $-V_{OUT} = 2.37V \left(1 + \dfrac{R3}{R4}\right)$

(EQ14) $V_{OUT2} = 2.37V \left(1 + \dfrac{R5}{R6}\right)$

LTAN46 • TA30
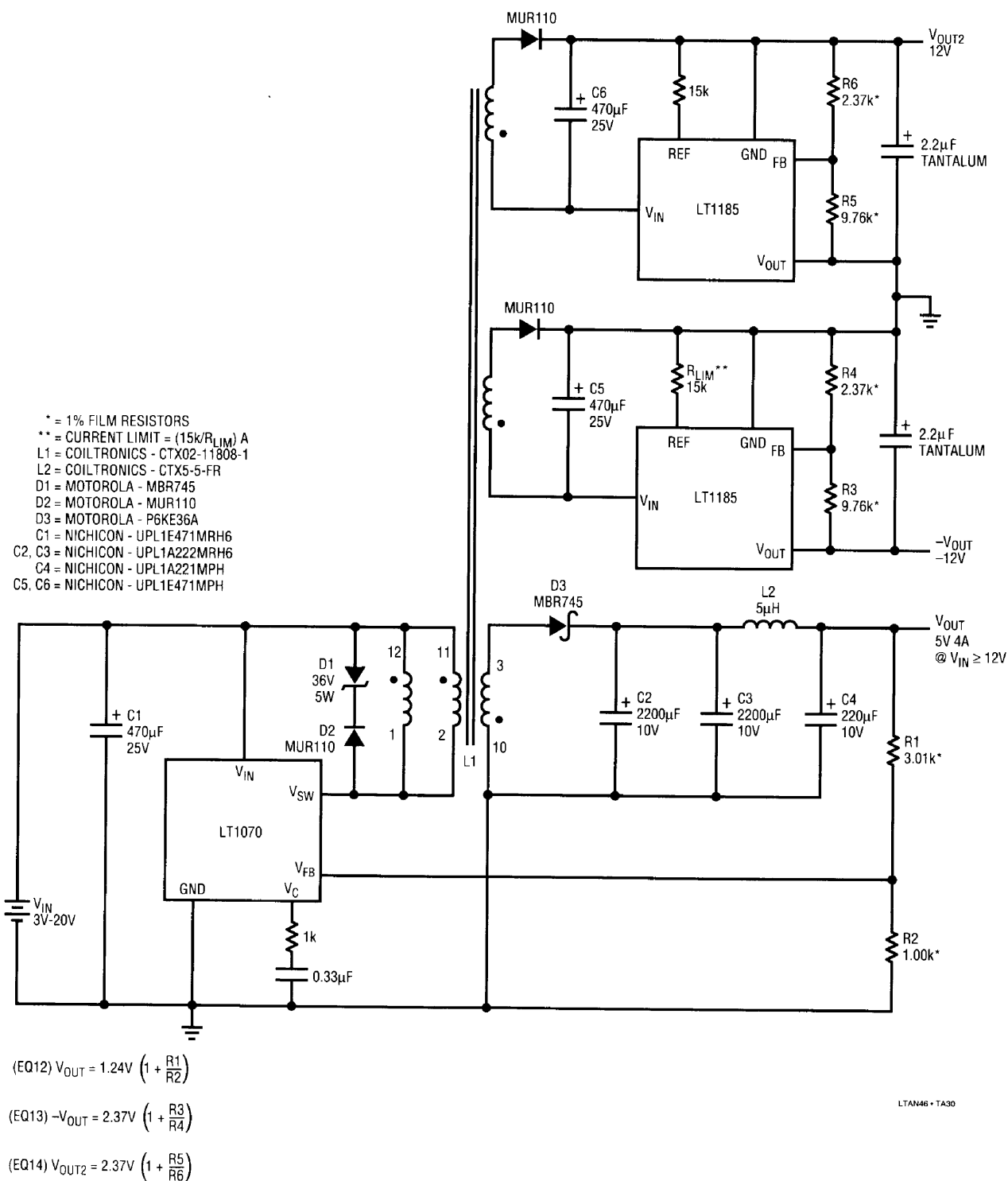
Figure 30. LT1070 Flyback Switching Regulator

A = 50V/DIV — LT1070 SWITCH VOLTAGE

B = 2A/DIV — LT1070 SWITCH CURRENT

C = 2A/DIV — PRIMARY CURRENT

D = 20V/DIV — 5V SECONDARY VOLTAGE

E = 10A/DIV — 5V SECONDARY CURRENT

F = 20V/DIV — LT1070 SWITCH VOLTAGE

G = 2A/DIV — SNUBBER DIODE CURRENT

A, B, C, D, E HORIZONTAL = 5μs/DIV
F, G HORIZONTAL = 1μs/DIV

LTAN46 • TA31
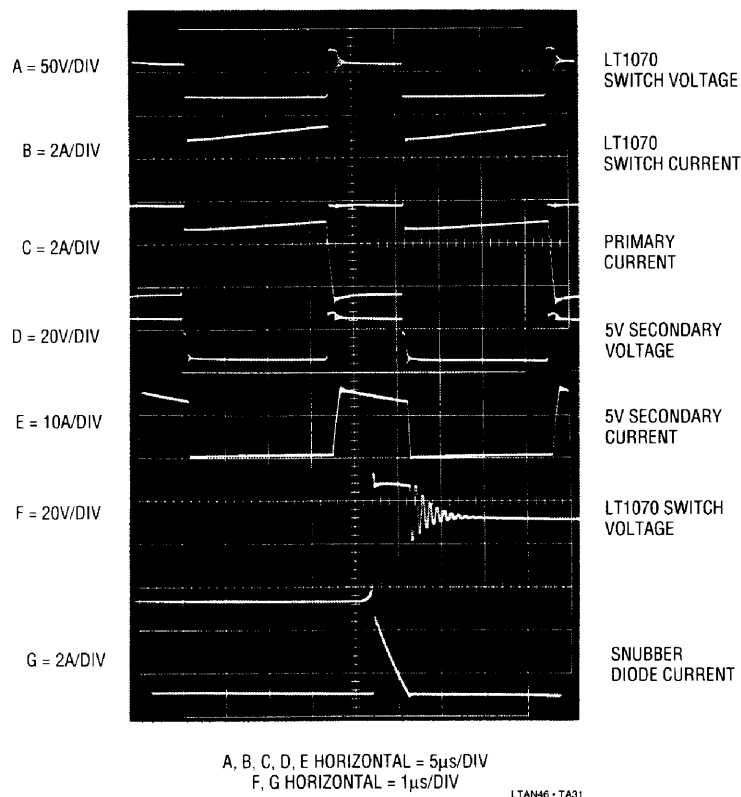
Figure 31. Waveforms for the LT1070 Flyback Converter

preventing excessive voltage at the LT1070's V_{SW} pin. When the primary current reaches zero, the V_{SW} pin's voltage settles to a potential related to the turns ratio, output and input voltage.

How well the unregulated outputs track each other, often referred to as cross regulation, depends upon how tightly they are magnetically coupled to one another. Post regulators are needed on the unregulated outputs if the cross regulation error is too great. Such error can be as much as 20% depending upon output loading conditions.

