



**THE DATASHEET OF  
LTC1779ES6#TRMPBF**



# 250mA Current Mode Step-Down DC/DC Converter in ThinSOT

## FEATURES

- High Efficiency: Up to 94%
- 250mA Output Current
- Wide  $V_{IN}$  Range: 2.5V to 9.8V
- 550kHz Constant Frequency Operation
- Burst Mode™ Operation at Light Load
- Low Dropout: 100% Duty Cycle
- 0.8V Reference Allows Low Output Voltages
- $\pm 2.5\%$  Reference Accuracy
- Current Mode Operation for Excellent Line and Load Transient Response
- Low Quiescent Current: 135 $\mu$ A
- Shutdown Mode Draws Only 8 $\mu$ A Supply Current
- Low Profile (1mm) ThinSOT™ Package

## APPLICATIONS

- 1- or 2-Cell Lithium-Ion-Powered Applications
- Cellular Telephones
- Wireless Modems
- Portable Computers
- Distributed 3.3V, 2.5V or 1.8V Power Systems
- Scanners

## DESCRIPTION

The LTC1779 is a constant frequency current mode step-down DC/DC converter in a 6-lead ThinSOT package. The part operates with a 2.5V to 9.8V input and can provide up to 250mA of output current. Current mode control provides excellent AC and DC load and line regulation. The device incorporates an accurate undervoltage lockout feature that shuts down the LTC1779 when the input voltage falls below 2V.

The LTC1779 boasts a  $\pm 2.5\%$  output voltage accuracy and consumes only 135 $\mu$ A of quiescent current. For applications where efficiency is a prime consideration, the LTC1779 is configured for Burst Mode operation, which enhances efficiency at low output current.

To further maximize the life of a battery source, the internal P-channel MOSFET is turned on continuously in dropout (100% duty cycle). In shutdown, the device draws a mere 8 $\mu$ A. High constant operating frequency of 550kHz allows the use of a small external inductor.

The LTC1779 is available in a low profile (1mm) ThinSOT package.

LT, LT, LTC, LTM, Burst Mode, Linear Technology and the Linear logo are registered trademarks and ThinSOT is a trademark of Linear Technology Corporation. All other trademarks are the property of their respective owners.

## TYPICAL APPLICATION

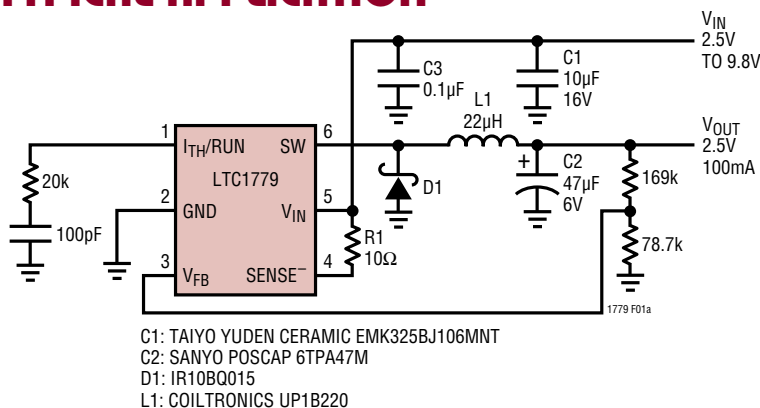
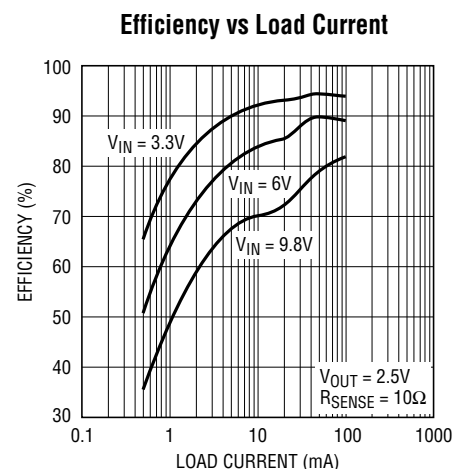


Figure 1. LTC1779 High Efficiency 2.5V/100mA Step-Down Converter



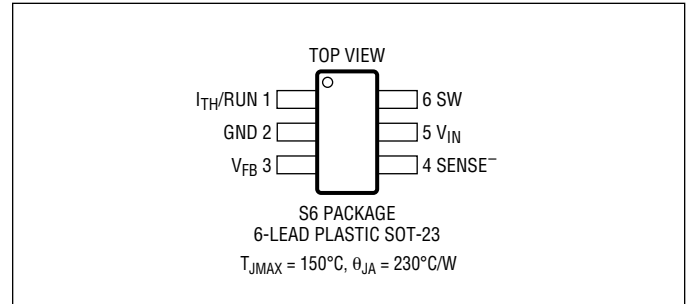
1779 F01b

## ABSOLUTE MAXIMUM RATINGS

(Note 1)

Input Supply Voltage ( $V_{IN}$ )	-0.3V to 10V
SENSE <sup>-</sup> , SW Voltages	-0.3V to ( $V_{IN} + 0.3V$ )
$V_{FB}$ , $I_{TH}/RUN$ Voltages	-0.3V to 2.4V
SW Peak Output Current (<10 $\mu$ s)	0.5A
Storage Ambient Temperature Range	-65°C to 150°C
Operating Temperature Range (Note 2)	-40°C to 85°C
Junction Temperature (Note 3)	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
LTC1779ES6#PBF	LTC1779ES6#TRPBF	LTLP	6-Lead Plastic SOT-23	-40°C to 85°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/tapeandreeel/>

