

# High Voltage, Current Mode Switching Regulator Controller with Programmable Operating Frequency

## FEATURES

- **High Voltage Operation: Up to 60V**
- **Output Voltages up to 36V (Step-Down)**
- **Programmable Constant Frequency: 100kHz to 500kHz**
- **Synchronizable up to 600kHz**
- **Burst Mode® Operation: 120μA Supply Current**
- **10μA Shutdown Supply Current**
- **±1.3% Reference Accuracy**
- **Drives N-Channel MOSFET**
- **Programmable Soft-Start**
- **Programmable Undervoltage Lockout**
- **Internal High Voltage Regulator for Gate Drive**
- **Thermal Shutdown**
- **Current Limit Unaffected by Duty Cycle**
- **16-Pin Thermally Enhanced TSSOP Package**

## APPLICATIONS

- Industrial Power Distribution
- 12V and 42V Automotive and Heavy Equipment
- High Voltage Single Board Systems
- Distributed Power Systems
- Avionics
- Telecom Power

## DESCRIPTION

The LT<sup>®</sup>3844 is a DC/DC controller used for medium power, low part count, high efficiency supplies. It offers a wide 4V-60V input range (7.5V minimum startup voltage) and can implement step-down, step-up, inverting and SEPIC topologies.

The LT3844 includes Burst Mode operation, which reduces quiescent current below 120μA and maintains high efficiency at light loads. An internal high voltage bias regulator allows for simple biasing.

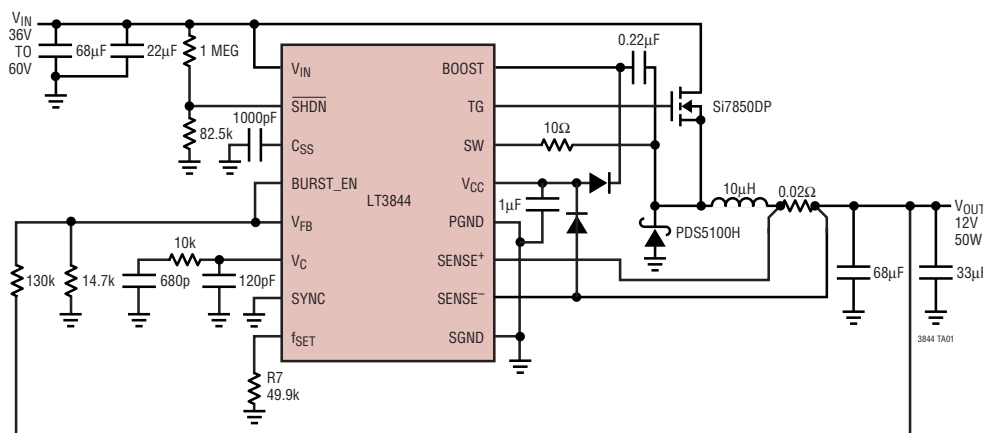
Additional features include current mode control for fast line and load transient response; programmable fixed operating frequency that can be synchronized to an external clock for noise sensitive applications; a gate driver capable of driving large N-channel MOSFETs; a precision undervoltage lockout function; 10μA shutdown current; short-circuit protection and a programmable soft-start function.

The LT3844 is available in a 16-lead thermally enhanced TSSOP package.

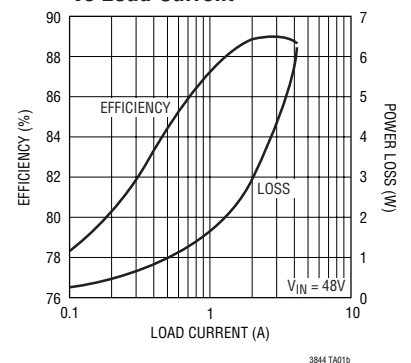
LT, LT, LTC and LTM are registered trademarks of Linear Technology Corporation. Burst Mode is a registered trademark of Linear Technology Corporation. All other trademarks are the property of their respective owners. Protected by U.S. Patents including 5731694, 6498466, 6611131.

## TYPICAL APPLICATION

**High Voltage Step-Down Regulator 48V to 12V, 50W**



**Efficiency and Power Loss  
vs Load Current**

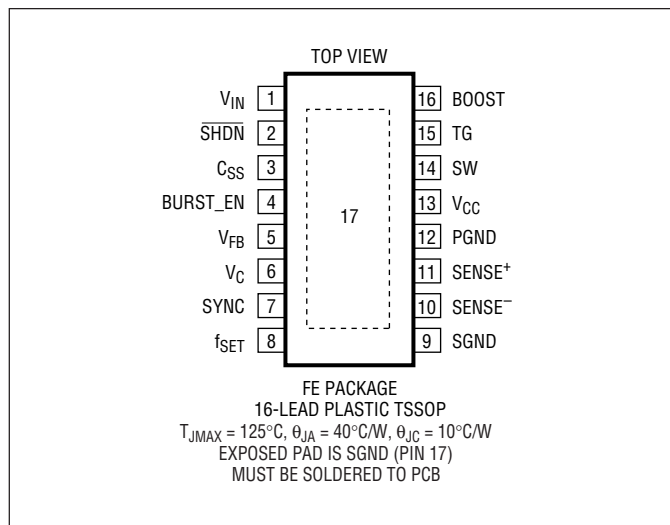


## ABSOLUTE MAXIMUM RATINGS

(Note 1)

Input Supply Voltage ( $V_{IN}$ )	65V to -0.3V
Boosted Supply Voltage (BOOST)	80V to -0.3V
Switch Voltage (SW) (Note 8)	65V to -1V
Differential Boost Voltage (BOOST to SW)	24V to -0.3V
Bias Supply Voltage ( $V_{CC}$ )	24V to -0.3V
SENSE <sup>+</sup> and SENSE <sup>-</sup> Voltages	40V to -0.3V
Differential Sense Voltage (SENSE <sup>+</sup> to SENSE <sup>-</sup> )	1V to -1V
BURST_EN Voltage	24V to -0.3V
SYNC, $V_C$ , $V_{FB}$ , $C_{SS}$ , and SHDN Voltages	5V to -0.3V
SHDN Pin Currents	1mA
Operating Junction Temperature Range (Note 2)	
LT3844E (Note 3)	-40°C to 125°C
LT3844I	-40°C to 125°C
Storage Temperature	-65°C to 150°C
Lead Temperature (Soldering, 10 sec)	300°C

## PIN CONFIGURATION



## ORDER INFORMATION

LEAD FREE FINISH	TAPE AND REEL	PART MARKING	PACKAGE DESCRIPTION	TEMPERATURE RANGE
LT3844EFE#PBF	LT3844EFE#TRPBF	3844EFE	16-Lead Plastic TSSOP	-40°C to 125°C
LT3844IFE#PBF	LT3844IFE#TRPBF	3844IFE	16-Lead Plastic TSSOP	-40°C to 125°C

Consult LTC Marketing for parts specified with wider operating temperature ranges.

Consult LTC Marketing for information on nonstandard lead based finish parts.

For more information on lead free part marking, go to: <http://www.linear.com/leadfree/>

For more information on tape and reel specifications, go to: <http://www.linear.com/tapeandreel/>

## ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the full operating temperature range, otherwise specifications are at  $T_A = 25^{\circ}\text{C}$ .  $V_{IN} = 20\text{V}$ ,  $V_{CC} = \text{BOOST} = \text{BURST\_EN} = 10\text{V}$ ,  $\text{SHDN} = 2\text{V}$ ,  $R_{SET} = 49.9\text{k}\Omega$ ,  $\text{SENSE}^- = \text{SENSE}^+ = 10\text{V}$ ,  $\text{SGND} = \text{PGND} = \text{SW} = \text{SYNC} = 0\text{V}$ , unless otherwise noted.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
$V_{IN}$ Operating Voltage Range (Note 4)	●	4		60	V
$V_{IN}$ Minimum Start Voltage	●			7.5	V
$V_{IN}$ UVLO Threshold (Falling)	●	3.6	3.8	4	V
$V_{IN}$ UVLO Threshold Hysteresis			670		mV
$V_{IN}$ Supply Current	$V_{CC} > 9\text{V}$		20		$\mu\text{A}$
$V_{IN}$ Burst Mode Current	$V_{BURST\_EN} = 0\text{V}$ , $V_{FB} = 1.35\text{V}$		20		$\mu\text{A}$
$V_{IN}$ Shutdown Current	$V_{SHDN} = 0\text{V}$	●	10	15	$\mu\text{A}$
BOOST Operating Voltage Range	●			75	V
BOOST Operating Voltage Range (Note 5)	●			20	V
BOOST UVLO Threshold (Rising)	$V_{BOOST} - V_{SW}$		5		V
BOOST UVLO Threshold Hysteresis	$V_{BOOST} - V_{SW}$		400		mV
BOOST Supply Current (Note 6)			1.4		mA
BOOST Burst Mode Current	$V_{BURST\_EN} = 0\text{V}$		0.1		$\mu\text{A}$
BOOST Shutdown Current	$V_{SHDN} = 0\text{V}$		0.1		$\mu\text{A}$

3844fa

**ELECTRICAL CHARACTERISTICS**

The ● denotes the specifications which apply over the full operating temperature range, otherwise specifications are at  $T_A = 25^\circ\text{C}$ .  $V_{IN} = 20\text{V}$ ,  $V_{CC} = \text{BOOST} = \text{BURST\_EN} = 10\text{V}$ ,  $\text{SHDN} = 2\text{V}$ ,  $R_{SET} = 49.9\text{k}\Omega$ ,  $\text{SENSE}^- = \text{SENSE}^+ = 10\text{V}$ ,  $\text{SGND} = \text{PGND} = \text{SW} = \text{SYNC} = 0\text{V}$ , unless otherwise noted.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
$V_{CC}$ Operating Voltage Range (Note 5)	Over Full Line and Load Range	●		20	V
$V_{CC}$ Output Voltage		●	8	8.3	V
$V_{CC}$ UVLO Threshold (Rising)			6.25		V
$V_{CC}$ UVLO Threshold Hysteresis			500		mV
$V_{CC}$ Supply Current (Note 6)	$V_{\text{BURST\_EN}} = 0\text{V}$ $V_{\text{SHDN}} = 0\text{V}$	●	1.7	2.1	mA
$V_{CC}$ Burst Mode Current			95		$\mu\text{A}$
$V_{CC}$ Shutdown Current			20		$\mu\text{A}$
$V_{CC}$ Current Limit		●	–40	–120	mA
Error Amp Reference Voltage	Measured at $V_{FB}$ Pin		1.224	1.231	V
		●	1.215	1.245	V
$V_{FB}$ Pin Input Current	$V_{FB} = 1.231\text{V}$		25		nA
$\text{SHDN}$ Enable Threshold (Rising)		●	1.3	1.35	V
$\text{SHDN}$ Threshold Hysteresis			120	1.4	mV
Sense Pins Common Mode Range	$V_{\text{SENSE}^+} - V_{\text{SENSE}^-}$	●	0	36	V
Current Limit Sense Voltage		●	90	115	mV
Input Current ( $I_{\text{SENSE}^+} + I_{\text{SENSE}^-}$ )	$V_{\text{SENSE}(\text{CM})} = 0\text{V}$ $V_{\text{SENSE}(\text{CM})} = 2\text{V}$ $V_{\text{SENSE}(\text{CM})} > 4\text{V}$		350 –25 –170		$\mu\text{A}$ $\mu\text{A}$ $\mu\text{A}$
Operating Frequency		●	290	310	kHz
			270	330	kHz
Minimum Programmable Frequency		●		100	kHz
Maximum Programmable Frequency		●	500		kHz
External Sync Frequency Range		●	100	600	kHz
SYNC Input Resistance			40		k $\Omega$
SYNC Voltage Threshold		●	1.4	2	V
Soft-Start Capacitor Control Current			2		$\mu\text{A}$
Error Amp Transconductance		●	270	340	$\mu\text{S}$
Error Amp DC Voltage Gain			62		dB
Error Amp Sink/Source Current			$\pm 30$		$\mu\text{A}$
TG Drive On Voltage (Note 7)	$C_{\text{LOAD}} = 2200\text{pF}$ $C_{\text{LOAD}} = 2200\text{pF}$		9.8		V
TG Drive Off Voltage			0.1		V
TG Drive Rise/Fall Time	10% to 90% or 90% to 10%, $C_{\text{LOAD}} = 2200\text{pF}$		40		ns
Minimum TG Off Time		●	350	500	ns
Minimum TG On Time		●	250	350	ns

**Note 1:** Stresses beyond those listed under Absolute Maximum Ratings may cause permanent damage to the device. Exposure to any Absolute Maximum Rating condition for extended periods may affect device reliability and lifetime.

