

LM25010/LM25010Q

42V, 1.0A Step-Down Switching Regulator

General Description

The LM25010 features all the functions needed to implement a low cost, efficient, buck regulator capable of supplying in excess of 1A load current. This high voltage regulator integrates an N-Channel Buck Switch, and is available in thermally enhanced LLP-10 and TSSOP-14EP packages. The constant on-time regulation scheme requires no loop compensation resulting in fast load transient response and simplified circuit implementation. The operating frequency remains constant with line and load variations due to the inverse relationship between the input voltage and the on-time. The valley current limit detection is set at 1.25A. Additional features include: VCC under-voltage lock-out, thermal shutdown, gate drive under-voltage lock-out, and maximum duty cycle limiter.

Features

- Wide 6V to 42V Input Voltage Range
- Valley Current Limiting At 1.25A
- Programmable Switching Frequency Up To 1 MHz
- Integrated N-Channel Buck Switch
- Integrated High Voltage Bias Regulator
- No Loop Compensation Required
- Ultra-Fast Transient Response
- Nearly Constant Operating Frequency With Line and Load Variations
- Adjustable Output Voltage
- 2.5V, $\pm 2\%$ Feedback Reference
- Programmable Soft-Start
- Thermal shutdown
- LM25010Q is AEC-Q100 Grade 1 & 0 qualified

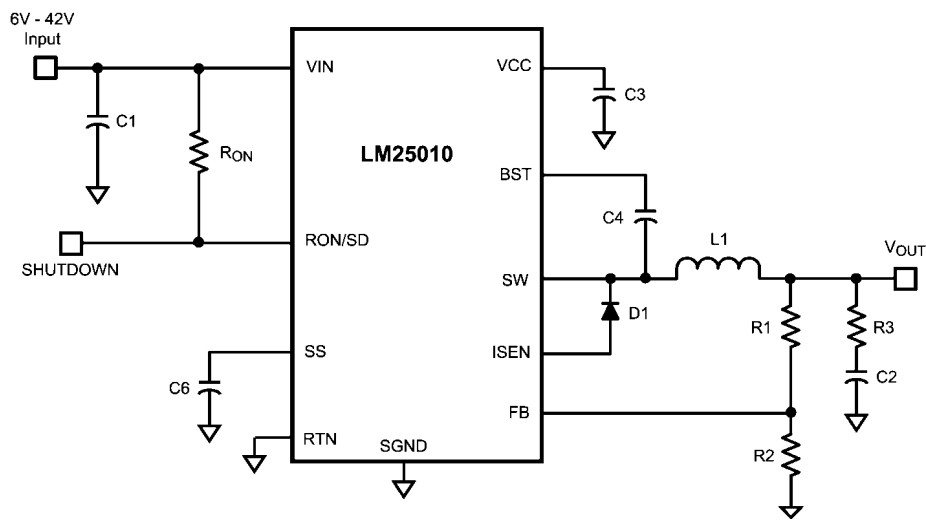
Typical Applications

- Non-Isolated Telecommunications Regulator
- Secondary Side Post Regulator
- Automotive Electronics

Package

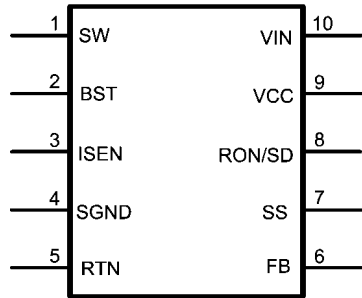
- LLP-10 (4 mm x 4 mm)
- TSSOP-14EP
- Both Packages Have Exposed Thermal Pad For Improved Heat Dissipation

Basic Step-Down Regulator

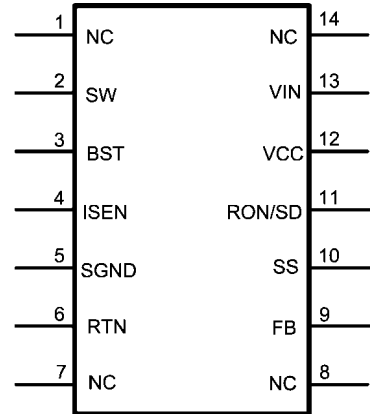


20172743

Connection Diagrams



20172702



20172703

Ordering Information

Order Number	Package Type	NSC Package Drawing	Supplied As	Automotive Grade*
LM25010SD	LLP-10 (4x4)	SDC10A	1000 Units on Tape and Reel	No
LM25010SDX	LLP-10 (4x4)	SDC10A	4500 Units on Tape and Reel	No
LM25010MH	TSSOP-14EP	MXA14A	94 Units in Rail	No
LM25010MHX	TSSOP-14EP	MXA14A	2500 Units on Tape and Reel	No
LM25010Q1MH	TSSOP-14EP	MXA14A	94 Units in Rail	Grade 1
LM25010Q1MHX	TSSOP-14EP	MXA14A	2500 Units on Tape and Reel	Grade 1
LM25010Q0MH	TSSOP-14EP	MXA14A	94 Units in Rail	Grade 0
LM25010Q0MHX	TSSOP-14EP	MXA14A	2500 Units on Tape and Reel	Grade 0

*Automotive Grade (Q) product incorporates enhanced manufacturing and support processes for the automotive market, including defect detection methodologies. Reliability qualification is compliant with the requirements and temperature grades defined in the AEC-Q100 standard. Automotive grade products are identified with the letter Q. For more information go to <http://www.national.com/automotive>.

Pin Descriptions

Pin Number		Name	Description	Application Information
LLP-10	TSSOP-14			
1	2	SW	Switching Node	Internally connected to the buck switch source. Connect to the inductor, free-wheeling diode, and bootstrap capacitor.
2	3	BST	Boost pin for bootstrap capacitor	Connect a capacitor from SW to the BST pin. The capacitor is charged from VCC via an internal diode during the buck switch off-time.
3	4	ISEN	Current sense	During the buck switch off-time, the inductor current flows through the internal sense resistor, and out of the ISEN pin to the free-wheeling diode. The current limit comparator keeps the buck switch off if the ISEN current exceeds 1.25A (typical).
4	5	SGND	Current Sense Ground	Re-circulating current flows into this pin to the current sense resistor.
5	6	RTN	Circuit Ground	Ground return for all internal circuitry other than the current sense resistor.
6	9	FB	Voltage feedback input from the regulated output	Input to both the regulation and over-voltage comparators. The FB pin regulation level is 2.5V.
7	10	SS	Softstart	An internal 11.5 μ A current source charges the SS pin capacitor to 2.5V to soft-start the reference input of the regulation comparator.
8	11	RON/SD	On-time control and shutdown	An external resistor from VIN to the RON/SD pin sets the buck switch on-time. Grounding this pin shuts down the regulator.
9	12	VCC	Output of the bias regulator	The voltage at VCC is nominally equal to V_{IN} for $V_{IN} < 8.9V$, and regulated at 7V for $V_{IN} > 8.9V$. Connect a 0.47 μ F, or larger capacitor from VCC to ground, as close as possible to the pins. An external voltage can be applied to this pin to reduce internal dissipation if V_{IN} is greater than 8.9V. MOSFET body diodes clamp VCC to VIN if $V_{CC} > V_{IN}$.
10	13	VIN	Input supply voltage	Nominal input range is 6V to 42V. Input bypass capacitors should be located as close as possible to the VIN pin and RTN pins.
	1,7,8,14	NC	No connection.	No internal connection. Can be connected to ground plane to improve heat dissipation.
		EP	Exposed Pad	Exposed metal pad on the underside of the device. It is recommended to connect this pad to the PC board ground plane to aid in heat dissipation.

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

VIN to RTN	-0.3V to 45V
BST to RTN	-0.3V to 59V
SW to RTN (Steady State)	-1.5V
BST to VCC	45V
BST to SW	14V
VCC to RTN	-0.3V to 14V
SGND to RTN	-0.3V to +0.3V
SS to RTN	-0.3V to 4V

VIN to SW	45V
All Other Inputs to RTN	-0.3V to 7V
ESD Rating (Note 2)	
Human Body Model	2kV
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 4 sec) (Note 4)	260°C

Operating Ratings (Note 1)

VIN Voltage	6.0V to 42V
Junction Temperature	
LM25010/LM25010Q1	-40°C to +125°C
LM25010Q0	-40°C to +150°C

Electrical Characteristics Specifications with standard type are for $T_J = 25^\circ\text{C}$ only; limits in boldface type apply over the full Operating Junction Temperature (T_J) range. Minimum and Maximum limits are guaranteed through test, design, or statistical correlation. Typical values represent the most likely parametric norm at $T_J = 25^\circ\text{C}$, and are provided for reference purposes only. Unless otherwise stated the following conditions apply: $V_{IN} = 24\text{V}$, $R_{ON} = 200\text{k}\Omega$. See (Note 5).

