LM2734Z/LM2734ZQ Thin SOT 1A Load Step-Down DC-DC Regulator

Check for Samples: LM2734Z

FEATURES

- **Thin SOT-6 Package, or 6 Lead WSON Package**
- **3.0V to 20V Input Voltage Range**
- **0.8V to 18V Output Voltage Range**
- **1A Output Current**
- **3MHz Switching Frequency**
- **300mΩ NMOS Switch**
- **30nA Shutdown Current**
- **0.8V, 2% Internal Voltage Reference**
- **Internal Soft-Start**
- **Current-Mode, PWM Operation**
- **Thermal Shutdown**
- **LM2734ZQ is AEC-Q100 Grade 1 Qualified and is Manufactured on an Automotive Grade Flow**

APPLICATIONS

- **DSL Modems**
- **Local Point of Load Regulation**
- **Battery Powered Devices**
- **USB Powered Devices**
- **Automotive**

DESCRIPTION

The LM2734Z regulator is a monolithic, high frequency, PWM step-down DC/DC converter assembled in a 6-pin Thin SOT and WSON non pull back package. It provides all the active functions to provide local DC/DC conversion with fast transient response and accurate regulation in the smallest possible PCB area.

With a minimum of external components and online design support through WEBENCH™, the LM2734Z is easy to use. The ability to drive 1A loads with an internal 300mΩ NMOS switch using state-of-the-art 0.5μm BiCMOS technology results in the best power density available. The world class control circuitry allows for on-times as low as 13ns, thus supporting exceptionally high frequency conversion over the entire 3V to 20V input operating range down to the minimum output voltage of 0.8V. Switching frequency is internally set to 3MHz, allowing the use of extremely small surface mount inductors and chip capacitors. Even though the operating frequency is very high, efficiencies up to 85% are easy to achieve. External shutdown is included, featuring an ultra-low stand-by current of 30nA. The LM2734Z utilizes current-mode control and internal compensation to provide high-performance regulation over a wide range of operating conditions. Additional features include internal soft-start circuitry to reduce inrush current, pulse-by-pulse current limit, thermal shutdown, and output over-voltage protection.


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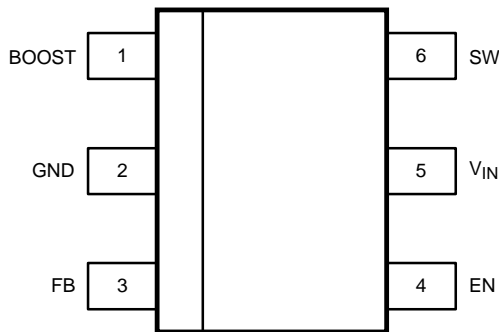
Typical Application Circuit

Efficiency vs Load Current
V_{IN} = 5V, V_{OUT} = 3.3V
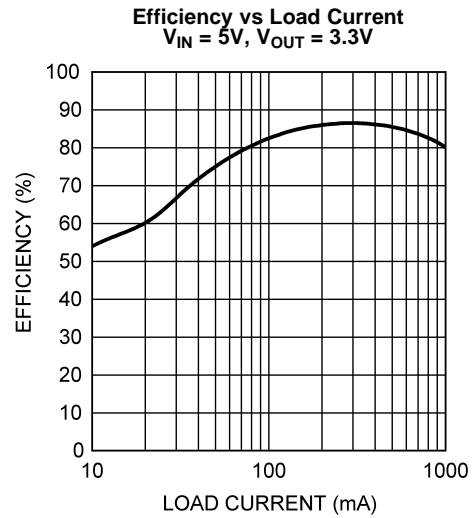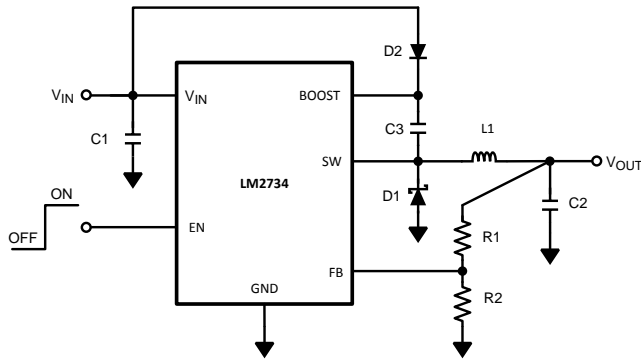
Connection Diagrams

Figure 1. 6-Lead SOT
See Package Number DDC (R-PDSO-G6)
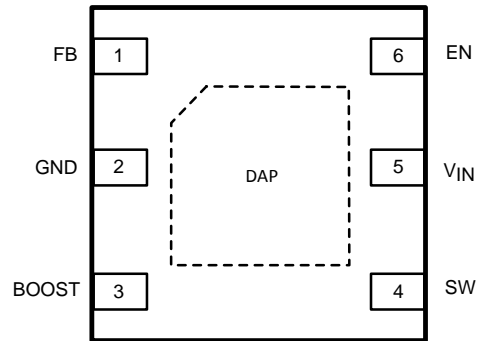
Figure 2. 6-Lead WSON (3mm x 3mm)
See Package Number NGG0006A

PIN DESCRIPTIONS

Pin	Name	Function
1	BOOST	Boost voltage that drives the internal NMOS control switch. A bootstrap capacitor is connected between the BOOST and SW pins.
2	GND	Signal and Power ground pin. Place the bottom resistor of the feedback network as close as possible to this pin for accurate regulation.
3	FB	Feedback pin. Connect FB to the external resistor divider to set output voltage.
4	EN	Enable control input. Logic high enables operation. Do not allow this pin to float or be greater than V_{IN} + 0.3V.
5	V_{IN}	Input supply voltage. Connect a bypass capacitor to this pin.
6	SW	Output switch. Connects to the inductor, catch diode, and bootstrap capacitor.
DAP	GND	The Die Attach Pad is internally connected to GND

These devices have limited built-in ESD protection. The leads should be shorted together or the device placed in conductive foam during storage or handling to prevent electrostatic damage to the MOS gates.

Absolute Maximum Ratings[(1)(2)]

V_{IN}		-0.5V to 24V
SW Voltage		-0.5V to 24V
Boost Voltage		-0.5V to 30V
Boost to SW Voltage		-0.5V to 6.0V
FB Voltage		-0.5V to 3.0V
EN Voltage		-0.5V to (V_{IN} + 0.3V)
Junction Temperature		150°C
ESD Susceptibility[(3)]		2kV
Storage Temp. Range		-65°C to 150°C
Soldering Information	Infrared/Convection Reflow (15sec)	220°C
	Wave Soldering Lead Temp. (10sec)	260°C

(1) Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but specific performance is not ensured. For specific specifications and the test conditions, see Electrical Characteristics.
(2) If Military/Aerospace specified devices are required, please contact the Texas Instruments Sales Office/ Distributors for availability and specifications.
(3) Human body model, 1.5kΩ in series with 100pF.

Operating Ratings[(1)]

V_{IN}		3V to 20V
SW Voltage		-0.5V to 20V
Boost Voltage		-0.5V to 25V
Boost to SW Voltage		1.6V to 5.5V
Junction Temperature Range		−40°C to +125°C
Thermal Resistance θ_{JA}[(2)]	SOT–6	118°C/W

(1) Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but specific performance is not ensured. For specific specifications and the test conditions, see Electrical Characteristics.
(2) Thermal shutdown will occur if the junction temperature exceeds 165°C. The maximum power dissipation is a function of $T_{J(MAX)}$, θ_{JA} and T_A. The maximum allowable power dissipation at any ambient temperature is $P_D = (T_{J(MAX)} − T_A)/\theta_{JA}$. All numbers apply for packages soldered directly onto a 3" x 3" PC board with 2oz. copper on 4 layers in still air. For a 2 layer board using 1 oz. copper in still air, θ_{JA} = 204°C/W.

Electrical Characteristics

Specifications with standard typeface are for T_J = 25°C, and those in **boldface type** apply over the full **Operating Temperature Range** (T_J = -40°C to 125°C). V_{IN} = 5V, V_{BOOST} - V_{SW} = 5V unless otherwise specified. Datasheet min/max specification limits are specified by design, test, or statistical analysis.

Symbol	Parameter	Conditions	Min[1]	Typ[2]	Max[1]	Units
V_{FB}	Feedback Voltage		**0.784**	0.800	**0.816**	V
$\Delta V_{FB}/\Delta V_{IN}$	Feedback Voltage Line Regulation	V_{IN} = 3V to 20V		0.01		% / V
I_{FB}	Feedback Input Bias Current	Sink/Source		10	**250**	nA
UVLO	Undervoltage Lockout	V_{IN} Rising		2.74	**2.90**	V
	Undervoltage Lockout	V_{IN} Falling	**2.0**	2.3		
	UVLO Hysteresis		**0.30**	0.44	**0.62**	
F_{SW}	Switching Frequency		**2.2**	3.0	**3.6**	MHz
D_{MAX}	Maximum Duty Cycle		**78**	85		%
D_{MIN}	Minimum Duty Cycle			8		%
$R_{DS(ON)}$	Switch ON Resistance	V_{BOOST} - V_{SW} = 3V (SOT Package)		300	**600**	mΩ
		V_{BOOST} - V_{SW} = 3V (WSON Package)		340	**650**	mΩ
I_{CL}	Switch Current Limit	V_{BOOST} - V_{SW} = 3V	**1.2**	1.7	**2.5**	A
I_Q	Quiescent Current	Switching		1.5	**2.5**	mA
	Quiescent Current (shutdown)	V_{EN} = 0V		30		nA
I_{BOOST}	Boost Pin Current	(Switching)		4.25	**6**	mA
V_{EN_TH}	Shutdown Threshold Voltage	V_{EN} Falling			**0.4**	V
	Enable Threshold Voltage	V_{EN} Rising	**1.8**			
I_{EN}	Enable Pin Current	Sink/Source		10		nA
I_{SW}	Switch Leakage			40		nA

(1) Specified to Texas Instruments' Average Outgoing Quality Level (AOQL).
(2) Typicals represent the most likely parametric norm.

Product Folder Links: *LM2734Z*

Typical Performance Characteristics

All curves taken at V_{IN} = 5V, V_{BOOST} - V_{SW} = 5V, L1 = 2.2 µH and T_A = 25°C, unless specified otherwise.