**Note 2:** The LT3844 includes overtemperature protection that is intended to protect the device during momentary overload conditions. Junction temperature will exceed  $125^\circ\text{C}$  when overtemperature protection is active. Continuous operation above the specified maximum operating junction temperature may impair device reliability.

**Note 3:** The LT3844E is guaranteed to meet performance specifications from  $0^\circ\text{C}$  to  $125^\circ\text{C}$  junction temperature. Specifications over the  $-40^\circ\text{C}$  to  $125^\circ\text{C}$  operating junction temperature range are assured by design, characterization and correlation with statistical process controls. The LT3844I is guaranteed

over the full  $-40^\circ\text{C}$  to  $125^\circ\text{C}$  operating junction temperature range.

**Note 4:**  $V_{IN}$  voltages below the start-up threshold (7.5V) are only supported when the  $V_{CC}$  is externally driven above 6.5V.

**Note 5:** Operating range is dictated by MOSFET absolute maximum  $V_{GS}$ .

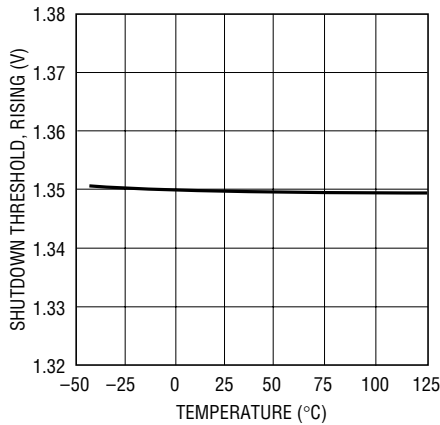
**Note 6:** Supply current specification does not include switch drive currents. Actual supply currents will be higher.

**Note 7:** DC measurement of gate drive output “ON” voltage is typically 8.6V. Internal dynamic bootstrap operation yields typical gate “ON” voltages of 9.8V during standard switching operation. Standard operation gate “ON” voltage is not tested but guaranteed by design.

**Note 8:** The  $-1\text{V}$  absolute maximum on the SW pin is a transient condition. It is guaranteed by design and not subject to test.

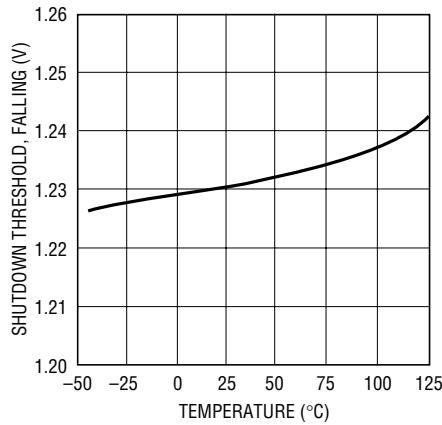
# TYPICAL PERFORMANCE CHARACTERISTICS

**Shutdown Threshold (Rising)  
vs Temperature**



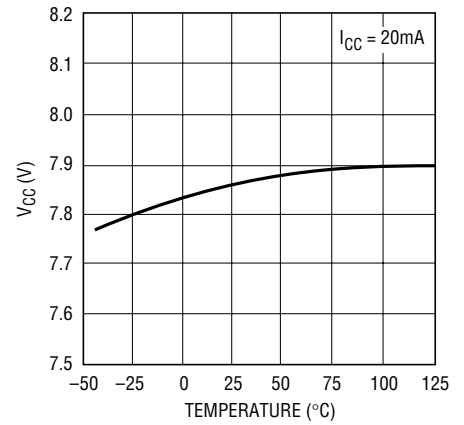
3844 G01

**Shutdown Threshold (Falling)  
vs Temperature**



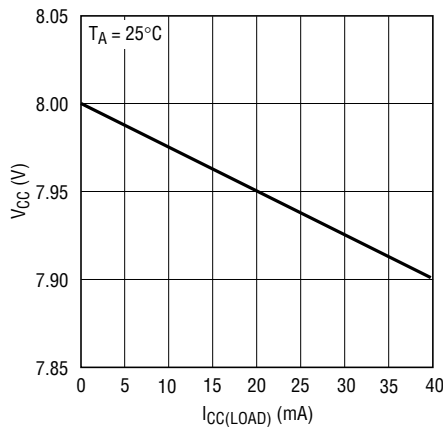
3844 G02

**V<sub>CC</sub> vs Temperature**



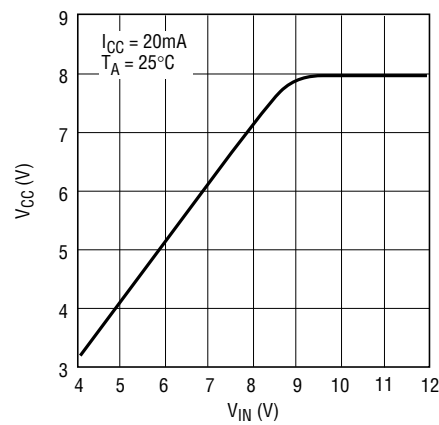
3844 G03

**V<sub>CC</sub> vs I<sub>CC</sub>(LOAD)**



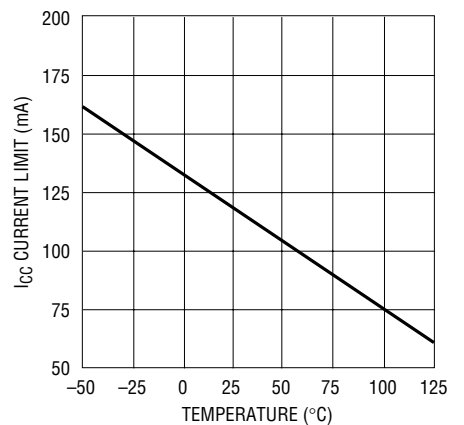
3844 G04

**V<sub>CC</sub> vs V<sub>IN</sub>**



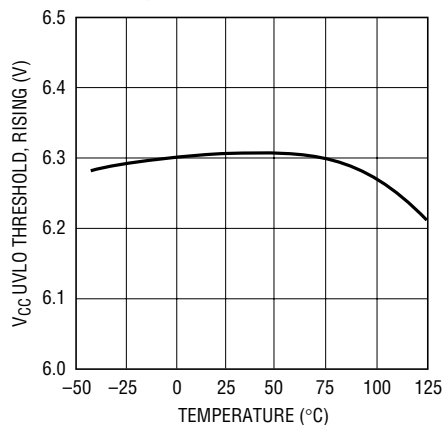
3844 G05

**I<sub>CC</sub> Current Limit vs Temperature**



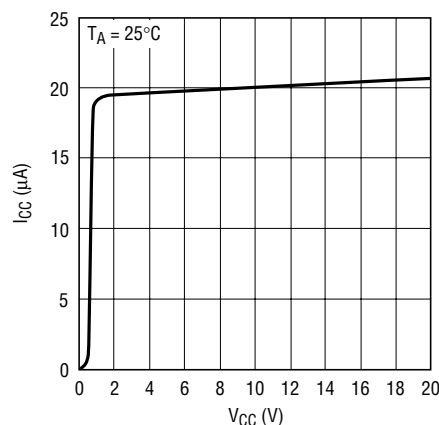
3844 G06

**V<sub>CC</sub> UVLO Threshold (Rising)  
vs Temperature**



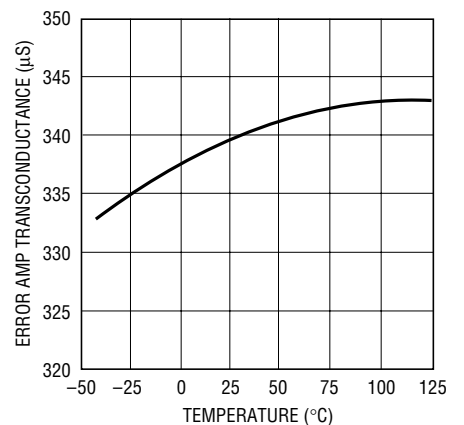
3844 G07

**I<sub>CC</sub> vs V<sub>CC</sub> (SHDN = 0V)**



3844 G08

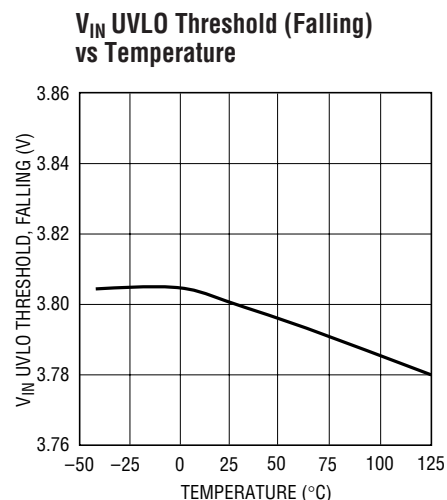
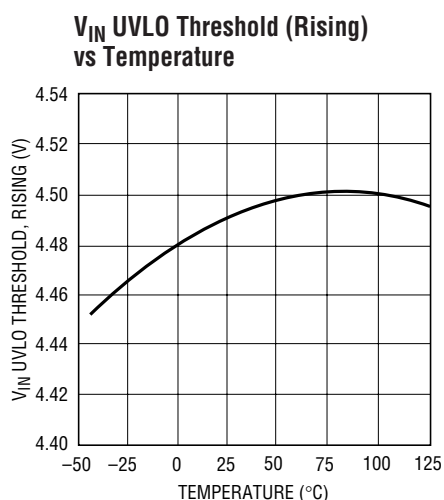
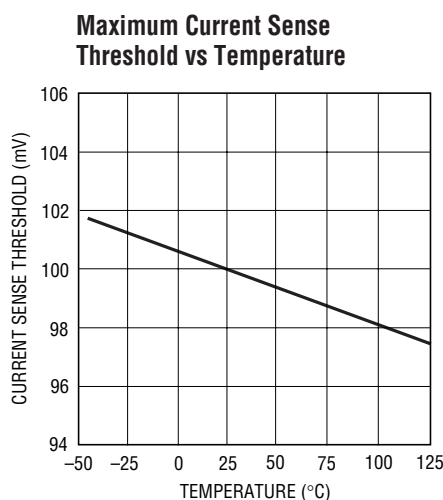
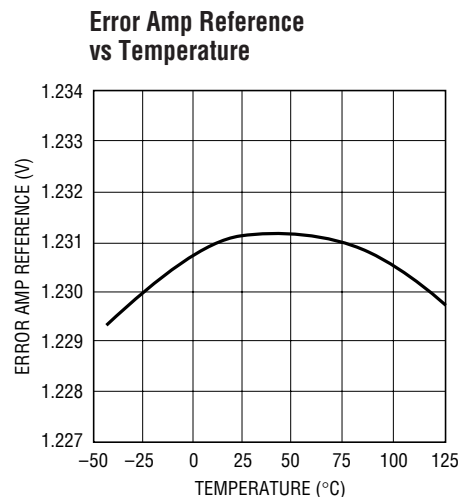
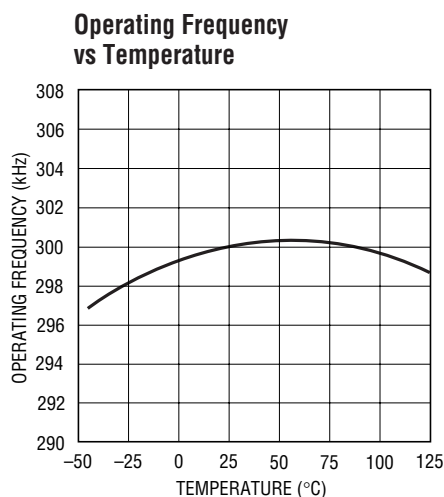
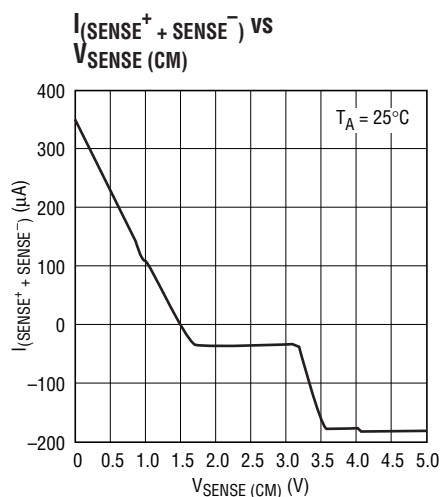
**Error Amp Transconductance  
vs Temperature**



3844 G09

3844fa

## TYPICAL PERFORMANCE CHARACTERISTICS



## PIN FUNCTIONS

**$V_{IN}$  (Pin 1):** The  $V_{IN}$  pin is the main supply pin and should be decoupled to SGND with a low ESR capacitor located close to the pin.