Symbol	Parameter	Conditions	Min	Typ	Max	Units
V_{CC} Regulator						
V _{CC} Reg	V _{CC} regulated output		6.6	7	7.4	Volts
	V _{IN} - V _{CC}	I _{CC} = 0 mA, F _S ≤ 200 kHz, 6.0V ≤ V _{IN} ≤ 8.5V		100		mV
	V _{CC} Bypass Threshold	V _{IN} Increasing		8.9		V
	V _{CC} Bypass Hysteresis	V _{IN} Decreasing		260		mV
	V _{CC} output impedance	V _{IN} = 6.0V		55		Ω
	(0 mA ≤ I _{CC} ≤ 5 mA)	V _{IN} = 8.0V		50		
		V _{IN} = 24V		0.21		
	V _{CC} current limit (Note 3)	V _{IN} = 24V, V _{CC} = 0V		15		mA
UVLO _{VCC}	V _{CC} under-voltage lock-out threshold	V _{CC} Increasing		5.25		V
	UVLO _{VCC} hysteresis	V _{CC} Decreasing		180		mV
	UVLO _{VCC} filter delay	100 mV overdrive		3		μs
	I _{IN} operating current	Non-switching, FB = 3V		645	920	μA
	I _{IN} shutdown current	RON/SD = 0V		90	170	μA
Switch Characteristics						
R _{DS(on)}	Buck Switch R _{DS(on)} @ f _{SW} = 200 mA	T _J ≤ 125°C T _J ≤ 150°C		0.35	0.80 0.85	Ω
UVLO _{GD}	Gate Drive UVLO	V _{BST} - V _{SW} Increasing	1.7	3.0	4.0	V
	UVLO _{GD} hysteresis			400		mV
SOFT-START Pin						
I _{SS}	Internal current source		8.0	11.5	15	μA
Current Limit						
I _{LIM}	Threshold	Current out of I _{SEN}	1	1.25	1.5	A
	Resistance from ISEN to SGND			130		mΩ
	Response time			150		ns
On Timer, RON/SD Pin						
t _{ON} - 1	On-time	V _{IN} = 10V, R _{ON} = 200 kΩ	2.1	2.75	3.4	μs
t _{ON} - 2	On-time	V _{IN} = 42V, R _{ON} = 200 kΩ	500	695	890	ns
	Shutdown threshold	Voltage at RON/SD rising	0.30	0.7	1.05	V
	Threshold hysteresis			40		mV

Symbol	Parameter	Conditions	Min	Typ	Max	Units
Off Timer						
t_{OFF}	Minimum Off-time			260		ns
Regulation and Over-Voltage Comparators (FB Pin)						
V_{REF}	FB regulation threshold	$T_J \leq 125^\circ\text{C}$	2.445	2.50	2.550	V
		$T_J \leq 150^\circ\text{C}$	2.435			
	FB over-voltage threshold			2.9		V
	FB bias current			1		nA
Thermal Shutdown						
T_{SD}	Thermal shutdown temperature			175		$^\circ\text{C}$
	Thermal shutdown hysteresis			20		$^\circ\text{C}$
Thermal Resistance						
θ_{JA}	Junction to Ambient, 0 LFPM Air Flow	SDC Package		40		$^\circ\text{C/W}$
		MXA Package		40		
θ_{JC}	Junction to Case	SDC Package		5.2		$^\circ\text{C/W}$
		MXA Package		5.2		

Note 1: Absolute Maximum Ratings are limits beyond which damage to the device may occur. Operating Ratings are conditions under which operation of the device is intended to be functional. For guaranteed specifications and test conditions, see the Electrical Characteristics.

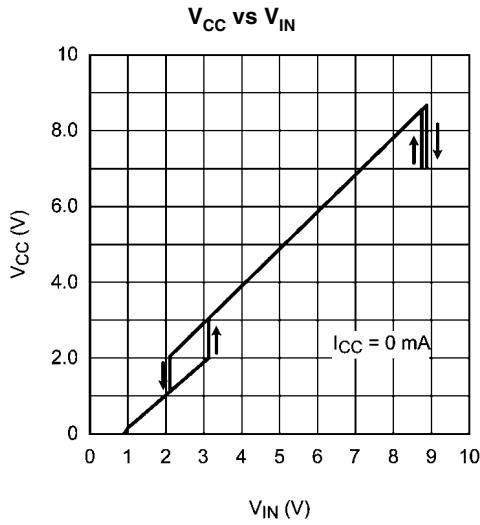
Note 2: The human body model is a 100pF capacitor discharged through a 1.5k Ω resistor into each pin.

Note 3: V_{CC} provides bias for the internal gate drive and control circuits. Device thermal limitations limit external loading.

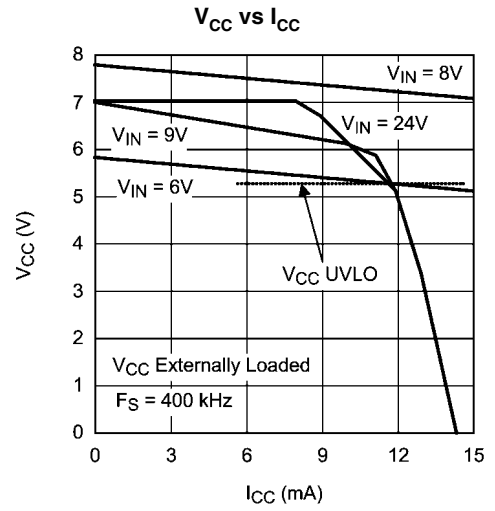
Note 4: For detailed information on soldering plastic TSSOP and LLP packages refer to the Packaging Data Book available from National Semiconductor Corporation.

Note 5: Typical specifications represent the most likely parametric norm at 25 $^\circ\text{C}$ operation.