## ELECTRICAL CHARACTERISTICS

The ● denotes specifications that apply over the full operating temperature range, otherwise specifications are at  $T_A = 25^\circ\text{C}$ .  $V_{IN} = 4.2\text{V}$  unless otherwise specified. (Note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Input DC Supply Current	Typicals at $V_{IN} = 4.2\text{V}$ (Note 4)				
Normal Operation	$2.5\text{V} \leq V_{IN} \leq 9.8\text{V}$		135	240	$\mu\text{A}$
Shutdown	$2.5\text{V} \leq V_{IN} \leq 9.8\text{V}$ , $V_{I\text{TH}/\text{RUN}} = 0\text{V}$		8	22	$\mu\text{A}$
UVLO	$V_{IN} < \text{UVLO Threshold}$		1	13	$\mu\text{A}$
Undervoltage Lockout Threshold	$V_{IN}$ Falling	● 1.60	2.0		V
	$V_{IN}$ Rising		2.1	2.5	V
Shutdown Threshold (at $I_{\text{TH}/\text{RUN}}$ )		● 0.15	0.325	0.5	V
Start-Up Current Source	$V_{I\text{TH}/\text{RUN}} = 0\text{V}$	0.25	0.5	0.85	$\mu\text{A}$
Regulated Feedback Voltage	(Note 5) $0^\circ\text{C}$ to $70^\circ\text{C}$	● 0.780	0.800	0.820	V
	(Note 5) $-40^\circ\text{C}$ to $85^\circ\text{C}$	● 0.770	0.800	0.830	V
Output Voltage Line Regulation	$2.5\text{V} \leq V_{IN} \leq 9.8\text{V}$ (Note 5)	-3	0	3	mV/V
Output Voltage Load Regulation	$I_{\text{TH}/\text{RUN}}$ Sinking $5\mu\text{A}$ (Note 5)		2.5		mV/ $\mu\text{A}$
	$I_{\text{TH}/\text{RUN}}$ Sourcing $5\mu\text{A}$ (Note 5)		2.5		mV/ $\mu\text{A}$
$V_{\text{FB}}$ Input Current	(Note 5)		5	25	nA
Overvoltage Protect Threshold	Measured at $V_{\text{FB}}$	0.820	0.860	0.895	V
Overvoltage Protect Hysteresis			30		mV
Overtemperature Protect Threshold			170		$^\circ\text{C}$
Overtemperature Protect Hysteresis			15		$^\circ\text{C}$
Oscillator Frequency	$V_{\text{FB}} = 0.8\text{V}$	500	550	650	kHz
	$V_{\text{FB}} = 0\text{V}$		100		kHz
$R_{\text{DS(ON)}}$ of Internal P-Channel FET	$V_{IN} = 4.2\text{V}$ , $I_{\text{SW}} = 100\text{mA}$		0.85	1.4	$\Omega$
Peak Current Sense Voltage	(Note 6)		120		mV

**Note 1:** Absolute Maximum Ratings are those values beyond which the life of a device may be impaired.

**Note 2:** The LTC1779E is guaranteed to meet performance specifications from  $0^\circ\text{C}$  to  $70^\circ\text{C}$ . Specifications over the  $-40^\circ\text{C}$  to  $85^\circ\text{C}$  operating temperature range are assured by design, characterization and correlation with statistical process controls.

**Note 3:**  $T_J$  is calculated from the ambient temperature  $T_A$  and power dissipation  $P_D$  according to the following formula:

$$T_J = T_A + (P_D \cdot \theta_J^\circ\text{C}/\text{W})$$

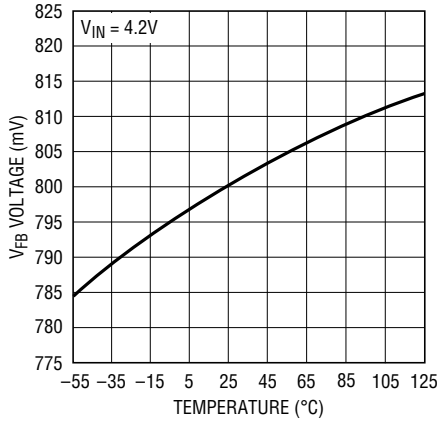
**Note 4:** Dynamic supply current is higher due to the gate charge being delivered at the switching frequency.

**Note 5:** The LTC1779 is tested in a feedback loop that servos  $V_{\text{FB}}$  to the output of the error amplifier.

**Note 6:** Peak current sense voltage is reduced dependent upon duty cycle to a percentage of value as given in Figure 2.

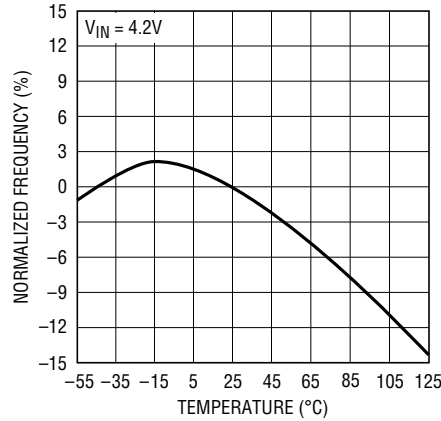
## TYPICAL PERFORMANCE CHARACTERISTICS

**Reference Voltage vs Temperature**



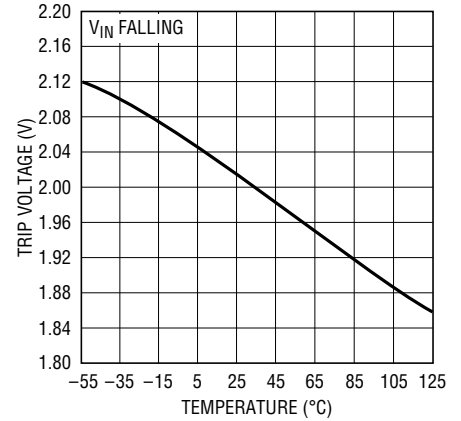
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**Normalized Oscillator Frequency vs Temperature**