**$\overline{\text{SHDN}}$  (Pin 2):** The  $\overline{\text{SHDN}}$  pin has a precision IC enable threshold of 1.35V (rising) with 120mV of hysteresis. It is used to implement an undervoltage lockout (UVLO) circuit. See Application Information section for implementing a UVLO function. When the  $\overline{\text{SHDN}}$  pin is pulled below a transistor  $V_{BE}$  (0.7V), a low current shutdown mode is entered, all internal circuitry is disabled and the  $V_{IN}$  supply

current is reduced to approximately 9 $\mu\text{A}$ . Typical pin input bias current is <10 $\mu\text{A}$  and the pin is internally clamped to 6V.

**$C_{SS}$  (Pin 3):** The soft-start pin is used to program the supply soft-start function. Use the following formula to calculate  $C_{SS}$  for a given output voltage slew rate:

$$C_{SS} = 2\mu\text{A}(t_{ss}/1.231\text{V})$$

The pin should be left unconnected when not using the soft-start function.

## PIN FUNCTIONS

**BURST\_EN (Pin 4):** The BURST\_EN pin is used to enable or disable Burst Mode operation. Connect the BURST\_EN pin to ground to enable the burst mode function. Connect the pin to  $V_{FB}$  or  $V_{CC}$  to disable the burst mode function.

**$V_{FB}$  (Pin 5):** The output voltage feedback pin,  $V_{FB}$ , is externally connected to the supply output voltage via a resistive divider. The  $V_{FB}$  pin is internally connected to the inverting input of the error amplifier. In regulation,  $V_{FB}$  is 1.231V.

**$V_C$  (Pin 6):** The  $V_C$  pin is the output of the error amplifier whose voltage corresponds to the maximum (peak) switch current per oscillator cycle. The error amplifier is typically configured as an integrator circuit by connecting an RC network from the  $V_C$  pin to SGND. This circuit creates the dominant pole for the converter regulation control loop. Specific integrator characteristics can be configured to optimize transient response. When Burst Mode operation is enabled (see Pin 4 description), an internal low impedance clamp on the  $V_C$  pin is set at 100mV below the burst threshold, which limits the negative excursion of the pin voltage. Therefore, this pin cannot be pulled low with a low impedance source. If the  $V_C$  pin must be externally manipulated, do so through a 1k $\Omega$  series resistance.

**SYNC (Pin 7):** The Sync pin provides an external clock input for synchronization of the internal oscillator.  $R_{SET}$  is set such that the internal oscillator frequency is 10% to 25% below the external clock frequency. If unused the Sync pin is connected to SGND. For more information see "Oscillator Sync" in the Application Information section of this datasheet.

**$f_{SET}$  (Pin 8):** The  $f_{SET}$  pin programs the oscillator frequency with an external resistor,  $R_{SET}$ . The resistor is required even when supplying external sync clock signal. See the Applications Information section for resistor value selection details.

**SGND (Pin 9, 17):** The SGND pin is the low noise ground reference. It should be connected to the  $-V_{OUT}$  side of the output capacitors. Careful layout of the PCB is necessary to keep high currents away from this SGND connection. See the Application Information section for helpful hints on PCB layout of grounds.

**SENSE<sup>-</sup> (Pin 10):** The SENSE<sup>-</sup> pin is the negative input for the current sense amplifier and is connected to the  $V_{OUT}$  side of the sense resistor for step-down applications. The sensed inductor current limit is set to 100mV across the SENSE inputs.

**SENSE<sup>+</sup> (Pin 11):** The SENSE<sup>+</sup> pin is the positive input for the current sense amplifier and is connected to the inductor side of the sense resistor for step-down applications. The sensed inductor current limit is set to 100mV across the SENSE inputs.

**PGND (Pin 12):** The PGND pin is the high-current ground reference for internal low side switch and the  $V_{CC}$  regulator circuit. Connect the pin directly to the negative terminal of the  $V_{CC}$  decoupling capacitor. See the Application Information section for helpful hints on PCB layout of grounds.

**$V_{CC}$  (Pin 13):** The  $V_{CC}$  pin is the internal bias supply decoupling node. Use a low ESR 1 $\mu$ F or greater ceramic capacitor to decouple this node to PGND. Most internal IC functions are powered from this bias supply. An external diode connected from  $V_{CC}$  to the BOOST pin charges the bootstrapped capacitor during the off-time of the main power switch. Back driving the  $V_{CC}$  pin from an external DC voltage source, such as the  $V_{OUT}$  output of the regulator supply, increases overall efficiency and reduces power dissipation in the IC. In shutdown mode this pin sinks 20 $\mu$ A until the pin voltage is discharged to 0V.

**SW (Pin 14):** In step-down applications the SW pin is connected to the cathode of an external clamping Schottky diode, the drain of the power MOSFET and the inductor. The SW node voltage swing is from  $V_{IN}$  during the on-time of the power MOSFET, to a Schottky voltage drop below ground during the off-time of the power MOSFET. In start-up and in operating modes where there is insufficient inductor current to freewheel the Schottky diode, an internal switch is turned on to pull the SW pin to ground so that the BOOST pin capacitor can be charged. Give careful consideration in choosing the Schottky diode to limit the negative voltage swing on the SW pin.

**TG (Pin 15):** The TG pin is the bootstrapped gate drive for the top N-Channel MOSFET. Since very fast high currents are driven from this pin, connect it to the gate of the power MOSFET with a short and wide, typically 0.02" width, PCB trace to minimize inductance.





## OPERATION (Refer to Functional Diagram)

The LT3844 is a PWM controller with a constant frequency, current mode control architecture. It is designed for low to medium power, switching regulator applications. Its high operating voltage capability allows it to step-up or down input voltages up to 60V without the need for a transformer. The LT3844 is used in nonsynchronous applications, meaning that a freewheeling rectifier diode (D1 of Function Diagram) is used instead of a bottom side MOSFET. For circuit operation, please refer to the Functional Diagram of the IC and Typical Application on the front page of the data sheet. The LT3800 is a similar part that uses synchronous rectification, replacing the diode with a MOSFET in a step-down application.

### Main Control Loop

During normal operation, the external N-channel MOSFET switch is turned on at the beginning of each cycle. The switch stays on until the current in the inductor exceeds a current threshold set by the DC control voltage,  $V_C$ , which is the output of the voltage control loop. The voltage control loop monitors the output voltage, via the  $V_{FB}$  pin voltage, and compares it to an internal 1.231V reference. It increases the current threshold when the  $V_{FB}$  voltage is below the reference voltage and decreases the current threshold when the  $V_{FB}$  voltage is above the reference voltage. For instance, when an increase in the load current occurs, the output voltage drops causing the  $V_{FB}$  voltage to drop relative to the 1.231V reference. The voltage control loop senses the drop and increases the current threshold. The peak inductor current is increased until the average inductor current equals the new load current and the output voltage returns to regulation.

### Current Limit/Short-Circuit

The inductor current is measured with a series sense resistor (see the Typical Application on the front page). When the voltage across the sense resistor reaches the maximum current sense threshold, typically 100mV, the TG MOSFET driver is disabled for the remainder of that cycle. If the maximum current sense threshold is still exceeded at the beginning of the next cycle, the entire cycle is skipped. Cycle skipping keeps the inductor currents to a reasonable value during a short-circuit, particularly when  $V_{IN}$  is high. Setting the sense resistor value is discussed in the “Application Information” section.

### $V_{CC}$ /Boosted Supply

An internal  $V_{CC}$  regulator provides  $V_{IN}$  derived gate-drive power for start-up under all operating conditions with MOSFET gate charge loads up to 90nC. The regulator can operate continuously in applications with  $V_{IN}$  voltages up to 60V, provided the power dissipation of the regulator does not exceed 250mW. The power dissipation is calculated as follows:

$$P_{d(REG)} = (V_{IN} - 8V) \cdot f_{SW} \cdot Q_G$$

where  $Q_G$  is the MOSFET gate charge.

In applications where these conditions are exceeded,  $V_{CC}$  must be derived from an external source after start-up. Maximum continuous regulator power dissipation may be exceeded for short duration  $V_{IN}$  transients.

For higher converter efficiency and less power dissipation in the IC,  $V_{CC}$  can also be supplied from an external supply such as the converter output. When an external supply back drives the internal  $V_{CC}$  regulator through an external diode and the  $V_{CC}$  voltage is pulled to a diode above its regulation voltage, the internal regulator is disabled and goes into a low current mode.  $V_{CC}$  is the bias supply for most of the internal IC functions and is also used to charge the bootstrapped capacitor ( $C_{BOOST}$ ) via an external diode. The external MOSFET switch is biased from the bootstrapped capacitor. While the external MOSFET switch is off, an internal BJT switch, whose collector is connected to the SW pin and emitter is connected to the PGND pin, is turned on to pull the SW node to PGND and recharge the bootstrap capacitor. The switch stays on until either the start of the next cycle or until the bootstrapped capacitor is fully charged.

### MOSFET Driver

The LT3844 contains a high speed boosted driver to turn on and off an external N-channel MOSFET switch. The MOSFET driver derives its power from the boost capacitor which is referenced to the SW pin and the source of the MOSFET. The driver provides a large pulse of current to turn on the MOSFET fast to minimize transition times. Multiple MOSFETs can be paralleled for higher current operation.



## OPERATION (Refer to Functional Diagram)

To eliminate the possibility of shoot through between the MOSFET and the internal SW pull-down switch, an adaptive nonoverlap circuit ensures that the internal pull-down switch does not turn on until the gate of the MOSFET is below its turn on threshold.

### Low Current Operation (Burst Mode Operation)

To increase low current load efficiency, the LT3844 is capable of operating in Linear Technology's proprietary Burst Mode operation where the external MOSFET operates intermittently based on load current demand. The Burst Mode function is disabled by connecting the BURST\_EN pin to  $V_{CC}$  or  $V_{FB}$  and enabled by connecting the pin to SGND.

When the required switch current, sensed via the  $V_C$  pin voltage, is below 15% of maximum, Burst Mode operation is employed and that level of sense current is latched onto the IC control path. If the output load requires less than this latched current level, the converter will overdrive the output slightly during each switch cycle. This overdrive condition is sensed internally and forces the voltage on the  $V_C$  pin to continue to drop. When the voltage on  $V_C$  drops 150mV below the 15% load level, switching is disabled, and the LT3844 shuts down most of its internal circuitry, reducing total quiescent current to 120 $\mu$ A. When the converter output begins to fall, the  $V_C$  pin voltage begins to climb. When the voltage on the  $V_C$  pin climbs back to the 15% load level, the IC returns to normal operation and switching resumes. An internal clamp on the  $V_C$  pin is set at 100mV below the output disable threshold, which limits the negative excursion of the pin voltage, minimizing the converter output ripple during Burst Mode operation.