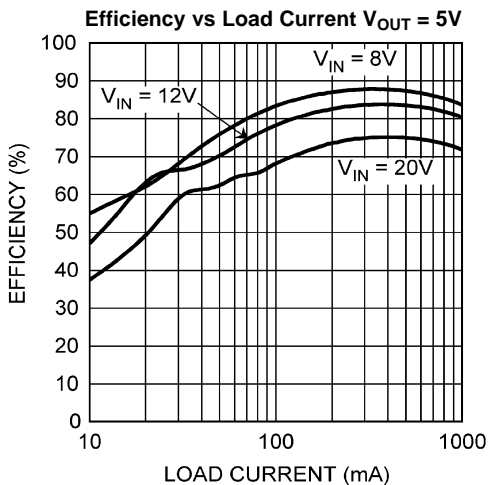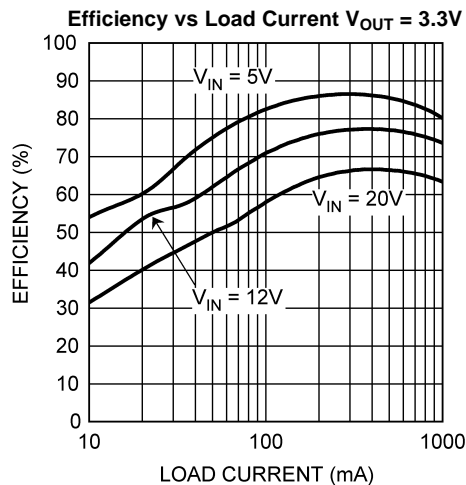**Efficiency vs Load Current V_{OUT} = 5V**

Figure 3.

Efficiency vs Load Current V_{OUT} = 3.3V

Figure 4.

Efficiency vs Load Current V_{OUT} = 1.5V

Figure 5.

Oscillator Frequency vs Temperature

Figure 6.

**Line Regulation
V_{OUT} = 1.5V, I_{OUT} = 500mA**
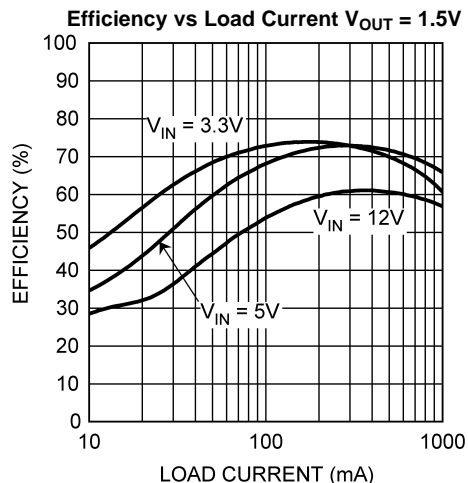
Figure 7.
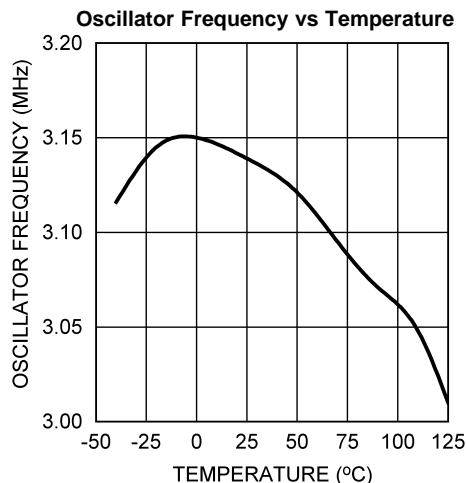
**Line Regulation
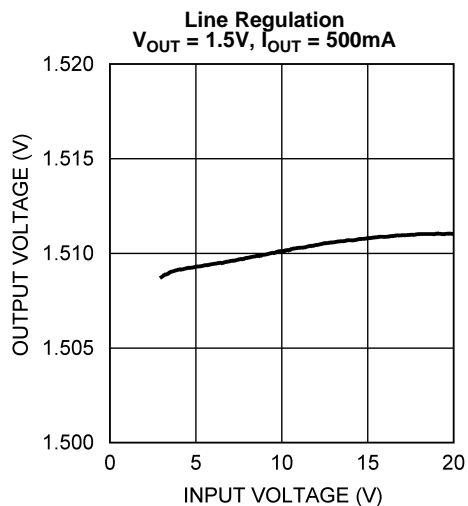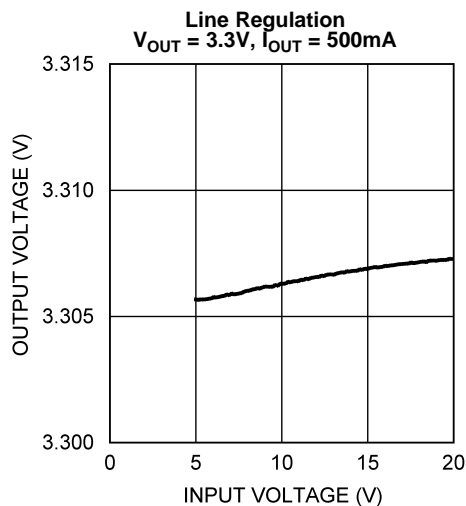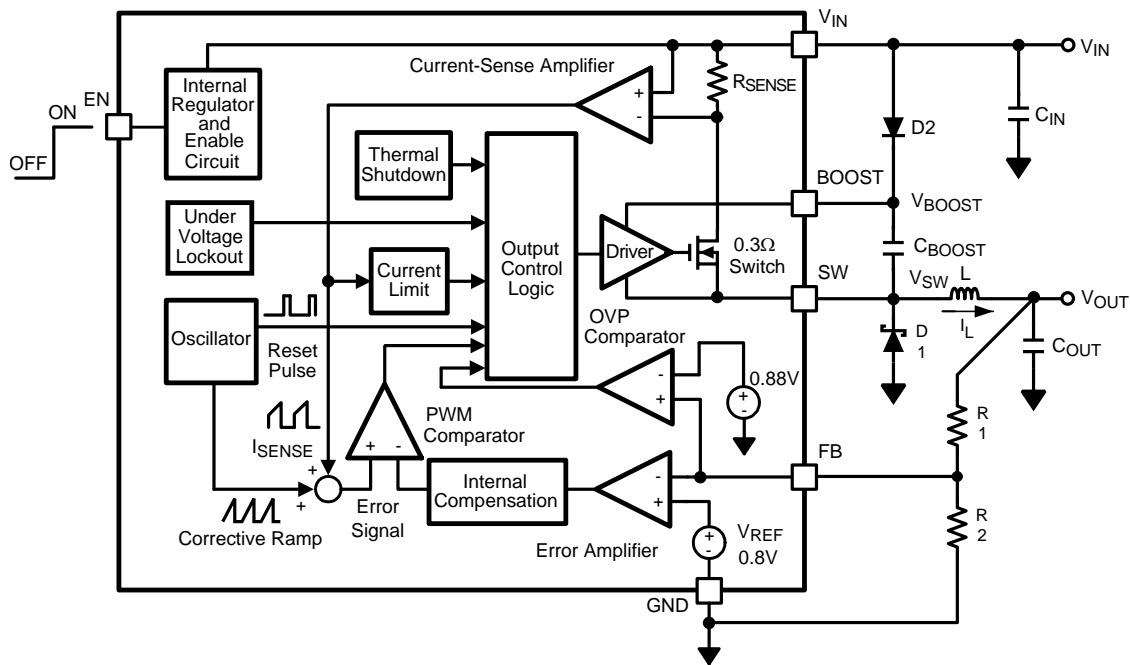V_{OUT} = 3.3V, I_{OUT} = 500mA**

Figure 8.

Block Diagram

APPLICATION INFORMATION

THEORY OF OPERATION

The LM2734Z is a constant frequency PWM buck regulator IC that delivers a 1A load current. The regulator has a preset switching frequency of 3MHz. This high frequency allows the LM2734Z to operate with small surface mount capacitors and inductors, resulting in a DC/DC converter that requires a minimum amount of board space. The LM2734Z is internally compensated, so it is simple to use, and requires few external components. The LM2734Z uses current-mode control to regulate the output voltage.

The following operating description of the LM2734Z will refer to the Simplified Block Diagram (Block Diagram) and to the waveforms in Figure 9. The LM2734Z supplies a regulated output voltage by switching the internal NMOS control switch at constant frequency and variable duty cycle. A switching cycle begins at the falling edge of the reset pulse generated by the internal oscillator. When this pulse goes low, the output control logic turns on the internal NMOS control switch. During this on-time, the SW pin voltage (V_{SW}) swings up to approximately V_{IN}, and the inductor current (I_L) increases with a linear slope. I_L is measured by the current-sense amplifier, which generates an output proportional to the switch current. The sense signal is summed with the regulator's corrective ramp and compared to the error amplifier's output, which is proportional to the difference between the feedback voltage and V_{REF}. When the PWM comparator output goes high, the output switch turns off until the next switching cycle begins. During the switch off-time, inductor current discharges through Schottky diode D1, which forces the SW pin to swing below ground by the forward voltage (V_D) of the catch diode. The regulator loop adjusts the duty cycle (D) to maintain a constant output voltage.
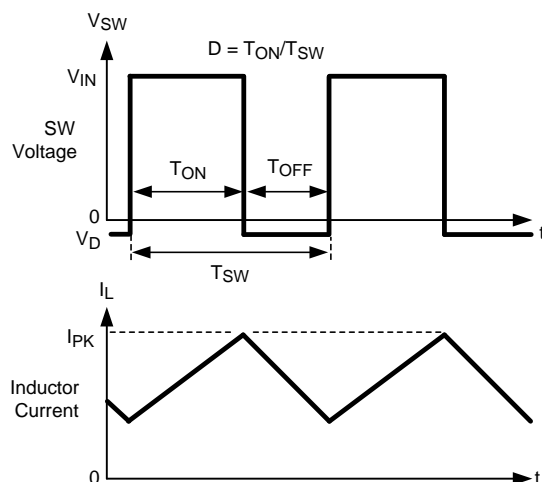
Figure 9. LM2734Z Waveforms of SW Pin Voltage and Inductor Current

BOOST FUNCTION

Capacitor C_{BOOST} and diode D2 in Figure 10 are used to generate a voltage V_{BOOST}. V_{BOOST} - V_{SW} is the gate drive voltage to the internal NMOS control switch. To properly drive the internal NMOS switch during its on-time, V_{BOOST} needs to be at least 1.6V greater than V_{SW}. Although the LM2734Z will operate with this minimum voltage, it may not have sufficient gate drive to supply large values of output current. Therefore, it is recommended that V_{BOOST} be greater than 2.5V above V_{SW} for best efficiency. V_{BOOST} − V_{SW} should not exceed the maximum operating limit of 5.5V.