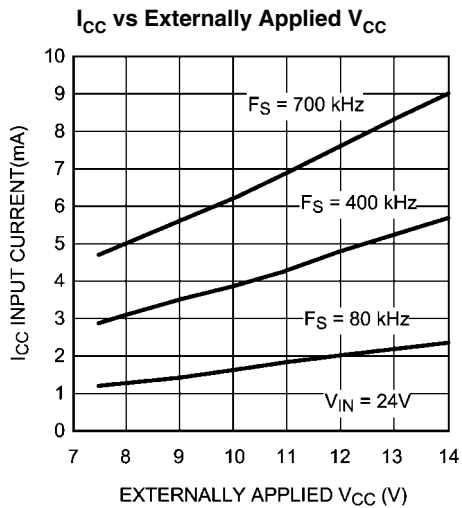
Typical Performance Characteristics



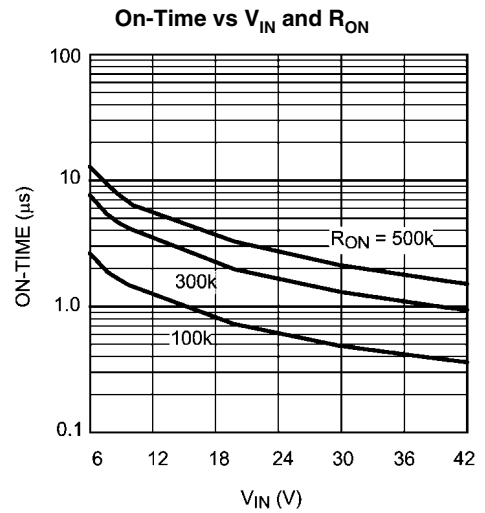
20172704



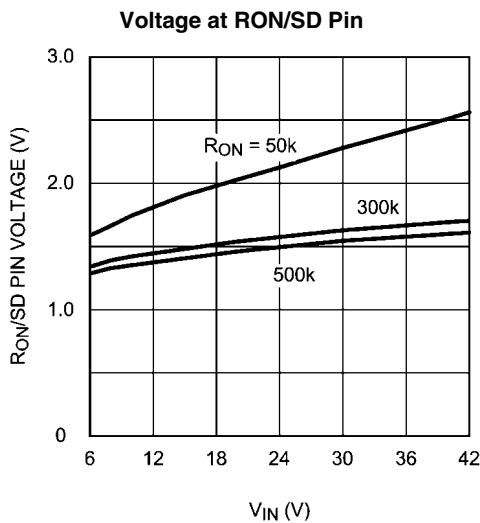
20172705



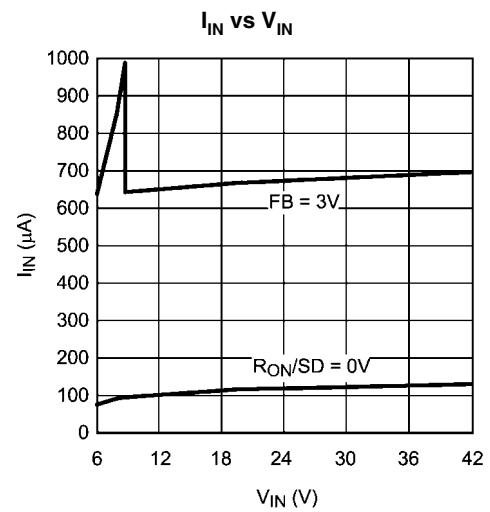
20172706



20172707

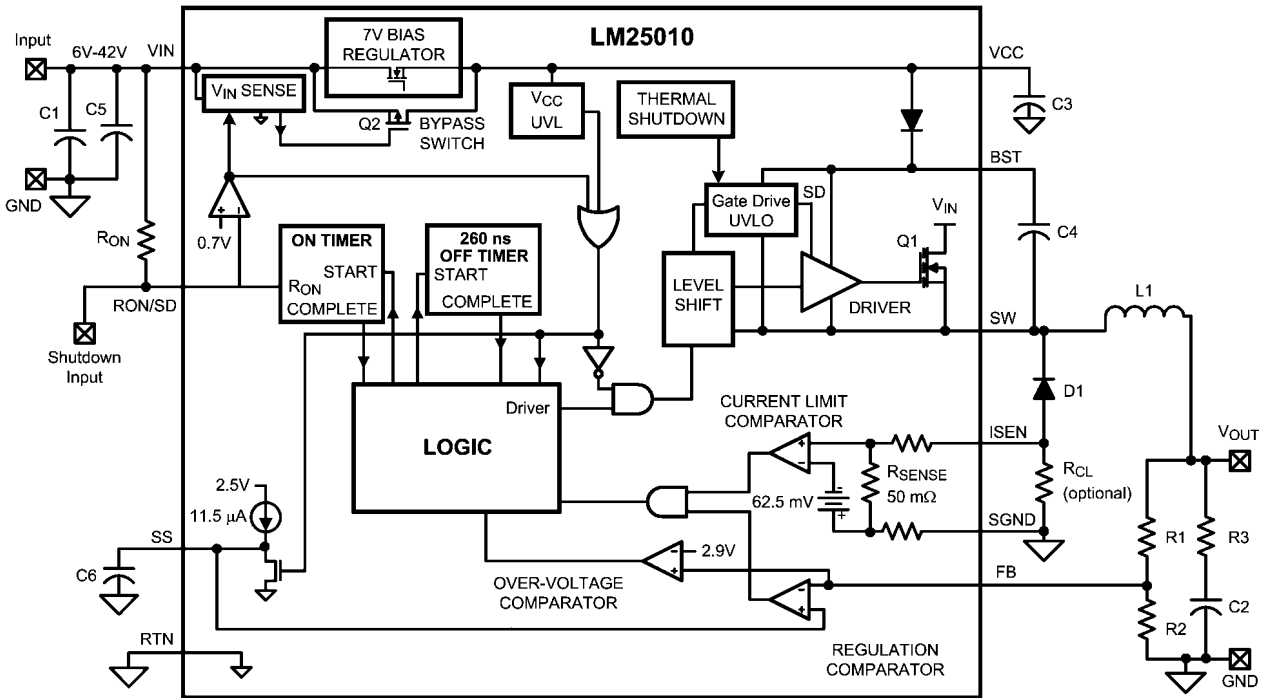


20172708

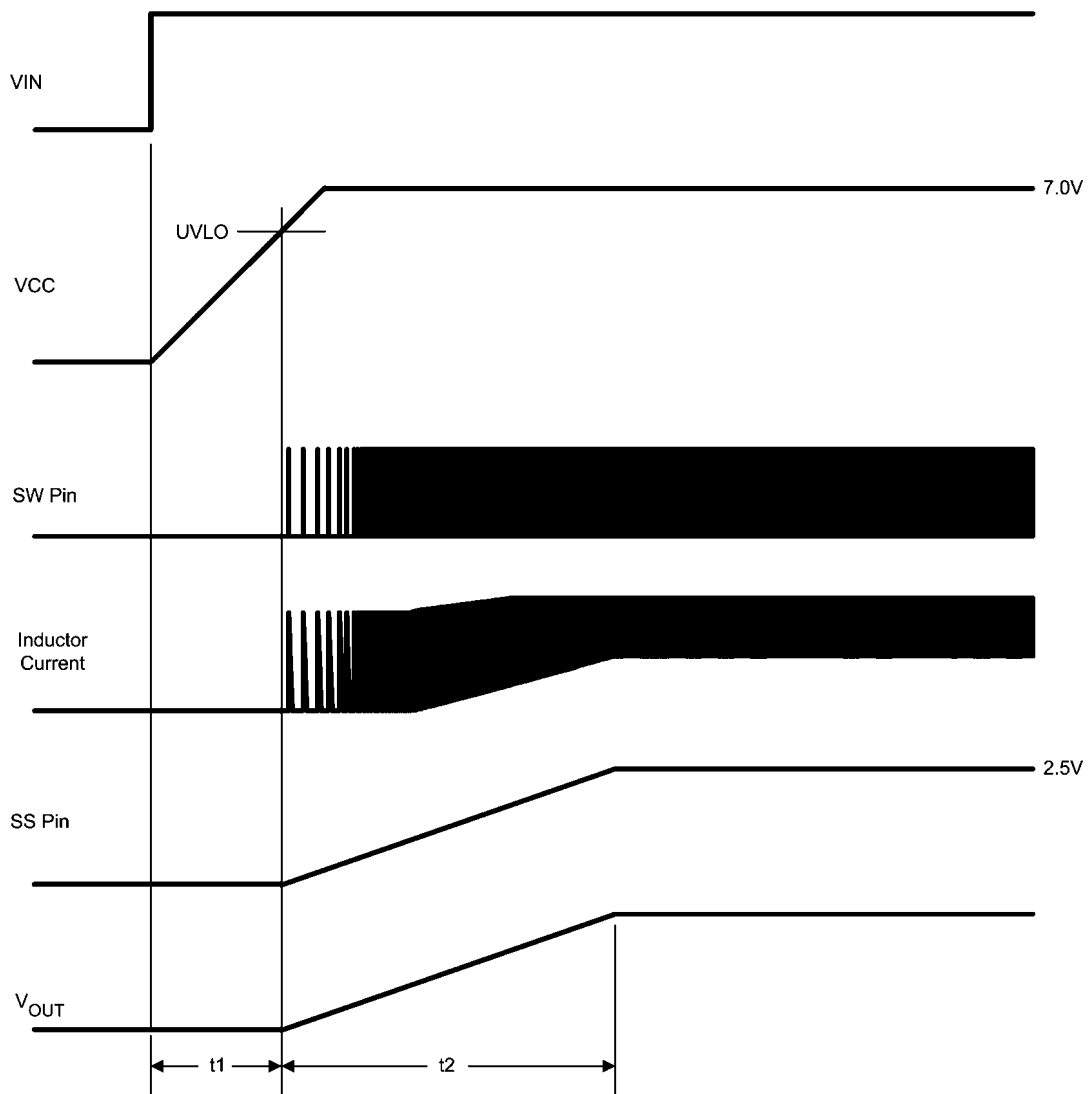


20172710

Block Diagram



20172744



20172711

FIGURE 1. Startup Sequence

Functional Description

The LM25010 Step-Down Switching Regulator features all the functions needed to implement a low cost, efficient buck DC-DC converter capable of supplying in excess of 1A to the load. This high voltage regulator integrates an N-Channel buck switch, with an easy to implement constant on-time controller. It is available in the thermally enhanced LLP-10 and TSSOP-14EP packages. The regulator compares the feedback voltage to a 2.5V reference to control the buck switch, and provides a switch on-time which varies inversely with VIN. This feature results in the operating frequency remaining relatively constant with load and input voltage variations. The switching frequency can range from less than 100 kHz to 1.0 MHz. The regulator requires no loop compensation resulting in very fast load transient response. The valley current limit circuit holds the buck switch off until the free-wheeling inductor current falls below the current limit threshold, nominally set at 1.25A.

The LM25010 can be applied in numerous applications to efficiently step-down higher DC voltages. Features include:

Thermal shutdown, VCC under-voltage lock-out, gate drive under-voltage lock-out, and maximum duty cycle limit.

Control Circuit Overview

The LM25010 employs a control scheme based on a comparator and a one-shot on-timer, with the output voltage feedback (FB) compared to an internal reference (2.5V). If the FB voltage is below the reference the buck switch is turned on for a time period determined by the input voltage and a programming resistor (R_{ON}). Following the on-time the switch remains off for a fixed 260 ns off-time, or until the FB voltage falls below the reference, whichever is longer. The buck switch then turns on for another on-time period. Referring to the Block Diagram, the output voltage is set by R1 and R2. The regulated output voltage is calculated as follows:

$$V_{OUT} = 2.5V \times (R1 + R2) / R2 \quad (1)$$

The LM25010 requires a minimum of 25 mV of ripple voltage at the FB pin for stable fixed-frequency operation. If the output capacitor's ESR is insufficient additional series resistance may be required (R3 in the Block Diagram).