During Burst Mode operation, the  $V_{IN}$  pin current is 20 $\mu$ A and the  $V_{CC}$  current is reduced to 100 $\mu$ A. If no external drive is provided for  $V_{CC}$ , all  $V_{CC}$  bias currents originate from the  $V_{IN}$  pin, giving a total  $V_{IN}$  current of 120 $\mu$ A. Burst current can be reduced further when  $V_{CC}$  is driven using an output derived source, as the  $V_{CC}$  component of  $V_{IN}$  current is then reduced by the converter duty cycle ratio.

### Start-Up

The following section describes the start-up of the supply and operation down to 4V once the step-down supply is up and running. For the protection of the LT3844 and the switching supply, there are internal undervoltage lockout (UVLO) circuits with hysteresis on  $V_{IN}$ ,  $V_{CC}$  and  $V_{BOOST}$ , as shown in the Electrical Characteristics table. Start-up and continuous operation require that all three of these undervoltage lockout conditions be satisfied because the TG MOSFET driver is disabled during any UVLO fault condition. In startup, for most applications,  $V_{CC}$  is powered from  $V_{IN}$  through the high voltage linear regulator of the LT3844. This requires  $V_{IN}$  to be high enough to drive the  $V_{CC}$  voltage above its undervoltage lockout threshold.  $V_{CC}$ , in turn, has to be high enough to charge the BOOST capacitor through an external diode so that the BOOST voltage is above its undervoltage lockout threshold. There is an NPN switch that pulls the SW node to ground each cycle during the TG power MOSFET off-time, ensuring the BOOST capacitor is kept fully charged. Once the supply is up and running, the output voltage of the supply can backdrive  $V_{CC}$  through an external diode. Internal circuitry disables the high voltage regulator to conserve  $V_{IN}$  supply current. Output voltages that are too low or too high to backdrive  $V_{CC}$  require additional circuitry such as a voltage doubler or linear regulator. Once  $V_{CC}$  is backdriven from a supply other than  $V_{IN}$ ,  $V_{IN}$  can be reduced to 4V with normal operation maintained.

### Soft-Start

The soft-start function controls the slew rate of the power supply output voltage during start-up. A controlled output voltage ramp minimizes output voltage overshoot, reduces inrush current from the  $V_{IN}$  supply, and facilitates supply sequencing. A capacitor,  $C_{SS}$ , connected from the  $C_{SS}$  pin to SGND, programs the slew rate. The capacitor is charged from an internal 2 $\mu$ A current source producing a ramped voltage. The capacitor voltage overrides the internal reference to the error amplifier. If the  $V_{FB}$  pin voltage

## OPERATION (Refer to Functional Diagram)

exceeds the  $C_{SS}$  pin voltage then the current threshold set by the DC control voltage,  $V_C$ , is decreased and the inductor current is lowered. This in turn decreases the output voltage slew rate allowing the  $C_{SS}$  pin voltage ramp to catch up to the  $V_{FB}$  pin voltage. An internal 100mV offset is added to the  $V_{FB}$  pin voltage relative to the  $C_{SS}$  pin voltage so that at start-up the soft-start circuit will discharge the  $V_C$  pin voltage below the DC control voltage equivalent to zero inductor current. This will reduce the input supply inrush current. The soft-start circuit is disabled once the  $C_{SS}$  pin voltage has been charged to 200mV above the internal reference of 1.231V.

During a  $V_{IN}$  UVLO,  $V_{CC}$  UVLO or  $\overline{SHDN}$  UVLO event, the  $C_{SS}$  pin voltage is discharged with a 50 $\mu$ A current source. In normal operation the  $C_{SS}$  pin voltage is clamped to a diode above the  $V_{FB}$  pin voltage. Therefore, the value of the  $C_{SS}$  capacitor is relevant in how long of a fault event will retrigger a soft-start. In other words, if any of the above UVLO conditions occur, the  $C_{SS}$  pin voltage will be discharged with a 50 $\mu$ A current source. There is a diode worth of voltage headroom to ride through the fault before the  $C_{SS}$  pin voltage enters its active region and the soft-start function is enabled.

Also, since the  $C_{SS}$  pin voltage is clamped to a diode above the  $V_{FB}$  pin voltage, during a short circuit the  $C_{SS}$  pin voltage is pulled low because the  $V_{FB}$  pin voltage is low. Once the short has been removed the  $V_{FB}$  pin voltage starts to recover. The soft-start circuit takes control of the output voltage slew rate once the  $V_{FB}$  pin voltage has exceeded the slowly ramping  $C_{SS}$  pin voltage, reducing the output voltage overshoot during a short circuit recovery.

## Slope/Antislope Compensation

The IC incorporates slope compensation to eliminate potential subharmonic oscillations in the current control loop. The IC's slope compensation circuit imposes an artificial ramp on the sensed current to increase the rising slope as duty cycle increases.

Typically, this additional ramp affects the sensed current value, thereby reducing the achievable current limit value by the same amount as the added ramp represents. As such, the current limit is typically reduced as the duty cycle increases. The LT3844, however, contains antislope compensation circuitry to eliminate the current limit reduction associated with slope compensation. As the slope compensation ramp is added to the sensed current, a similar ramp is added to the current limit threshold. The end result is that the current limit is not compromised so the LT3844 can provide full power regardless of required duty cycle.

## Shutdown

The LT3844 includes a shutdown mode where all the internal IC functions are disabled and the  $V_{IN}$  current is reduced to less than 10 $\mu$ A. The shutdown pin can be used for undervoltage lockout with hysteresis, micropower shutdown or as a general purpose on/off control of the converter output. The shutdown function has two thresholds. The first threshold, a precision 1.23V threshold with 120mV of hysteresis, disables the converter from switching. The second threshold, approximately a 0.7V referenced to SGND, completely disables all internal circuitry and reduces the  $V_{IN}$  current to less than 10 $\mu$ A. See the Application Information section for more information.

## APPLICATIONS INFORMATION

The basic LT3844 step-down (buck) application, shown in the Typical Application on the front page, converts a larger positive input voltage to a lower positive or negative output voltage. This Application Information section assists selection of external components for the requirements of the power supply.

## $R_{SENSE}$ Selection

The current sense resistor,  $R_{SENSE}$ , monitors the inductor current of the supply (See Typical Application on front page). Its value is chosen based on the maximum required output load current. The LT3844 current sense amplifier has a maximum voltage threshold of, typically, 100mV.

## APPLICATIONS INFORMATION

Therefore, the peak inductor current is  $100\text{mV}/R_{\text{SENSE}}$ . The maximum output load current,  $I_{\text{OUT(MAX)}}$ , is the peak inductor current minus half the peak-to-peak ripple current,  $\Delta I_L$ .

Allowing adequate margin for ripple current and external component tolerances,  $R_{\text{SENSE}}$  can be calculated as follows:

$$R_{\text{SENSE}} = \frac{70\text{mV}}{I_{\text{OUT(MAX)}}}$$

Typical values for  $R_{\text{SENSE}}$  are in the range of  $0.005\Omega$  to  $0.05\Omega$ .

### Operating Frequency

The choice of operating frequency and inductor value is a trade off between efficiency and component size. Low frequency operation improves efficiency by reducing MOSFET switching losses and gate charge losses. However, lower frequency operation requires more inductance for a given amount of ripple current, resulting in a larger inductor size and higher cost. If the ripple current is allowed to increase, larger output capacitors may be required to maintain the same output ripple. For converters with high step-down  $V_{\text{IN}}$  to  $V_{\text{OUT}}$  ratios, another consideration is the minimum on-time of the LT3844 (see the Minimum On-time Considerations section). A final consideration for operating frequency is that in noise-

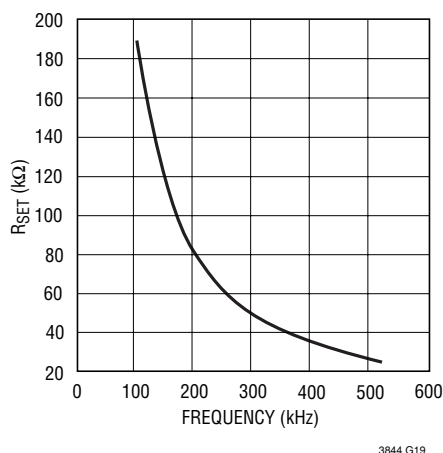


Figure 1. Timing Resistor ( $R_{\text{SET}}$ ) Value

sensitive communications systems, it is often desirable to keep the switching noise out of a sensitive frequency band. The LT3844 uses a constant frequency architecture that can be programmed over a 100kHz to 500kHz range with a single resistor from the  $f_{\text{SET}}$  pin to ground, as shown in Figure 1. The nominal voltage on the  $f_{\text{SET}}$  pin is 1V and the current that flows from this pin is used to charge an internal oscillator capacitor. The value of  $R_{\text{SET}}$  for a given operating frequency can be chosen from Figure 4 or from the following equation:

$$R_{\text{SET}}(\text{k}\Omega) = 8.4 \cdot 10^4 \cdot f_{\text{SW}}^{(-1.31)}$$

The following table lists typical resistor values for common operating frequencies:

#### Recommended 1% Standard Values

$R_{\text{SET}}$	$f_{\text{SW}}$
191kΩ	100kHz
118kΩ	150kHz
80.6kΩ	200kHz
63.4kΩ	250kHz
49.9kΩ	300kHz
40.2kΩ	350kHz
33.2kΩ	400kHz
27.4kΩ	450kHz
23.2kΩ	500kHz

### Step-Down Converter: Inductor Selection

The critical parameters for selection of an inductor are minimum inductance value, volt-second product, saturation current and/or RMS current.

For a given  $\Delta I_L$ , The minimum inductance value is calculated as follows:

$$L \geq V_{\text{OUT}} \cdot \frac{V_{\text{IN(MAX)}} - V_{\text{OUT}}}{f_{\text{SW}} \cdot V_{\text{IN(MAX)}} \cdot \Delta I_L}$$

$f_{\text{SW}}$  is the switch frequency.

## APPLICATIONS INFORMATION

The typical range of values for  $\Delta I_L$  is  $(0.2 \cdot I_{OUT(MAX)})$  to  $(0.5 \cdot I_{OUT(MAX)})$ , where  $I_{OUT(MAX)}$  is the maximum load current of the supply. Using  $\Delta I_L = 0.3 \cdot I_{OUT(MAX)}$  yields a good design compromise between inductor performance versus inductor size and cost. A value of  $\Delta I_L = 0.3 \cdot I_{OUT(MAX)}$  produces a  $\pm 15\%$  of  $I_{OUT(MAX)}$  ripple current around the DC output current of the supply. Lower values of  $\Delta I_L$  require larger and more costly magnetics. Higher values of  $\Delta I_L$  will increase the peak currents, requiring more filtering on the input and output of the supply. If  $\Delta I_L$  is too high, the slope compensation circuit is ineffective and current mode instability may occur at duty cycles greater than 50%. To satisfy slope compensation requirements the minimum inductance is calculated as follows:

$$L > V_{OUT} \cdot \frac{2DC_{MAX} - 1}{DC_{MAX}} \cdot \frac{R_{SENSE} \cdot 8.33}{f_{SW}}$$

Some magnetics vendors specify a volt-second product in their datasheet. If they do not, consult the magnetics vendor to make sure the specification is not being exceeded by your design. The volt-second product is calculated as follows:

$$\text{Volt-second } (\mu\text{sec}) = \frac{(V_{IN(MAX)} - V_{OUT}) \cdot V_{OUT}}{V_{IN(MAX)} \cdot f_{SW}}$$

The magnetics vendors specify either the saturation current, the RMS current or both. When selecting an inductor based on inductor saturation current, use the peak current through the inductor,  $I_{OUT(MAX)} + \Delta I_L/2$ . The inductor saturation current specification is the current at which the inductance, measured at zero current, decreases by a specified amount, typically 30%.

When selecting an inductor based on RMS current rating, use the average current through the inductor,  $I_{OUT(MAX)}$ . The RMS current specification is the RMS current at which the part has a specific temperature rise, typically 40°C, above 25°C ambient.

After calculating the minimum inductance value, the volt-second product, the saturation current and the RMS current for your design, select an off-the-shelf inductor. Contact the Application group at Linear Technology for further support.