The isolated secondaries allow a negative voltage regulator to be used to regulate the +12V output. The advantage of the LT1185 over standard linear regulators is its ability to control current limit to typically within 4%, between its 1A-3A range, allowing the use of smaller rectifier diodes, secondary windings wire size, and core size. The isolated secondary windings' allows the input of the LT1185 to float below ground. The LT1185 negative voltage regulators maintain both positive and negative outputs.

Figure 32 represents the total available output power for various input voltages and power devices. For simplicity the available output power is summed into the 5V output. If the auxiliary outputs are used, the maximum available current from the 5V output is reduced.

This flyback circuit's efficiency approaches 80% (see Figure 33). Power is primarily dissipated in the LT1070, catch diode D3, and zener diode D1. The LT1070 and catch diode impose losses in the same manner as they did in the previous circuits. The zener clamp diode is a power loss element commonly found in flyback designs. It dissipates energy stored in the transformer's leakage inductance. To keep leakage inductance losses to a reasonable level, leakage inductance should be kept to less than 1% of the primary inductance.

For the flyback topology, the output filter capacitor takes a real beating at high output currents. This is due to the transformer providing current to the output capacitor in pulses. In many flyback designs the turns ratio is less than
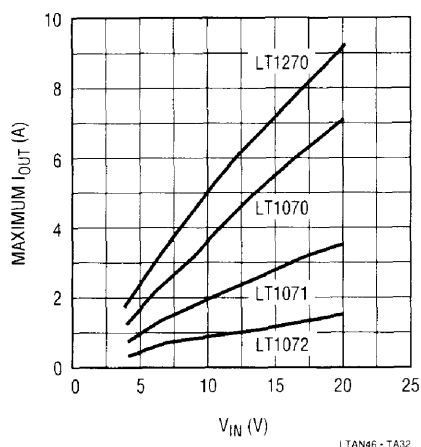
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Figure 32. V_{IN} vs I_{OUT} for Flyback Converter

Figure 33. LT1070 Flyback Converter Efficiency for Various Input Voltages and Load Currents

one and the secondary current is 1/N times higher than the primary current (see Figure 31). The primary winding peak current (trace C) is 5A, whereas the secondary winding peak current (trace E) is 15A. In this case two capacitors connected in parallel are needed to handle the RMS ripple current.
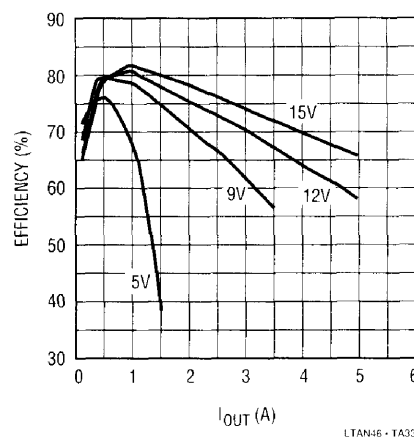
References

1. Coiltronics
   984 S.W. 13th Court
   Pompano Beach, FL 33069
   Phone: (305) 781-8900
   Fax: (305) 782-4163

APPENDIX A

Thermal Considerations for Aluminum Electrolytic Filter Capacitors

Aluminum electrolytic capacitors often fail in switching regulators because many designers do not view them as power components. Like any power device, capacitors have thermal limitations which must be observed for acceptable performance and reliability. Excessive capacitor temperature can cause an open or short circuit, capacitance drop, electrolyte leakage, increased leakage current or safety venting.

Increased temperature causes a gradual evaporation of the electrolyte through the capacitor's seal. As the electrolyte is lost, the capacitance is reduced and Effective Series Resistance (ESR) rises, causing increased power dissipation. If this regenerative process continues it can cause the capacitor to exceed its maximum thermal rating. In poorly designed power supplies it is not uncommon to have early field failures because the electrolyte in a filter capacitor has dried up.

Filter capacitors are chosen by physical size rather than capacitance value, since larger size capacitors have higher ripple current ratings, lower ESR and more heat dissipation capability. Selecting the appropriate size for a given application depends upon several factors:

- Ripple current rating/ESR
- Position on the PC board
- Maximum ambient temperature
- Load life

Additional factors include output voltage ripple and loop stability. These secondary considerations are not treated here.

The capacitor's operating ripple current sets the minimum acceptable capacitance size. Maximum allowable ripple current is selected to limit temperature rise in the capacitor. This internal temperature rise is proportional to the square of the capacitor's ripple current. Typical core to ambient temperature rise is between 5°C to 10°C. For reliable operation the capacitor must operate below the

maximum allowable ripple current. Appendix B explains how to measure operating ripple current.

There is a tendency for designers to select filter capacitors based on capacitor value instead of ripple current. This approach can be catastrophic because ripple current ratings vary widely between capacitor technologies, manufacturers and voltage ratings. Figure A1 shows how varied the ripple current rating is for similar value capacitors with different voltage ratings. In this example, the 63V part can handle over twice the RMS current of the 6.3V device because the higher voltage capacitor is physically larger.

Voltage Rating	6.3V	10V	16V	25V	35V	50V	63V
I_{RMS} (mA)	950	1060	1410	1660	1989	2120	2370

Nichicon PL series 105µF-105°C

Figure A1. Ripple Current Ratings for Different Voltage Ratings

High ripple currents require capacitors to have low ESR, greater surface area, and a high heat transfer constant. Figure A2 shows the relationship between volume and ripple current rating. Tall capacitors of equal volume as short, fat ones tend to be able to dissipate more heat since they can transfer the heat to the case surface more easily.
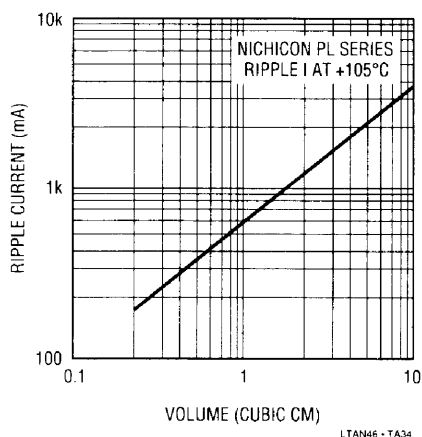
Figure A2. Ripple Current vs Volume

When the required ripple current is greater than the maximum rated ripple current, it becomes necessary to parallel capacitors. Paralleling allows sharing of the ripple current between capacitors. The ESR of each capacitor acts as a current ballasting impedance. In some instances it may be preferable to parallel capacitors even when a single device of higher ripple current rating is available. This allows smaller size capacitors to be utilized. Heat flow increases from multiple capacitors when compared to a single device with a higher current rating because the heat is spread over a greater area.