5.5V > V_{BOOST} − V_{SW} > 2.5V for best performance.
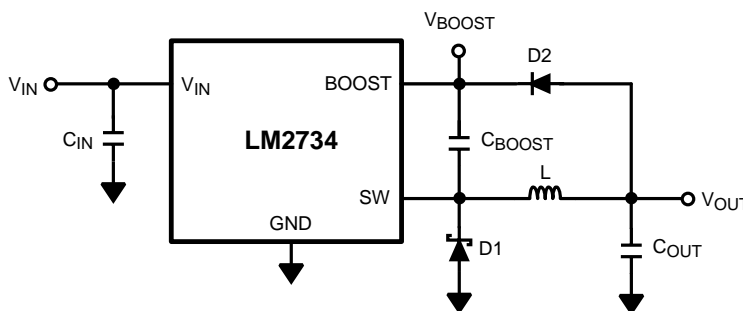
Figure 10. V$_{OUT}$ Charges C$_{BOOST}$

When the LM2734Z starts up, internal circuitry from the BOOST pin supplies a maximum of 20mA to C_{BOOST}. This current charges C_{BOOST} to a voltage sufficient to turn the switch on. The BOOST pin will continue to source current to C_{BOOST} until the voltage at the feedback pin is greater than 0.76V.

There are various methods to derive V_{BOOST}:

1. From the input voltage (V_{IN})
2. From the output voltage (V_{OUT})
3. From an external distributed voltage rail (V_{EXT})
4. From a shunt or series zener diode

In the Simplifed Block Diagram of Block Diagram, capacitor C_{BOOST} and diode D2 supply the gate-drive current for the NMOS switch. Capacitor C_{BOOST} is charged via diode D2 by V_{IN}. During a normal switching cycle, when the internal NMOS control switch is off (T_{OFF}) (refer to Figure 9), V_{BOOST} equals V_{IN} minus the forward voltage of D2 (V_{FD2}), during which the current in the inductor (L) forward biases the Schottky diode D1 (V_{FD1}). Therefore the voltage stored across C_{BOOST} is

$$V_{BOOST} - V_{SW} = V_{IN} - V_{FD2} + V_{FD1} \tag{1}$$

When the NMOS switch turns on (T_{ON}), the switch pin rises to

$$V_{SW} = V_{IN} - (R_{DSON} \times I_L), \tag{2}$$

forcing V_{BOOST} to rise thus reverse biasing D2. The voltage at V_{BOOST} is then

$$V_{BOOST} = 2V_{IN} - (R_{DSON} \times I_L) - V_{FD2} + V_{FD1} \tag{3}$$

which is approximately

$$2V_{IN} - 0.4V \tag{4}$$

for many applications. Thus the gate-drive voltage of the NMOS switch is approximately

$$V_{IN} - 0.2V \tag{5}$$

An alternate method for charging C_{BOOST} is to connect D2 to the output as shown in Figure 10. The output voltage should be between 2.5V and 5.5V, so that proper gate voltage will be applied to the internal switch. In this circuit, C_{BOOST} provides a gate drive voltage that is slightly less than V_{OUT}.

In applications where both V_{IN} and V_{OUT} are greater than 5.5V, or less than 3V, C_{BOOST} cannot be charged directly from these voltages. If V_{IN} and V_{OUT} are greater than 5.5V, C_{BOOST} can be charged from V_{IN} or V_{OUT} minus a zener voltage by placing a zener diode D3 in series with D2, as shown in Figure 11. When using a series zener diode from the input, ensure that the regulation of the input supply doesn't create a voltage that falls outside the recommended V_{BOOST} voltage.

$$(V_{INMAX} - V_{D3}) < 5.5V \tag{6}$$
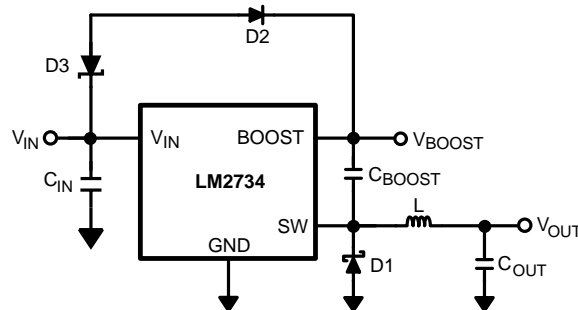$$(V_{INMIN} - V_{D3}) > 1.6V \tag{7}$$

Figure 11. Zener Reduces Boost Voltage from V_{IN}

An alternative method is to place the zener diode D3 in a shunt configuration as shown in Figure 12. A small 350mW to 500mW 5.1V zener in a SOT or SOD package can be used for this purpose. A small ceramic capacitor such as a 6.3V, 0.1µF capacitor (C4) should be placed in parallel with the zener diode. When the internal NMOS switch turns on, a pulse of current is drawn to charge the internal NMOS gate capacitance. The 0.1 µF parallel shunt capacitor ensures that the V_{BOOST} voltage is maintained during this time.

Resistor R3 should be chosen to provide enough RMS current to the zener diode (D3) and to the BOOST pin. A recommended choice for the zener current (I_{ZENER}) is 1 mA. The current I_{BOOST} into the BOOST pin supplies the gate current of the NMOS control switch and varies typically according to the following formula:

$$I_{BOOST} = (D + 0.5) \times (V_{ZENER} - V_{D2}) \text{ mA} \tag{8}$$

where D is the duty cycle, V_{ZENER} and V_{D2} are in volts, and I_{BOOST} is in milliamps. V_{ZENER} is the voltage applied to the anode of the boost diode (D2), and V_{D2} is the average forward voltage across D2. Note that this formula for I_{BOOST} gives typical current. For the worst case I_{BOOST}, increase the current by 25%. In that case, the worst case boost current will be

$$I_{BOOST-MAX} = 1.25 \times I_{BOOST} \tag{9}$$

R3 will then be given by

$$R3 = (V_{IN} - V_{ZENER}) / (1.25 \times I_{BOOST} + I_{ZENER}) \qquad (10)$$

For example, let V_{IN} = 10V, V_{ZENER} = 5V, V_{D2} = 0.7V, I_{ZENER} = 1mA, and duty cycle D = 50%. Then

$$I_{BOOST} = (0.5 + 0.5) \times (5 - 0.7) \text{ mA} = 4.3\text{mA} \qquad (11)$$

$$R3 = (10V - 5V) / (1.25 \times 4.3\text{mA} + 1\text{mA}) = 787\Omega \qquad (12)$$
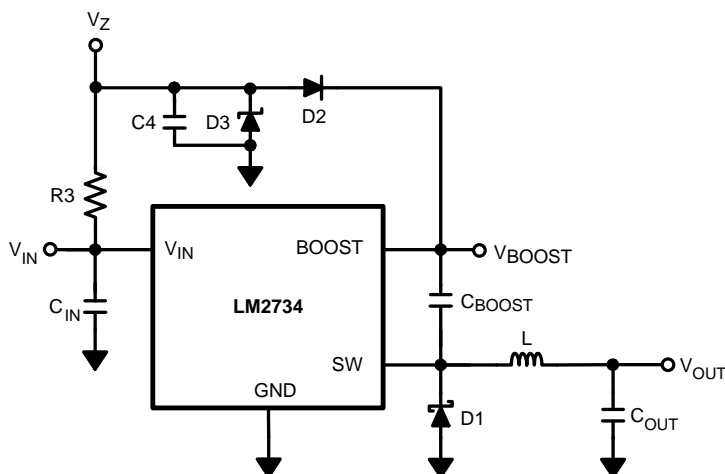
Figure 12. Boost Voltage Supplied from the Shunt Zener on V$_{IN}$

ENABLE PIN / SHUTDOWN MODE

The LM2734Z has a shutdown mode that is controlled by the enable pin (EN). When a logic low voltage is applied to EN, the part is in shutdown mode and its quiescent current drops to typically 30nA. Switch leakage adds another 40nA from the input supply. The voltage at this pin should never exceed V_{IN} + 0.3V.

SOFT-START

This function forces V_{OUT} to increase at a controlled rate during start up. During soft-start, the error amplifier's reference voltage ramps from 0V to its nominal value of 0.8V in approximately 200μs. This forces the regulator output to ramp up in a more linear and controlled fashion, which helps reduce inrush current.

OUTPUT OVERVOLTAGE PROTECTION

The overvoltage comparator compares the FB pin voltage to a voltage that is 10% higher than the internal reference Vref. Once the FB pin voltage goes 10% above the internal reference, the internal NMOS control switch is turned off, which allows the output voltage to decrease toward regulation.

UNDERVOLTAGE LOCKOUT

Undervoltage lockout (UVLO) prevents the LM2734Z from operating until the input voltage exceeds 2.74V(typ).

The UVLO threshold has approximately 440mV of hysteresis, so the part will operate until V_{IN} drops below 2.3V(typ). Hysteresis prevents the part from turning off during power up if V_{IN} is non-monotonic.

CURRENT LIMIT

The LM2734Z uses cycle-by-cycle current limiting to protect the output switch. During each switching cycle, a current limit comparator detects if the output switch current exceeds 1.7A (typ), and turns off the switch until the next switching cycle begins.

THERMAL SHUTDOWN

Thermal shutdown limits total power dissipation by turning off the output switch when the IC junction temperature exceeds 165°C. After thermal shutdown occurs, the output switch doesn't turn on until the junction temperature drops to approximately 150°C.