For more detailed information on selecting an inductor, please see the “Inductor Selection” section of Linear Technology Application Note 44.

### Step-Down Converter: MOSFET Selection

The selection criteria of the external N-channel standard level power MOSFET include on resistance ( $R_{DS(ON)}$ ), reverse transfer capacitance ( $C_{RSS}$ ), maximum drain source voltage ( $V_{DSS}$ ), total gate charge ( $Q_G$ ) and maximum continuous drain current.

For maximum efficiency, minimize  $R_{DS(ON)}$  and  $C_{RSS}$ . Low  $R_{DS(ON)}$  minimizes conduction losses while low  $C_{RSS}$  minimizes transition losses. The problem is that  $R_{DS(ON)}$  is inversely related to  $C_{RSS}$ . Balancing the transition losses with the conduction losses is a good idea in sizing the MOSFET. Select the MOSFET to balance the two losses.

Calculate the maximum conduction losses of the MOSFET:

$$P_{COND} = (I_{OUT(MAX)})^2 \left( \frac{V_{OUT}}{V_{IN}} \right) (R_{DS(ON)})$$

Note that  $R_{DS(ON)}$  has a large positive temperature dependence. The MOSFET manufacturer's data sheet contains a curve,  $R_{DS(ON)}$  vs Temperature.

Calculate the maximum transition losses:

$$P_{TRAN} = (k)(V_{IN})^2 (I_{OUT(MAX)})(C_{RSS})(f_{SW})$$

where  $k$  is a constant inversely related to the gate driver current, approximated by  $k = 2$  for LT3844 applications.

The total maximum power dissipation of the MOSFET is the sum of these two loss terms:

$$P_{FET(TOTAL)} = P_{COND} + P_{TRAN}$$

To achieve high supply efficiency, keep the  $P_{FET(TOTAL)}$  to less than 3% of the total output power. Also, complete a thermal analysis to ensure that the MOSFET junction temperature is not exceeded.

$$T_J = T_A + P_{FET(TOTAL)} \cdot \theta_{JA}$$

where  $\theta_{JA}$  is the package thermal resistance and  $T_A$  is the ambient temperature. Keep the calculated  $T_J$  below the maximum specified junction temperature, typically 150°C.



## APPLICATIONS INFORMATION

Note that when  $V_{IN}$  is high and  $f_{SW}$  is high, the transition losses may dominate. A MOSFET with higher  $R_{DS(ON)}$  and lower  $C_{RSS}$  may provide higher efficiency. MOSFETs with higher voltage  $V_{DSS}$  specification usually have higher  $R_{DS(ON)}$  and lower  $C_{RSS}$ .

Choose the MOSFET  $V_{DSS}$  specification to exceed the maximum voltage across the drain to the source of the MOSFET, which is  $V_{IN(MAX)}$  plus any additional ringing on the switch node. Ringing on the switch node can be greatly reduced with good PCB layout and, if necessary, an RC snubber.

The internal  $V_{CC}$  regulator is capable of sourcing up to 40mA which limits the maximum total MOSFET gate charge,  $Q_G$ , to  $40mA/f_{SW}$ . The  $Q_G$  vs  $V_{GS}$  specification is typically provided in the MOSFET data sheet. Use  $Q_G$  at  $V_{GS}$  of 8V. If  $V_{CC}$  is back driven from an external supply, the MOSFET drive current is not sourced from the internal regulator of the LT3844 and the  $Q_G$  of the MOSFET is not limited by the IC. However, note that the MOSFET drive current is supplied by the internal regulator when the external supply back driving  $V_{CC}$  is not available such as during startup or short-circuit.

The manufacturer's maximum continuous drain current specification should exceed the peak switch current,  $I_{OUT(MAX)} + \Delta I_L/2$ .

During the supply startup, the gate drive levels are set by the  $V_{CC}$  voltage regulator, which is approximately 8V. Once the supply is up and running, the  $V_{CC}$  can be back driven by an auxiliary supply such as  $V_{OUT}$ . It is important not to exceed the manufacturer's maximum  $V_{GS}$  specification. A standard level threshold MOSFET typically has a  $V_{GS}$  maximum of 20V.

### Step-Down Converter: Rectifier Selection

The rectifier diode (D1 on the Functional Diagram) in a buck converter generates a current path for the inductor current when the main power switch is turned off. The rectifier is selected based upon the forward voltage, reverse voltage and maximum current. A Schottky diode is recommended. Its low forward voltage yields the lowest power loss and highest efficiency. The maximum reverse voltage that the diode will see is  $V_{IN(MAX)}$ .

In continuous mode operation, the average diode current is calculated at maximum output load current and maximum  $V_{IN}$ :

$$I_{DIODE(AVG)} = I_{OUT(MAX)} \frac{V_{IN(MAX)} - V_{OUT}}{V_{IN(MAX)}}$$

To improve efficiency and to provide adequate margin for short-circuit operation, a diode rated at 1.5 to 2 times the maximum average diode current,  $I_{DIODE(AVG)}$ , is recommended.

### Step-Down Converter: Input Capacitor Selection

A local input bypass capacitor is required for buck converters because the input current is pulsed with fast rise and fall times. The input capacitor selection criteria are based on the bulk capacitance and RMS current capability. The bulk capacitance will determine the supply input ripple voltage. The RMS current capability is used to keep from overheating the capacitor.

The bulk capacitance is calculated based on maximum input ripple,  $\Delta V_{IN}$ :

$$C_{IN(BULK)} = \frac{I_{OUT(MAX)} \cdot V_{OUT}}{\Delta V_{IN} \cdot f_{SW} \cdot V_{IN(MIN)}}$$

$\Delta V_{IN}$  is typically chosen at a level acceptable to the user. 100mV-200mV is a good starting point. Aluminum electrolytic capacitors are a good choice for high voltage, bulk capacitance due to their high capacitance per unit area.

The capacitor's RMS current is:

$$I_{CIN(RMS)} = I_{OUT} \sqrt{\frac{V_{OUT}(V_{IN} - V_{OUT})}{(V_{IN})^2}}$$

If applicable, calculate it at the worst case condition,  $V_{IN} = 2V_{OUT}$ . The RMS current rating of the capacitor is specified by the manufacturer and should exceed the calculated  $I_{CIN(RMS)}$ . Due to their low ESR (Equivalent Series Resistance), ceramic capacitors are a good choice for high voltage, high RMS current handling. Note that the ripple current ratings from aluminum electrolytic capacitor manufacturers are based on 2000 hours of life. This makes it

## APPLICATIONS INFORMATION

advisable to further derate the capacitor or to choose a capacitor rated at a higher temperature than required.

The combination of aluminum electrolytic capacitors and ceramic capacitors is an economical approach to meeting the input capacitor requirements. The capacitor voltage rating must be rated greater than  $V_{IN(MAX)}$ . Multiple capacitors may also be paralleled to meet size or height requirements in the design. Locate the capacitor very close to the MOSFET switch and use short, wide PCB traces to minimize parasitic inductance.

### Step-Down Converter: Output Capacitor Selection

The output capacitance,  $C_{OUT}$ , selection is based on the design's output voltage ripple,  $\Delta V_{OUT}$  and transient load requirements.  $\Delta V_{OUT}$  is a function of  $\Delta I_L$  and the  $C_{OUT}$  ESR. It is calculated by:

$$\Delta V_{OUT} = \Delta I_L \cdot \left( ESR + \frac{1}{(8 \cdot f_{SW} \cdot C_{OUT})} \right)$$

The maximum ESR required to meet a  $\Delta V_{OUT}$  design requirement can be calculated by:

$$ESR(MAX) = \frac{(\Delta V_{OUT})(L)(f_{SW})}{V_{OUT} \cdot \left( 1 - \frac{V_{OUT}}{V_{IN(MAX)}} \right)}$$

Worst-case  $\Delta V_{OUT}$  occurs at highest input voltage. Use paralleled multiple capacitors to meet the ESR requirements. Increasing the inductance is an option to lower the ESR requirements. For extremely low  $\Delta V_{OUT}$ , an additional LC filter stage can be added to the output of the supply. Application Note 44 has some good tips on sizing an additional output filter.

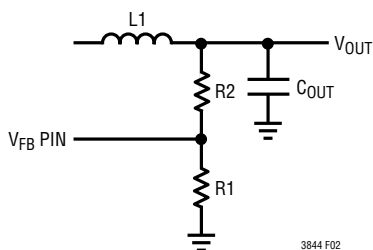


Figure 2. Output Voltage Feedback Divider

### Output Voltage Programming

A resistive divider sets the DC output voltage according to the following formula:

$$R2 = R1 \left( \frac{V_{OUT}}{1.231V} - 1 \right)$$

The external resistor divider is connected to the output of the converter as shown in Figure 2. Tolerance of the feedback resistors will add additional error to the output voltage.

Example:  $V_{OUT} = 12V$ ;  $R1 = 10k\Omega$

$$R2 = 10k\Omega \left( \frac{12V}{1.231V} - 1 \right) = 87.48k\Omega - \text{use } 86.6k\Omega \text{ } 1\%$$

The  $V_{FB}$  pin input bias current is typically 25nA, so use of extremely high value feedback resistors could cause a converter output that is slightly higher than expected. Bias current error at the output can be estimated as:

$$\Delta V_{OUT(BIAS)} = 25nA \cdot R2$$

### Supply UVLO and Shutdown

The  $\overline{SHDN}$  pin has a precision voltage threshold with hysteresis which can be used as an undervoltage lockout threshold (UVLO) for the power supply. Undervoltage lockout keeps the LT3844 in shutdown until the supply input voltage is above a certain voltage programmed by the user. The hysteresis voltage prevents noise from falsely tripping UVLO.

Resistors are chosen by first selecting  $R_B$ . Then

$$R_A = R_B \cdot \left( \frac{V_{SUPPLY(ON)}}{1.35V} - 1 \right)$$

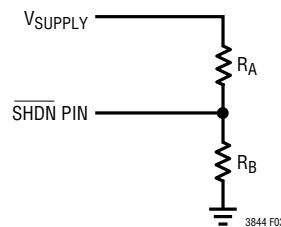


Figure 3. Undervoltage Lockout Circuit



## APPLICATIONS INFORMATION

$V_{\text{SUPPLY(ON)}}$  is the input voltage at which the undervoltage lockout is disabled and the supply turns on.

Example: Select  $R_B = 49.9\text{k}\Omega$ ,  $V_{\text{SUPPLY(ON)}} = 14.5\text{V}$  (based on a 15V minimum input voltage)

$$R_A = 49.9\text{k}\Omega \cdot \left( \frac{14.5\text{V}}{1.35\text{V}} - 1 \right) \\ = 486.1\text{k}\Omega \text{ (499k}\Omega \text{ resistor is selected)}$$

If low supply current in standby mode is required, select a higher value of  $R_B$ .

The supply turn off voltage is 9% below turn on. In the example the  $V_{\text{SUPPLY(OFF)}}$  would be 13.2V.

If additional hysteresis is desired for the enable function, an external positive feedback resistor can be used from the LT3844 regulator output.

The shutdown function can be disabled by connecting the  $\overline{\text{SHDN}}$  pin to the  $V_{\text{IN}}$  through a large value pull-up resistor. This pin contains a low impedance clamp at 6V, so the  $\overline{\text{SHDN}}$  pin will sink current from the pull-up resistor ( $R_{\text{PU}}$ ):

$$I_{\text{SHDN}} = \frac{V_{\text{IN}} - 6\text{V}}{R_{\text{PU}}}$$

Because this arrangement will clamp the  $\overline{\text{SHDN}}$  pin to the 6V, it will violate the 5V absolute maximum voltage rating of the pin. This is permitted, however, as long as the absolute maximum input current rating of 1mA is not exceeded. Input  $\overline{\text{SHDN}}$  pin currents of  $<100\mu\text{A}$  are recommended: a  $1\text{M}\Omega$  or greater pull-up resistor is typically used for this configuration.