The ESR of the capacitor is the predominant cause of internal temperature rise above ambient. The power loss in the capacitor is determined by:

$$P_{CAP} = (I_{RMS})^2 \bullet ESR$$
$$ESR = \text{Effective Series Resistance}$$
$$I_{RMS} = \text{Capacitor Ripple Current}$$

Capacitor size versus ESR rating varies widely for different capacitor technologies and between manufacturers. Figure A3 shows ESR vs Volume for Different Capacitor Manufacturers.
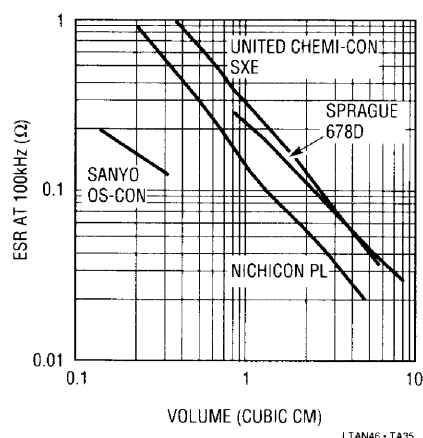
Figure A3. ESR vs Volume for Different Capacitor Manufacturers

Printed circuit board layout can have a dramatic effect on a capacitor's operating temperature. For example, one terminal of an output capacitor is often connected to a catch diode. During full load conditions the diode can reach junction temperatures in excess of 100°C, dissipating several watts. The diode's leads conduct the heat into the PC board traces, where it can be transferred into the filter capacitor, elevating its internal temperature. Wide PC board traces at the diode will dissipate this heat, reducing capacitor heating.

Another frequently overlooked layout consideration is the capacitor's location relative to heatsinks. Heatsinks radiate heat, increasing an adjacent capacitor's temperature.

Capacitor load life includes both parametric and catastrophic failures. A capacitor is considered failed if capacitance, ESR or leakage current exceeds maximum allowable variation. ESR variation is the parameter of primary concern for switching regulators because, as ESR increases, output voltage ripple and thermal dissipation go up. The load life of a capacitor is rated in hours, typically between 1,000 hours and 10,000 hours, with full rated voltage applied across its terminals at a specified temperature, usually 105°C. Load life definitions vary for different type of capacitors.

Load life increases for larger case diameters; refer to Figure A4. Load life is primarily limited by electrolyte loss. One reason for the increased load life between case sizes is that larger devices hold more electrolyte; it simply takes longer for the electrolyte to dry up.

Case Diameter (mm)	0.5	0.6	0.8
Load Life (hours)	2,000	3,000	5,000

Nichicon PL series 105μF-105°C

Figure A4. Load Life Rating for Different Case Diameters

The load life specification of a capacitor can be a little confusing. For example, a Nichicon PL series capacitor with a case diameter of between 8mm-10mm has a load life of 3,000 hours while being operated at 105°C. This information as stated is not very useful, since most products are designed to last longer than 18 weeks (3,000 hours), and will not be subject to 105°C.

The relationship between thermal stress and expected life of the capacitor can be predicted by:

$$L_X = L_0 * 2^{\frac{105°C - (T_X + \Delta T_X)}{10}} \qquad (A1)$$

L_X: Life expectancy in actual operation (hrs)
L_0: Load life at maximum operating temperature (hrs)
T_X: Ambient temperature in actual operation (°C)
ΔT_X: Temperature rise produced by ripple current (°C)

$$\Delta T_X = \frac{I_X^2 \bullet ESR}{B \bullet A} \qquad (A2)$$

I_X: Operating ripple current (A)
ESR: Effective Series Resistance (Ω)

B: Heat transfer constant — determined by case size (W/cm²/°C)
A: Case surface area (cm²)

$$A = \frac{\pi \bullet \phi D \left(\phi D + 4L \right)}{4} \qquad (A3)$$

ϕD: Case diameter
L: Case length

Heat transfer constants are not normal data sheet parameters, but can be obtained from the manufacturer. Figure A5 shows the heat transfer constant for United Chemi-Con electrolytic capacitors. The heat transfer constant is predominately affected by the thermal characteristics of the capacitor's aluminum case and its surface area. Since the aluminum material is the same for all series capacitors, the heat transfer constant depends only on surface area. These thermal parameters assume that the case is completely filled with the foil winding. Not all capacitors' cans are full, and should be checked by disassembling a sample.

Equation A1 shows that with aluminum electrolytic capacitors, load life doubles for every 10°C drop in operating temperature. The equation includes the effects of operating ripple current, ambient temperature and heat transfer constant of the package.

For example, a United Chemi-Con SXE25VB471M10X20LL capacitor operating at a ripple current of 860mA and an ambient temperature of 60°C would have an calculated life time of 3.1 years.

$I_X = 860mA$, $ESR = 0.14Ω$, $L_0 = 2,000$ hrs, $T_X = 60°C$

$\phi D \times L = 10 \times 20$ (mm), $B = 0.0019$, $A = 7.1cm^2$

$$\Delta T_X = \frac{0.860A^2 \bullet 0.14Ω}{0.0019 \bullet 7.1\,cm^2} = 7.06°C$$

$$L_X = 2,000\,hrs \bullet 2^{\frac{105°C - (60°C + 7.06°C)}{10}}$$

$L_X = 27,665\,hrs = 3.1\,yrs$

The total lifetime and operating duty cycle of a product must first be defined in order to generate a capacitor's actual total operating hours. Then the lifetime equation (A1) is used to select a filter capacitor that can meet these operating conditions. Lap top computers, for instance,

φD × L (mm × mm)	A (cm²)	B (W/cm²/°C)	B • A (W/°C)	φD × L (mm × mm)	A (cm²)	B (W/cm²/°C)	B • A (W/°C)
5 × 11	1.9	0.00210	0.00399	30 × 40	44.8	0.00097	0.0435
6.3 × 11	2.5	0.00208	0.00520	30 × 50	54.2	0.00090	0.0488
8 × 11.5	3.3	0.00206	0.00680	30 × 60	63.6	0.00085	0.0541
8 × 14	4.0	0.00200	0.00800	35 × 40	53.6	0.00090	0.0482
10 × 12.5	4.7	0.00201	0.00945	35 × 50	64.6	0.00084	0.0543
10 × 16	5.8	0.00198	0.0115	35 × 60	75.6	0.00080	0.0605
10 × 20	7.1	0.00190	0.0135	35 × 70	86.6	0.00076	0.0658
12.5 × 20	9.1	0.00182	0.0166	35 × 80	97.5	0.00074	0.0722
12.5 × 25	11.0	0.00178	0.0196	35 × 100	119.5	0.00070	0.0837
13 × 20	9.5	0.00182	0.0173	40 × 50	74.5	0.00080	0.0603
13 × 25	11.5	0.00178	0.0205	40 × 60	88.0	0.00075	0.0660
13 × 30	13.5	0.00170	0.0230	40 × 70	100.6	0.00074	0.0744
16 × 25	14.6	0.00164	0.0240	40 × 80	113.1	0.00072	0.0814
16 × 31.5	17.8	0.00156	0.0276	40 × 90	125.7	0.00070	0.0880
16 × 35.5	19.9	0.00146	0.0291	40 × 100	138.2	0.00070	0.0967
16 × 40	22.1	0.00140	0.0310	40 × 110	150.8	0.00070	0.106
18 × 31.5	20.3	0.00146	0.0297	50 × 60	113.8	0.00072	0.0819
18 × 35.5	22.6	0.00140	0.0317	50 × 70	129.6	0.00070	0.0907
18 × 40	25.1	0.00130	0.0327	50 × 80	145.3	0.00070	0.102
18 × 45	26.0	0.00122	0.0342	50 × 90	161.0	0.00070	0.113
22.4 × 30	25.0	0.00130	0.0325	50 × 100	176.7	0.00070	0.124
22.4 × 40	32.1	0.00112	0.0360	50 × 110	192.4	0.00070	0.135
22.4 × 50	39.1	0.00102	0.0399	50 × 120	208.1	0.00070	0.146
25 × 30	28.5	0.00120	0.0342				
25 × 40	36.3	0.00106	0.0385				
25 × 50	44.2	0.00097	0.0429				

(Courtesy of United Chemi-Con)

Figure A5. Heat Transfer Constants for Various Case Sizes

might be expected to operate no more than four hours a day on an average, so a ten year life is only 15,000 actual operating hours.