The LM25010 operates in continuous conduction mode at heavy load currents, and discontinuous conduction mode at light load currents. In continuous conduction mode current always flows through the inductor, never decaying to zero during the off-time. In this mode the operating frequency remains relatively constant with load and line variations. The minimum load current for continuous conduction mode is one-half the inductor's ripple current amplitude. The operating frequency in the continuous conduction mode is calculated as follows:

$$F_S = \frac{V_{OUT} \times (V_{IN} - 1.4V)}{1.18 \times 10^{-10} \times (R_{ON} + 1.4 \text{ k}\Omega) \times V_{IN}} \quad (2)$$

The buck switch duty cycle is equal to:

$$DC = \frac{t_{ON}}{t_{ON} + t_{OFF}} = t_{ON} \times F_S = \frac{V_{OUT}}{V_{IN}} \quad (3)$$

Under light load conditions, the LM25010 operates in discontinuous conduction mode, with zero current flowing through the inductor for a portion of the off-time. The operating frequency is always lower than that of the continuous conduction mode, and the switching frequency varies with load current. Conversion efficiency is maintained at a relatively high level at light loads since the switching losses diminish as the power delivered to the load is reduced. The discontinuous mode operating frequency is approximately:

$$F_S = \frac{V_{OUT}^2 \times L1 \times 1.4 \times 10^{20}}{R_L \times R_{ON}^2} \quad (4)$$

where R_L = the load resistance.

Start-Up Bias Regulator (V_{CC})

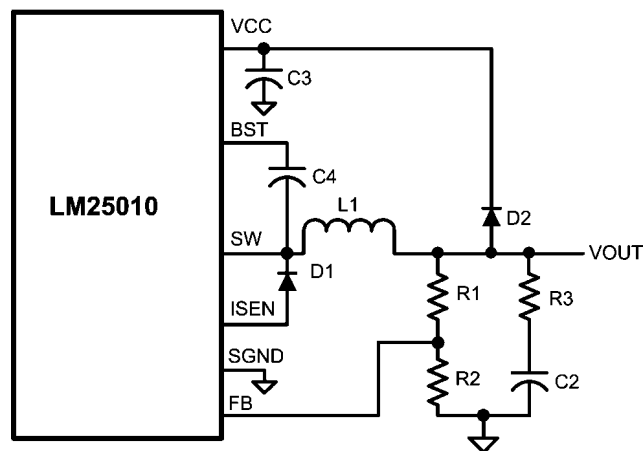
A high voltage bias regulator is integrated within the LM25010. The input pin (V_{IN}) can be connected directly to

line voltages between 6V and 42V. Referring to the block diagram and the graph of V_{CC} vs. V_{IN} , when V_{IN} is between 6V and the bypass threshold (nominally 8.9V), the bypass switch (Q2) is on, and V_{CC} tracks V_{IN} within 100 mV to 150 mV. The bypass switch on-resistance is approximately 50 Ω , with inherent current limiting at approximately 100 mA. When V_{IN} is above the bypass threshold, Q2 is turned off, and V_{CC} is regulated at 7V. The V_{CC} regulator output current is limited at approximately 15 mA. When the LM25010 is shutdown using the RON/SD pin, the V_{CC} bypass switch is shut off, regardless of the voltage at V_{IN} .

When V_{IN} exceeds the bypass threshold, the time required for Q2 to shut off is approximately 2 - 3 μ s. The capacitor at VCC (C3) must be a minimum of 0.47 μ F to prevent the voltage at VCC from rising above its absolute maximum rating in response to a step input applied at V_{IN} . C3 must be located as close as possible to the LM25010 pins.

In applications with a relatively high input voltage, power dissipation in the bias regulator is a concern. An auxiliary voltage of between 7.5V and 14V can be diode connected to the VCC pin (D2 in Figure 2) to shut off the VCC regulator, reducing internal power dissipation. The current required into the VCC pin is shown in the Typical Performance Characteristics. Internally a diode connects VCC to V_{IN} requiring that the auxiliary voltage be less than V_{IN} .

The turn-on sequence is shown in Figure 1. When VCC exceeds the under-voltage lock-out (UVLO) of 5.25V (t_1 in Figure 1), the buck switch is enabled, and the SS pin is released to allow the soft-start capacitor (C6) to charge up. The output voltage V_{OUT} is regulated at a reduced level which increases to the desired value as the soft-start voltage increases (t_2 in Figure 1).



20172716

FIGURE 2. Self Biased Configuration

Regulation Comparator

The feedback voltage at the FB pin is compared to the voltage at the SS pin (2.5V, $\pm 2\%$). In normal operation an on-time period is initiated when the voltage at FB falls below 2.5V. The buck switch conducts for the on-time programmed by R_{ON} ,

causing the FB voltage to rise above 2.5V. After the on-time period the buck switch remains off until the FB voltage falls below 2.5V. Input bias current at the FB pin is less than 5 nA over temperature.

Over-Voltage Comparator

The feedback voltage at FB is compared to an internal 2.9V reference. If the voltage at FB rises above 2.9V the on-time is immediately terminated. This condition can occur if the input voltage, or the output load, changes suddenly. The buck switch remains off until the voltage at FB falls below 2.5V.

ON-Time Control

The on-time of the internal buck switch is determined by the R_{ON} resistor and the input voltage (V_{IN}), and is calculated as follows:

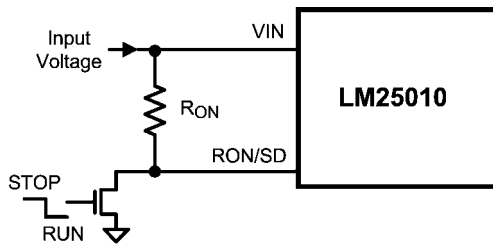
$$t_{ON} = \frac{1.18 \times 10^{-10} \times (R_{ON} + 1.4k)}{(V_{IN} - 1.4V)} + 67 \text{ ns} \quad (5)$$

The R_{ON} resistor can be determined from the desired on-time by re-arranging Equation 5 to the following:

$$R_{ON} = \frac{(t_{ON} - 67 \text{ ns}) \times (V_{IN} - 1.4V)}{1.18 \times 10^{-10}} - 1.4 \text{ k}\Omega \quad (6)$$

To set a specific continuous conduction mode switching frequency (F_S), the R_{ON} resistor is determined from the following:

$$R_{ON} = \frac{V_{OUT} \times (V_{IN} - 1.4V)}{V_{IN} \times F_S \times 1.18 \times 10^{-10}} - 1.4 \text{ k}\Omega \quad (7)$$



20172718

FIGURE 3. Shutdown Implementation

Current Limit

Current limit detection occurs during the off-time by monitoring the recirculating current through the internal current sense resistor (R_{SENSE}). The detection threshold is 1.25A, $\pm 0.25A$. Referring to the Block Diagram, if the current into SGND during the off-time exceeds the threshold level the current limit comparator delays the start of the next on-time period. The next on-time starts when the current into SGND is below the threshold **and** the voltage at FB is below 2.5V. *Figure 4* illustrates the inductor current waveform during normal operation and during current limit. The output current I_O is the average of the inductor ripple current waveform. The Low Load Current waveform illustrates continuous conduction mode operation with peak and valley inductor currents below the current limit threshold. When the load current is increased (High Load Current), the ripple waveform maintains the same amplitude and frequency since the current falls below the current limit threshold at the valley of the ripple waveform. Note the average current in the High Load Current portion of *Figure 4* is above the current limit threshold. Since the current reduces below the threshold in the normal off-time each cycle, the start

of each on-time is not delayed, and the circuit's output voltage is regulated at the correct value. When the load current is further increased such that the lower peak would be above the threshold, the off-time is lengthened to allow the current to decrease to the threshold before the next on-time begins (Current Limited portion of *Figure 4*). Both V_{OUT} and the switching frequency are reduced as the circuit operates in a constant current mode. The load current (I_{OCL}) is equal to the current limit threshold plus half the ripple current ($\Delta I/2$). The ripple amplitude (ΔI) is calculated from:

$$t_{ON(min)} = \frac{V_{OUT} \times 300 \text{ ns}}{(V_{IN(min)} - V_{OUT})} \quad (8)$$

Shutdown

The LM25010 can be remotely shut down by forcing the RON/SD pin below 0.7V with a switch or open drain device. See *Figure 3*. In the shutdown mode the SS pin is internally grounded, the on-time one-shot is disabled, the input current at V_{IN} is reduced, and the V_{CC} bypass switch is turned off. The V_{CC} regulator is not disabled in the shutdown mode. Releasing the RON/SD pin allows normal operation to resume. The nominal voltage at RON/SD is shown in the Typical Performance Characteristics. When switching the RON/SD pin, the transition time should be faster than one to two cycles of the regulator's nominal switching frequency.

of each on-time is not delayed, and the circuit's output voltage is regulated at the correct value. When the load current is further increased such that the lower peak would be above the threshold, the off-time is lengthened to allow the current to decrease to the threshold before the next on-time begins (Current Limited portion of *Figure 4*). Both V_{OUT} and the switching frequency are reduced as the circuit operates in a constant current mode. The load current (I_{OCL}) is equal to the current limit threshold plus half the ripple current ($\Delta I/2$). The ripple amplitude (ΔI) is calculated from:

$$\Delta I = \frac{(V_{IN} - V_{OUT}) \times t_{ON}}{L1} \quad (9)$$

The current limit threshold can be increased by connecting an external resistor (R_{CL}) between SGND and ISEN. R_{CL} typically is less than 1 Ω , and the calculation of its value is explained in the Applications Information section. If the current limit threshold is increased by adding R_{CL} , the maximum continuous load current should not exceed 1.5A, and the peak current out of the SW pin should not exceed 2A.

