LTC1779
250mA Current Mode Step-Down DC/DC Converter in ThinSOT

FEATURES

- High Efficiency: Up to 94%
- 250mA Output Current
- Wide VIN 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
- ±2.5% Reference Accuracy
- Current Mode Operation for Excellent Line and Load Transient Response
- Low Quiescent Current: 135µA
- Shutdown Mode Draws Only 8µA Supply Current
- Low Profile (1mm) ThinSOT™ Package

DESCRIPTION

The LTC®-1779 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 ±2.5% output voltage accuracy and consumes only 135µA 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 8µA. 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.

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

TYPICAL APPLICATION

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

### ABSOLUTE MAXIMUM RATINGS

(Not 1)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Conditions</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Units</th>
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<td>Input Supply Voltage ( (V_{IN}) )</td>
<td>(-0.3V \text{ to } 10V)</td>
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<td>SENSE(^-), SW Voltages</td>
<td>(-0.3V \text{ to } (V_{IN} + 0.3V))</td>
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<tr>
<td>(V_{FB}, I_{TH/RUN} ) Voltages</td>
<td>(-0.3V \text{ to } 2.4V)</td>
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<td>SW Peak Output Current (&lt;10\mu s)</td>
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<td>Storage Ambient Temperature Range</td>
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<tr>
<td>Operating Temperature Range (Note 2)</td>
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<td>Junction Temperature (Note 3)</td>
<td>150(^\circ C)</td>
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<tr>
<td>Lead Temperature (Soldering, 10 sec)</td>
<td>300(^\circ C)</td>
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**PACKAGE/ORDER INFORMATION**

<table>
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<tr>
<th>ORDER PART NUMBER</th>
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<td>S6 PART MARKING</td>
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<tr>
<td>LTLP</td>
<td>LTLP</td>
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</table>

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

### ELECTRICAL CHARACTERISTICS

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

**PARAMETER**

**CONDITIONS**

**MIN** | **TYP** | **MAX** | **UNITS**

| Input DC Supply Current | Typicals at \(V_{IN} = 4.2V\) (Note 4) | 135 | 240 | \(\mu A\) |
| Normal Operation | \(2.5V \leq V_{IN} \leq 9.8V\) | 8 | 22 | \(\mu A\) |
| Shutdown | \(2.5V \leq V_{IN} \leq 9.8V, V_{TH/RUN} = 0V\) | 7 | 13 | \(\mu A\) |
| Undervoltage Lockout Threshold | \(V_{IN}\) Falling | 1.60 | 2.0 | \(V\) |
| | \(V_{IN}\) Rising | 2.1 | 2.5 | \(V\) |
| Shutdown Threshold (at \(I_{TH/RUN}\)) | \(V_{TH/RUN} = 0V\) | 0.25 | 0.5 | 0.85 | \(\mu A\) |
| Start-Up Current Source | \(V_{FB} = 0V\) | 0.780 | 0.800 | 0.820 | \(V\) |
| Regulated Feedback Voltage | \(V_{FB} = 0V\) | 0.780 | 0.800 | 0.830 | \(V\) |
| Output Voltage Line Regulation | \(2.5V \leq V_{IN} \leq 9.8V\) (Note 5) | \(-3\) | 0 | 3 | \(mV/V\) |
| Output Voltage Load Regulation | \(I_{TH/RUN}\) Sinking 5\(\mu A\) (Note 5) | 2.5 | | | \(mV/\mu A\) |
| | \(I_{TH/RUN}\) Sourcing 5\(\mu A\) (Note 5) | 2.5 | | | \(mV/\mu A\) |
| \(V_{FB}\) Input Current | (Note 5) | 5 | 25 | nA |
| Overvoltage Protect Threshold | Measured at \(V_{FB}\) | 0.820 | 0.860 | 0.895 | \(V\) |
| Overvoltage Protect Hysteresis | 30 | | | \(mV\) |
| Overtemperature Protect Threshold | 170 | | | \(^\circ C\) |
| Overtemperature Protect Hysteresis | 15 | | | \(^\circ C\) |
| Oscillator Frequency | \(V_{FB} = 0.8V\) | 500 | 550 | 650 | \(kHz\) |
| | \(V_{FB} = 0V\) | 500 | 550 | 650 | \(kHz\) |
| \(R_{DS(ON)}\) of Internal P-Channel FET | \(V_{IN} = 4.2V, I_{SW} = 100mA\) (Note 6) | 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\(^\circ C\) to 70\(^\circ C\). Specifications over the –40\(^\circ C\) to 85\(^\circ C\) 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:

\[
T_J = T_A + (P_D \times \theta_J^\circ C/W)
\]

**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 \(V_{FB}\) 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**

**I_{TH}/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_{FB} (Pin 3):** Receives the feedback voltage from an external resistive divider across the output.