### Soft-Start

The desired soft-start time ( $t_{\text{SS}}$ ) is programmed via the  $C_{\text{SS}}$  capacitor as follows:

$$C_{\text{SS}} = \frac{2\mu\text{A} \cdot t_{\text{SS}}}{1.23\text{V}}$$

The amount of time in which the power supply can withstand a  $V_{\text{IN}}$ ,  $V_{\text{CC}}$  or  $\overline{\text{VSHDN}}$  UVLO fault condition ( $t_{\text{FAULT}}$ ) before the  $C_{\text{SS}}$  pin voltage enters its active region is approximated by the following formula:

$$t_{\text{FAULT}} = \frac{C_{\text{SS}} \cdot 0.65\text{V}}{50\mu\text{A}}$$

### Oscillator SYNC

The oscillator can be synchronized to an external clock. Set the  $R_{\text{SET}}$  resistor at least 10% below the desired sync frequency.

It is recommended that the SYNC pin be driven with a square wave that has amplitude greater than 2V, pulse width greater than  $1\mu\text{s}$  and rise time less than 500ns. The rising edge of the sync wave form triggers the discharge of the internal oscillator capacitor.

### Efficiency Considerations

The efficiency of a switching regulator is equal to the output power divided by the input power times 100%. Express percent efficiency as:

$$\% \text{ Efficiency} = 100\% - (L1 + L2 + L3 + \dots)$$

where L1, L2, etc. are individual loss terms as a percentage of input power.

Although all dissipative elements in the circuit produce losses, four main contributors usually account for most of the losses in LT3844 circuits:

1. LT3844  $V_{\text{IN}}$  and  $V_{\text{CC}}$  current loss
2.  $I^2R$  conduction losses
3. MOSFET transition loss
4. Schottky diode conduction loss

1. The  $V_{\text{IN}}$  and  $V_{\text{CC}}$  currents are the sum of the quiescent currents of the LT3844 and the MOSFET drive currents. The quiescent currents are in the LT3844 Electrical Characteristics table. The MOSFET drive current is a result of charging the gate capacitance of the power MOSFET each cycle with a packet of charge,  $Q_G$ .  $Q_G$  is found in the MOSFET data sheet. The average charging current is calculated as  $Q_G \cdot f_{\text{SW}}$ . The power loss term due to these currents can be reduced by backdriving  $V_{\text{CC}}$  with a lower voltage than  $V_{\text{IN}}$  such as  $V_{\text{OUT}}$ .

## APPLICATIONS INFORMATION

2.  $I^2R$  losses are calculated from the DC resistances of the MOSFET, the inductor, the sense resistor and the input and output capacitors. In continuous conduction mode the average output current flows through the inductor and  $R_{SENSE}$  but is chopped between the MOSFET and the Schottky diode. The resistances of the MOSFET ( $R_{DS(ON)}$ ) and the  $R_{SENSE}$  multiplied by the duty cycle can be summed with the resistances of the inductor and  $R_{SENSE}$  to obtain the total series resistance of the circuit. The total conduction power loss is proportional to this resistance and usually accounts for between 2% to 5% loss in efficiency.

3. Transition losses of the MOSFET can be substantial with input voltages greater than 20V. See MOSFET Selection section.

4. The Schottky diode can be a major contributor of power loss especially at high input to output voltage ratios (low duty cycles) where the diode conducts for the majority of the switch period. Lower  $V_f$  reduces the losses. Note that oversizing the diode does not always help because as the diode heats up the  $V_f$  is reduced and the diode loss term is decreased.

$I^2R$  losses and the Schottky diode loss dominate at high load currents. Other losses including  $C_{IN}$  and  $C_{OUT}$  ESR dissipative losses and inductor core losses generally account for less than 2% total additional loss in efficiency.

### PCB Layout Checklist

When laying out the printed circuit board, the following checklist should be used to ensure proper operation. These items are illustrated graphically in the layout diagram of Figure 3.

1. Keep the signal and power grounds separate. The signal ground consists of the LT3844 SGND pin, the exposed pad on the backside of the LT3844 IC and the (–) terminal of  $V_{OUT}$ . The signal ground is the quiet ground and does not contain any high, fast currents. The power ground consists of the Schottky diode anode, the (–) terminal of the input capacitor and the ground return of the  $V_{CC}$  capacitor.

This ground has very fast high currents and is considered the noisy ground. The two grounds are connected to each other only at the (–) terminal of  $V_{OUT}$ .

2. Use short wide traces in the loop formed by the MOSFET, the Schottky diode and the input capacitor to minimize high frequency noise and voltage stress from parasitic inductance. Surface mount components are preferred.

3. Connect the  $V_{FB}$  pin directly to the feedback resistors independent of any other nodes, such as the  $SENSE^-$  pin. Connect the feedback resistors between the (+) and (–) terminals of  $C_{OUT}$ . Locate the feedback resistors in close proximity to the LT3844 to keep the high impedance node,  $V_{FB}$ , as short as possible.

4. Route the  $SENSE^-$  and  $SENSE^+$  traces together and keep as short as possible.

5. Locate the  $V_{CC}$  and BOOST capacitors in close proximity to the IC. These capacitors carry the MOSFET driver's high peak currents. Place the small signal components away from high frequency switching nodes (BOOST, SW and TG). In the layout shown in Figure 3, place all the small signal components on one side of the IC and all the power components on the other. This helps to keep the signal and power grounds separate.

6. A small decoupling capacitor (100pF) is sometimes useful for filtering high frequency noise on the feedback and sense nodes. If used, locate as close to the IC as possible.

7. The LT3844 packaging will efficiently remove heat from the IC through the exposed pad on the backside of the part. The exposed pad is soldered to a copper footprint on the PCB. Make this footprint as large as possible to improve the thermal resistance of the IC case to ambient air. This helps to keep the LT3844 at a lower temperature.

8. Make the trace connecting the gate of MOSFET M1 to the TG pin of the LT3844 short and wide.

# APPLICATIONS INFORMATION

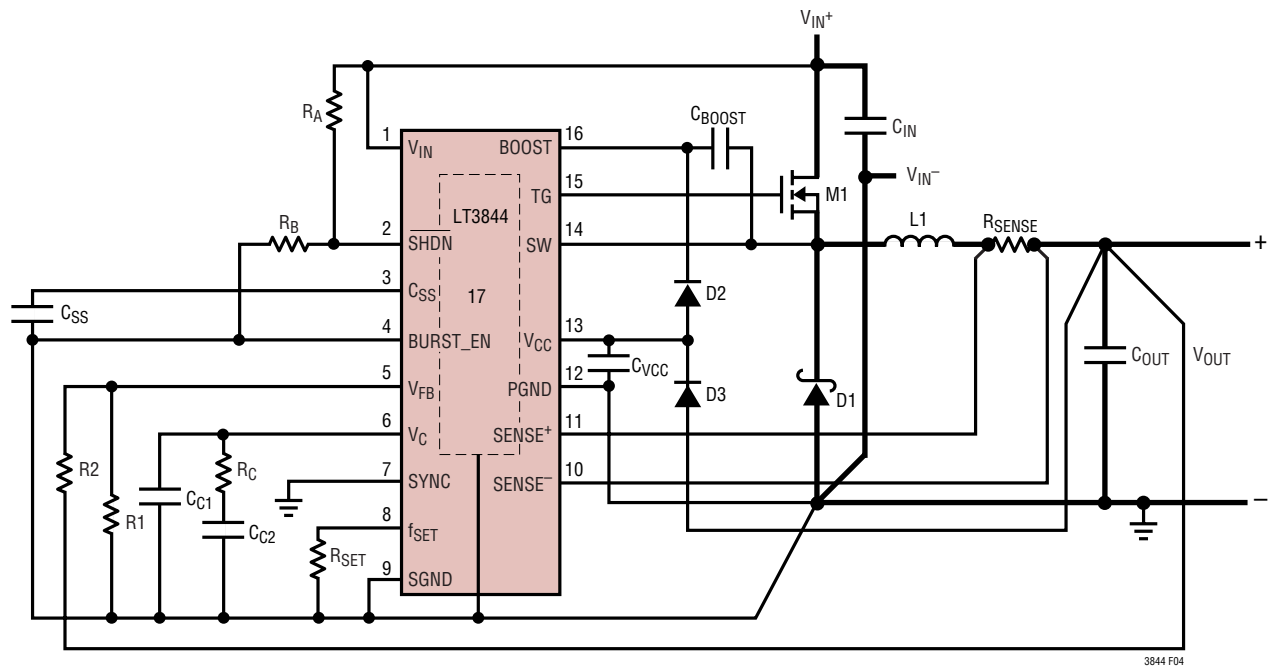


Figure 4. LT3844 Layout Diagram (See PCB Layout Checklist).

## APPLICATIONS INFORMATION

### Minimum On-Time Considerations (Buck Mode)

Minimum on-time  $t_{ON(MIN)}$  is the smallest amount of time that the LT3844 is capable of turning the top MOSFET on and off again. It is determined by internal timing delays and the amount of gate charge required turning on the top MOSFET. Low duty cycle applications may approach this minimum on-time limit and care should be taken to ensure that:

$$t_{ON} = \frac{V_{OUT}}{V_{IN} \cdot f_{SW}} > t_{ON(MIN)}$$

where  $t_{ON(MIN)}$  is typically 350ns worst case.

If the duty cycle falls below what can be accommodated by the minimum on-time, the LT3844 will begin to skip cycles. The output will be regulated, but the ripple current and ripple voltage will increase. If lower frequency operation is acceptable, the on-time can be increased above  $t_{ON(MIN)}$  for the same step-down ratio.

### Boost Converter Design

The LT3844 can be used to configure a boost converter to step-up voltages to as high as hundreds of volts. An example of a boost converter circuit schematic is shown in the Typical Applications section. The following sections are a guide to designing a boost converter:

The maximum duty cycle of the main switch is:

$$DC_{MAX} = \frac{V_{OUT} - V_{IN(MIN)}}{V_{OUT}}$$

### Boost Converter: Inductor Selection

The critical parameters for selection of an inductor are minimum inductance value and saturation current. The minimum inductance value is calculated as follows:

$$L_{MIN} = \frac{V_{IN(MIN)}}{\Delta I_L \cdot f_{SW}} \cdot DC_{MAX}$$

$f_{SW}$  is the switch frequency.

Similar to the buck converter, the typical range of values for  $\Delta I_L$  is  $(0.2 \cdot I_{L(MAX)})$  to  $(0.5 \cdot I_{L(MAX)})$ , where  $I_{L(MAX)}$  is the maximum average inductor current.

$$I_{L(MAX)} = I_{OUT(MAX)} \cdot \frac{V_{OUT}}{V_{IN(MIN)}}$$

Using  $\Delta I_L = 0.3 \cdot I_{L(MAX)}$  yields a good design compromise between inductor performance versus inductor size and cost.

The inductor must not saturate at the peak operating current,  $I_{L(MAX)} + \Delta I_L/2$ . The inductor saturation current specification is the current at which the inductance, measured at zero current, decreases by a specified amount, typically 30%.

One drawback of boost regulators is that they cannot be current limited for output shorts because the current steering diode makes a direct connection between input and output. Therefore, the inductor current during an output short circuit is only limited by the available current of the input supply.

After calculating the minimum inductance value and the saturation current for your design, select an off-the-shelf inductor. For more detailed information on selecting an inductor, please see the "Inductor Selection" section of Linear Technology Application Note 19.

### Boost Converter: MOSFET Selection

The selection criteria of the external N-channel standard level power MOSFET include on resistance ( $R_{DS(ON)}$ ), reverse transfer capacitance ( $C_{RSS}$ ), maximum drain source voltage ( $V_{DSS}$ ), total gate charge ( $Q_G$ ) and maximum continuous drain current.