Design Guide

INDUCTOR SELECTION

The Duty Cycle (D) can be approximated quickly using the ratio of output voltage (V_O) to input voltage (V_{IN}):

$$D = \frac{V_O}{V_{IN}}$$

(13)

The catch diode (D1) forward voltage drop and the voltage drop across the internal NMOS must be included to calculate a more accurate duty cycle. Calculate D by using the following formula:

$$D = \frac{V_O + V_D}{V_{IN} + V_D - V_{SW}}$$

(14)

V_{SW} can be approximated by:

$$V_{SW} = I_O \times R_{DS(ON)}$$

(15)

The diode forward drop (V_D) can range from 0.3V to 0.7V depending on the quality of the diode. The lower V_D is, the higher the operating efficiency of the converter.

The inductor value determines the output ripple current. Lower inductor values decrease the size of the inductor, but increase the output ripple current. An increase in the inductor value will decrease the output ripple current. The ratio of ripple current (Δi_L) to output current (I_O) is optimized when it is set between 0.3 and 0.4 at 1A. The ratio r is defined as:

$$r = \frac{\Delta i_L}{I_O}$$

(16)

One must also ensure that the minimum current limit (1.2A) is not exceeded, so the peak current in the inductor must be calculated. The peak current (I_{LPK}) in the inductor is calculated by:

$$I_{LPK} = I_O + \Delta I_L / 2$$

(17)

If r = 0.5 at an output of 1A, the peak current in the inductor will be 1.25A. The minimum specified current limit over all operating conditions is 1.2A. One can either reduce r to 0.4 resulting in a 1.2A peak current, or make the engineering judgement that 50mA over will be safe enough with a 1.7A typical current limit and 6 sigma limits. When the designed maximum output current is reduced, the ratio r can be increased. At a current of 0.1A, r can be made as high as 0.9. The ripple ratio can be increased at lighter loads because the net ripple is actually quite low, and if r remains constant the inductor value can be made quite large. An equation empirically developed for the maximum ripple ratio at any current below 2A is:

$$r = 0.387 \times I_{OUT}^{-0.3667}$$

(18)

Note that this is just a guideline.

The LM2734Z operates at frequencies allowing the use of ceramic output capacitors without compromising transient response. Ceramic capacitors allow higher inductor ripple without significantly increasing output ripple. See the OUTPUT CAPACITOR section for more details on calculating output voltage ripple.

Now that the ripple current or ripple ratio is determined, the inductance is calculated by:

$$L = \frac{V_O + V_D}{I_O \times r \times f_S} \times (1-D)$$

(19)

where f_s is the switching frequency and I_O is the output current. When selecting an inductor, make sure that it is capable of supporting the peak output current without saturating. Inductor saturation will result in a sudden reduction in inductance and prevent the regulator from operating correctly. Because of the speed of the internal current limit, the peak current of the inductor need only be specified for the required maximum output current. For example, if the designed maximum output current is 0.5A and the peak current is 0.7A, then the inductor should be specified with a saturation current limit of >0.7A. There is no need to specify the saturation or peak current of the inductor at the 1.7A typical switch current limit. The difference in inductor size is a factor of 5. Because of the operating frequency of the LM2734Z, ferrite based inductors are preferred to minimize core losses. This presents little restriction since the variety of ferrite based inductors is huge. Lastly, inductors with lower series resistance (DCR) will provide better operating efficiency. For recommended inductors see Example Circuits.

INPUT CAPACITOR

An input capacitor is necessary to ensure that V_{IN} does not drop excessively during switching transients. The primary specifications of the input capacitor are capacitance, voltage, RMS current rating, and ESL (Equivalent Series Inductance). The recommended input capacitance is 10µF, although 4.7µF works well for input voltages below 6V. The input voltage rating is specifically stated by the capacitor manufacturer. Make sure to check any recommended deratings and also verify if there is any significant change in capacitance at the operating input voltage and the operating temperature. The input capacitor maximum RMS input current rating (I_{RMS-IN}) must be greater than:

$$I_{RMS-IN} = I_O \times \sqrt{D \times \left(1\text{-}D + \frac{r^2}{12}\right)}$$

(20)

It can be shown from the above equation that maximum RMS capacitor current occurs when D = 0.5. Always calculate the RMS at the point where the duty cycle, D, is closest to 0.5. The ESL of an input capacitor is usually determined by the effective cross sectional area of the current path. A large leaded capacitor will have high ESL and a 0805 ceramic chip capacitor will have very low ESL. At the operating frequencies of the LM2734Z, certain capacitors may have an ESL so large that the resulting impedance (2πfL) will be higher than that required to provide stable operation. As a result, surface mount capacitors are strongly recommended. Sanyo POSCAP, Tantalum or Niobium, Panasonic SP or Cornell Dubilier ESR, and multilayer ceramic capacitors (MLCC) are all good choices for both input and output capacitors and have very low ESL. For MLCCs it is recommended to use X7R or X5R dielectrics. Consult capacitor manufacturer datasheet to see how rated capacitance varies over operating conditions.

OUTPUT CAPACITOR

The output capacitor is selected based upon the desired output ripple and transient response. The initial current of a load transient is provided mainly by the output capacitor. The output ripple of the converter is:

$$\Delta V_O = \Delta i_L \times (R_{ESR} + \frac{1}{8 \times f_S \times C_O})$$

(21)

When using MLCCs, the ESR is typically so low that the capacitive ripple may dominate. When this occurs, the output ripple will be approximately sinusoidal and 90° phase shifted from the switching action. Given the availability and quality of MLCCs and the expected output voltage of designs using the LM2734Z, there is really no need to review any other capacitor technologies. Another benefit of ceramic capacitors is their ability to bypass high frequency noise. A certain amount of switching edge noise will couple through parasitic capacitances in the inductor to the output. A ceramic capacitor will bypass this noise while a tantalum will not. Since the output capacitor is one of the two external components that control the stability of the regulator control loop, most applications will require a minimum at 10 µF of output capacitance. Capacitance can be increased significantly with little detriment to the regulator stability. Like the input capacitor, recommended multilayer ceramic capacitors are X7R or X5R. Again, verify actual capacitance at the desired operating voltage and temperature.

Check the RMS current rating of the capacitor. The RMS current rating of the capacitor chosen must also meet the following condition:

$$I_{RMS-OUT} = I_O \times \frac{r}{\sqrt{12}}$$

(22)

CATCH DIODE

The catch diode (D1) conducts during the switch off-time. A Schottky diode is recommended for its fast switching times and low forward voltage drop. The catch diode should be chosen so that its current rating is greater than:

$$I_{D1} = I_O \times (1-D) \tag{23}$$

The reverse breakdown rating of the diode must be at least the maximum input voltage plus appropriate margin. To improve efficiency choose a Schottky diode with a low forward voltage drop.

BOOST DIODE

A standard diode such as the 1N4148 type is recommended. For V_{BOOST} circuits derived from voltages less than 3.3V, a small-signal Schottky diode is recommended for greater efficiency. A good choice is the BAT54 small signal diode.

BOOST CAPACITOR

A ceramic 0.01µF capacitor with a voltage rating of at least 6.3V is sufficient. The X7R and X5R MLCCs provide the best performance.

OUTPUT VOLTAGE

The output voltage is set using the following equation where R2 is connected between the FB pin and GND, and R1 is connected between V_O and the FB pin. A good value for R2 is 10kΩ.

$$R1 = \left(\frac{V_O}{V_{REF}} - 1 \right) \times R2 \tag{24}$$

PCB Layout Considerations

When planning layout there are a few things to consider when trying to achieve a clean, regulated output. The most important consideration when completing the layout is the close coupling of the GND connections of the C_{IN} capacitor and the catch diode D1. These ground ends should be close to one another and be connected to the GND plane with at least two through-holes. Place these components as close to the IC as possible. Next in importance is the location of the GND connection of the C_{OUT} capacitor, which should be near the GND connections of C_{IN} and D1.

There should be a continuous ground plane on the bottom layer of a two-layer board except under the switching node island.

The FB pin is a high impedance node and care should be taken to make the FB trace short to avoid noise pickup and inaccurate regulation. The feedback resistors should be placed as close as possible to the IC, with the GND of R2 placed as close as possible to the GND of the IC. The V_{OUT} trace to R1 should be routed away from the inductor and any other traces that are switching.

High AC currents flow through the V_{IN}, SW and V_{OUT} traces, so they should be as short and wide as possible. However, making the traces wide increases radiated noise, so the designer must make this trade-off. Radiated noise can be decreased by choosing a shielded inductor.

The remaining components should also be placed as close as possible to the IC. Please see Application Note AN-1229 for further considerations and the LM2734Z demo board as an example of a four-layer layout.

Calculating Efficiency, and Junction Temperature

The complete LM2734Z DC/DC converter efficiency can be calculated in the following manner.

$$\eta = \frac{P_{OUT}}{P_{IN}} \tag{25}$$

Or

$$\eta = \frac{P_{OUT}}{P_{OUT} + P_{LOSS}} \tag{26}$$

Calculations for determining the most significant power losses are shown below. Other losses totaling less than 2% are not discussed.

Power loss (P_{LOSS}) is the sum of two basic types of losses in the converter, switching and conduction. Conduction losses usually dominate at higher output loads, where as switching losses remain relatively fixed and dominate at lower output loads. The first step in determining the losses is to calculate the duty cycle (D).

$$D = \frac{V_{OUT} + V_D}{V_{IN} + V_D - V_{SW}}$$

(27)

V_{SW} is the voltage drop across the internal NFET when it is on, and is equal to:

$$V_{SW} = I_{OUT} \times R_{DSON}$$

(28)

V_D is the forward voltage drop across the Schottky diode. It can be obtained from the Electrical Characteristics section. If the voltage drop across the inductor (V_{DCR}) is accounted for, the equation becomes:

$$D = \frac{V_O + V_D + V_{DCR}}{V_{IN} + V_D - V_{SW}}$$

(29)

This usually gives only a minor duty cycle change, and has been omitted in the examples for simplicity.

The conduction losses in the free-wheeling Schottky diode are calculated as follows:

$$P_{DIODE} = V_D \times I_{OUT}(1-D)$$

(30)

Often this is the single most significant power loss in the circuit. Care should be taken to choose a Schottky diode that has a low forward voltage drop.