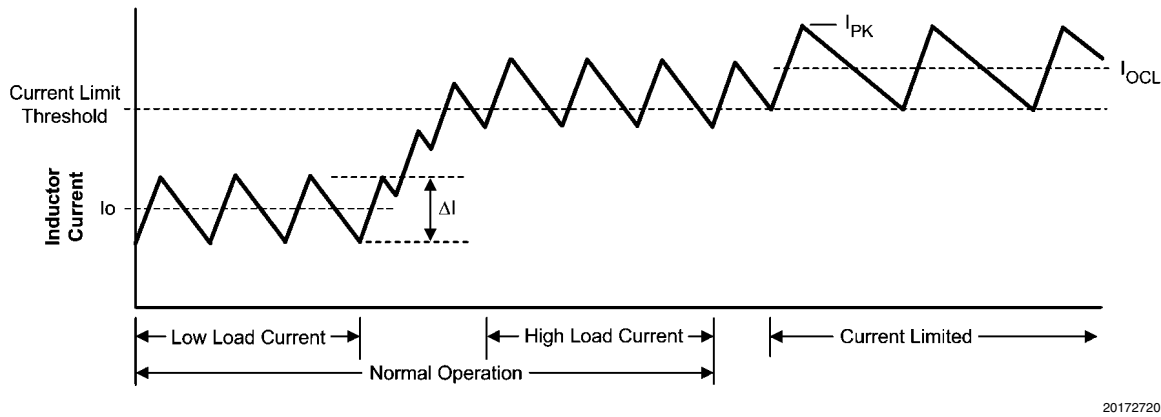


FIGURE 4. Inductor Current - Current Limit Operation

N - Channel Buck Switch and Driver

The LM25010 integrates an N-Channel buck switch and associated floating high voltage gate driver. The peak current through the buck switch should not exceed 2A, and the load current should not exceed 1.5A. The gate driver circuit is powered by the external bootstrap capacitor between BST and SW (C4), which is recharged each off-time from V_{CC} through the internal high voltage diode. The minimum off-time, nominally 260 ns, ensures sufficient time during each cycle to recharge the bootstrap capacitor. A 0.022 μF ceramic capacitor is recommended for C4.

Soft-Start

The soft-start feature allows the regulator to gradually reach a steady state operating point, thereby reducing startup stresses and current surges. At turn-on, while V_{CC} is below the under-voltage threshold (t_1 in Figure 1), the SS pin is internally grounded, and V_{OUT} is held at 0V. When V_{CC} exceeds the under-voltage threshold (UVLO) an internal 11.5 μA current source charges the external capacitor (C6) at the SS pin to 2.5V (t_2 in Figure 1). The increasing SS voltage at the non-inverting input of the regulation comparator gradually increases the output voltage from zero to the desired value. The soft-start feature keeps the load inductor current from reaching the current limit threshold during start-up, thereby reducing inrush currents.

An internal switch grounds the SS pin if V_{CC} is below the under-voltage lock-out threshold, or if the circuit is shutdown using the RON/SD pin.

Thermal Shutdown

The LM25010 should be operated below the Maximum Operating Junction Temperature rating. If the junction temperature increases during a fault or abnormal operating condition, the internal Thermal Shutdown circuit activates typically at 175°C. The Thermal Shutdown circuit reduces power dissipation by disabling the buck switch and the on-timer. This feature helps prevent catastrophic failures from accidental device overheating. When the junction temperature reduces below approximately 155°C (20°C typical hysteresis), normal operation resumes.

Applications Information

EXTERNAL COMPONENTS

The procedure for calculating the external components is illustrated with a design example. Referring to the Block Diagram, the circuit is to be configured for the following specifications:

- $V_{OUT} = 5\text{V}$
- $V_{IN} = 6\text{V to } 40\text{V}$
- $F_S = 175\text{ kHz}$
- Minimum load current = 200 mA
- Maximum load current = 1.0A
- Softstart time = 5 ms.

R1 and R2: These resistors set the output voltage, and their ratio is calculated from:

$$R1/R2 = (V_{OUT}/2.5V) - 1 \quad (10)$$

R1/R2 calculates to 1.0. The resistors should be chosen from standard value resistors in the range of 1.0 k Ω - 10 k Ω . A value of 1.0 k Ω will be used for R1 and for R2.

R_{ON}, F_S: R_{ON} can be chosen using Equation 7 to set the nominal frequency, or from Equation 6 if the on-time at a particular V_{IN} is important. A higher frequency generally means a smaller inductor and capacitors (value, size and cost), but higher switching losses. A lower frequency means a higher efficiency, but with larger components. Generally, if PC board space is tight, a higher frequency is better. The resulting on-time and frequency have a $\pm 25\%$ tolerance. Using equation 7 at a nominal V_{IN} of 8V,

$$R_{ON} = \frac{5V \times (8V - 1.4V)}{8V \times 175\text{ kHz} \times 1.18 \times 10^{-10}} - 1.4\text{ k}\Omega = 198\text{ k}\Omega$$

A value of 200 k Ω will be used for R_{ON}, yielding a nominal frequency of 161 kHz at $V_{IN} = 6\text{V}$, and 203 kHz at $V_{IN} = 40\text{V}$.

L1: The guideline for choosing the inductor value in this example is that it must keep the circuit's operation in continuous conduction mode at minimum load current. This is not a strict requirement since the LM25010 regulates correctly when in discontinuous conduction mode, although at a lower frequency. However, to provide an initial value for L1 the above guideline will be used.

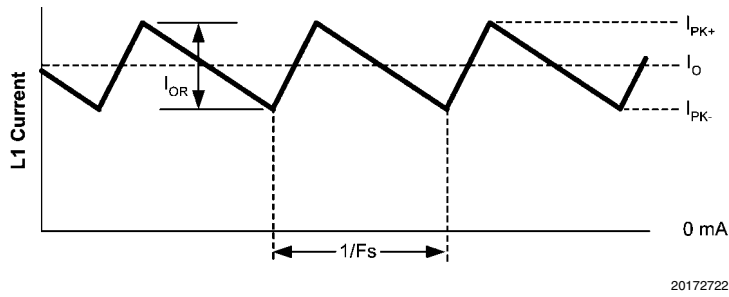


FIGURE 5. Inductor Current

To keep the circuit in continuous conduction mode, the maximum allowed ripple current is twice the minimum load current, or 400 mAp-p. Using this value of ripple current, the inductor (L1) is calculated using the following:

$$L1 = \frac{V_{OUT} \times (V_{IN(max)} - V_{OUT})}{I_{OR} \times F_{S(min)} \times V_{IN(max)}} \quad (11)$$

where $F_{S(min)}$ is the minimum frequency of 152 kHz (203 kHz - 25%) at $V_{IN(max)}$.

$$L1 = \frac{5V \times (40V - 5V)}{0.40A \times 152 \text{ kHz} \times 40V} = 72 \mu\text{H}$$

This provides a minimum value for L1 - the next higher standard value (100 μH) will be used. To prevent saturation, and possible destructive current levels, L1 must be rated for the peak current which occurs if the current limit and maximum ripple current are reached simultaneously (I_{PK} in Figure 4). The maximum ripple amplitude is calculated by re-arranging Equation 11 using $V_{IN(max)}$, $F_{S(min)}$, and the minimum inductor value, based on the manufacturer's tolerance. Assume, for this exercise, the inductor's tolerance is $\pm 20\%$.

$$I_{OR(max)} = \frac{V_{OUT} \times (V_{IN(max)} - V_{OUT})}{L1_{min} \times F_{S(min)} \times V_{IN(max)}} \quad (12)$$

$$I_{OR(max)} = \frac{5V \times (40V - 5V)}{80 \mu\text{H} \times 152 \text{ kHz} \times 40V} = 360 \text{ mAp-p}$$

$$I_{PK} = I_{LIM} + I_{OR(max)} = 1.5A + 0.36A = 1.86A$$

where I_{LIM} is the maximum guaranteed current limit threshold. At the nominal maximum load current of 1.0A, the peak inductor current is 1.18A.

R_{CL}: Since it is obvious that the lower peak of the inductor current waveform does not exceed 1.0A at maximum load current (see Figure 5), it is not necessary to increase the current limit threshold. Therefore R_{CL} is not needed for this exercise. For applications where the lower peak exceeds 1.0A, see the section entitled Increasing The Current Limit Threshold.

C1: This capacitor limits the ripple voltage at V_{IN} resulting from the source impedance of the supply feeding this circuit, and the on/off nature of the switch current into V_{IN} . At maximum load current, when the buck switch turns on, the current into V_{IN} steps up from zero to the lower peak of the inductor current waveform (I_{PK-} in Figure 5), ramps up to the peak val-

ue (I_{PK+}), then drops to zero at turn-off. The average current into V_{IN} during this on-time is the load current. For a worst case calculation, C1 must supply this average current during the maximum on-time. The maximum on-time is calculated at $V_{IN} = 6V$ using Equation 5, with a 25% tolerance added:

$$t_{ON(max)} = \left[\frac{1.18 \times 10^{-10} \times (200k + 1.4k)}{6V - 1.4V} + 67 \text{ ns} \right] \times 1.25 = 6.5 \mu\text{s}$$

The voltage at V_{IN} should not be allowed to drop below 5.5V in order to maintain V_{CC} above its UVLO.

$$C1 = \frac{I_O \times t_{ON}}{\Delta V} = \frac{1.0A \times 6.5 \mu\text{s}}{0.5V} = 13 \mu\text{F}$$

Normally a lower value can be used for C1 since the above calculation is a worst case calculation which assumes the power source has a high source impedance. A quality ceramic capacitor with a low ESR should be used for C1.