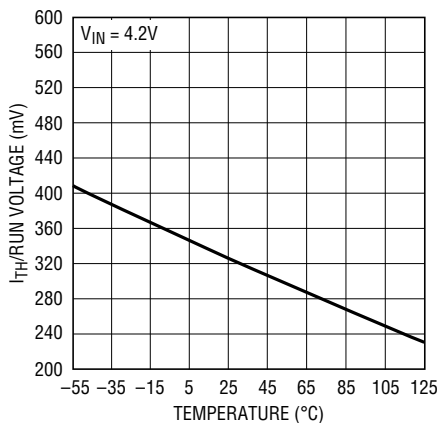
1779 G02

**Undervoltage Lockout Trip Voltage vs Temperature**



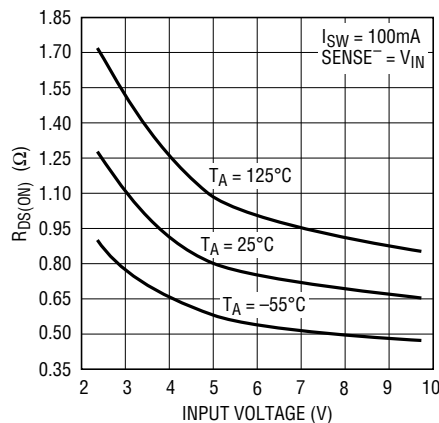
1779 G03

**Shutdown Threshold vs Temperature**



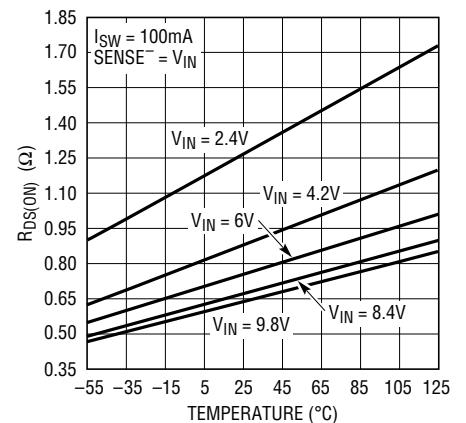
1779 G04

**R<sub>DS(ON)</sub> of Internal P-Channel FET vs Input Voltage**



1779 G05

**R<sub>DS(ON)</sub> of Internal P-Channel FET vs Temperature**



1779 G06

## PIN FUNCTIONS

**I<sub>TH</sub>/RUN (Pin 1):** This pin performs two functions. It serves as the error amplifier compensation point as well as the run control input. The current comparator threshold increases with this control voltage. Nominal voltage range for this pin is 0.7V to 1.9V. Forcing this pin below 0.325V causes the device to be shut down. In shutdown all functions are disabled and the internal P-channel MOSFET is turned off. The SW pin will be high impedance.

**GND (Pin 2):** Ground Pin.

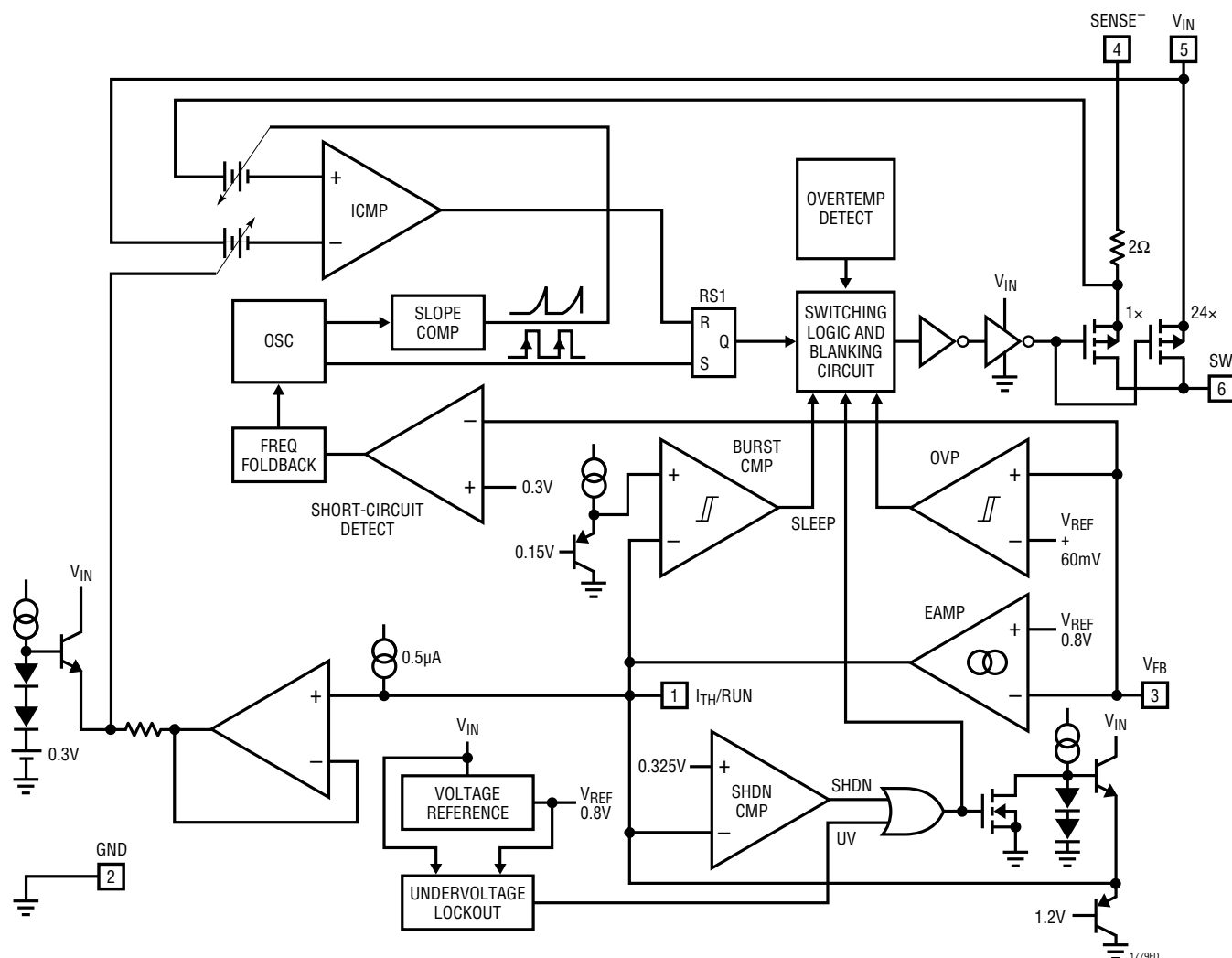
**V<sub>FB</sub> (Pin 3):** Receives the feedback voltage from an external resistive divider across the output.

**SENSE<sup>-</sup> (Pin 4):** The Negative Input to the Current Comparator. Can be connected to V<sub>IN</sub> for default minimum peak current of 250mA. Connecting a resistor between SENSE<sup>-</sup> and V<sub>IN</sub> specifies a lower peak current. (See Applications Information for specifying resistor value.)

**V<sub>IN</sub> (Pin 5):** Supply Pin. Must be closely decoupled to GND Pin 2.

**SW (Pin 6):** Switching Node and Drain of Internal P-Channel Power MOSFET. Connects to external inductor and catch diode.

## FUNCTIONAL DIAGRAM



## OPERATION (Refer to Functional Diagram)

## Main Control Loop

The LTC1779 is a constant frequency current mode switching regulator. During normal operation, the internal P-channel power MOSFET is turned on each cycle when the oscillator sets the RS latch (RS1) and turned off when the current comparator (ICMP) resets the latch. The peak inductor current at which ICMP resets the RS latch is controlled by the voltage on the  $I_{TH}/RUN$  pin, which is the output of the error amplifier EAMP. An external resistive divider connected between  $V_{OUT}$  and ground allows the EAMP to receive an output feedback voltage  $V_{FB}$ . When the load current increases, it causes a slight decrease in

$V_{FB}$  relative to the 0.8V reference, which in turn causes the  $I_{TH}/RUN$  voltage to increase until the average inductor current matches the new load current.

