



## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### General Description

The MAX15036/MAX15037 high-frequency, DC-DC converters with an integrated n-channel power MOSFET provide up to 3A of load current. The MAX15036 includes an internal power MOSFET to enable the design of a nonsynchronous buck or boost topology power supply. The MAX15037 is for the design of a synchronous buck topology power supply. These devices operate from a 4.5V to 5.5V or 5.5V to 23V input voltage and offer the ability to set the switching frequency from 200kHz to 2.2MHz with an external resistor. The voltage-mode architecture with a peak switch current-limit scheme provides stable operation up to a 2.2MHz switching frequency. The MAX15036 includes a clock output for driving a second DC-DC converter 180° out-of-phase and a power-on-reset (RESET) output. The MAX15037 includes a power-good output and a synchronous rectifier driver to drive an external low-side MOSFET in the buck converter configuration for high efficiency.

The MAX15036/MAX15037 protect against overcurrent conditions by utilizing a peak current limit as well as overtemperature shutdown providing a very reliable and compact power source for point-of-load regulation applications. Additional features include synchronization, internal digital soft-start, and an enable input. The MAX15036/MAX15037 are available in a thermally enhanced, space-saving 16-pin TQFN (5mm x 5mm) package and operate over the -40°C to +125°C temperature range.

### Applications

- xDSL Modem Power Supplies
- Automotive Radio Power Supplies
- Servers and Networks
- IP Phones/WLAN Access Points

### Selector Guide

PART	CONFIGURATION	FEATURES
MAX15036ATE	Nonsynchronous Buck or Boost	RESET Output, Clock Output
MAX15037ATE	Synchronous Buck	PGOOD Output, Synchronous FET Driver

*Pin Configurations continued at end of data sheet.*

### Features

- ◆ 4.5V to 5.5V or 5.5V to 23V Input Voltage Range
- ◆ Output Voltage Adjustable Down to 0.6V (Buck) or Up to 28V (Boost)
- ◆ 3A Output Current
- ◆ Synchronous Rectifier Driver Output (MAX15037) for Higher Efficiency
- ◆ Resistor-Programmable Switching Frequency from 200kHz to 2.2MHz
- ◆ External Synchronization and Enable (On/Off) Inputs
- ◆ Clock Output for Driving Second Converter 180° Out-Of-Phase (MAX15036)
- ◆ Integrated 150mΩ High-Side n-Channel Power MOSFET
- ◆ Power-On-Reset Output (MAX15036)/Power-Good Output (MAX15037)
- ◆ Short-Circuit Protection (Buck)/Maximum Duty-Cycle Limit (Boost)
- ◆ Thermal-Shutdown Protection
- ◆ Thermally Enhanced 16-Pin TQFN Package Dissipates 2.7W

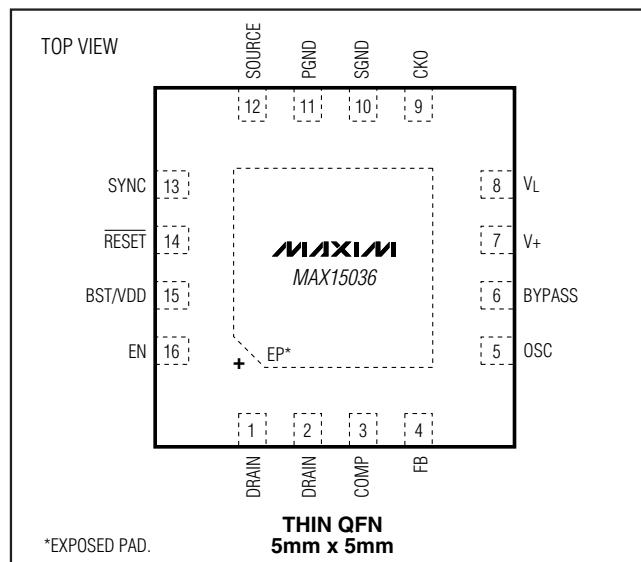
### Ordering Information

PART	TEMP RANGE	PIN-PACKAGE
MAX15036ATE+	-40°C to +125°C	16 TQFN-EP*
MAX15037ATE+	-40°C to +125°C	16 TQFN-EP*

+Denotes a lead-free/RoHS-compliant package.

\*EP = Exposed pad.