C2 and R3: Since the LM25010 requires a minimum of 25 mVp-p of ripple at the FB pin for proper operation, the required ripple at V_{OUT} is increased by R1 and R2, and is equal to:

$$V_{RIPPLE} = 25 \text{ mVp-p} \times (R1 + R2)/R2 = 50 \text{ mVp-p}$$

This necessary ripple voltage is created by the inductor ripple current acting on C2's ESR + R3. First, the minimum ripple current, which occurs at minimum V_{IN} , maximum inductor value, and maximum frequency, is determined.

$$I_{OR(min)} = \frac{V_{OUT} \times (V_{IN(min)} - V_{OUT})}{L1_{max} \times F_{S(max)} \times V_{IN(min)}} \\ = \frac{5V \times (6V - 5V)}{120 \mu\text{H} \times 201 \text{ kHz} \times 6V} = 34.5 \text{ mAp-p} \quad (13)$$

The minimum ESR for C2 is then equal to:

$$ESR_{(min)} = \frac{50 \text{ mV}}{34.5 \text{ mA}} = 1.45\Omega$$

If the capacitor used for C2 does not have sufficient ESR, R3 is added in series as shown in the Block Diagram. The value chosen for C2 is application dependent, and it is recommended that it be no smaller than 3.3 μF . C2 affects the ripple at V_{OUT} , and transient response. Experimentation is usually necessary to determine the optimum value for C2.

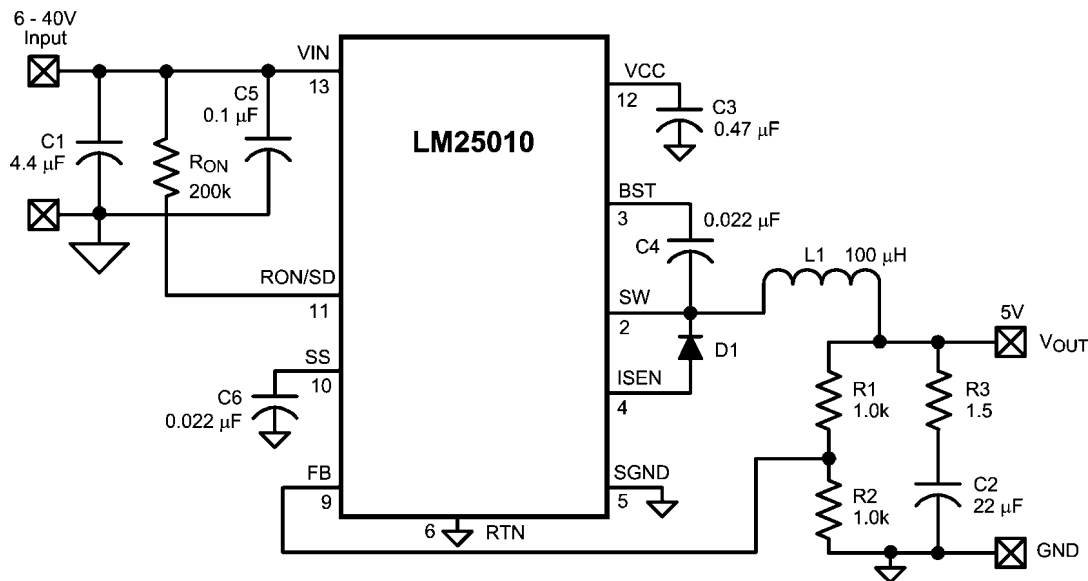
C3: The capacitor at the VCC pin provides noise filtering and stability, prevents false triggering of the V_{CC} UVLO at the buck switch on/off transitions, and limits the peak voltage at V_{CC} when a high voltage with a short rise time is initially applied at V_{IN} . C3 should be no smaller than $0.47 \mu\text{F}$, and should be a good quality, low ESR, ceramic capacitor, physically close to the IC pins.

C4: The recommended value for C4 is $0.022 \mu\text{F}$. A high quality ceramic capacitor with low ESR is recommended as C4 supplies the surge current to charge the buck switch gate at each turn-on. A low ESR also ensures a complete recharge during each off-time.

C5: This capacitor suppresses transients and ringing due to lead inductance at V_{IN} . A low ESR, $0.1 \mu\text{F}$ ceramic chip capacitor is recommended, located physically close to the LM25010.

C6: The capacitor at the SS pin determines the soft-start time, i.e. the time for the reference voltage at the regulation comparator, and the output voltage, to reach their final value. The capacitor value is determined from the following:

$$C6 = \frac{t_{SS} \times 11.5 \mu\text{A}}{2.5\text{V}}$$



20172733

FIGURE 6. Example Circuit

For a 5 ms softstart time, C6 calculates to $0.022 \mu\text{F}$.

D1: A Schottky diode is recommended. Ultra-fast recovery diodes are not recommended as the high speed transitions at the SW pin may inadvertently affect the IC's operation through external or internal EMI. The diode should be rated for the maximum V_{IN} (40V), the maximum load current (1A), and the peak current which occurs when current limit and maximum ripple current are reached simultaneously (I_{PK} in Figure 4), previously calculated to be 1.86A. The diode's forward voltage drop affects efficiency due to the power dissipated during the off-time. The average power dissipation in D1 is calculated from:

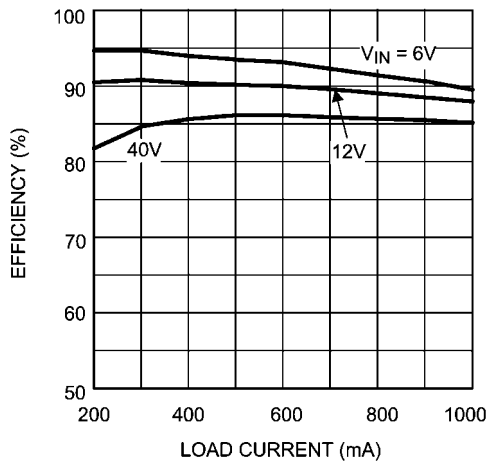
$$P_{D1} = V_F \times I_O \times (1 - D)$$

where I_O is the load current, and D is the duty cycle.

FINAL CIRCUIT

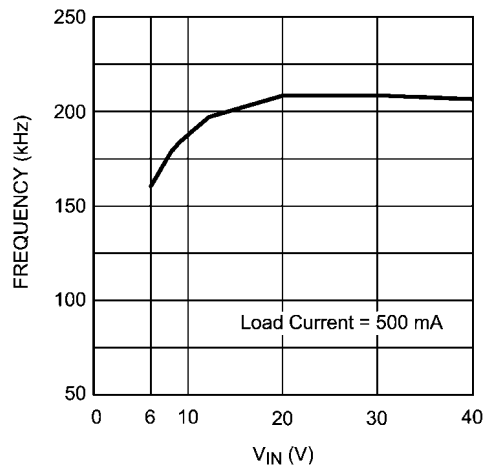
The final circuit is shown in Figure 6, and its performance is shown in Figures 7 & 8. Current limit measured approximately 1.3A.

Item	Description	Value
C1	Ceramic Capacitor	(2) 2.2 μ F, 50V
C2	Ceramic Capacitor	22 μ F, 16V
C3	Ceramic Capacitor	0.47 μ F, 16V
C4, C6	Ceramic Capacitor	0.022 μ F, 16V
C5	Ceramic Capacitor	0.1 μ F, 50V
D1	Schottky Diode	60V, 2A
L1	Inductor	100 μ H
R1	Resistor	1.0 k Ω
R2	Resistor	1.0 k Ω
R3	Resistor	1.5 Ω
R _{ON}	Resistor	200 k Ω
U1	National Semi LM25010	



20172735

FIGURE 7. Efficiency vs Load Current and V_{IN}
Circuit of Figure 6



20172737

FIGURE 8. Frequency vs V_{IN}
Circuit of Figure 6

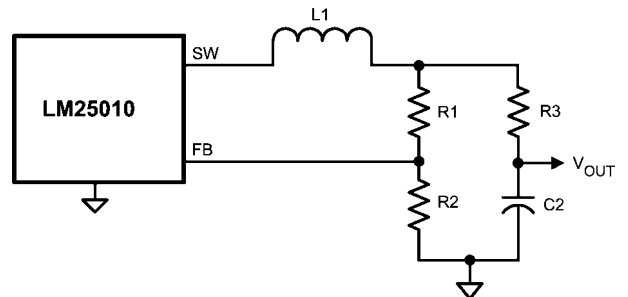
MINIMUM LOAD CURRENT

The LM25010 requires a minimum load current of 500 μ A. If the load current falls below that level, the bootstrap capacitor (C4) may discharge during the long off-time, and the circuit

will either shutdown, or cycle on and off at a low frequency. If the load current is expected to drop below 500 μ A in the application, R1 and R2 should be chosen low enough in value so they provide the minimum required current at nominal V_{OUT} .