**SENSE– (Pin 4):** The Negative Input to the Current Comparator. Can be connected to \( V_{IN} \) for default minimum peak current of 250mA. Connecting a resistor between SENSE– and \( V_{IN} \) specifies a lower peak current. (See Applications Information for specifying resistor value.)

**V_{IN} (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.

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**PIN FUNCTIONS**
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 ITH/RUN pin, which is the output of the error amplifier EAMP. An external resistive divider connected between VOUT and ground allows the EAMP to receive an output feedback voltage VFB. When the load current increases, it causes a slight decrease in VFB relative to the 0.8V reference, which in turn causes the ITH/RUN voltage to increase until the average inductor current matches the new load current.

The main control loop is shut down by pulling the ITH/RUN pin low. Releasing ITH/RUN allows an internal 0.5µA current source to charge up the external compensation network. When the ITH/RUN pin reaches 325mV, the main control loop is enabled with the ITH/RUN voltage then pulled up to its zero current level of approximately 0.7V. As the external compensation network continues to charge...
**OPERATION** (Refer to Functional Diagram)

up, the corresponding output current trip level follows, allowing normal operation.

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 \( V_{ITH/RUN} = 1V \) (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:

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

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](image-url)
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 RSENSE (= R1). Next, the output diode D1 is selected followed by Cin (= C1) and Cout (= C2).

**Inductor Value Calculation**

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

\[
I_{\text{RIPPLE}} = \left( \frac{V_{\text{IN}} - V_{\text{OUT}}}{V_{\text{IN}} \cdot f \cdot L} \right) \frac{V_{\text{OUT}}}{V_{\text{IN}}}
\]

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 at maximum input voltage. Use the following equations to calculate \( L \):

\[
I_{\text{RIPPLE}} = 0.4 \cdot I_{\text{OUT}(\text{MAX})}
\]

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

\[
I_{\text{L}(\text{MAX})} = I_{\text{OUT}(\text{MAX})} + \frac{I_{\text{RIPPLE}}}{2}
\]

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:

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

This implies a minimum inductance of:

\[
L_{\text{MIN}} = f \left( \frac{M(0.030)}{(R_{\text{SENSE}} + 2\Omega)} \right) \left( \frac{V_{\text{OUT}} + V_{\text{D}}}{V_{\text{IN}} + V_{\text{D}}} \right)
\]

(Use \( V_{\text{IN}(\text{MAX})} = V_{\text{IN}} \))
A smaller value than $L_{\text{MIN}}$ could be used in the circuit; however, the inductor current will not be continuous during burst periods.

**R\textsubscript{SENSE} Selection for Output Current**

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

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

where SF and M are as defined in the previous section, Figures 2 and 3. Typically, $R_{\text{SENSE}}$ is chosen between 0Ω and 20Ω. 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.

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

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 $R_{\text{SENSE}}$ 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 $V_{\text{IN}}$ approaches $V_{\text{OUT}}$ 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.
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:

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

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 $V_{OUT}$ on $V_{REF}$ as $V_{IN}$ goes below 2.3V.
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 VIN which is typically much larger than the DC supply current. In continuous mode, \( I_{GATECHG} = f(Q_p) \).

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 RSENSE and the output diode. The MOSFET RDS(ON) plus RSENSE multiplied by duty cycle can be summed with the resistances of L and RSENSE 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:

\[
\text{Transition Loss} = 2(V_{IN})^2I_{O(MAX)}CRSS(f)
\]

Other losses including C_IN and C_OUT ESR dissipative losses, and inductor core losses, generally account for less than 2% total additional loss.

### 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:

\[
V_{OUT} = 0.8 \left( 1 + \frac{R_2}{R_1} \right)
\]

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

### 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:

\[
\text{Efficiency} = 100\% - (\eta_1 + \eta_2 + \eta_3 + \ldots)
\]

where \( \eta_1, \eta_2, \text{etc.} \) are the individual losses as a percentage of input power.
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 DFB1 and DFB2 between the output and the ITH/RUN pin as shown in Figure 6. In a hard short (VOUT = 0V), the current will be reduced to approximately 50% of the maximum output current.

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 CIN, 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 VIN (Pin 5) and ground (Pin 2)?