The main control loop is shut down by pulling the  $I_{TH}/RUN$  pin low. Releasing  $I_{TH}/RUN$  allows an internal  $0.5\mu A$  current source to charge up the external compensation network. When the  $I_{TH}/RUN$  pin reaches 325mV, the main control loop is enabled with the  $I_{TH}/RUN$  voltage then pulled up to its zero current level of approximately 0.7V. As the external compensation network continues to charge up, the corresponding output current trip level follows, allowing normal operation.

## OPERATION (Refer to Functional Diagram)

Comparator OVP guards against transient overshoots >7.5% by turning off the internal P-channel power MOSFET and keeping it off until the fault is removed.

### Burst Mode Operation

The LTC1779 enters Burst Mode operation at low load currents. In this mode, the peak current of the inductor is set as if  $V_{I_{TH}/RUN} = 1V$  (at low duty cycles) even though the voltage at the  $I_{TH}/RUN$  pin is at a lower value. If the inductor's average current is greater than the load requirement, the voltage at the  $I_{TH}/RUN$  pin will drop. When the  $I_{TH}/RUN$  voltage goes below 0.85V, the sleep signal goes high, turning off the internal MOSFET. The sleep signal goes low when the  $I_{TH}/RUN$  voltage goes above 0.925V and the LTC1779 resumes normal operation. The next oscillator cycle will turn the internal MOSFET on and the switching cycle repeats.

### Dropout Operation

When the input supply voltage decreases towards the output voltage, the rate of change of inductor current during the ON cycle decreases. This reduction means that the internal P-channel MOSFET will remain on for more than one oscillator cycle since the inductor current has not ramped up to the threshold set by EAMP. Further reduction in input supply voltage will eventually cause the P-channel MOSFET to be turned on 100%, i.e., DC. The output voltage will then be determined by the input voltage minus the voltage drop across the MOSFET, the sense resistor and the inductor.

### Undervoltage Lockout

To prevent operation of the P-channel MOSFET below safe input voltage levels, an undervoltage lockout is incorporated into the LTC1779. When the input supply voltage drops below approximately 2.0V, the P-channel MOSFET and all circuitry is turned off except the undervoltage block, which draws only several microamperes.

### Short-Circuit Protection

When the output is shorted to ground, the frequency of the oscillator will be reduced to about 100kHz. This lower frequency allows the inductor current to safely discharge, thereby preventing current runaway. The oscillator's frequency will gradually increase to its designed rate when the feedback voltage again approaches 0.8V.

### Overvoltage Protection

As a further protection, the overvoltage comparator in the LTC1779 will turn the internal MOSFET off when the feedback voltage has risen 7.5% above the reference voltage of 0.8V. This comparator has a typical hysteresis of 30mV.

### Slope Compensation and Inductor's Peak Current

The inductor's peak current is determined by:

$$I_{PK} = \frac{M(V_{I_{TH}/RUN} - 0.7)}{10(R_{SENSE} + 2\Omega)}$$

when the LTC1779 is operating below 40% duty cycle. However, once the duty cycle exceeds 40%, slope compensation begins and effectively reduces the peak inductor current. The amount of reduction is given by the curves in Figure 2.

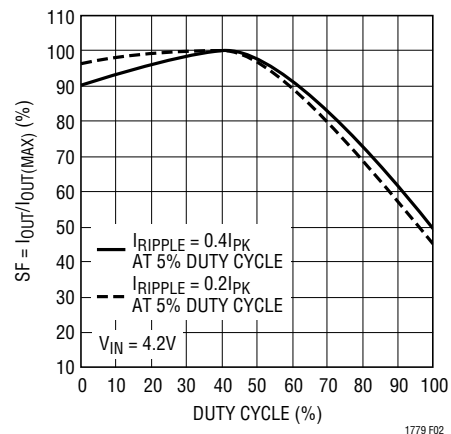


Figure 2. Maximum Output Current vs Duty Cycle

## OPERATION (Refer to Functional Diagram)

The variable  $M$  is the ratio of the total switch current to that portion of the switch current that flows through  $R_{SENSE}$ .  $M$  is a function of both  $R_{SENSE}$  and  $R_{OUT}$  of the internal power switch, which in turn, is a strong function of supply voltage. For values of  $M$  refer to Figure 3. In order to guarantee the desired  $I_{PK}$  over the full range of supply voltage, the minimum value of  $M$ , corresponding to the minimum supply voltage seen in the application, should be chosen. Note that the selection of  $R_{SENSE}$ , and hence the resulting  $M$ , is an iterative process. For most applications, a value of  $R_{SENSE}$  between  $0\Omega$  and  $20\Omega$  will be chosen.

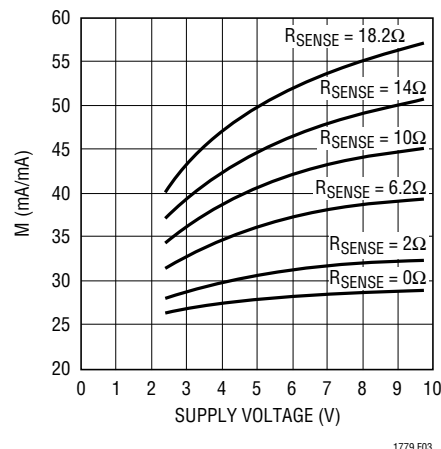


Figure 3.  $M$  vs Supply Voltage

## APPLICATIONS INFORMATION

The basic LTC1779 application circuit is shown in Figure 1. External component selection is driven by the load requirement and begins with the selection of  $L1$  and  $R_{SENSE}$  ( $= R1$ ). Next, the output diode  $D1$  is selected followed by  $C_{IN}$  ( $= C1$ ) and  $C_{OUT}$  ( $= C2$ ).

### Inductor Value Calculation

The inductance value has a direct effect on ripple current. The ripple current,  $I_{RIPPLE}$ , decreases with higher inductance or frequency and increases with higher  $V_{IN}$ . The inductor's peak-to-peak ripple current is given by:

$$I_{RIPPLE} = \frac{(V_{IN} - V_{OUT}) V_{OUT}}{V_{IN} \cdot f \cdot L}$$

where  $f$  is the operating frequency fixed at 550kHz in the LTC1779.