LOW OUTPUT RIPPLE CONFIGURATIONS

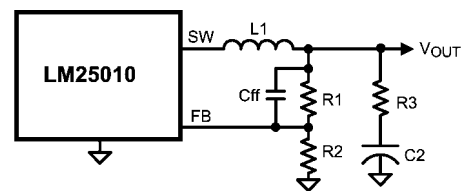
For applications where low output voltage ripple is required the output can be taken directly from the low ESR output capacitor (C2) as shown in Figure 9. However, R3 slightly degrades the load regulation. The specific component values, and the application determine if this is suitable.



20172715

FIGURE 9. Low Ripple Output

Where the circuit of Figure 9 is not suitable, the circuits of Figure 10 or Figure 11 can be used.

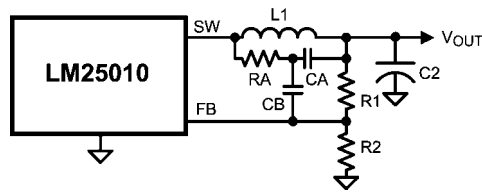


20172748

FIGURE 10. Low Output Ripple Using a Feedforward Capacitor

In Figure 10, Cff is added across R1 to AC-couple the ripple at V_{OUT} directly to the FB pin. This allows the ripple at V_{OUT} to be reduced, in some cases considerably, by reducing R3. In the circuit of Figure 6, the ripple at V_{OUT} ranged from 50 mVp-p at $V_{IN} = 6V$ to 285 mVp-p at $V_{IN} = 40V$. By adding a

1000 pF capacitor at C_{ff} and reducing R₃ to 0.75Ω, the V_{OUT} ripple was reduced by 50%, ranging from 25 mVp-p to 142 mVp-p.



20172749

FIGURE 11. Low Output Ripple Using Ripple Injection

To reduce V_{OUT} ripple further, the circuit of *Figure 11* can be used. R₃ has been removed, and the output ripple amplitude is determined by C₂'s ESR and the inductor ripple current. RA and CA are chosen to generate a 40-50 mVp-p sawtooth at their junction, and that voltage is AC-coupled to the FB pin via CB. In selecting RA and CA, V_{OUT} is considered a virtual ground as the SW pin switches between V_{IN} and -1V. Since the on-time at SW varies inversely with V_{IN}, the waveform amplitude at the RA/CA junction is relatively constant. R₁ and R₂ must typically be increased to more than 5k each to not significantly attenuate the signal provided to FB through CB. Typical values for the additional components are RA = 200k, CA = 680 pF, and CB = 0.01 μF.

INCREASING THE CURRENT LIMIT THRESHOLD

The current limit threshold is nominally 1.25A, with a minimum guaranteed value of 1.0A. If, at maximum load current, the lower peak of the inductor current (I_{PK-} in *Figure 5*) exceeds 1.0A, resistor R_{CL} must be added between SGND and ISEN to increase the current limit threshold to equal or exceed that lower peak current. This resistor diverts some of the recirculating current from the internal sense resistor so that a higher current level is needed to switch the internal current limit comparator. I_{PK-} is calculated from:

$$I_{PK-} = I_{O(max)} - \frac{I_{OR(min)}}{2} \quad (14)$$

where I_{O(max)} is the maximum load current, and I_{OR(min)} is the minimum ripple current calculated using Equation 13. R_{CL} is calculated from:

$$R_{CL} = \frac{1.0A \times 0.11\Omega}{I_{PK-} - 1.0A} \quad (15)$$

where 0.11Ω is the minimum value of the internal resistance from SGND to ISEN. The next smaller standard value resistor should be used for R_{CL}. With the addition of R_{CL}, and when the circuit is in current limit, the upper peak current out of the SW pin (I_{PK} in *Figure 4*) can be as high as:

$$I_{PK} = \frac{1.5A \times (150\text{ m}\Omega + R_{CL})}{R_{CL}} + I_{OR(MAX)}$$

where I_{OR(max)} is calculated using Equation 12. The inductor L₁ and diode D₁ must be rated for this current. If I_{PK} exceeds 2A, the inductor value must be increased to reduce the ripple amplitude. This will necessitate recalculation of I_{OR(min)}, I_{PK-}, and R_{CL}.

Increasing the circuit's current limit will increase power dissipation and the junction temperature within the LM25010. See the next section for guidelines on this issue.

PC BOARD LAYOUT and THERMAL CONSIDERATIONS

The LM25010 regulation, over-voltage, and current limit comparators are very fast, and will respond to short duration noise pulses. Layout considerations are therefore critical for optimum performance. The layout must be as neat and compact as possible, and all the components must be as close as possible to their associated pins. The two major current loops have currents which switch very fast, and so the loops should be as small as possible to minimize conducted and radiated EMI. The first loop is that formed by C₁, through the VIN to SW pins, L₁, C₂, and back to C₁. The second loop is that formed by D₁, L₁, C₂, and the SGND and ISEN pins. The ground connection from C₂ to C₁ should be as short and direct as possible, preferably without going through vias. Directly connect the SGND and RTN pin to each other, and they should be connected as directly as possible to the C₁/C₂ ground line without going through vias. The power dissipation within the IC can be approximated by determining the total conversion loss (P_{IN} - P_{OUT}), and then subtracting the power losses in the free-wheeling diode and the inductor. The power loss in the diode is approximately:

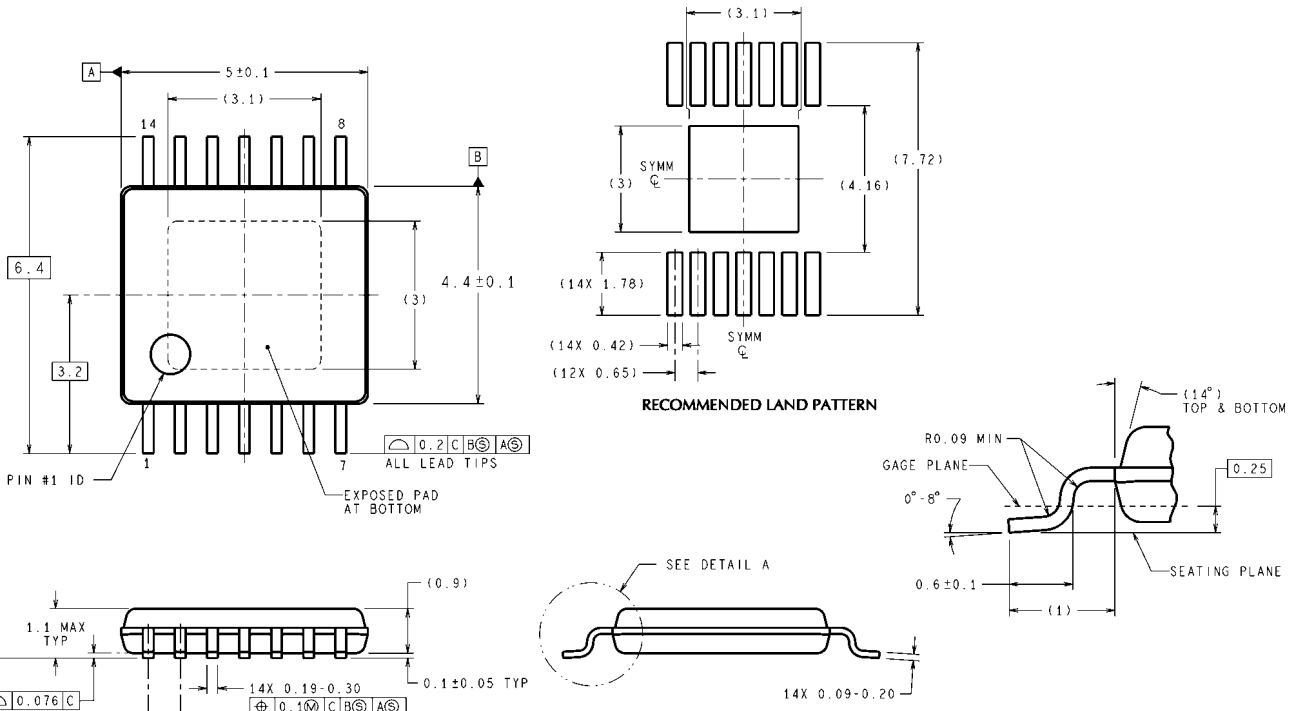
$$P_{D1} = I_O \times V_F \times (1-D)$$

where I_o is the load current, V_F is the diode's forward voltage drop, and D is the duty cycle. The power loss in the inductor is approximately:

$$P_{L1} = I_O^2 \times R_L \times 1.1$$

where R_L is the inductor's DC resistance, and the 1.1 factor is an approximation for the AC losses. If it is expected that the internal dissipation of the LM25010 will produce high junction temperatures during normal operation, good use of the PC board's ground plane can help considerably to dissipate heat. The exposed pad on the IC package bottom should be soldered to a ground plane, and that plane should both extend from beneath the IC, and be connected to exposed ground plane on the board's other side using as many vias as possible. The use of wide PC board traces at the pins, where possible, can help conduct heat away from the IC. The four No Connect pins on the TSSOP package are not electrically connected to any part of the IC, and may be connected to ground plane to help dissipate heat from the package. Judicious positioning of the PC board within the end product, along with the use of any available air flow (forced or natural convection) can help reduce the junction temperature.