A capacitor's ripple current multiplier can be used to increase the maximum allowable ripple current, allowing smaller size capacitors to be used. However, this method of increasing the maximum ripple current rating assumes load life is kept constant, so it must be used with extreme caution. For example, a Nichicon PL series capacitor rated at 1A RMS at 105°C has a load life of 3,000 hours and a ripple current multiplier of 2.2 at 65°C. If the ripple current

multiplier is used, the capacitor's ripple current rating can be increased to 2.2A RMS as long as the operating temperature does not exceed 65°C; however, the load life of the capacitor is still 3,000 hours. Eighteen weeks is a rather short operating life.

References

1. United Chemi-Con Inc., "Understanding Aluminum Electrolytic Capacitors."

2. Nichicon (America) Corp., "Technical Notes on Aluminum Electrolytic Capacitors."

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Capacitor Manufacturers

1. Nichicon (America) Corporation
   927 East State Parkway
   Schaumburg, IL 60195
   (708) 843-7500

2) Sanyo Video Components (USA) Corporation
   1201 Sanyo Avenue
   San Diego, CA 92073
   (619) 661-6322

3) United Chemi-Con, Inc.
   9801 West Higgins Road
   Rosemount, IL 60018
   (708) 696-2000

4) Sprague Electric Company
   Aluminum Electrolytic Div.
   9800 Kincey Ave. Suite 100
   Huntersville, NC 28078
   (704) 875-8070

5) Kemet Electronics
   P.O. Box 5928
   Greenville, SC 29606
   (803) 675-1760

6) Marcon
   998 Forest Edge Drive
   Vernon Hills, IL 60061
   (708) 913-9980

7) Wima
   2269 Saw Mill River Rd
   Bldg. 4C
   P.O. Box 217
   Elmsford, NY 10523
   (914) 347-2474

APPENDIX B

Measuring RMS Current in Switching Regulator Filter Capacitors

One of the most critical parameters on a capacitor's data sheet is its ripple current rating, specified in RMS current. The operating ripple current must be accurately determined in order to select the proper size capacitor. The current waveforms are usually square or triangular and their RMS value can be determined by either measurement or an analytical approach.

The preferred method of determining ripple current is to measure it. This can easily be accomplished using a true-RMS voltmeter (HP3400A or Fluke 8920A or equivalent) and a current probe (Tektronix P6021).

Thermally based RMS voltmeters provide the high bandwidth and crest factor capability necessary to accurately measure the RMS current through a filter capacitor. The RMS voltmeter's bandwidth must exceed 1MHz, since current transients can exceed 100A/μs. Do not use average responding or logarithmic based RMS voltmeters.

Most hand held voltmeters are average responding and only good for low frequency sinewaves, typically under 10kHz. The logarithmic approach measures true RMS, but bandwidths are limited to well below 1MHz.

There are two types of current probes available: the traditional AC only probe and the true DC Hall Effect type. AC only current probes (P6021) use a transformer to convert current flux into AC signals and have a frequency response from a few hundred hertz to 60MHz. Therefore, do not use a P6021 if the ripple current waveform has a low frequency component. Hall Effect current probes (P6042) include semiconductors to provide a frequency response from DC to 50MHz. Both types have saturation limitations which, when exceeded, cause erroneous results.

The following procedure will accurately determine the maximum RMS current for a capacitor no matter how complex the ripple current waveforms are. The current probe is clipped on the capacitor's lead and the other end of the probe is connected to the RMS voltmeter. Set the P6021 terminator to its 10mA/mV scale. It is always a

good idea to view the current waveform on an oscilloscope to verify that the converter is working properly before measuring the RMS voltage. Next, apply maximum load current to the output of the regulator. The RMS current can be calculated by multiplying the current probe scale factor by the RMS voltmeter reading:
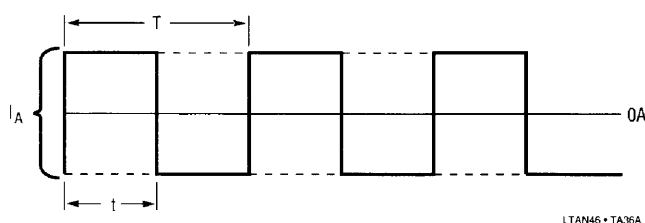
I_{RMS} = Scale Factor * V_{RMS} Reading
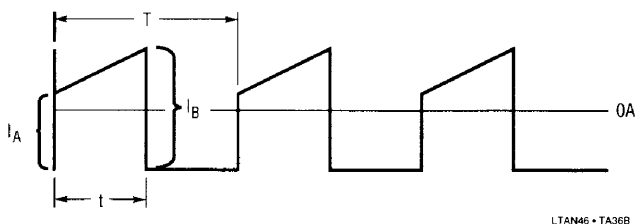10mA/mV * 100mV = 1A RMS

Vary the switching regulator's input voltage over its entire operating range. The maximum RMS voltage reading will be the worst case operating condition for the capacitor. Select the capacitor based on this RMS current reading.

If a true RMS voltmeter is not available, the RMS current can be estimated by analyzing the capacitor ripple current waveform. Capacitor RMS current waveforms vary for different converter topologies, and between input and output capacitors.
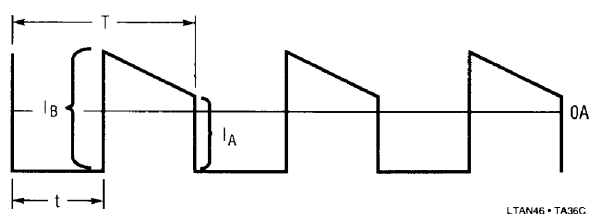
Figure B1 shows some common filter capacitor waveforms and equations used to derive the values of I_{RMS}. Current waveforms generally fall into one of four cases. Case 1 is not an actual ripple current waveform, but it is often used to approximate the RMS current since only two variables are used. Here, worst case ripple occurs at 50% duty cycle. Case 2 is the ripple current waveform for an
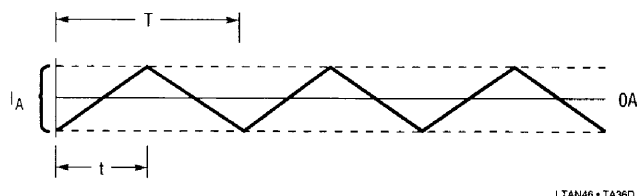
$$I_{RMS} = I_A \sqrt{DC(1-DC)}$$