Another significant external power loss is the conduction loss in the output inductor. The equation can be simplified to:

$$P_{IND} = I_{OUT}^2 \times R_{DCR}$$

(31)

The LM2734Z conduction loss is mainly associated with the internal NFET:

$$P_{COND} = I_{OUT}^2 \times R_{DSON} \times D$$

(32)

Switching losses are also associated with the internal NFET. They occur during the switch on and off transition periods, where voltages and currents overlap resulting in power loss. The simplest means to determine this loss is to empirically measuring the rise and fall times (10% to 90%) of the switch at the switch node:

$$P_{SWF} = 1/2(V_{IN} \times I_{OUT} \times freq \times T_{FALL})$$

(33)

$$P_{SWR} = 1/2(V_{IN} \times I_{OUT} \times freq \times T_{RISE})$$

(34)

$$P_{SW} = P_{SWF} + P_{SWR}$$

(35)

Table 1. Typical Rise and Fall Times vs Input Voltage

V_{IN}	T_{RISE}	T_{FALL}
5V	8ns	4ns
10V	9ns	6ns
15V	10ns	7ns

Another loss is the power required for operation of the internal circuitry:

$$P_Q = I_Q \times V_{IN}$$

(36)

I_Q is the quiescent operating current, and is typically around 1.5mA. The other operating power that needs to be calculated is that required to drive the internal NFET:

$$P_{BOOST} = I_{BOOST} \times V_{BOOST}$$

(37)

V_{BOOST} is normally between 3VDC and 5VDC. The I_{BOOST} rms current is approximately 4.25mA. Total power losses are:

$$\Sigma P_{COND} + P_{SW} + P_{DIODE} + P_{IND} + P_Q + P_{BOOST} = P_{LOSS}$$

(38)

Design Example 1:

Operating Conditions			
V_{IN}	5.0V	P_{OUT}	2.5W
V_{OUT}	2.5V	P_{DIODE}	151mW
I_{OUT}	1.0A	P_{IND}	75mW
V_D	0.35V	P_{SWF}	53mW
Freq	3MHz	P_{SWR}	53mW
I_Q	1.5mA	P_{COND}	187mW
T_{RISE}	8ns	P_Q	7.5mW
T_{FALL}	8ns	P_{BOOST}	21mW
R_{DSON}	330mΩ	P_{LOSS}	**548mW**
IND_{DCR}	75mΩ		
D	**0.568**		

η = 82%

Calculating the LM2734Z Junction Temperature

Thermal Definitions:

T_J = Chip junction temperature

T_A = Ambient temperature

$R_{\theta JC}$ = Thermal resistance from chip junction to device case

$R_{\theta JA}$ = Thermal resistance from chip junction to ambient air
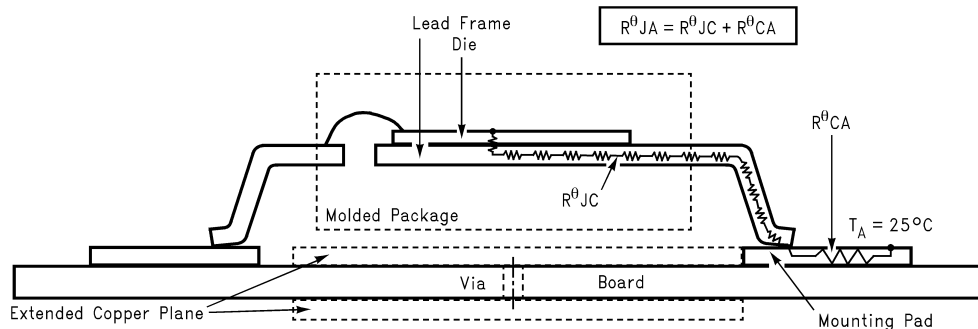
Figure 13. Cross-Sectional View of Integrated Circuit Mounted on a Printed Circuit Board

Heat in the LM2734Z due to internal power dissipation is removed through conduction and/or convection.

Conduction: Heat transfer occurs through cross sectional areas of material. Depending on the material, the transfer of heat can be considered to have poor to good thermal conductivity properties (insulator vs conductor).

Heat Transfer goes as:

silicon→package→lead frame→PCB.

Convection: Heat transfer is by means of airflow. This could be from a fan or natural convection. Natural convection occurs when air currents rise from the hot device to cooler air.

Thermal impedance is defined as:

$$R_\theta = \frac{\Delta T}{Power}$$

(39)

Thermal impedance from the silicon junction to the ambient air is defined as:

$$R_{\theta JA} = \frac{T_J - T_A}{Power}$$

(40)

This impedance can vary depending on the thermal properties of the PCB. This includes PCB size, weight of copper used to route traces and ground plane, and number of layers within the PCB. The type and number of thermal vias can also make a large difference in the thermal impedance. Thermal vias are necessary in most applications. They conduct heat from the surface of the PCB to the ground plane. Four to six thermal vias should be placed under the exposed pad to the ground plane if the WSON package is used. If the Thin SOT-6 package is used, place two to four thermal vias close to the ground pin of the device.

The datasheet specifies two different $R_{\theta JA}$ numbers for the Thin SOT–6 package. The two numbers show the difference in thermal impedance for a four-layer board with 2oz. copper traces, vs. a four-layer board with 1oz. copper. $R_{\theta JA}$ equals 120°C/W for 2oz. copper traces and GND plane, and 235°C/W for 1oz. copper traces and GND plane.

Method 1:

To accurately measure the silicon temperature for a given application, two methods can be used. The first method requires the user to know the thermal impedance of the silicon junction to case. ($R_{\theta JC}$) is approximately 80°C/W for the Thin SOT-6 package. Knowing the internal dissipation from the efficiency calculation given previously, and the case temperature, which can be empirically measured on the bench we have:

$$R_{\theta JA} = \frac{T_J - T_C}{Power}$$

(41)

Therefore:

$$T_J = (R_{\theta JC} \times P_{LOSS}) + T_C \tag{42}$$

Design Example 2:			
Operating Conditions			
V_{IN}	5.0V	P_{OUT}	2.5W
V_{OUT}	2.5V	P_{DIODE}	151mW
I_{OUT}	1.0A	P_{IND}	75mW
V_D	0.35V	P_{SWF}	53mW
Freq	3MHz	P_{SWR}	53mW
I_Q	1.5mA	P_{COND}	187mW
T_{RISE}	8ns	P_Q	7.5mW
T_{FALL}	8ns	P_{BOOST}	21mW
R_{DSON}	330mΩ	P_{LOSS}	**548mW**
IND_{DCR}	75mΩ		
D	**0.568**		

$$\Sigma P_{COND} + P_{SWF} + P_{SWR} + P_Q + P_{BOOST} = P_{INTERNAL}$$

$$P_{INTERNAL} = \textbf{322 mW}$$

$$T_J = (R_{\theta JC} \times Power) + T_C = 80^oC/W \times 322\ mW + T_C$$

(43)

The second method can give a very accurate silicon junction temperature. The first step is to determine $R_{\theta JA}$ of the application. The LM2734Z has over-temperature protection circuitry. When the silicon temperature reaches 165°C, the device stops switching. The protection circuitry has a hysteresis of 15°C. Once the silicon temperature has decreased to approximately 150°C, the device will start to switch again. Knowing this, the $R_{\theta JA}$ for any PCB can be characterized during the early stages of the design by raising the ambient temperature in the given application until the circuit enters thermal shutdown. If the SW-pin is monitored, it will be obvious when the internal NFET stops switching indicating a junction temperature of 165°C. Knowing the internal power dissipation from the above methods, the junction temperature and the ambient temperature, $R_{\theta JA}$ can be determined.

$$R_{\theta JA} = \frac{165^oC - T_A}{P_{INTERNAL}}$$

(44)

Once this is determined, the maximum ambient temperature allowed for a desired junction temperature can be found.

Design Example 3:			
Operating Conditions			
Package	SOT-6		
V_{IN}	12.0V	P_{OUT}	2.475W
V_{OUT}	3.30V	P_{DIODE}	523mW
I_{OUT}	750mA	P_{IND}	56.25mW
V_D	0.35V	P_{SWF}	108mW
Freq	3MHz	P_{SWR}	108mW
I_Q	1.5mA	P_{COND}	68.2mW
I_{BOOST}	4mA	P_Q	18mW
V_{BOOST}	5V	P_{BOOST}	20mW
T_{RISE}	8ns	P_{LOSS}	**902mW**
T_{FALL}	8ns		
R_{DSON}	400mΩ		
IND_{DCR}	75mΩ		
D	**30.3%**		

$$\Sigma P_{COND} + P_{SWF} + P_{SWR} + P_Q + P_{BOOST} = P_{INTERNAL}$$

$$P_{INTERNAL} = \textbf{322 mW}$$

(45)

Using a standard Texas Instruments Thin SOT-6 demonstration board to determine the $R_{\theta JA}$ of the board. The four layer PCB is constructed using FR4 with 1/2oz copper traces. The copper ground plane is on the bottom layer. The ground plane is accessed by two vias. The board measures 2.5cm x 3cm. It was placed in an oven with no forced airflow.

The ambient temperature was raised to 94°C, and at that temperature, the device went into thermal shutdown.

$$R_{\theta JA} = \frac{165^oC - 94^oC}{322\ mW} = 220^oC/W$$

(46)

If the junction temperature was to be kept below 125°C, then the ambient temperature cannot go above 54.2°C.

$$T_J - (R_{\theta JA} \times P_{LOSS}) = T_A$$

(47)

The method described above to find the junction temperature in the Thin SOT-6 package can also be used to calculate the junction temperature in the WSON package. The 6 pin WSON package has a $R_{\theta JC}$ = 20°C/W, and $R_{\theta JA}$ can vary depending on the application. $R_{\theta JA}$ can be calculated in the same manner as described in method #2 (see example 3).