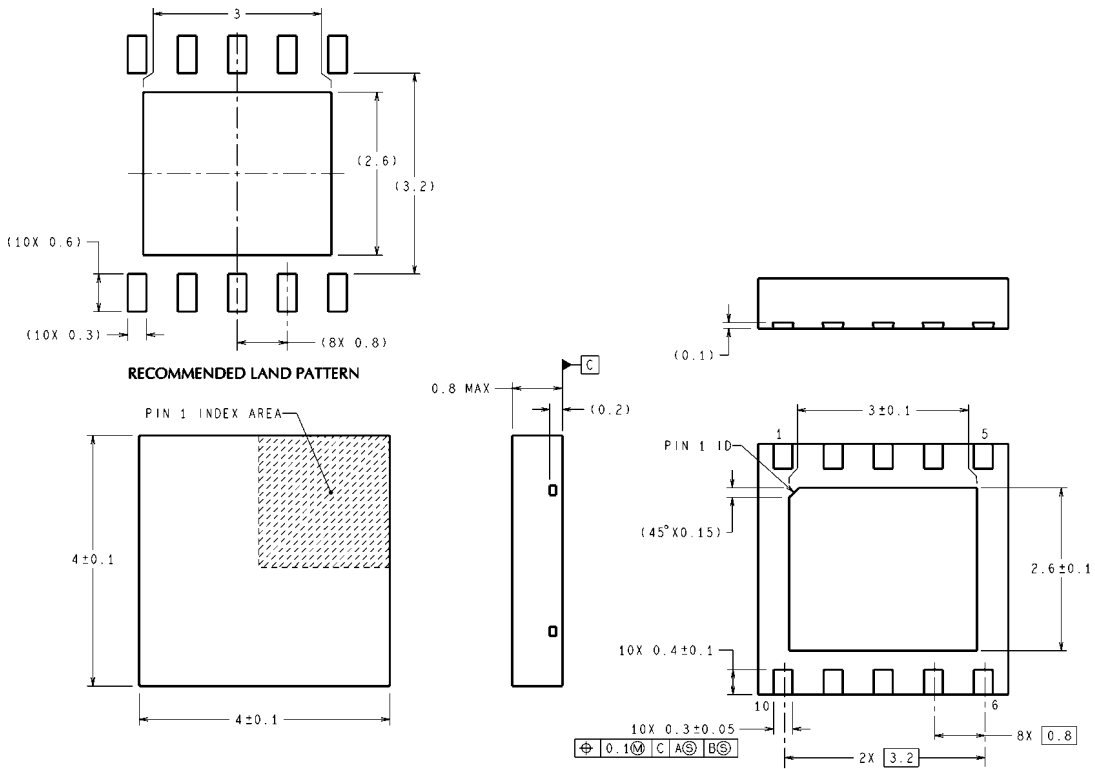
Physical Dimensions inches (millimeters) unless otherwise noted



DIMENSIONS ARE IN MILLIMETERS

14-Lead TSSOP Package
NS Package Number MXA14A

MXA14A (Rev A)



DIMENSIONS ARE IN MILLIMETERS

10-Lead LLP Package
NS Package Number SDC10A

SDC10A (Rev A)

Notes

LM25010/LM25010Q

Notes

For more National Semiconductor product information and proven design tools, visit the following Web sites at:

Products		Design Support	
Amplifiers	www.national.com/amplifiers	WEBENCH® Tools	www.national.com/webench
Audio	www.national.com/audio	App Notes	www.national.com/appnotes
Clock and Timing	www.national.com/timing	Reference Designs	www.national.com/refdesigns
Data Converters	www.national.com/adc	Samples	www.national.com/samples
Interface	www.national.com/interface	Eval Boards	www.national.com/evalboards
LVDS	www.national.com/lvds	Packaging	www.national.com/packaging
Power Management	www.national.com/power	Green Compliance	www.national.com/quality/green
Switching Regulators	www.national.com/switchers	Distributors	www.national.com/contacts
LDOs	www.national.com/ldo	Quality and Reliability	www.national.com/quality
LED Lighting	www.national.com/led	Feedback/Support	www.national.com/feedback
Voltage Reference	www.national.com/vref	Design Made Easy	www.national.com/easy
PowerWise® Solutions	www.national.com/powerwise	Solutions	www.national.com/solutions
Serial Digital Interface (SDI)	www.national.com/sdi	Mil/Aero	www.national.com/milaero
Temperature Sensors	www.national.com/tempsensors	Solar Magic®	www.national.com/solarmagic
Wireless (PLL/VCO)	www.national.com/wireless	Analog University®	www.national.com/AU

THE CONTENTS OF THIS DOCUMENT ARE PROVIDED IN CONNECTION WITH NATIONAL SEMICONDUCTOR CORPORATION ("NATIONAL") PRODUCTS. NATIONAL MAKES NO REPRESENTATIONS OR WARRANTIES WITH RESPECT TO THE ACCURACY OR COMPLETENESS OF THE CONTENTS OF THIS PUBLICATION AND RESERVES THE RIGHT TO MAKE CHANGES TO SPECIFICATIONS AND PRODUCT DESCRIPTIONS AT ANY TIME WITHOUT NOTICE. NO LICENSE, WHETHER EXPRESS, IMPLIED, ARISING BY ESTOPPEL OR OTHERWISE, TO ANY INTELLECTUAL PROPERTY RIGHTS IS GRANTED BY THIS DOCUMENT.

TESTING AND OTHER QUALITY CONTROLS ARE USED TO THE EXTENT NATIONAL DEEMS NECESSARY TO SUPPORT NATIONAL'S PRODUCT WARRANTY. EXCEPT WHERE MANDATED BY GOVERNMENT REQUIREMENTS, TESTING OF ALL PARAMETERS OF EACH PRODUCT IS NOT NECESSARILY PERFORMED. NATIONAL ASSUMES NO LIABILITY FOR APPLICATIONS ASSISTANCE OR BUYER PRODUCT DESIGN. BUYERS ARE RESPONSIBLE FOR THEIR PRODUCTS AND APPLICATIONS USING NATIONAL COMPONENTS. PRIOR TO USING OR DISTRIBUTING ANY PRODUCTS THAT INCLUDE NATIONAL COMPONENTS, BUYERS SHOULD PROVIDE ADEQUATE DESIGN, TESTING AND OPERATING SAFEGUARDS.

EXCEPT AS PROVIDED IN NATIONAL'S TERMS AND CONDITIONS OF SALE FOR SUCH PRODUCTS, NATIONAL ASSUMES NO LIABILITY WHATSOEVER, AND NATIONAL DISCLAIMS ANY EXPRESS OR IMPLIED WARRANTY RELATING TO THE SALE AND/OR USE OF NATIONAL PRODUCTS INCLUDING LIABILITY OR WARRANTIES RELATING TO FITNESS FOR A PARTICULAR PURPOSE, MERCHANTABILITY, OR INFRINGEMENT OF ANY PATENT, COPYRIGHT OR OTHER INTELLECTUAL PROPERTY RIGHT.

LIFE SUPPORT POLICY

NATIONAL'S PRODUCTS ARE NOT AUTHORIZED FOR USE AS CRITICAL COMPONENTS IN LIFE SUPPORT DEVICES OR SYSTEMS WITHOUT THE EXPRESS PRIOR WRITTEN APPROVAL OF THE CHIEF EXECUTIVE OFFICER AND GENERAL COUNSEL OF NATIONAL SEMICONDUCTOR CORPORATION. As used herein:

Life support devices or systems are devices which (a) are intended for surgical implant into the body, or (b) support or sustain life and whose failure to perform when properly used in accordance with instructions for use provided in the labeling can be reasonably expected to result in a significant injury to the user. A critical component is any component in a life support device or system whose failure to perform can be reasonably expected to cause the failure of the life support device or system or to affect its safety or effectiveness.

National Semiconductor and the National Semiconductor logo are registered trademarks of National Semiconductor Corporation. All other brand or product names may be trademarks or registered trademarks of their respective holders.

Copyright© 2008 National Semiconductor Corporation

For the most current product information visit us at www.national.com



National Semiconductor Americas Technical Support Center
Email: support@nsc.com
Tel: 1-800-272-9959

National Semiconductor Europe Technical Support Center
Email: europe.support@nsc.com
German Tel: +49 (0) 180 5010 771
English Tel: +44 (0) 870 850 4288

National Semiconductor Asia Pacific Technical Support Center
Email: ap.support@nsc.com

National Semiconductor Japan Technical Support Center
Email: jpn.feedback@nsc.com