$$DC = \frac{t}{T}$$

LTAN46 • TA36A

Case 1. Rectangular

$$I_{RMS} = \sqrt{DC\left[\frac{I_A{}^2 + I_A \bullet I_B + I_B{}^2}{3} - \frac{DC}{4}(I_A + I_B)^2\right]}$$

$$DC = \frac{t}{T}$$

LTAN46 • TA36B

Case 2. Trapezoid

$$I_{RMS} = \sqrt{(1-DC)\left[\frac{I_A{}^2 + I_A \bullet I_B + I_B{}^2}{3} - \frac{(1-DC)}{4}(I_A + I_B)^2\right]}$$

$$DC = \frac{t}{T}$$

LTAN46 • TA36C

Case 3. Trapezoid

$$I_{RMS} = \frac{I_A}{\sqrt{12}}$$

LTAN46 • TA36D

Case 4. Saw Tooth

Figure B1. Typical Filter Capacitor Ripple Current Waveforms

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input capacitor for buck, buck-boost, or flyback topology. Case 3 is for an output capacitor for boost, buck-boost, or flyback topology. Case 4's waveform is for the input capacitor for boost and output capacitor for buck mode.

Figure B2 summarizes these equations and shows equations that can be used to make a quick approximation of ripple current. Here the RMS current is calculated from output current and duty cycle. As long as the ratio of I_B over I_A is < 2, these equations approximate the RMS current to within 10% of the actual value.

There exists a little confusion about where zero current is located on the capacitor's current waveform. Figure B3 shows an input capacitor's ripple current for a buck, buck-boost and flyback topology. The zero point is always in the middle since the average current through the capacitor must be zero, assuming negligible leakage current. The two shaded areas are equal because the average current flowing into the capacitor equals the average outgoing current.
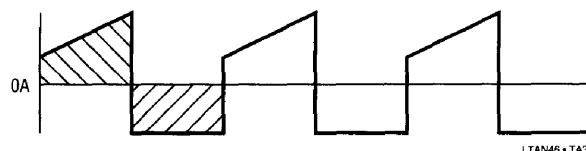
**Figure B3. The Average Current Through the Filter Capacitor is Zero**

In many switching regulator applications the input supply consists of a 60Hz single phase step-down transformer followed by a rectifier whose output is smoothed by a filter capacitor. Figure B4 shows the filter capacitor's ripple current waveform for a full wave bridge rectifier. The ripple current flowing through the input capacitor consists of the high frequency switching ripple current superimposed on the 120Hz ripple (trace A). Trace B details the content of the 100kHz ripple current which is produced by a LT1074 buck switching regulator.

To select the proper size input capacitor, the effects of the 120Hz and the 100kHz ripple current waveforms must be considered. The best way to determine the input capacitor's

TOPOLOGY	FLYBACK	BOOST	BUCK	BUCK-BOOST
C_{IN} $I_{RMS}*$	$\sqrt{DC\left[\dfrac{I_A^2+I_A\cdot I_B+I_B^2}{3}-\dfrac{(DC)}{4}(I_A+I_B)^2\right]}$	$\dfrac{I_A}{\sqrt{12}}$	$\sqrt{DC\left[\dfrac{I_A^2+I_A\cdot I_B+I_B^2}{3}-\dfrac{(DC)}{4}(I_A+I_B)^2\right]}$	
C_{OUT} $I_{RMS}*$	$\sqrt{(1-DC)\left[\dfrac{I_A^2+I_A\cdot I_B+I_B^2}{3}-\dfrac{(1-DC)}{4}(I_A+I_B)^2\right]}$		$\dfrac{I_A}{\sqrt{12}}$	$\sqrt{(1-DC)\left[\dfrac{I_A^2+I_A\cdot I_B+I_B^2}{3}-\dfrac{(1-DC)}{4}(I_A+I_B)^2\right]}$
C_{IN} $\approx I_{RMS}*$	$\dfrac{I_{OUT}\cdot V_{OUT}}{V_{IN}}\sqrt{\dfrac{1-DC}{DC}}$	$0.3\Delta I$	$I_{OUT}\sqrt{DC-(DC)^2}$	$\dfrac{I_{OUT}\cdot V_{OUT}}{V_{IN}}\sqrt{\dfrac{1-DC}{DC}}$
C_{OUT} $\approx I_{RMS}*$	$I_{OUT}\sqrt{\dfrac{DC}{1-DC}}$	$I_{OUT}\sqrt{\dfrac{DC}{1-DC}}$	$0.3\Delta I$	$I_{OUT}\sqrt{\dfrac{DC}{1-DC}}$
ΔI		$\dfrac{V_{IN}\cdot DC}{L\cdot f}$	$\dfrac{(V_{IN}-V_{OUT})\cdot DC}{L\cdot f}$	
DC	$\dfrac{V_{OUT}}{V_{OUT}+NV_{IN}}$	$\dfrac{V_{OUT}-V_{IN}}{V_{OUT}}$	$\dfrac{V_{OUT}}{V_{IN}}$	$\dfrac{V_{OUT}}{V_{OUT}+V_{IN}}$

*Refer to Figure B1 for details on I_A and I_B

Figure B2. Filter Capacitor RMS Current Equations

A HORIZONTAL = 2ms/DIV
B HORIZONTAL = 5µs/DIV

LTAN46 • TA38
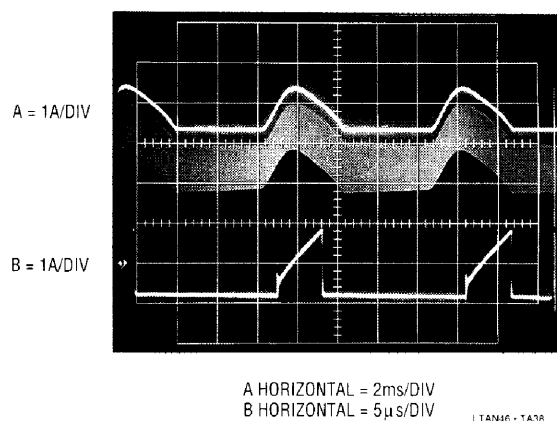
Figure B4. Input Capacitor's Ripple Current Waveform When Input Supply is a 60Hz Full Wave Bridge Circuit

ripple current is to measure it. A DC current probe must be used to measure the 120Hz content of the waveform. An AC current probe can measure the 100kHz ripple but can not accurately measure the 120Hz ripple since its lower bandwidth limit is a few hundred hertz. The HP3400A RMS voltmeter can be used here since its frequency range extends from 10Hz to10MHz.

APPENDIX C

Bipolar Power Switch Conduction Losses

Power transistor conduction losses limit power conversion efficiency. It can be a substantial limitation when input voltage is low. Conduction losses include both switch driver and switch on losses. Switch driver losses occur because bipolar devices require base current to turn on and switch on losses are a product of the power transistor saturation characteristic.