3. Keep the switching node SW away from sensitive small signal nodes.

4. Does the VFB pin connect directly to the feedback resistors? The resistive divider R1 and R2 must be connected between the (+) plate of COUT and signal ground. Locate R1 and R2 close to the VFB pin.
## 6-Lead Plastic SOT-23

### S6 Package

- **SOT-23 (Original)**
- **SOT-23 (ThinSOT)**

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<th>SOT-23 (Original)</th>
<th>SOT-23 (ThinSOT)</th>
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<td>A</td>
<td>.90 – 1.45 (.035 – .057)</td>
<td>1.00 MAX (.039 MAX)</td>
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<tr>
<td>A1</td>
<td>.00 – 0.15 (.00 – .006)</td>
<td>.01 – .10 (.0004 – .004)</td>
</tr>
<tr>
<td>A2</td>
<td>.90 – 1.30 (.035 – .051)</td>
<td>.89 – .90 (.031 – .035)</td>
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<tr>
<td>L</td>
<td>35 – 55 (.014 – .021)</td>
<td>30 – 50 REF (.012 – .019 REF)</td>
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</table>

### Notes:
1. CONTROLLING DIMENSION: MILLIMETERS
2. DIMENSIONS ARE IN MILLIMETERS
3. DRAWING NOT TO SCALE
4. DIMENSIONS ARE INCLUSIVE OF PLATING
5. DIMENSIONS ARE EXCLUSIVE OF MOLD FLASH AND METAL BURR
6. MOLD FLASH SHALL NOT EXCEED .254mm
7. PACKAGE EIAJ REFERENCE IS:
   - SC-74A (EIAJ) FOR ORIGINAL
   - JEDEL MO-193 FOR THIN
TYPICAL APPLICATIONS

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

Efficiency vs Load Current

PART NUMBER DESCRIPTION COMMENTS

LT®1375/LT1376 1.5A, 500kHz Step-Down Switching Regulators High Frequency, Small Inductor, High Efficiency
LT1616 600mA Step-Down Switching Regulator 1.4MHz, 4V to 25V Input, ThinSOT Package
LT1624 High Efficiency SO-8 N-Channel Switching Regulator Controller 8-Pin N-Channel Drive, 3.5V
LT1625 No RSENSE™ Synchronous Step-Down Regulator High Efficiency, No Sense Resistor
LT1627 Low Voltage, Monolithic Synchronous Step-Down Regulator Low Supply Voltage Range: 2.65V to 8V, IOUT = 0.5A
LT1676/LT1776 Wide Input Range Step-Down Switching Regulators 60V Input, 700mA Internal Switches
LT1735 Single, High Efficiency, Low Noise Synchronous Switching Controller High Efficiency 5V to 3.3V Conversion at up to 15A
LT1767 1.5A, 1.4MHz Step-Down DC/DC Converter Higher Current, 8-Lead MSOP Package
LT1771 Ultralow Supply Current Step-Down DC/DC Controller 10µA IQ, 93% Efficiency, 1.23V ≤ VOUT ≤ 18V, 2.8V ≤ VIN ≤ 20V
LT1772 Constant Frequency Current Mode Step-Down DC/DC Controller VIN = 2.5V to 9.8V, IOUT Up to 2A, ThinSOT Package
LT1773 95% Efficient Synchronous Step-Down Controller 2.65V ≤ VIN ≤ 8.5V, 0.8V ≤ VOUT ≤ VIN, Current Mode, 550kHz
LT1877 High Efficiency Monolithic Step-Down Regulator 550kHz, MS8, VIN Up to 10V, IQ = 10µA, IOUT to 600mA at VIN = 5V
LT1878 High Efficiency Monolithic Step-Down Regulator 550kHz, MS8, VIN Up to 6V, IQ = 10µA, IOUT to 600mA at VIN = 5V
LT1879 1.2MHz Synchronous Step-Up DC/DC Converter in ThinSOT 92% Efficiency, VIN = 0.5V to 6V, VOUT = 2.6V to 5V
LTC3401 Single Cell, High Current (1A), Micropower, Synchronous 3MHz Step-Up DC/DC Converter VIN = 0.5V to 5V, Up to 97% Efficiency Synchronizable Oscillator from 100kHz to 3MHz
LTC3402 Single Cell, High Current (2A), Micropower, Synchronous 3MHz Step-Up DC/DC Converter VIN = 0.7V to 5V, Up to 95% Efficiency Synchronizable Oscillator from 100kHz to 3MHz
LTC3404 1.4MHz High Efficiency, Monolithic Synchronous Step-Down Regulator Up to 95% Efficiency, 100% Duty Cycle, IQ = 10µA, VOUT = 2.65V to 6V

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