A smaller value of  $L$  results in higher current ripple and output voltage ripple as well as greater core losses. Larger values of  $L$  decrease the ripple, but require finding physically larger inductors since maximum DC current rating decreases significantly as inductance increases within inductor product types. Generally, by choosing the desired ripple current based on the maximum output current, the inductor value can be calculated from the previous equation. It is typical to choose the inductor so that the ripple current is about 40% of the maximum output current at

maximum input voltage. Use the following equations to calculate  $L$ :

$$I_{RIPPLE} = 0.4 \cdot I_{OUT(MAX)}$$

$$L = \frac{(V_{IN(MAX)} - V_{OUT}) \cdot V_{OUT}}{V_{IN(MAX)} \cdot f \cdot I_{RIPPLE}}$$

$$I_{L(MAX)} = I_{OUT(MAX)} + \frac{I_{RIPPLE}}{2}$$

and then choose an appropriate  $L$  and recalculate the ripple current.

In Burst Mode operation on the LTC1779, the ripple current is normally set such that the inductor current is continuous during the burst periods. Therefore, the peak-to-peak ripple current must not exceed:

$$I_{RIPPLE} \leq \frac{M(0.030)}{(R_{SENSE} + 2\Omega)}$$

This implies a minimum inductance of:

$$L_{MIN} = \frac{V_{IN} - V_{OUT}}{f \left( \frac{M(0.030)}{(R_{SENSE} + 2\Omega)} \right)} \left( \frac{V_{OUT} + V_D}{V_{IN} + V_D} \right)$$

(Use  $V_{IN(MAX)} = V_{IN}$ )

## APPLICATIONS INFORMATION

A smaller value than  $L_{MIN}$  could be used in the circuit; however, the inductor current will not be continuous during burst periods.

### $R_{SENSE}$ Selection for Output Current

The selection of  $R_{SENSE}$  determines the output current limit, the maximum possible output current before the internal current limit threshold is reached.  $I_{OUT(MAX)}$ , the maximum specified output current in a design, must be less than  $I_{CL}$ . With the current comparator monitoring the voltage developed across  $R_{SENSE}$ , the threshold of the comparator determines the inductor's peak current. The maximum output current,  $I_{CL}$ , the LTC1779 can provide is given by:

$$I_{CL} = M \left( \frac{SF}{100} \right) \left( \frac{0.12V}{R_{SENSE} + 2\Omega} \right) - \frac{I_{RIPPLE}}{2}$$

where SF and M are as defined in the previous section, Figures 2 and 3. Typically,  $R_{SENSE}$  is chosen between  $0\Omega$  and  $20\Omega$ . Current limit is at a minimum at minimum input voltage and maximum at maximum input voltage. Both conditions should be considered in a design where current limit is important.

To calculate several current limit conditions and choose the best sense resistor for your design, first use minimum input voltage. Calculate the duty cycle at minimum input voltage.

$$DC = \frac{V_{OUT}}{V_{IN(MIN)}}$$

Choose the slope factor, SF, from Figure 2 based on the duty cycle. The ripple current calculated at minimum input voltage and the chosen L should be used in the current limit equation (see Inductor Value Calculation). Figure 3 provides several values of  $R_{SENSE}$  and their corresponding M values at different input voltages. Select the minimum input voltage and calculate the resulting minimum current limit settings.

The process must be repeated for maximum current limit using duty cycle, slope factor, ripple current and mirror ratio based on *maximum* input voltage in order to choose the best sense resistor for a particular design

and to understand how it is going to work over the entire input voltage range.

### Inductor Core Selection

Once the value for L is known, the type of inductor must be selected. High efficiency converters generally cannot afford the core loss found in low cost powdered iron cores, forcing the use of more expensive ferrite, molypermalloy or Kool Mu® cores. Actual core loss is independent of core size for a fixed inductor value, but it is very dependent on inductance selected. As inductance increases, core losses go down. Unfortunately, increased inductance requires more turns of wire and therefore copper losses will increase. Ferrite designs have very low core losses and are preferred at high switching frequencies, so design goals can concentrate on copper loss and preventing saturation. Ferrite core material saturates "hard," which means that inductance collapses abruptly when the peak design current is exceeded. This results in an abrupt increase in inductor ripple current and consequent output voltage ripple. Do not allow the core to saturate!

Molypermalloy (from Magnetics, Inc.) is a very good, low loss core material for toroids, but it is more expensive than ferrite. A reasonable compromise from the same manufacturer is Kool Mu. Toroids are very space efficient, especially when you can use several layers of wire. Because they generally lack a bobbin, mounting is more difficult. However, new designs for surface mount that do not increase the height significantly are available.

### Output Diode Selection

The catch diode carries load current during the off-time. The average diode current is therefore dependent on the internal P-channel switch duty cycle. At high input voltages the diode conducts most of the time. As  $V_{IN}$  approaches  $V_{OUT}$  the diode conducts only a small fraction of the time. The most stressful condition for the diode is when the output is short-circuited. Under this condition the diode must safely handle  $I_{PK}$  at close to 100% duty cycle. Therefore, it is important to adequately specify the diode peak current and average power dissipation so as not to exceed the diode ratings.

Kool Mu is a registered trademark of Magnetics, Inc.

## APPLICATIONS INFORMATION

Under normal load conditions, the average current conducted by the diode is:

$$I_D = \left( \frac{V_{IN} - V_{OUT}}{V_{IN} + V_D} \right) I_{OUT}$$

The allowable forward voltage drop in the diode is calculated from the maximum short-circuit current as:

$$V_F \approx \frac{P_D}{I_{CL(MAX)}}$$

where  $P_D$  is the allowable power dissipation and will be determined by efficiency and/or thermal requirements.

A fast switching diode must also be used to optimize efficiency. Schottky diodes are a good choice for low forward drop and fast switching times. Remember to keep lead length short and observe proper grounding (see Board Layout Checklist) to avoid ringing and increased dissipation.