### Pin Configurations



MAX15036/MAX15037

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### ABSOLUTE MAXIMUM RATINGS

V <sub>+</sub> to PGND	-0.3V to +25V
BST/VDD, DRAIN to SGND	-0.3V to +30V
SGND to PGND	-0.3V to +0.3V
BST/VDD to SOURCE	-0.3V to +6V
SOURCE to SGND	-0.6V to +25V
SOURCE or DRAIN Maximum Peak Current	5A for 1ms
V <sub>L</sub> to SGND	-0.3V to the lower of +6V and (V <sub>+</sub> + 0.3V)
SYNC, EN, DL, CKO, OSC, COMP, FB to SGND	-0.3V to (V <sub>L</sub> + 0.3V)
BYPASS, CKO, OSC, COMP, FB, EN, SYNC, RESET, PGOOD Maximum Input Current	±50mA
RESET, PGOOD to SGND	-0.3V to +6V

BYPASS to SGND	-0.3V to +2.2V
V <sub>L</sub> and BYPASS Short-Circuit Duration to SGND	Continuous
Continuous Power Dissipation (T <sub>A</sub> = +70°C)	
16-Pin TQFN (derate 33mW/°C above +70°C)	2666mW
Junction-to-Case Thermal Resistance (θ <sub>JC</sub> ) (Note 1)	
16-Pin TQFN	1.7°C/W
Junction-to-Ambient Thermal Resistance (θ <sub>JA</sub> ) (Note 1)	
16-Pin TQFN	30°C/W
Operating Temperature Range	-40°C to +125°C
Junction Temperature Range	-65°C to +150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (soldering, 10s)	+300°C

**Note 1:** Package thermal resistances were obtained using the method described in JEDEC specification JESD51-7, using a four-layer board. For detailed information on package thermal considerations, refer to <http://www.maxim-ic.com/thermal-tutorial>.

Stresses beyond those listed under "Absolute Maximum Ratings" may cause permanent damage to the device. These are stress ratings only, and functional operation of the device at these or any other conditions beyond those indicated in the operational sections of the specifications is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

### ELECTRICAL CHARACTERISTICS

(V<sub>+</sub> = V<sub>L</sub> = 5V or V<sub>+</sub> = 5.5V to 23V, V<sub>EN</sub> = 5V, T<sub>A</sub> = T<sub>J</sub> = -40°C to +125°C, unless otherwise noted. Circuits of Figures 5 and 6. Typical values are at T<sub>A</sub> = T<sub>J</sub> = +25°C.) (Note 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
<b>SYSTEM SPECIFICATIONS</b>						
Input Voltage Range	V <sub>+</sub>		5.5	23.0		V
		V <sub>+</sub> = V <sub>L</sub>	4.5	5.5		
V <sub>+</sub> Operating Supply Current	I <sub>Q</sub>	V <sub>+</sub> = 12V, V <sub>FB</sub> = 0.8V ROSC = 10kΩ, no switching		1.8	2.5	mA
V <sub>+</sub> Standby Supply Current	I <sub>STBY</sub>	V <sub>+</sub> = 12V, V <sub>EN</sub> = 0V, PGOOD (MAX15037), RESET, CKO unconnected (MAX15036), ROSC = 10kΩ		1	1.4	mA
Efficiency	η	Nonsynchronous (MAX15036), f <sub>SW</sub> = 1.25MHz, V <sub>+</sub> = 12V, I <sub>OUT</sub> = 1.5A, V <sub>OUT</sub> = 3.3V		79		%
		Synchronous (MAX15037), f <sub>SW</sub> = 300kHz, V <sub>+</sub> = 12V, I <sub>OUT</sub> = 1.5A, V <sub>OUT</sub> = 3.3V		90		
<b>V<sub>L</sub> REGULATOR (V<sub>L</sub>)/BYPASS OUTPUT (BYPASS)</b>						
V <sub>L</sub> Undervoltage Lockout	V <sub>UVLO</sub>	V <sub>L</sub> falling	4.1	4.3		V
V <sub>L</sub> Undervoltage Lockout Hysteresis	V <sub>HYST</sub>			137		mV
V <sub>L</sub> Output Voltage	V <sub>L</sub>	V <sub>+</sub> = 5.5V to 23V, I <sub>VL</sub> = 0 to 40mA	5.0	5.2	5.5	V
V <sub>L</sub> Regulator Short-Circuit Current	I <sub>VLSHORT</sub>	V <sub>IN</sub> = 5.5V		110		mA
BYPASS Output Voltage	V <sub>BYPASS</sub>	V <sub>+</sub> = V <sub>L</sub> = 5.2V, I <sub>BYPASS</sub> = 0	1.98	2	2.02	V
BYPASS Load Regulation	ΔV <sub>BYPASS</sub>	I <sub>BYPASS</sub> steps from 0 to 50μA, V <sub>+</sub> = V <sub>L</sub> = 5.2V	0	1.2	5	mV