Figure C1 shows the type and location of the power switch used in the LT1070 step-up (Case 1) and LT1074 step-down (Case 2) converters. These two switching configurations have different power transistor architectures and saturation characteristics. The boost circuit, implemented with an LT1070, uses a ground referred NPN transistor as the switch device, whereas the buck circuit, implemented with an LT1074, uses a supply referred composite PNP high side switch.

Figure C2 shows the saturation characteristics for each power transistor. The switch voltage drop for the composite PNP in the LT1074 (Case 2) is noticeably higher than the NPN used in the LT1070 (Case 1) because of the way the switch is configured.

NPN Switch

Figure C2, Case 1 shows Current vs Voltage Characteristics of the LT1070 NPN power switch. In saturation, the NPN switch can be modeled as resistance. The slope of the curve indicates switch on resistance, which is found by dividing the collector to emitter voltage by the collector current. The on resistance of the NPN determines both its voltage drop and power dissipation. The NPN switch conduction loss can be calculated from:

$$P_C \approx (I_{RMS})^2 \bullet R_{ON} \qquad (C1)$$

Another dissipation factor associated with the NPN transistor is the base drive loss. Driving the base requires current, resulting in power loss in a driver transistor.

To optimize efficiency, the LT1070 uses a constant beta drive circuit to control base current. This scheme provides a base drive that is proportional to the collector current. The LT1070 operates with a constant beta drive of 40, which means with a collector current of 5A, the base drive current would be 125mA, but at I_C=1A, I_B is only 25mA. This technique is very efficient over a wide range of collector currents.
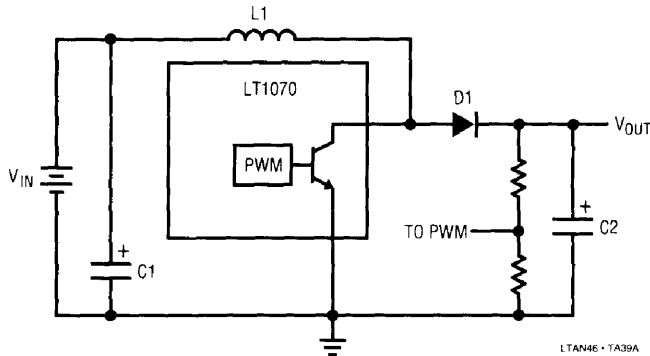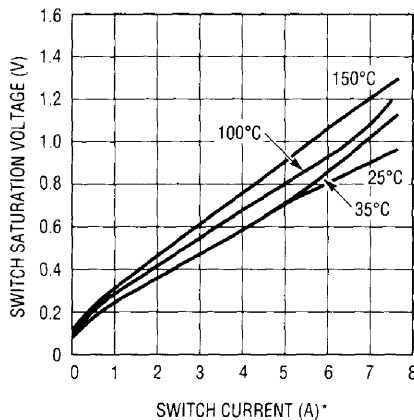
Case 1. LT1070 Step-Up (Boost) Converter
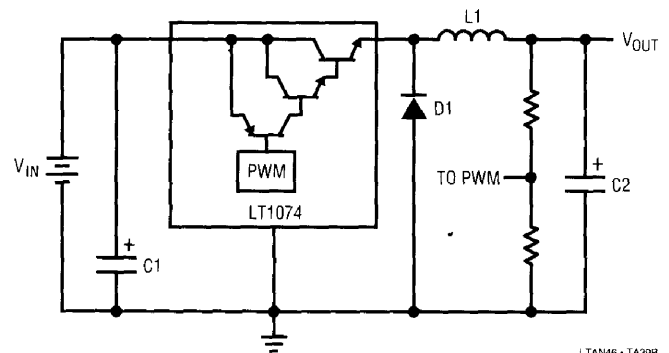
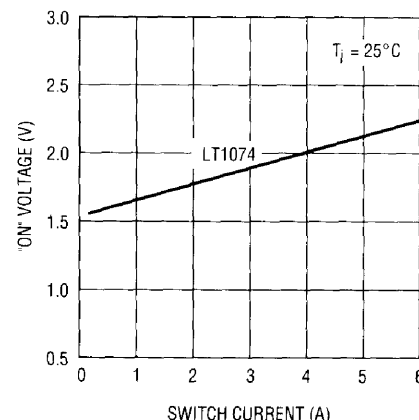Case 2. LT1074 Step-Down (Buck) Converter

Figure C1. Types and Location of Bipolar Power Switches

*DIVIDE CURRENT BY 2 FOR LT1071

Case 1. LT1070 Switch Current vs Voltage Characteristics

Case 2. LT1074 Switch Current vs Voltage Characteristics

Figure C2. Switch Current vs Voltage Characteristics of Bipolar Power Switches

The base drive current is drawn from the input pin of the LT1070 when the power transistor is on; therefore, driver losses are duty cycle dependent. The LT1070 dissipation because of base drive losses becomes:

$$P_{DRV} = V_{IN} \cdot I_{SW}/40 \cdot DC \qquad (C2)$$

The total LT1070 power dissipation is the sum of the switch conduction and driver losses and is given by:

$$P_{TOT} = (I_{RMS})^2 \cdot R_{ON} + V_{IN} \cdot I_{AVG}/40 \qquad (C3)$$

At low switch currents and high input voltages, driver losses dominate; switch losses dominate at low input voltages and high switch currents.

Composite PNP Switch Conduction Losses

Figure C2, Case 2 shows the Current vs Voltage Characteristics of the LT1074 composite PNP power switch. Its power switch can be modeled by a resistance and a series offset voltage, typically 1.8V. The fixed 1.8V drop of the composite switch is made up of $2V_{BE}$ ($\approx 0.75V$ ea.) drops across the Darlington-connected NPN and a PNP saturation drop ($\approx 0.3V$). The composite PNP power dissipation can be predicted by the following formula:

$$P_{D\ TOT} = 1.8V \cdot I_{AVG} + 0.1\Omega \cdot (I_{RMS})^2 \qquad (C4)$$

In this case, both the average and RMS current are needed to calculate power loss. The average current can be used here since the fixed voltage drop is independent of switch current.

At low input voltages the switch's fixed voltage loss degrades efficiency because it makes up a higher percentage of the available input supply. Higher input voltages make the fixed loss a smaller percentage, improving efficiency.

The composite PNP high gain configuration needs only 5mA of driver current to fully saturate the switch. This small current introduces negligible loss.

Determining RMS and Average Switch Currents

In order to calculate efficiency, the switch's average and RMS currents must be determined. Switch current waveforms generally look like those of Figure C3. The associated RMS and average current equations are also given. Figure C3, Case 1 can be used to make a quick approximation of switch RMS and average current. Figure C3, Case 2 is the switch current waveform for continuous mode and Figure C3, Case 3 for discontinuous mode.