WSON PACKAGE

The LM2734Z is packaged in a Thin SOT-6 package and the 6–pin WSON. The WSON package has the same footprint as the Thin SOT-6, but is thermally superior due to the exposed ground paddle on the bottom of the package.
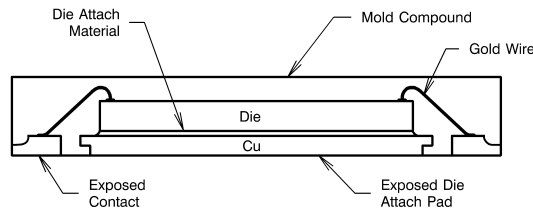
Figure 14. No Pullback WSON Configuration

$R_{\theta JA}$ of the WSON package is normally two to three times better than that of the Thin SOT-6 package for a similar PCB configuration (area, copper weight, thermal vias).
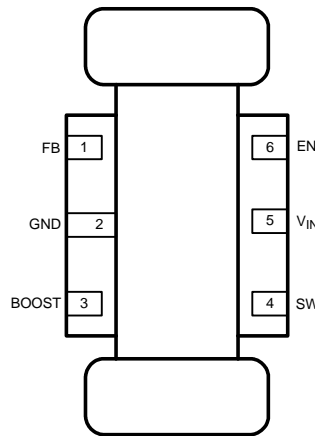
Figure 15. Dog Bone

For certain high power applications, the PCB land may be modified to a "dog bone" shape (see Figure 15). By increasing the size of ground plane, and adding thermal vias, the $R_{\theta JA}$ for the application can be reduced.

Design Example 4:			
Operating Conditions			
Package	WSON-6		
V_{IN}	12.0V	P_{OUT}	2.475W
V_{OUT}	3.3V	P_{DIODE}	523mW
I_{OUT}	750mA	P_{IND}	56.25mW
V_D	0.35V	P_{SWF}	108mW
Freq	3MHz	P_{SWR}	108mW
I_Q	1.5mA	P_{COND}	68.2mW
I_{BOOST}	4mA	P_Q	18mW
V_{BOOST}	5V	P_{BOOST}	20mW
T_{RISE}	8ns	P_{LOSS}	**902mW**
T_{FALL}	8ns		
R_{DSON}	400mΩ		
IND_{DCR}	75mΩ		
D	**30.3%**		

$$\Sigma P_{COND} + P_{SWF} + P_{SWR} + P_Q + P_{BOOST} = P_{INTERNAL}$$

$$P_{INTERNAL} = \textbf{322 mW}$$

(48)

This example follows example 2, but uses the WSON package. Using a standard Texas Instruments WSON-6 demonstration board, use Method 2 to determine $R_{\theta JA}$ of the board. The four layer PCB is constructed using FR4 with 1/2oz copper traces. The copper ground plane is on the bottom layer. The ground plane is accessed by four vias. The board measures 2.5cm x 3cm. It was placed in an oven with no forced airflow.

The ambient temperature was raised to 113°C, and at that temperature, the device went into thermal shutdown.

$$R_{\theta JA}a = \frac{165^oC - 113^oC}{322 \text{ mW}} = 161^oC/W$$

(49)

If the junction temperature is to be kept below 125°C, then the ambient temperature cannot go above 73.2°C.

$$T_J - (R_{\theta JA} \times P_{LOSS}) = T_A$$ (50)

Package Selection

To determine which package you should use for your specific application, variables need to be known before you can determine the appropriate package to use.

1. Maximum ambient system temperature
2. Internal LM2734Z power losses
3. Maximum junction temperature desired
4. $R_{\theta JA}$ of the specific application, or $R_{\theta JC}$ (WSON or Thin SOT-6)

The junction temperature must be less than 125°C for the worst-case scenario.

LM2734Z Design Examples
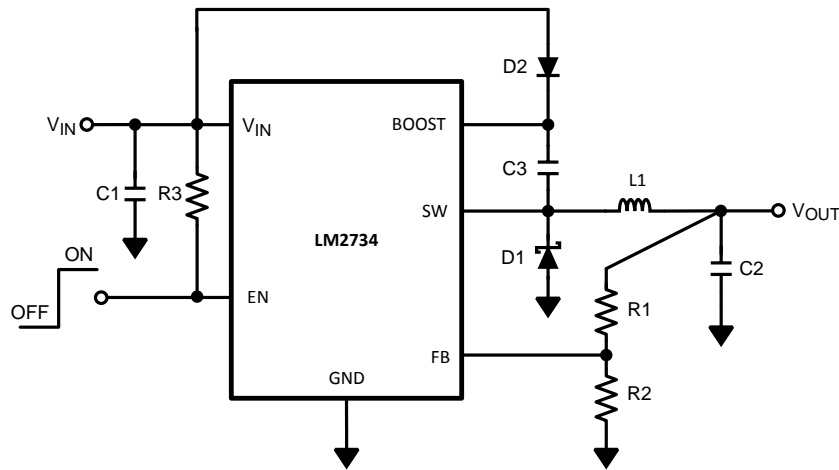
Figure 16. V_{BOOST} Derived from V_{IN}
Operating Conditions: 5V to 1.5V/1A

Table 2. Bill of Materials for Figure 16

Part ID	Part Value	Part Number	Manufacturer
U1	1A Buck Regulator	LM2734ZX	Texas Instruments
C1, Input Cap	10µF, 6.3V, X5R	C3216X5ROJ106M	TDK
C2, Output Cap	10µF, 6.3V, X5R	C3216X5ROJ106M	TDK
C3, Boost Cap	0.01uF, 16V, X7R	C1005X7R1C103K	TDK
D1, Catch Diode	$0.3V_F$ Schottky 1A, 10VR	MBRM110L	ON Semi
D2, Boost Diode	$1V_F$ @ 50mA Diode	1N4148W	Diodes, Inc.
L1	2.2µH, 1.8A	ME3220–222MX	Coilcraft
R1	8.87kΩ, 1%	CRCW06038871F	Vishay
R2	10.2kΩ, 1%	CRCW06031022F	Vishay
R3	100kΩ, 1%	CRCW06031003F	Vishay
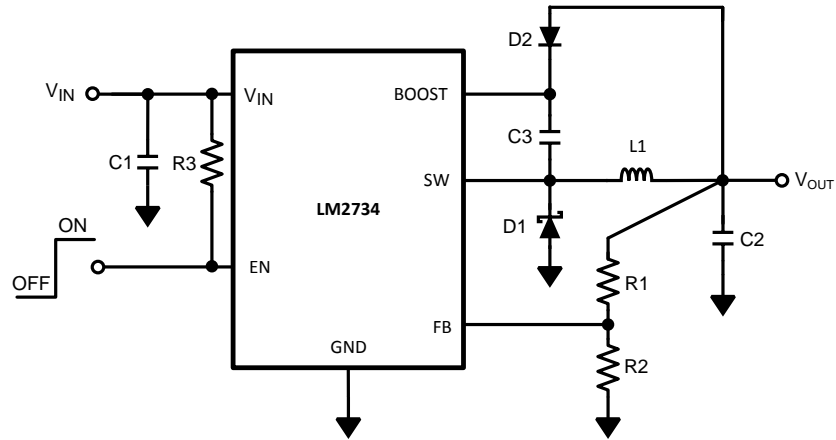
**Figure 17. V$_{BOOST}$ Derived from V$_{OUT}$
12V to 3.3V/1A**

Table 3. Bill of Materials for Figure 17

Part ID	Part Value	Part Number	Manufacturer
U1	1A Buck Regulator	LM2734ZX	Texas Instruments
C1, Input Cap	10µF, 25V, X7R	C3225X7R1E106M	TDK
C2, Output Cap	22µF, 6.3V, X5R	C3216X5ROJ226M	TDK
C3, Boost Cap	0.01µF, 16V, X7R	C1005X7R1C103K	TDK
D1, Catch Diode	0.34V$_F$ Schottky 1A, 30VR	SS1P3L	Vishay
D2, Boost Diode	0.6V$_F$ @ 30mA Diode	BAT17	Vishay
L1	3.3µH, 1.3A	ME3220–332MX	Coilcraft
R1	31.6kΩ, 1%	CRCW06033162F	Vishay
R2	10.0 kΩ, 1%	CRCW06031002F	Vishay
R3	100kΩ, 1%	CRCW06031003F	Vishay

Product Folder Links: *LM2734Z*
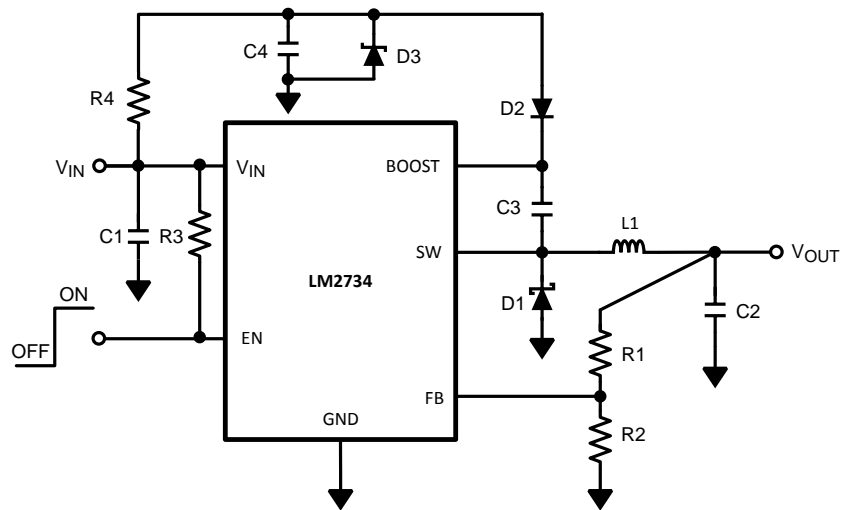
Figure 18. V$_{BOOST}$ Derived from V$_{SHUNT}$
18V to 1.5V/1A

Table 4. Bill of Materials for Figure 18

Part ID	Part Value	Part Number	Manufacturer
U1	1A Buck Regulator	LM2734ZX	Texas Instruments
C1, Input Cap	10μF, 25V, X7R	C3225X7R1E106M	TDK
C2, Output Cap	22μF, 6.3V, X5R	C3216X5ROJ226M	TDK
C3, Boost Cap	0.01μF, 16V, X7R	C1005X7R1C103K	TDK
C4, Shunt Cap	0.1μF, 6.3V, X5R	C1005X5R0J104K	TDK
D1, Catch Diode	0.4V$_F$ Schottky 1A, 30VR	SS1P3L	Vishay
D2, Boost Diode	1V$_F$ @ 50mA Diode	1N4148W	Diodes, Inc.
D3, Zener Diode	5.1V 250Mw SOT	BZX84C5V1	Vishay
L1	3.3μH, 1.3A	ME3220−332MX	Coilcraft
R1	8.87kΩ, 1%	CRCW06038871F	Vishay
R2	10.2kΩ, 1%	CRCW06031022F	Vishay
R3	100kΩ, 1%	CRCW06031003F	Vishay
R4	4.12kΩ, 1%	CRCW06034121F	Vishay