For maximum efficiency, minimize  $R_{DS(ON)}$  and  $C_{RSS}$ . Low  $R_{DS(ON)}$  minimizes conduction losses while low  $C_{RSS}$  minimizes transition losses. The problem is that  $R_{DS(ON)}$  is inversely related to  $C_{RSS}$ . Balancing the transition losses with the conduction losses is a good idea in sizing the MOSFET. Select the MOSFET to balance the two losses. Calculate the maximum conduction losses of the MOSFET:

## APPLICATIONS INFORMATION

$$P_{\text{COND}} = DC_{\text{MAX}} \left( \frac{I_{\text{OUT(MAX)}}}{1 - DC_{\text{MAX}}} \right) \cdot R_{\text{DS(ON)}}$$

Note that  $R_{\text{DS(ON)}}$  has large positive temperature dependence. The MOSFET manufacturer's data sheet contains a curve,  $R_{\text{DS(ON)}}$  vs Temperature. Calculate the maximum transition losses:

$$P_{\text{TRAN}} = \frac{(k)(V_{\text{OUT}})^2 (I_{\text{OUT(MAX)}})(C_{\text{RSS}})(f_{\text{SW}})}{(1 - DC_{\text{MAX}})}$$

where  $k$  is a constant inversely related to the gate driver current, approximated by  $k = 2$  for LT3844 applications. The total maximum power dissipation of the MOSFET is the sum of these two loss terms:

$$P_{\text{FET(TOTAL)}} = P_{\text{COND}} + P_{\text{TRAN}}$$

To achieve high supply efficiency, keep the  $P_{\text{FET(TOTAL)}}$  to less than 3% of the total output power. Also, complete a thermal analysis to ensure that the MOSFET junction temperature is not exceeded.

$$T_J = T_A + P_{\text{FET(TOTAL)}} \cdot \theta_{JA}$$

where  $\theta_{JA}$  is the package thermal resistance and  $T_A$  is the ambient temperature. Keep the calculated  $T_J$  below the maximum specified junction temperature, typically 150°C. Note that when  $V_{\text{OUT}}$  is high (>20V), the transition losses may dominate. A MOSFET with higher  $R_{\text{DS(ON)}}$  and lower  $C_{\text{RSS}}$  may provide higher efficiency. MOSFETs with higher voltage  $V_{\text{DSS}}$  specification usually have higher  $R_{\text{DS(ON)}}$  and lower  $C_{\text{RSS}}$ .

Choose the MOSFET  $V_{\text{DSS}}$  specification to exceed the maximum voltage across the drain to the source of the MOSFET, which is  $V_{\text{OUT}}$  plus the forward voltage of the rectifier, typically less than 1V.

The internal  $V_{\text{CC}}$  regulator is capable of sourcing up to 40mA which limits the maximum total MOSFET gate charge,  $Q_G$ , to 40mA /  $f_{\text{SW}}$ . The  $Q_G$  vs  $V_{\text{GS}}$  specification is typically provided in the MOSFET data sheet. Use  $Q_G$  at  $V_{\text{GS}}$  of 8V. If  $V_{\text{CC}}$  is back driven from an external supply, the MOSFET drive current is not sourced from the internal regulator of the LT3844 and the  $Q_G$  of the MOSFET is not

limited by the IC. However, note that the MOSFET drive current is supplied by the internal regulator when the external supply back driving  $V_{\text{CC}}$  is not available such as during startup or short-circuit.

The manufacturer's maximum continuous drain current specification should exceed the peak switch current which is the same as the inductor peak current,  $I_{\text{L(MAX)}} + \Delta I_L/2$ .

During the supply startup, the gate drive levels are set by the  $V_{\text{CC}}$  voltage regulator, which is approximately 8V. Once the supply is up and running, the  $V_{\text{CC}}$  can be back driven by an auxiliary supply such as  $V_{\text{OUT}}$ . It is important not to exceed the manufacturer's maximum  $V_{\text{GS}}$  specification. A standard level threshold MOSFET typically has a  $V_{\text{GS}}$  maximum of 20V.

### Boost Converter: Rectifier Selection

The rectifier is selected based upon the forward voltage, reverse voltage and maximum current. A Schottky diode is recommended for its low forward voltage and yields the lowest power loss and highest efficiency. The maximum reverse voltage that the diode will see is  $V_{\text{OUT}}$ . The average diode current is equal to the maximum output load current,  $I_{\text{OUT(MAX)}}$ . A diode rated at 1.5 to 2 times the maximum average diode current is recommended. Remember boost converters are not short circuit protected.

### Boost Converter: Output Capacitor Selection

In boost mode, the output capacitor requirements are more demanding due to the fact that the current waveform is pulsed instead of continuous as in a buck converter. The choice of component(s) is driven by the acceptable ripple voltage which is affected by the ESR, ESL and bulk capacitance. The total output ripple voltage is:

$$\Delta V_{\text{OUT}} = I_{\text{OUT(MAX)}} \left( \frac{1}{f_{\text{SW}} \cdot C_{\text{OUT}}} + \frac{\text{ESR}}{1 - DC_{\text{MAX}}} \right)$$

where the first term is due to the bulk capacitance and the second term due to the ESR.

The choice of output capacitor is also driven by the RMS ripple current requirement. The RMS ripple current is:

## APPLICATIONS INFORMATION

$$I_{\text{RMS(COUT)}} = I_{\text{OUT(MAX)}} \cdot \sqrt{\frac{V_{\text{OUT}} - V_{\text{IN(MIN)}}}{V_{\text{IN(MIN)}}}}$$

At lower output voltages (<30V) it may be possible to satisfy both the output ripple voltage and RMS requirements with one or more capacitors of a single type. However, at output voltages above 30V where capacitors with both low ESR and high bulk capacitance are hard to find, the best approach is to use a combination of aluminum electrolytic and ceramic capacitors. The low ESR ceramic capacitor will minimize the ESR while the Aluminum Electrolytic capacitor will supply the required bulk capacitance.

### Boost Converter: Input Capacitor Selection

The input capacitor of a boost converter is less critical than the output capacitor, due to the fact that the inductor is in series with the input and the input current waveform is continuous. The input voltage source impedance determines the size of the input capacitor, which is typically in the range of 10μF to 100μF. A low ESR capacitor is recommended though not as critical as with the output capacitor. The RMS input capacitor ripple current for a boost converter is:

$$I_{\text{RMS(CIN)}} = 0.3 \cdot \frac{V_{\text{IN(MIN)}}}{L \cdot f_{\text{SW}}} \cdot \text{DC}_{\text{MAX}}$$

Please note that the input capacitor can see a very high surge current when a battery is suddenly connected to the input of the converter and solid tantalum capacitors can fail catastrophically under these conditions. Be sure to specify surge-tested capacitors.

### Boost Converter: R<sub>SENSE</sub> Selection

The boost application in the “Typical Applications” section has the location of the current sense resistor in series with the inductor with one side referenced to V<sub>IN</sub>. This location was chosen for two reasons. Firstly, the circulating current is always monitored so in the case of an output overvoltage or input over current condition the main switch will skip cycles to protect the circuitry. Secondly, the V<sub>IN</sub> node can be considered low noise since it is heavily filtered and the input current is not pulsed but continuous.

In the case where the input voltage exceeds the voltage limits on the LT3844 Sense pins, the sense resistor can be moved to the source of the MOSFET. In both cases the resistor value is the calculated using the same formula.

The LT3844 current comparator has a maximum threshold of 100mV/R<sub>SENSE</sub>. The current comparator threshold sets the peak of the inductor current. Allowing adequate margin for ripple current and external component tolerances, R<sub>SENSE</sub> can be calculated as follows:

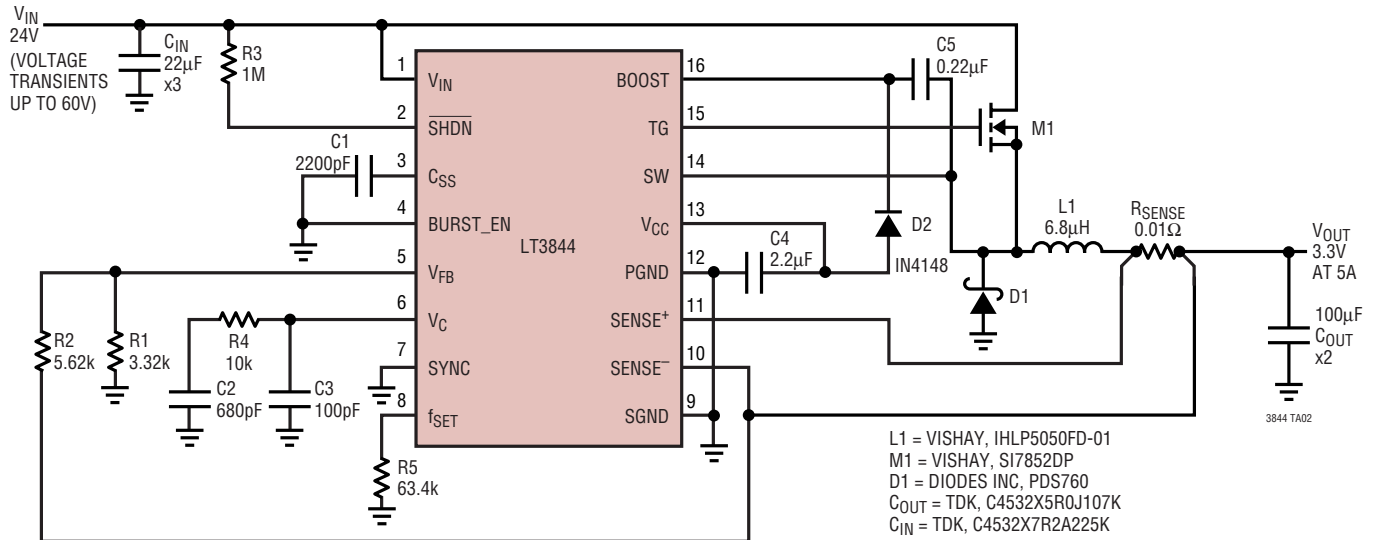
$$R_{\text{SENSE}} = \frac{70\text{mV}}{I_{\text{L(MAX)}}}$$

Where I<sub>L(MAX)</sub> is the maximum average inductor current as calculated in the “Boost Converter: Inductor Selection” section.