### $C_{IN}$ and $C_{OUT}$ Selection

In continuous mode, the source current of the internal P-channel MOSFET is a square wave of duty cycle  $(V_{OUT} + V_D)/(V_{IN} + V_D)$ . To prevent large voltage transients, a low ESR input capacitor sized for the maximum RMS current must be used. The maximum RMS capacitor current is given by:

$$C_{IN} \text{ Required } I_{RMS} \approx I_{MAX} \frac{[V_{OUT}(V_{IN} - V_{OUT})]^{1/2}}{V_{IN}}$$

This formula has a maximum at  $V_{IN} = 2V_{OUT}$ , where  $I_{RMS} = I_{OUT}/2$ . This simple worst-case condition is commonly used for design because even significant deviations do not offer much relief. Note that capacitor manufacturer's ripple current ratings are often based on 2000 hours of life. This makes it advisable to further derate the capacitor, or to choose a capacitor rated at a higher temperature than required. Several capacitors may be paralleled to meet the size or height requirements in the design. Due to the high operating frequency of the LTC1779, ceramic capacitors can also be used for  $C_{IN}$ . Always consult the manufacturer if there is any question.

The selection of  $C_{OUT}$  is driven by the required effective series resistance (ESR). Typically, once the ESR requirement is satisfied, the capacitance is adequate for filtering. The output ripple ( $\Delta V_{OUT}$ ) is approximated by:

$$\Delta V_{OUT} \approx I_{RIPPLE} \left( ESR + \frac{1}{8fC_{OUT}} \right)$$

where  $f$  is the operating frequency,  $C_{OUT}$  is the output capacitance and  $I_{RIPPLE}$  is the ripple current in the inductor. The output ripple is highest at maximum input voltage since  $\Delta I_L$  increases with input voltage.

Manufacturers such as Nichicon, United Chemicon and Sanyo should be considered for high performance through-hole capacitors. The OS-CON semiconductor dielectric capacitor available from Sanyo has the lowest ESR (size) product of any aluminum electrolytic at a somewhat higher price. Once the ESR requirement for  $C_{OUT}$  has been met, the RMS current rating generally far exceeds the  $I_{RIPPLE(P-P)}$  requirement.

In surface mount applications, multiple capacitors may have to be paralleled to meet the ESR or RMS current handling requirements of the application. Aluminum electrolytic and dry tantalum capacitors are both available in surface mount configurations. In the case of tantalum, it is critical that the capacitors are surge tested for use in switching power supplies. An excellent choice is the AVX TPS, AVX TPSV and KEMET T510 series of surface mount tantalum, available in case heights ranging from 2mm to 4mm. Other capacitor types include Sanyo OS-CON, Nichicon PL series and Panasonic SP.

### Low Supply Operation

Although the LTC1779 can function down to approximately 2.0V, the maximum allowable output current is reduced when  $V_{IN}$  decreases below 3V. Figure 3 shows the amount of change as the supply is reduced down to 2V. Also shown in Figure 4 is the effect of  $V_{IN}$  on  $V_{REF}$  as  $V_{IN}$  goes below 2.3V.

## APPLICATIONS INFORMATION

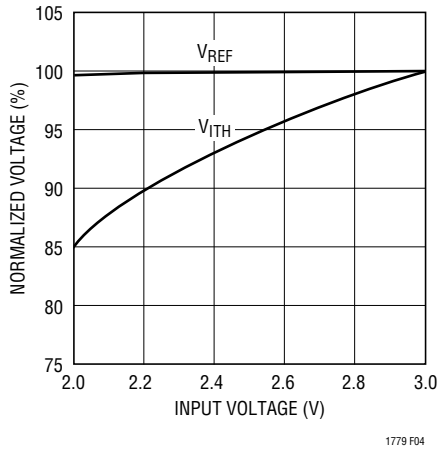


Figure 4. Line Regulation of V<sub>REF</sub> and V<sub>ITH</sub>

### Setting Output Voltage

The LTC1779 develops a 0.8V reference voltage between the feedback (Pin 3) terminal and ground (see Figure 5). By selecting resistor R1, a constant current is caused to flow through R1 and R2 to set the overall output voltage. The regulated output voltage is determined by:

$$V_{OUT} = 0.8 \left( 1 + \frac{R2}{R1} \right)$$

For most applications, an 80k resistor is suggested for R1. To prevent stray pickup, locate resistors R1 and R2 close to LTC1779.

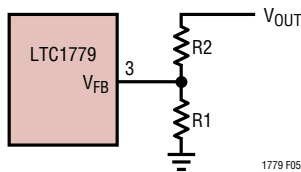


Figure 5. Setting Output Voltage

### Efficiency Considerations

The efficiency of a switching regulator is equal to the output power divided by the input power times 100%. It is often useful to analyze individual losses to determine what is limiting the efficiency and which change would produce the most improvement. Efficiency can be expressed as:

$$\text{Efficiency} = 100\% - (\eta_1 + \eta_2 + \eta_3 + \dots)$$

where  $\eta_1$ ,  $\eta_2$ , etc. are the individual losses as a percentage of input power.

Although all dissipative elements in the circuit produce losses, four main sources usually account for most of the losses in LTC1779 circuits: 1) LTC1779 DC bias current, 2) MOSFET gate charge current, 3)  $I^2R$  losses and 4) voltage drop of the output diode.

1. The  $V_{IN}$  current is the DC supply current, given in the electrical characteristics, that excludes MOSFET driver and control currents.  $V_{IN}$  current results in a small loss which increases with  $V_{IN}$ .
2. MOSFET gate charge current results from switching the gate capacitance of the internal power MOSFET. Each time the MOSFET gate is switched from low to high to low again, a packet of charge  $dQ$  moves from  $V_{IN}$  to ground. The resulting  $dQ/dt$  is a current out of  $V_{IN}$  which is typically much larger than the DC supply current. In continuous mode,  $I_{GATECHG} = f(Qp)$ .
3.  $I^2R$  losses are predicted from the DC resistances of the internal MOSFET, inductor and current shunt. In continuous mode the average output current flows through L but is “chopped” between the internal P-channel MOSFET in series with  $R_{SENSE}$  and the output diode. The MOSFET  $R_{DS(ON)}$  plus  $R_{SENSE}$  multiplied by duty cycle can be summed with the resistances of L and  $R_{SENSE}$  to obtain  $I^2R$  losses.
4. The output diode is a major source of power loss at high currents and gets worse at high input voltages. The diode loss is calculated by multiplying the forward voltage times the diode duty cycle multiplied by the load current. For example, assuming a duty cycle of 50% with a Schottky diode forward voltage drop of 0.4V, the loss increases from 0.5% to 8% as the load current increases from 0.5A to 2A.