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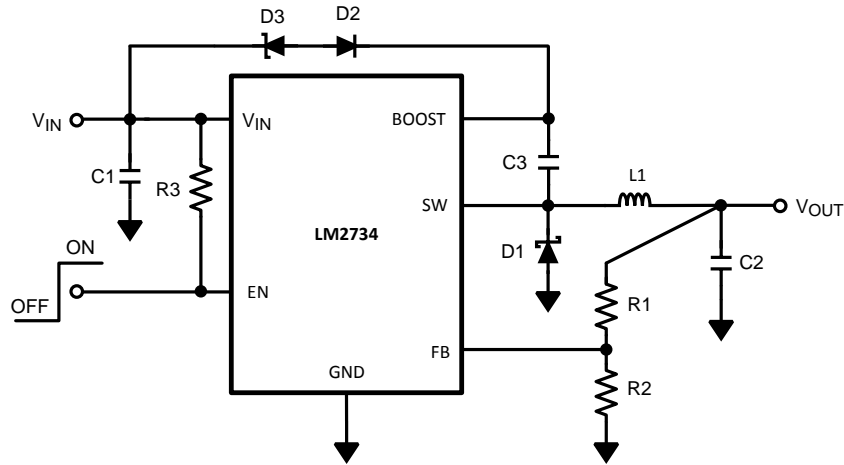
**Figure 19. V_{BOOST} Derived from Series Zener Diode (V_{IN})
15V to 1.5V/1A**

Table 5. Bill of Materials for Figure 19

Part ID	Part Value	Part Number	Manufacturer
U1	1A Buck Regulator	LM2734ZX	Texas Instruments
C1, Input Cap	10µF, 25V, X7R	C3225X7R1E106M	TDK
C2, Output Cap	22µF, 6.3V, X5R	C3216X5ROJ226M	TDK
C3, Boost Cap	0.01µF, 16V, X7R	C1005X7R1C103K	TDK
D1, Catch Diode	0.4V$_F$ Schottky 1A, 30VR	SS1P3L	Vishay
D2, Boost Diode	1V$_F$ @ 50mA Diode	1N4148W	Diodes, Inc.
D3, Zener Diode	11V 350Mw SOT	BZX84C11T	Diodes, Inc.
L1	3.3µH, 1.3A	ME3220–332MX	Coilcraft
R1	8.87kΩ, 1%	CRCW06038871F	Vishay
R2	10.2kΩ, 1%	CRCW06031022F	Vishay
R3	100kΩ, 1%	CRCW06031003F	Vishay

Product Folder Links: *LM2734Z*
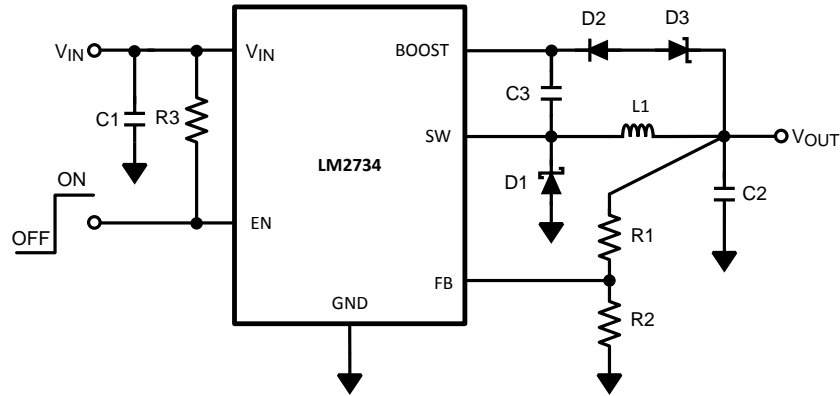
Figure 20. V$_{BOOST}$ Derived from Series Zener Diode (V$_{OUT}$)
15V to 9V/1A

Table 6. Bill of Materials for Figure 20

Part ID	Part Value	Part Number	Manufacturer
U1	1A Buck Regulator	LM2734ZX	Texas Instruments
C1, Input Cap	10µF, 25V, X7R	C3225X7R1E106M	TDK
C2, Output Cap	22µF, 16V, X5R	C3216X5R1C226M	TDK
C3, Boost Cap	0.01µF, 16V, X7R	C1005X7R1C103K	TDK
D1, Catch Diode	0.4V$_F$ Schottky 1A, 30VR	SS1P3L	Vishay
D2, Boost Diode	1V$_F$ @ 50mA Diode	1N4148W	Diodes, Inc.
D3, Zener Diode	4.3V 350mw SOT	BZX84C4V3	Diodes, Inc.
L1	2.2µH, 1.8A	ME3220–222MX	Coilcraft
R1	102kΩ, 1%	CRCW06031023F	Vishay
R2	10.2kΩ, 1%	CRCW06031022F	Vishay
R3	100kΩ, 1%	CRCW06031003F	Vishay

Product Folder Links: LM2734Z

REVISION HISTORY

Changes from Revision D (April 2013) to Revision E   **Page**

Product Folder Links: *LM2734Z*

PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead/Ball Finish	MSL Peak Temp (3)	Op Temp (°C)	Top-Side Markings (4)	Samples
LM2734ZMK	ACTIVE	SOT	DDC	6	1000	TBD	Call TI	Call TI	-40 to 125	SFTB	Samples
LM2734ZMK/NOPB	ACTIVE	SOT	DDC	6	1000	Green (RoHS & no Sb/Br)	CU SN	Level-1-260C-UNLIM	-40 to 125	SFTB	Samples
LM2734ZMKX	ACTIVE	SOT	DDC	6	3000	TBD	Call TI	Call TI	-40 to 125	SFTB	Samples
LM2734ZMKX/NOPB	ACTIVE	SOT	DDC	6	3000	Green (RoHS & no Sb/Br)	CU SN	Level-1-260C-UNLIM	-40 to 125	SFTB	Samples
LM2734ZQMK/NOPB	ACTIVE	SOT	DDC	6	1000	Green (RoHS & no Sb/Br)	CU SN	Level-1-260C-UNLIM	-40 to 125	SVBB	Samples
LM2734ZQMKE/NOPB	ACTIVE	SOT	DDC	6	250	Green (RoHS & no Sb/Br)	CU SN	Level-1-260C-UNLIM	-40 to 125	SVBB	Samples
LM2734ZQMKX/NOPB	ACTIVE	SOT	DDC	6	3000	Green (RoHS & no Sb/Br)	CU SN	Level-1-260C-UNLIM	-40 to 125	SVBB	Samples
LM2734ZQSD/NOPB	ACTIVE	WSON	NGG	6	1000	Green (RoHS & no Sb/Br)	CU SN	Level-3-260C-168 HR	-40 to 125	L238B	Samples
LM2734ZQSDE/NOPB	ACTIVE	WSON	NGG	6	250	Green (RoHS & no Sb/Br)	CU SN	Level-3-260C-168 HR	-40 to 125	L238B	Samples
LM2734ZQSDX/NOPB	ACTIVE	WSON	NGG	6	4500	Green (RoHS & no Sb/Br)	CU SN	Level-3-260C-168 HR	-40 to 125	L238B	Samples
LM2734ZSD	ACTIVE	WSON	NGG	6	1000	TBD	Call TI	Call TI	-40 to 125	L163B	Samples
LM2734ZSD/NOPB	ACTIVE	WSON	NGG	6	1000	Green (RoHS & no Sb/Br)	CU SN	Level-3-260C-168 HR	-40 to 125	L163B	Samples
LM2734ZSDX	ACTIVE	WSON	NGG	6	4500	TBD	Call TI	Call TI	-40 to 125	L163B	Samples
LM2734ZSDX/NOPB	ACTIVE	WSON	NGG	6	4500	Green (RoHS & no Sb/Br)	CU SN	Level-3-260C-168 HR	-40 to 125	L163B	Samples

(1) The marketing status values are defined as follows:
ACTIVE: Product device recommended for new designs.
LIFEBUY: TI has announced that the device will be discontinued, and a lifetime-buy period is in effect.
NRND: Not recommended for new designs. Device is in production to support existing customers, but TI does not recommend using this part in a new design.
PREVIEW: Device has been announced but is not in production. Samples may or may not be available.
OBSOLETE: TI has discontinued the production of the device.

(2) Eco Plan - The planned eco-friendly classification: Pb-Free (RoHS), Pb-Free (RoHS Exempt), or Green (RoHS & no Sb/Br) - please check http://www.ti.com/productcontent for the latest availability information and additional product content details.

TBD: The Pb-Free/Green conversion plan has not been defined.

Pb-Free (RoHS): TI's terms "Lead-Free" or "Pb-Free" mean semiconductor products that are compatible with the current RoHS requirements for all 6 substances, including the requirement that lead not exceed 0.1% by weight in homogeneous materials. Where designed to be soldered at high temperatures, TI Pb-Free products are suitable for use in specified lead-free processes.

Pb-Free (RoHS Exempt): This component has a RoHS exemption for either 1) lead-based flip-chip solder bumps used between the die and package, or 2) lead-based die adhesive used between the die and leadframe. The component is otherwise considered Pb-Free (RoHS compatible) as defined above.

Green (RoHS & no Sb/Br): TI defines "Green" to mean Pb-Free (RoHS compatible), and free of Bromine (Br) and Antimony (Sb) based flame retardants (Br or Sb do not exceed 0.1% by weight in homogeneous material)

(3) MSL, Peak Temp. -- The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

(4) Multiple Top-Side Markings will be inside parentheses. Only one Top-Side Marking contained in parentheses and separated by a "~" will appear on a device. If a line is indented then it is a continuation of the previous line and the two combined represent the entire Top-Side Marking for that device.