Using the thermal RMS voltmeter to measure switch RMS current as discussed in appendix B will work here, but **the RMS meter must be able to measure DC. The DC current probe must be properly zeroed!** For example, the Fluke 8920A RMS voltmeter will work and the HP 3400A will not because it has an AC-coupled input.
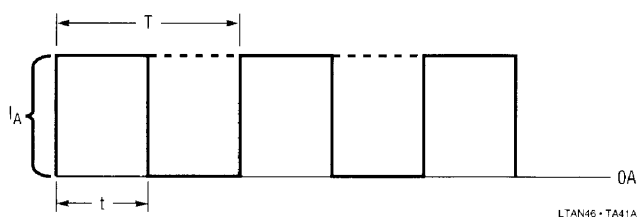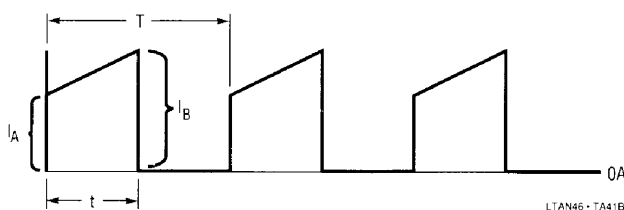

$$I_{RMS} = I_A \sqrt{DC}$$
$$I_{AVG} = DC \cdot I_A$$
$$DC = \frac{t}{T}$$

Case 1. Rectangular Approximation
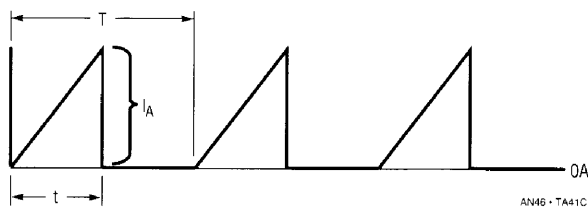

$$I_{RMS} = \sqrt{DC \left[\frac{I_A^2 + I_A \cdot I_B + I_B^2}{3} \right]}$$
$$I_{AVG} = \frac{DC(I_A + I_B)}{2}$$
$$DC = \frac{t}{T}$$

Case 2. Trapezoid Continuous Mode

$$I_{RMS} = I_A \sqrt{\frac{DC}{3}}$$
$$I_{AVG} = \frac{DC \cdot I_A}{2}$$
$$DC = \frac{t}{T}$$

Case 3. Trapezoid Discontinuous Mode

Figure C3. Typical Switch Current Waveforms

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APPENDIX D

Diode Conduction Losses

The output diode conduction loss is often the major source of power loss in switching regulators. It exhibits a forward voltage drop (V_f) which limits efficiency. Efficiency loss due to V_f is most significant at low output voltages, and should be as low as possible to optimize efficiency. At high output voltages, the forward voltage drop's effect on efficiency is small, because it makes up a very small percentage of output voltage.

For low output voltages, Schottky diodes are recommended over silicon diodes because they have a lower forward voltage drop for the same current rating. In a flyback topology with 5V output, a Schottky diode with a V_f of 0.6V introduces a loss of 12% of the output power; whereas a silicon diode with a V_f of 1.0V contributes a 20% loss. However, Schottkys are limited to low voltage applications since they have low Peak Inverse Voltage (PIV) limits.

Diode conduction loss can be approximated by the following formula:

$$P_D \approx I_{D\ AVG} \cdot V_f \qquad (D1)$$

$I_{D\ AVG}$: Average Diode Current
$\quad V_f$: Diode forward voltage drop at average peak diode current ($I_{f\ AVG}$)

Power loss due to diode leakage current is assumed to be negligible.

The expression for P_D looks simple, but is deceiving because V_f is a function of the diode's instantaneous forward current (I_f), not average diode current ($I_{D\ AVG}$). The dependence of V_f on I_f is shown in Figure D1.

The value to use for I_f can be determined by examining the diode's current waveform. Figure D2 shows a typical diode current waveform. During the diode on-time, the diode current (I_f) is not constant; however, it can be approximated by taking the average peak diode current ($I_{f\ AVG}$)
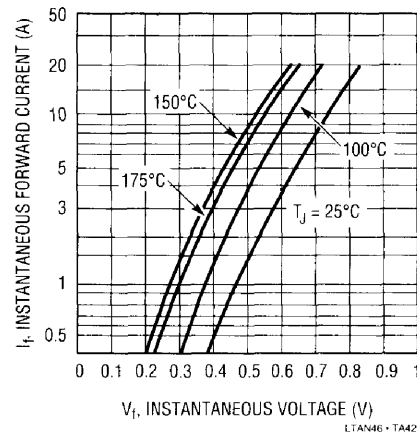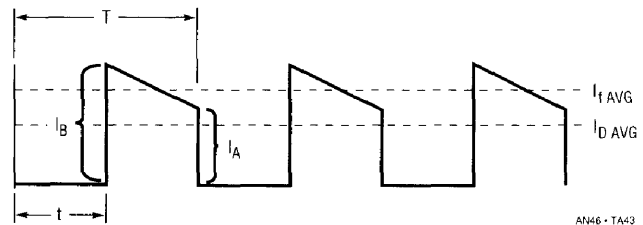

Figure D1. Typical Forward Voltage

$$I_{f\ AVG} = \frac{I_A + I_B}{2} \qquad (D2)$$

$$I_{D\ AVG} = (1 - DC)\frac{I_A + I_B}{2} \qquad (D3)$$

$$DC = \frac{t}{T} \qquad (D4)$$

Figure D2. Typical Diode Current Waveform

during this period and is given by Equation D2. This approximation assumes that the diode forward voltage drop is relatively linear for the different peak current values and is reasonably accurate if the ratio of I_B/I_A is less than three.

The average diode current ($I_{D\ AVG}$) can be determined by using Equation D3. In the boost, flyback and buck-boost topology the average diode current is equal to the output current (I_{OUT}), since the diode has to conduct the full output current. With the buck topology, the average diode current is only a fraction of the output current; therefore, it has lower conduction losses then the other converter topologies for the same output current level.

The most stressful condition for the output diode is overload or short circuit conditions. The internal current limit of the LT1070 is typically 8A at low switch duty cycles. This is almost a factor of 1.6 higher than the 5A rated switch current. If full load output current requires only a fraction of the 5A rated switch current, the ratio of diode short circuit current to full load current may be much higher than 2:1. **A regulator designed to withstand continuous short conditions must either use diodes rated for full short circuit current or it must incorporate some form of external current limiting.**

The boost topology does not provide short circuit protection, therefore the output diode and inductor can fail if the output is shorted to ground. If the converter must survive a short without blowing a fuse, other circuit techniques must be used.

LT1432 High Efficiency Step-down Switching Regulator Controller

PRELIMINARY

October 1991

FEATURES

- Accurate preset +5 Volt output
- Up to 90% efficiency
- Optional burst mode for light loads
- Can be used with many LTC switching ICs
- Accurate ultra-low-loss current limit
- Operates with inputs from 6V to 30V
- Shutdown mode draws only 15µA
- Uses small 50µH inductor

APPLICATIONS

- Lap-top and palm-top computers
- Portable data gathering instruments
- DC bus distribution systems
- Battery powered digital widgets

DESCRIPTION

The LT1432 is a control chip designed to operate with the LT1070 family of switching regulators to make a very high efficiency 5V step-down (buck) switching regulator. A minimum of external components is needed.