5. Transition losses apply to the internal MOSFET and increase at higher operating frequencies and input voltages. Transition losses can be estimated from:

$$\text{Transition Loss} = 2(V_{IN})^2 I_{O(MAX)} C_{RSS}(f)$$

Other losses including  $C_{IN}$  and  $C_{OUT}$  ESR dissipative losses, and inductor core losses, generally account for less than 2% total additional loss.

## APPLICATIONS INFORMATION

### Foldback Current Limiting

As described in the Output Diode Selection, the worst-case dissipation occurs with a short-circuited output when the diode conducts the current limit value almost continuously. To prevent excessive heating in the diode, foldback current limiting can be added to reduce the current in proportion to the severity of the fault.

Foldback current limiting is implemented by adding diodes  $D_{FB1}$  and  $D_{FB2}$  between the output and the  $I_{TH}/RUN$  pin as shown in Figure 6. In a hard short ( $V_{OUT} = 0V$ ), the current will be reduced to approximately 50% of the maximum output current.

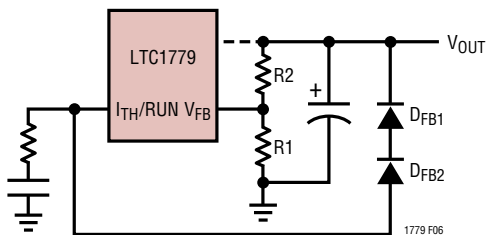


Figure 6. Foldback Current Limiting

### PC Board Layout Checklist

When laying out the printed circuit board, the following checklist should be used to ensure proper operation of the LTC1779. These items are illustrated graphically in the layout diagram in Figure 7. Check the following in your layout:

1. Large switch currents flow into the input capacitor  $C_{IN}$ , the power switch and the Schottky diode D1. The loop formed by these components should be as small as possible.
2. Is the input decoupling capacitor ( $0.1\mu F$ ) connected closely between  $V_{IN}$  (Pin 5) and ground (Pin 2)?
3. Keep the switching node SW away from sensitive small signal nodes.
4. Does the  $V_{FB}$  pin connect directly to the feedback resistors? The resistive divider R1 and R2 must be connected between the (+) plate of  $C_{OUT}$  and signal ground. Locate R1 and R2 close to the  $V_{FB}$  pin.

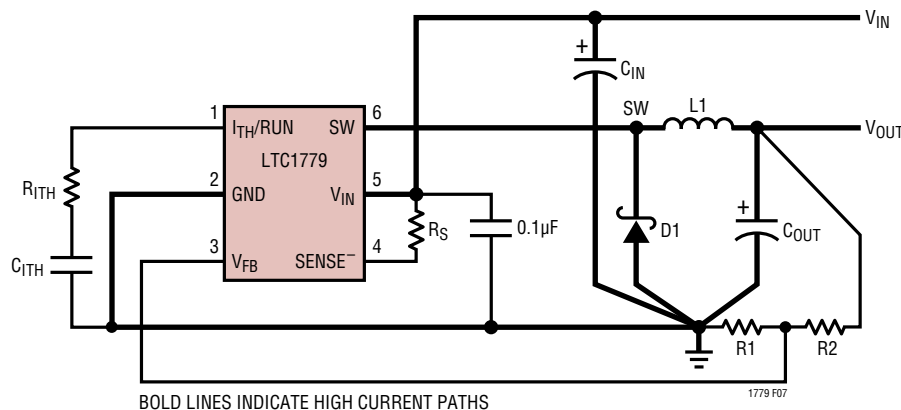
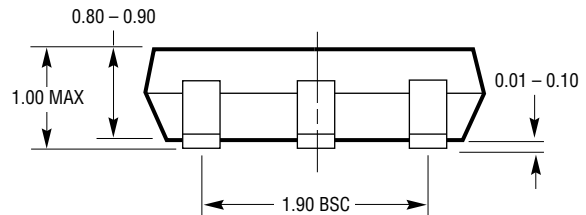
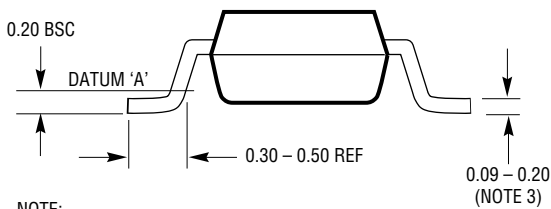
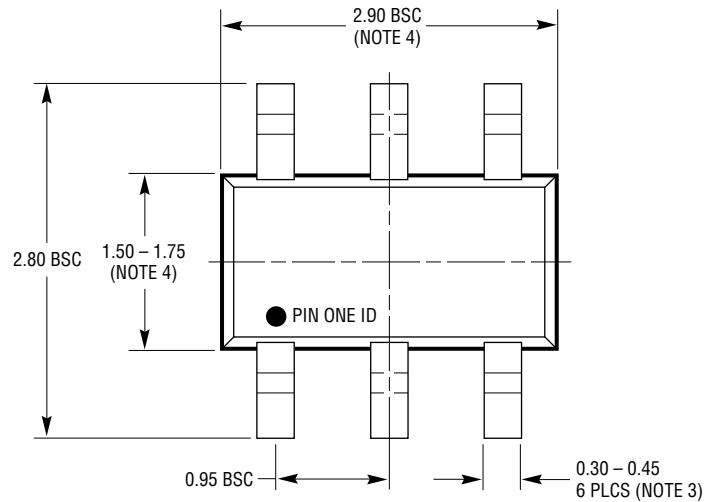
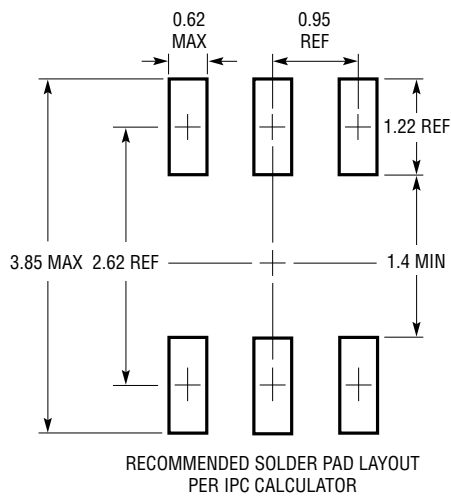


Figure 7. LTC1779 Layout Diagram (See PC Board Layout Checklist)

## PACKAGE DESCRIPTION

Please refer to <http://www.linear.com/designtools/packaging/> for the most recent package drawings.