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OTHER QUALIFIED VERSIONS OF LM2734Z, LM2734Z-Q1 :

• Catalog: LM2734Z

• Automotive: LM2734Z-Q1

NOTE: Qualified Version Definitions:

   • Catalog - TI's standard catalog product

   • Automotive - Q100 devices qualified for high-reliability automotive applications targeting zero defects

TAPE AND REEL INFORMATION

REEL DIMENSIONS

TAPE DIMENSIONS

A0	Dimension designed to accommodate the component width
B0	Dimension designed to accommodate the component length
K0	Dimension designed to accommodate the component thickness
W	Overall width of the carrier tape
P1	Pitch between successive cavity centers

QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE

*All dimensions are nominal

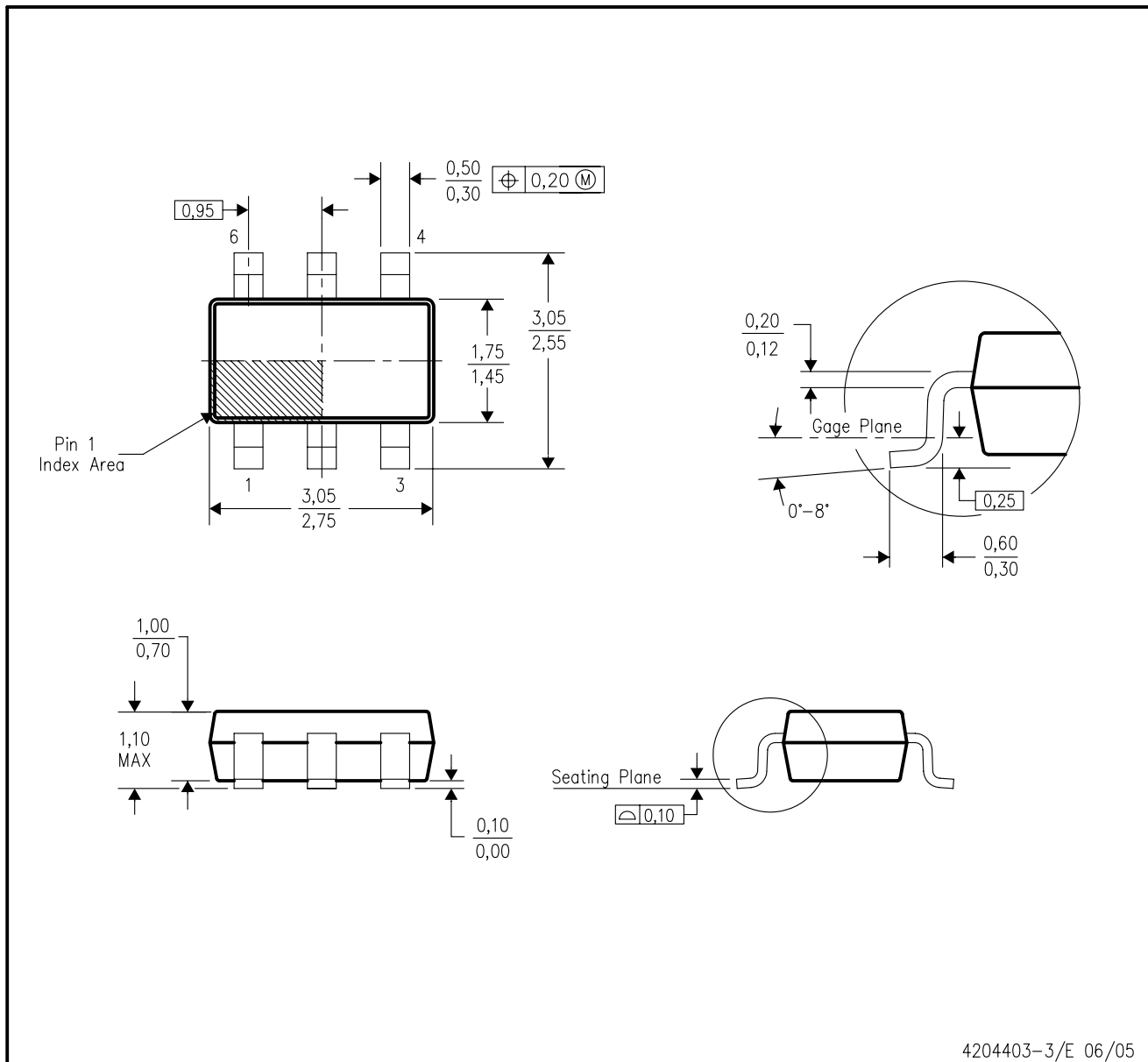| Device	Package Type	Package Drawing	Pins	SPQ	Reel Diameter (mm)	Reel Width W1 (mm)	A0 (mm)	B0 (mm)	K0 (mm)	P1 (mm)	W (mm)	Pin1 Quadrant
LM2734ZMK	SOT	DDC	6	1000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZMK/NOPB	SOT	DDC	6	1000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZMKX	SOT	DDC	6	3000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZMKX/NOPB	SOT	DDC	6	3000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZQMK/NOPB	SOT	DDC	6	1000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZQMKE/NOPB	SOT	DDC	6	250	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZQMKX/NOPB	SOT	DDC	6	3000	178.0	8.4	3.2	3.2	1.4	4.0	8.0	Q3
LM2734ZQSD/NOPB	WSON	NGG	6	1000	178.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZQSDE/NOPB	WSON	NGG	6	250	178.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZQSDX/NOPB	WSON	NGG	6	4500	330.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZSD	WSON	NGG	6	1000	178.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZSD/NOPB	WSON	NGG	6	1000	178.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZSDX	WSON	NGG	6	4500	330.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1
LM2734ZSDX/NOPB	WSON	NGG	6	4500	330.0	12.4	3.3	3.3	1.0	8.0	12.0	Q1

TAPE AND REEL BOX DIMENSIONS

*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
LM2734ZMK	SOT	DDC	6	1000	210.0	185.0	35.0
LM2734ZMK/NOPB	SOT	DDC	6	1000	210.0	185.0	35.0
LM2734ZMKX	SOT	DDC	6	3000	210.0	185.0	35.0
LM2734ZMKX/NOPB	SOT	DDC	6	3000	210.0	185.0	35.0
LM2734ZQMK/NOPB	SOT	DDC	6	1000	210.0	185.0	35.0
LM2734ZQMKE/NOPB	SOT	DDC	6	250	210.0	185.0	35.0
LM2734ZQMKX/NOPB	SOT	DDC	6	3000	210.0	185.0	35.0
LM2734ZQSD/NOPB	WSON	NGG	6	1000	213.0	191.0	55.0
LM2734ZQSDE/NOPB	WSON	NGG	6	250	213.0	191.0	55.0
LM2734ZQSDX/NOPB	WSON	NGG	6	4500	367.0	367.0	35.0
LM2734ZSD	WSON	NGG	6	1000	203.0	190.0	41.0
LM2734ZSD/NOPB	WSON	NGG	6	1000	203.0	190.0	41.0
LM2734ZSDX	WSON	NGG	6	4500	367.0	367.0	35.0
LM2734ZSDX/NOPB	WSON	NGG	6	4500	367.0	367.0	35.0

DDC (R−PDSO−G6)   PLASTIC SMALL−OUTLINE

0,95

6   4
0,50
0,30   ⊕ 0,20 Ⓜ

3,05
2,55

1,75
1,45

Pin 1
Index Area

1   3
3,05
2,75

0,20
0,12

Gage Plane

0°−8°

0,25

0,60
0,30

1,00
0,70

1,10
MAX

0,10
0,00

Seating Plane
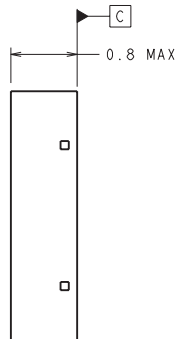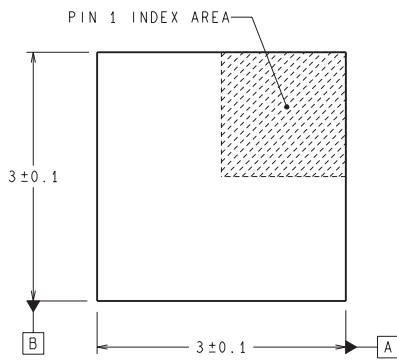
⌒ 0,10

4204403−3/E 06/05

NOTES: A.   All linear dimensions are in millimeters.
         B.   This drawing is subject to change without notice.
         C.   Body dimensions do not include mold flash or protrusion.
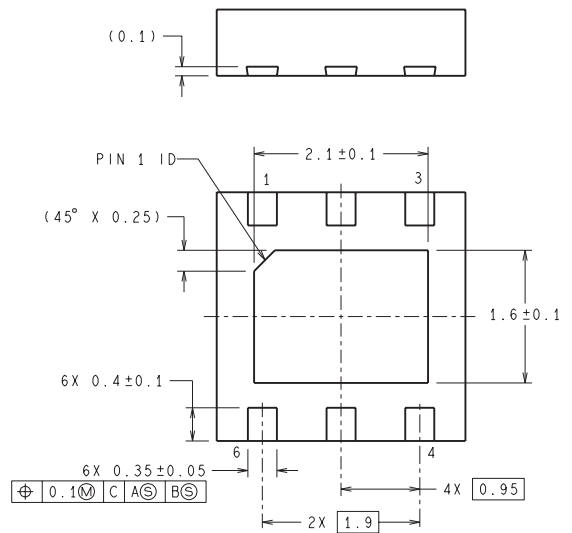   D.   Falls within JEDEC MO−193 variation AA (6 pin).

NGG0006A

DIMENSIONS ARE IN MILLIMETERS
DIMENSION IN ( ) FOR REFERENCE ONLY
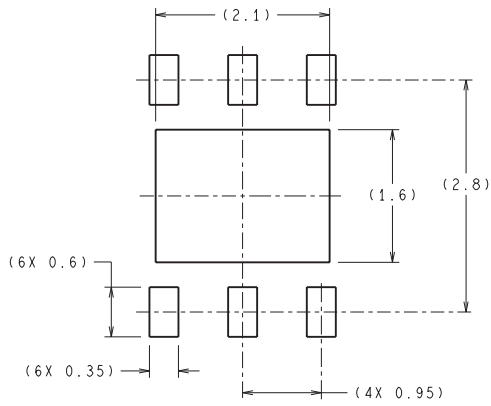
RECOMMENDED LAND PATTERN

PIN 1 INDEX AREA

PIN 1 ID

SDE06A (Rev A)

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