Included is an accurate current limit which uses only 60mV sense voltage and uses "free" pc board trace material for the sense resistor. Logic controlled electronic shutdown mode draws only 15µA battery current. The switching regulator operates down to 6 volts input.

The LT1432 has a logic controlled "burst" mode to achieve high efficiency at very light load currents (0 to 100mA) such as memory keep alive. In normal switching mode, the standby power loss is about 60mW, limiting efficiency at light loads. In burst mode, standby loss is reduced to approximately 15mW. Output current in this mode is typically in the 5-100mA range.

The LT1432 is available in 8-pin surface mount and DIP packages. The LT1070 family will also be available in a surface mount version of the 5-pin TO-220 package.

TYPICAL APPLICATION   High Efficiency 5V Buck Convertor
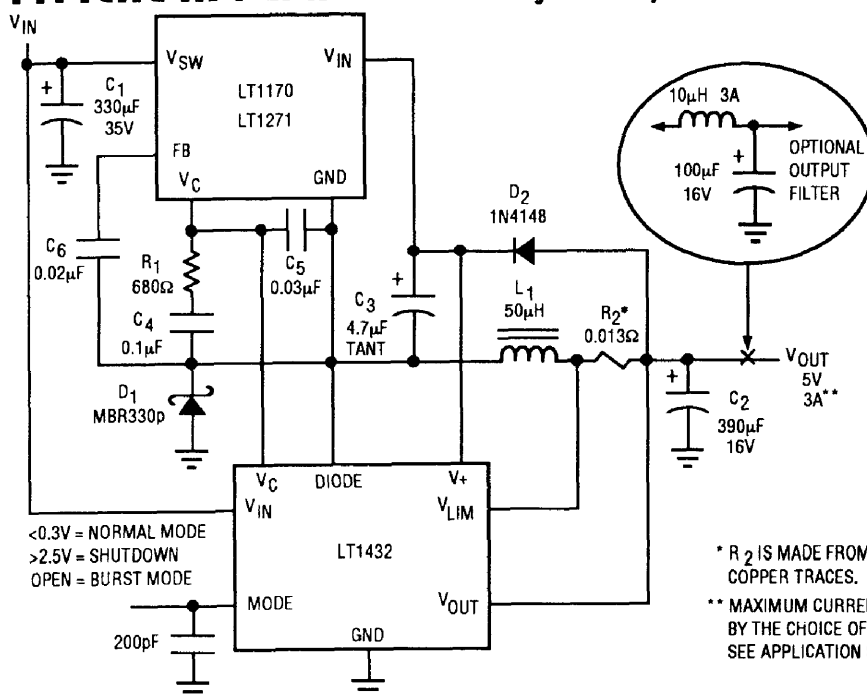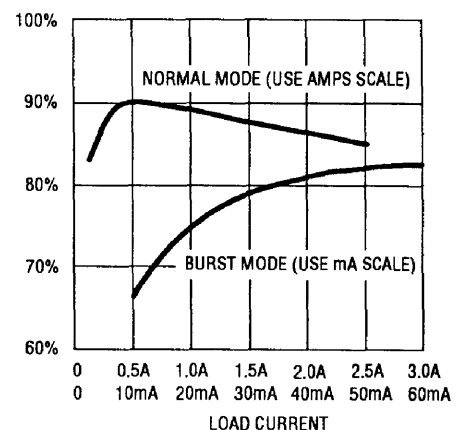
* R$_2$ IS MADE FROM PC BOARD COPPER TRACES.

** MAXIMUM CURRENT IS DETERMINED BY THE CHOICE OF LT1070 FAMILY. SEE APPLICATION SECTION.

<0.3V = NORMAL MODE
>2.5V = SHUTDOWN
OPEN = BURST MODE

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ABSOLUTE MAXIMUM RATINGS

V_{IN} Pin 30V
$V+$ Pin 40V
V_C 35V
V_{LIM} and V_{OUT} Pins 7V
V_{LIM} and V_{OUT} Pin Differential Voltage 0.5V
Diode Pin Voltage 30V
Mode Pin Current (Note 2) 1mA
Operating Ambient Temperature Range .... 0°C to 70°C
Storage Temperature Range .... -65°C to 150°C
Lead Temperature (Soldering, 10 sec.) .... 300°C

PACKAGE/ORDER INFORMATION

TOP VIEW

```
         ┌────────┐
V_LIM  1 │        │ 8  MODE
V_OUT  2 │        │ 7  GND
V_IN   3 │        │ 6  V_C
V+     4 │        │ 5  DIODE
         └────────┘
```

N8 PACKAGE
8-LEAD PLASTIC DIP

S8 PACKAGE
8-LEAD PLASTIC SOIC

ORDER PART NUMBER
LT1432CN8 LT1432CS8

ELECTRICAL CHARACTERISTICS

$V_C = 6V$, $V_{IN} = 6V$, $V+ = 10V$, $V_{DIODE} = 5V$, $I_C = 220\mu A$, $V_{LIM} = V_{OUT}$, $V_{MODE} = 0V$, $T_J = 25°C$
Device is in standard test loop unless otherwise noted.

PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
Regulated output voltage		●	4.9	5.0	5.1	V
Output voltage line regulation	$V_{IN} = 6V$ to 30V	●		5	20	mV
Input supply current (Note 1)	$V_{IN} = 6V$ to 30V, $V+ = Vin + 4V$	●		0.3	0.5	mA
Quiescent output load current		●		0.9	1.2	mA
Mode pin current	$V_{MODE} = 0V$, (Current is out of pin)	●		30	50	μA
	$V_{MODE} = 5V$ (shutdown)	●		15	30	μA
Mode pin threshold voltage (normal to burst)	$I_{MODE} = 1\mu A$ out of pin	●	0.6	0.9	1.5	V
V_C pin saturation voltage	$V_{OUT} = 5.5V$ (forced)	●		0.25	0.45	V
V_C pin maximum sink current	$V_{OUT} = 5.5V$ (forced)	●	0.45	0.8	1.5	mA
V_C pin source current	$V_{OUT} = 4.5V$ (forced)	●	40	60	100	μA
Current limit sense voltage	Device in current limit loop		56	60	64	mV
V_{LIM} pin current	Device in current limit loop (Current is out of pin)	●	30	45	70	μA
Supply current in shutdown	Vmode > 3V, Vin < 30V	●		15	40	μA
Burst mode output ripple	Device in burst test circuit			100		mV$_{P-P}$
Burst mode average output voltage	Device in burst test circuit	●	4.8	5	5.2	V
Clamp diode forward voltage	$I_F = 1mA$	●		0.5	0.65	V
Startup drive current	$V_{OUT} = 4.5V$ (forced), $V+ = 5V$ to 25V	●	30	45		mA

The ● denotes specifications which apply over the operating temperature range.

Note 1: Does not include current drawn by the LT1070 IC. See operating parameters in standard circuit.

Note 2: Breakdown voltage on the mode pin is 7V. External current must be limited to value shown.

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