### S6 Package 6-Lead Plastic TSOT-23 (Reference LTC DWG # 05-08-1636)



S6 TSOT-23 0302

- NOTE:
1. DIMENSIONS ARE IN MILLIMETERS
  2. DRAWING NOT TO SCALE
  3. DIMENSIONS ARE INCLUSIVE OF PLATING
  4. DIMENSIONS ARE EXCLUSIVE OF MOLD FLASH AND METAL BURR
  5. MOLD FLASH SHALL NOT EXCEED 0.254mm
  6. JEDEC PACKAGE REFERENCE IS MO-193

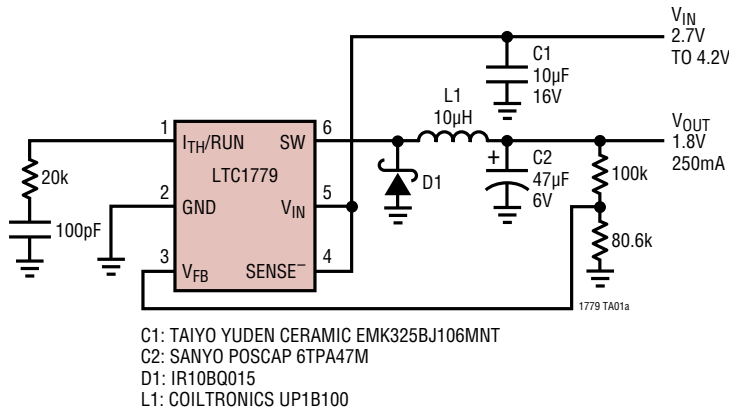
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## REVISION HISTORY

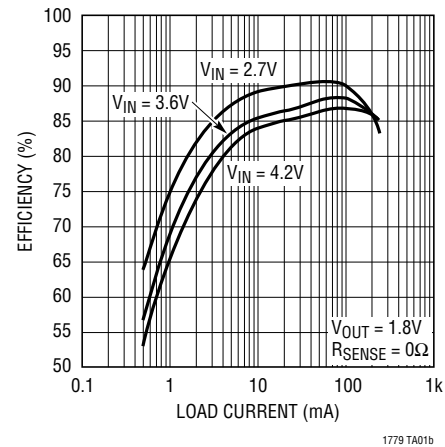
REV	DATE	DESCRIPTION	PAGE NUMBER
A	09/15	Revised package drawing	12

## TYPICAL APPLICATION

### LTC1779 Minimal Component Count, Single Li-Ion to 1.8V/250mA Step-Down Converter



### Efficiency vs Load Current



## RELATED PARTS

PART NUMBER	DESCRIPTION	COMMENTS
<a href="#">LT®1375/LT1376</a>	1.5A, 500kHz Step-Down Switching Regulators	High Frequency, Small Inductor, High Efficiency
<a href="#">LT1616</a>	600mA Step-Down Switching Regulator	1.4MHz, 4V to 25V Input, ThinSOT Package
<a href="#">LTC1624</a>	High Efficiency SO-8 N-Channel Switching Regulator Controller	8-Pin N-Channel Drive, 3.5V ≤ V <sub>IN</sub> ≤ 36V
<a href="#">LTC1625</a>	No R <sub>SENSE</sub> ™ Synchronous Step-Down Regulator	High Efficiency, No Sense Resistor
<a href="#">LTC1627</a>	Low Voltage, Monolithic Synchronous Step-Down Regulator	Low Supply Voltage Range: 2.65V to 8V, I <sub>OUT</sub> = 0.5A
<a href="#">LT1676/LT1776</a>	Wide Input Range Step-Down Switching Regulators	60V Input, 700mA Internal Switches
<a href="#">LTC1735</a>	Single, High Efficiency, Low Noise Synchronous Switching Controller	High Efficiency 5V to 3.3V Conversion at up to 15A
<a href="#">LT1767</a>	1.5A, 1.4MHz Step-Down DC/DC Converter	Higher Current, 8-Lead MSOP Package
<a href="#">LTC1771</a>	Ultralow Supply Current Step-Down DC/DC Controller	10µA I <sub>Q</sub> , 93% Efficiency, 1.23V ≤ V <sub>OUT</sub> ≤ 18V, 2.8V ≤ V <sub>IN</sub> ≤ 20V
<a href="#">LTC1772</a>	Constant Frequency Current Mode Step-Down DC/DC Controller	V <sub>IN</sub> = 2.5V to 9.8V, I <sub>OUT</sub> Up to 2A, ThinSOT Package
<a href="#">LTC1773</a>	95% Efficient Synchronous Step-Down Controller	2.65V ≤ V <sub>IN</sub> ≤ 8.5V, 0.8V ≤ V <sub>OUT</sub> ≤ V <sub>IN</sub> , Current Mode, 550kHz
<a href="#">LTC1877</a>	High Efficiency Monolithic Step-Down Regulator	550kHz, MS8, V <sub>IN</sub> Up to 10V, I <sub>Q</sub> = 10µA, I <sub>OUT</sub> to 600mA at V <sub>IN</sub> = 5V
<a href="#">LTC1878</a>	High Efficiency Monolithic Step-Down Regulator	550kHz, MS8, V <sub>IN</sub> Up to 6V, I <sub>Q</sub> = 10µA, I <sub>OUT</sub> to 600mA at V <sub>IN</sub> = 3.3V
<a href="#">LTC3400</a>	1.2MHz Synchronous Step-Up DC/DC Converter in ThinSOT	92% Efficiency, V <sub>IN</sub> = 0.5V to 6V, V <sub>OUT</sub> = 2.6V to 5V
<a href="#">LTC3401</a>	Single Cell, High Current (1A), Micropower, Synchronous 3MHz Step-Up DC/DC Converter	V <sub>IN</sub> = 0.5V to 5V, Up to 97% Efficiency Synchronizable Oscillator from 100kHz to 3MHz
<a href="#">LTC3402</a>	Single Cell, High Current (2A), Micropower, Synchronous 3MHz Step-Up DC/DC Converter	V <sub>IN</sub> = 0.7V to 5V, Up to 95% Efficiency Synchronizable Oscillator from 100kHz to 3MHz
<a href="#">LTC3404</a>	1.4MHz High Efficiency, Monolithic Synchronous Step-Down Regulator	Up to 95% Efficiency, 100% Duty Cycle, I <sub>Q</sub> = 10µA, V <sub>IN</sub> = 2.65V to 6V

No R<sub>SENSE</sub> is a trademark of Linear Technology Corporation.

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