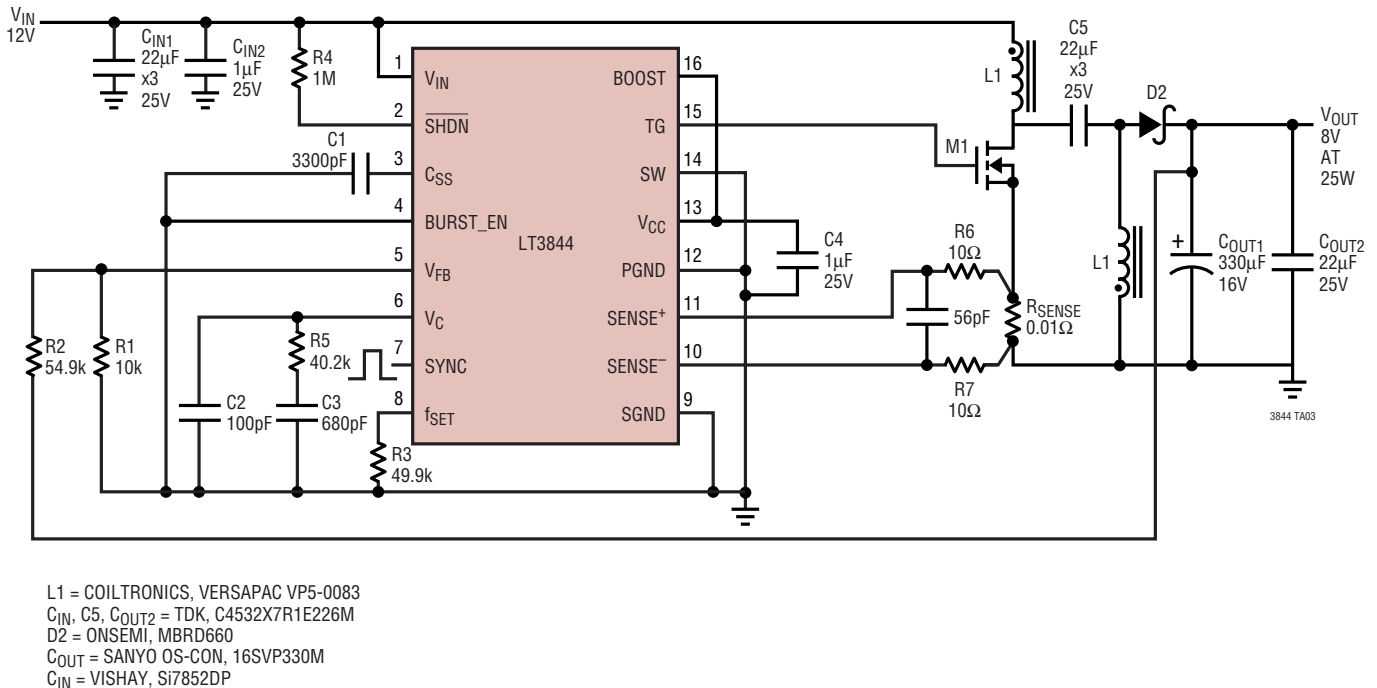


# TYPICAL APPLICATIONS

All Ceramic Capacitor Application, 24V to 3.3V at 5A,  $f_{SW} = 250kHz$

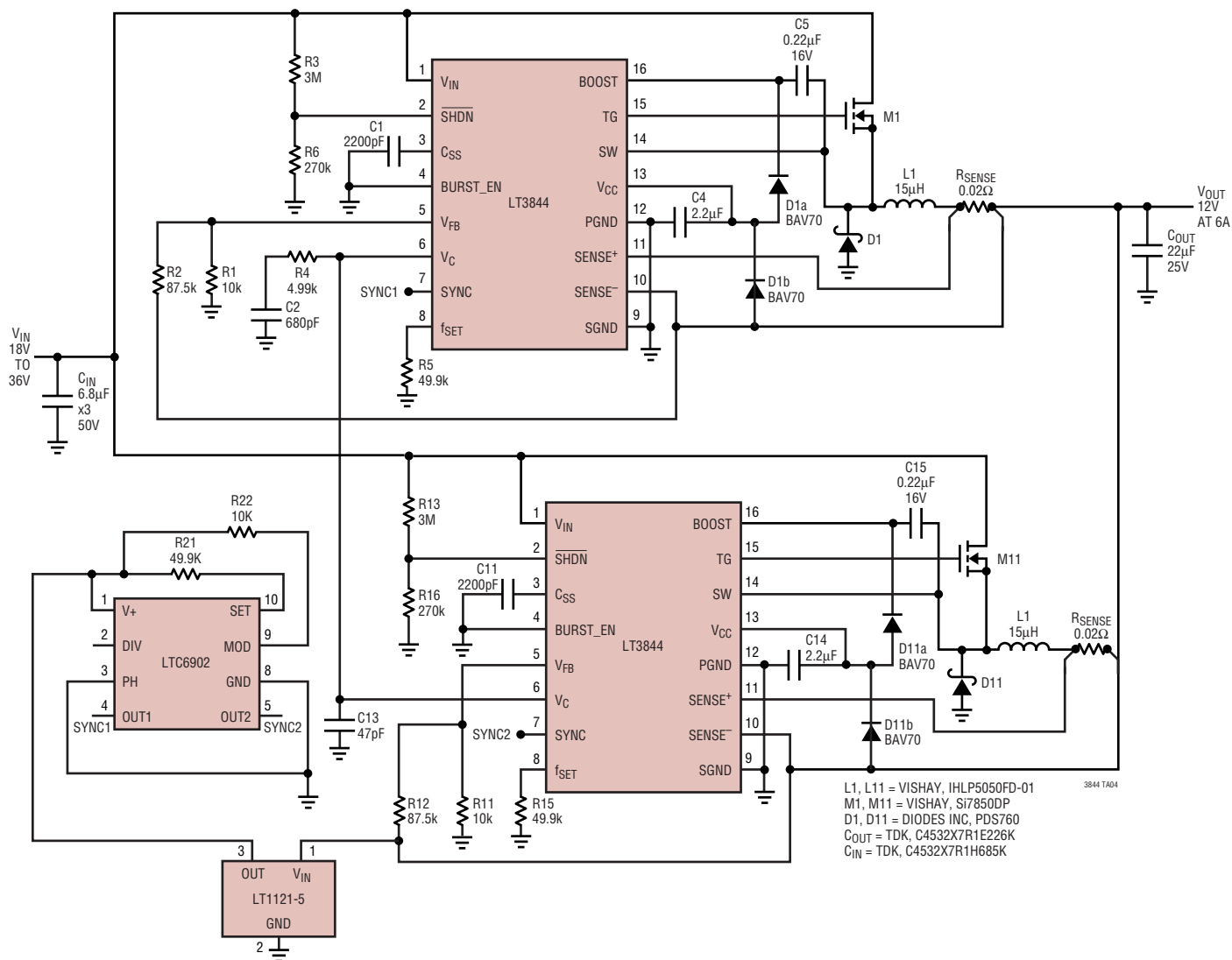


8V to 20V to 8V, 25W SEPIC Application



## TYPICAL APPLICATIONS

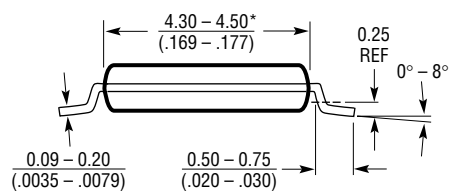
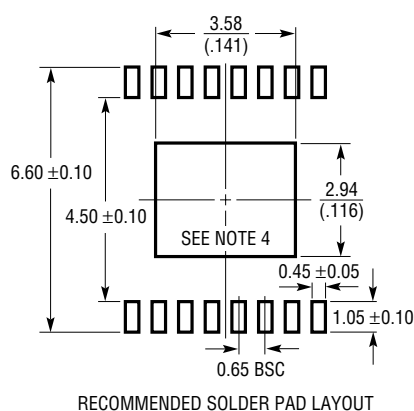
Two Phase Spread Spectrum 24V Input to 12V, 6A Output



# PACKAGE DESCRIPTION

## FE Package 16-Lead Plastic TSSOP (4.4mm) (Reference LTC DWG # 05-08-1663)

### Exposed Pad Variation BC



#### NOTE:

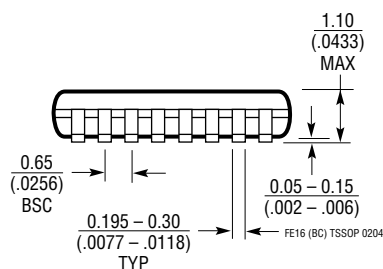
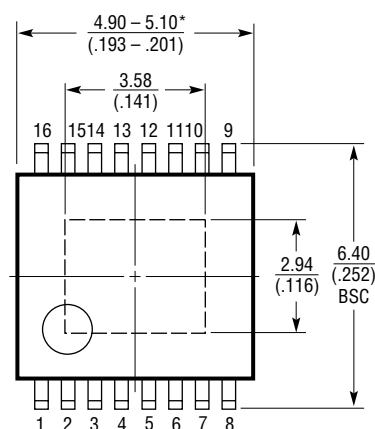
1. CONTROLLING DIMENSION: MILLIMETERS

2. DIMENSIONS ARE IN  $\frac{\text{MILLIMETERS}}{\text{(INCHES)}}$

3. DRAWING NOT TO SCALE

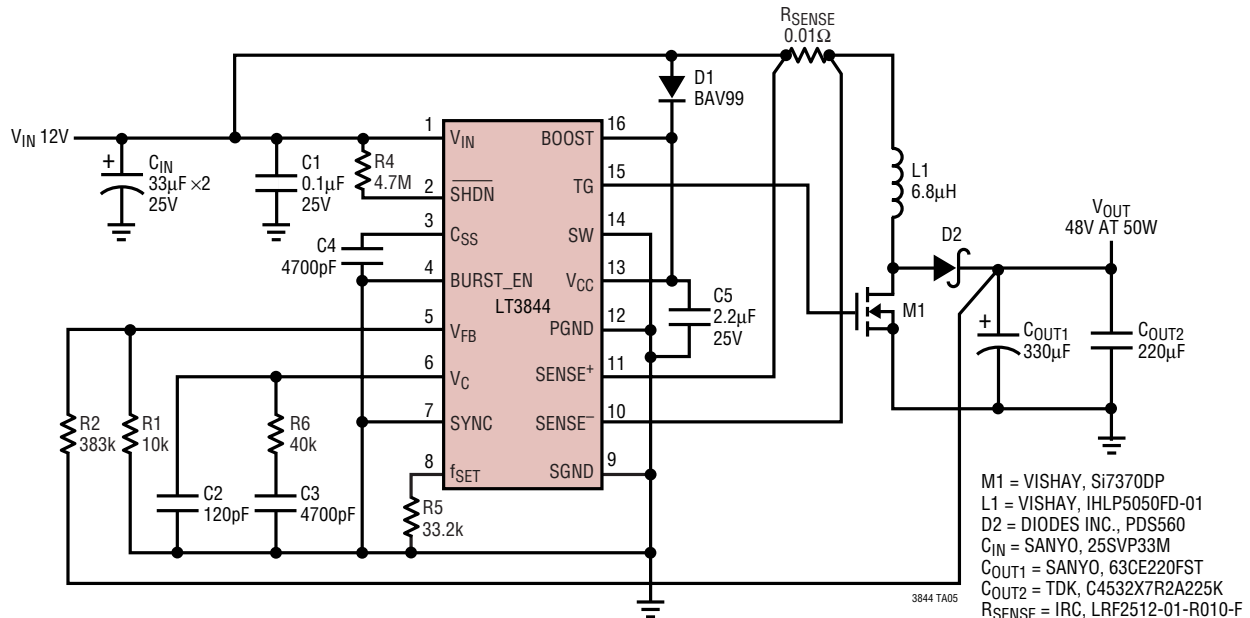
4. RECOMMENDED MINIMUM PCB METAL SIZE FOR EXPOSED PAD ATTACHMENT

\*DIMENSIONS DO NOT INCLUDE MOLD FLASH. MOLD FLASH SHALL NOT EXCEED 0.150mm (.006") PER SIDE



## TYPICAL APPLICATIONS

### 12V to 48V 50W Step Up Converter with 400kHz Switching Frequency



## RELATED PARTS

PART NUMBER	DESCRIPTION	COMMENTS
LT1339	High Power Synchronous DC/DC Controller	V <sub>IN</sub> up to 60V, Drivers 10000pF Gate Capacitance, I <sub>OUT</sub> = <20A
LTC1624	Switching Controller	Buck, Boost, SEPIC, 3.5V ≤ V <sub>IN</sub> ≤ 36V; 8-Lead SO Package
LTC1702A	Dual 2-Phase Synchronous DC/DC Controller	550kHz Operation, No R <sub>SENSE</sub> , 3V = <V <sub>IN</sub> = <7V, I <sub>OUT</sub> = <20A
LTC1735	Synchronous Step-Down DC/DC Controller	3.5V = <V <sub>IN</sub> = <36V, 0.8V = <V <sub>OUT</sub> = <6V, Current Mode, I <sub>OUT</sub> = <20A
LTC1778	No R <sub>SENSE</sub> Synchronous DC/DC Controller	4V = <V <sub>IN</sub> = <36V, Fast Transient Response, Current Mode, I <sub>OUT</sub> = <20A
LT3010	50mA, 3V to 80V Linear Regulator	1.275V = <V <sub>OUT</sub> = <60V, No Protection Diode Required, 8-Lead MSOP Package
LT3430/LT3431	Monolithic 3A, 200kHz/500kHz Step-Down Regulator	5.5V = <V <sub>IN</sub> = <60V, 0.1Ω Saturation Switch, 16-Lead SSOP Package
LTC3703/LTC3703-5	100V Synchronous Switching Regulator Controllers	No R <sub>SENSE</sub> , Voltage Mode Control, GN16 Package
LT3724	High Voltage Current Mode Switching Regulator Controllers	V <sub>IN</sub> up to 60V, I <sub>OUT</sub> ≤ 5A, 16-Lead TSSOP FE Package, On Board Bias Regulator, Burst Mode Operation, 200kHz Operation
LT3800	High Voltage Synchronous Regulator Controller	V <sub>IN</sub> up to 60V, I <sub>OUT</sub> ≤ 20A, Current Mode, On Board Bias Regulator, Burst Mode Operation, 16-Lead TSSOP FE Package