# 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

## ELECTRICAL CHARACTERISTICS (continued)

( $V_+ = V_L = 5V$  or  $V_+ = 5.5V$  to  $23V$ ,  $V_{EN} = 5V$ ,  $T_A = T_J = -40^\circ C$  to  $+125^\circ C$ , unless otherwise noted. Circuits of Figures 5 and 6. Typical values are at  $T_A = T_J = +25^\circ C$ .) (Note 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
<b>SOFT-START</b>						
Digital Soft-Start Period		Internal 6-bit DAC		4096		Clock periods
Soft-Start Steps				64		Steps
<b>ERROR AMPLIFIER (FB and COMP)</b>						
FB to COMP Transconductance	$g_M$		1.20	1.8	2.75	mS
FB Input Bias Current	$I_{FB}$			250		nA
FB Input Voltage Set Point	$V_{FB}$		0.591	0.600	0.609	V
COMP Sink-and-Source Current Capability	$I_{COMP}$		100	150		$\mu A$
<b>INTERNAL MOSFETs</b>						
On-Resistance n-Channel Power MOSFET	$R_{ON}$	$V_+ = V_L = 5.2V$ , $I_{SINK} = 100mA$	0.150	0.302		$\Omega$
Leakage Current	$I_{LEAK}$	$V_{EN} = 0V$ , $V_{DRAIN} = 23V$ , SOURCE = PGND		20		$\mu A$
Minimum Output Current	$I_{OUT}$	$V_{OUT} = 3.3V$ , $V_+ = 12V$ (Note 3)		3		A
Peak Current Limit	$I_{LIMIT}$		3.56	4.6	5.6	A
On-Resistance Internal Low-Side Switch	$R_{ONLSW}$	$I_{SWITCH} = 50mA$ , $V_+ = V_L = 5.2V$	20	40		$\Omega$
<b>SYNCHRONOUS RECTIFIER DRIVER (DL) (MAX15037 Only)</b>						
On-Resistance nMOS	$R_{ONDLN}$	$I_{SINK} = 10mA$	1	4		$\Omega$
On-Resistance pMOS	$R_{ONDLP}$	$I_{SOURCE} = 10mA$	1.9	5		$\Omega$
Peak Sink Current	$I_{IDL\_SINK}$			1		A
Peak Source Current	$I_{IDL\_SOURCE}$			0.75		A
<b>CLOCK OUTPUT (CKO) (MAX15036 Only)</b>						
Clock Output-High Level	$V_{CKOH}$	$V_L = 5.2V$ , $I_{SOURCE} = 5mA$	3.54			V
Clock Output-Low Level	$V_{CKOL}$	$V_L = 5.2V$ , $I_{SINK} = 5mA$		0.4		V
Clock Output Phase Delay With Respect to SOURCE Waveform	$CKOPHASE$	$R_{OSC} = 10k\Omega$ , SYNC = GND (Note 4)		115		Degrees
<b>OSCILLATOR (OSC)/SYNCHRONIZATION (SYNC)</b>						
Switching Frequency	$f_{SW}$	$V_+ = V_L = 5.2V$	$R_{OSC} = 5.62k\Omega$	2100		kHz
			$R_{OSC} = 41.2k\Omega$	312		
			$R_{OSC} = 10k\Omega$	1130	1250	1380
Minimum Controllable On-Time	$t_{ON\_MIN}$			120		ns
Maximum Duty Cycle	$D_{MAX}$	$f_{SW} = 2.2MHz$	82	87.5		%

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### ELECTRICAL CHARACTERISTICS (continued)

( $V_+ = V_L = 5V$  or  $V_+ = 5.5V$  to  $23V$ ,  $V_{EN} = 5V$ ,  $T_A = T_J = -40^\circ C$  to  $+125^\circ C$ , unless otherwise noted. Circuits of Figures 5 and 6. Typical values are at  $T_A = T_J = +25^\circ C$ .) (Note 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
SYNC Frequency Range		(Note 5)	200	2200		kHz
SYNC Input to SOURCE Rising-Edge Phase Delay	SYNC <sub>PHASE</sub>	$R_{OSC} = 10k\Omega$ , $f_{SYNC} = 1.2MHz$ (Note 6)		65		Degrees
SYNC High Threshold	V <sub>SYNCH</sub>		2.0			V
SYNC Low Threshold	V <sub>SYNCL</sub>			0.8		V
SYNC Input Bias Current	I <sub>SYNC</sub>			250		nA
Minimum SYNC High Pulse Width	t <sub>SYNC_H</sub>		100			ns
<b>EN, RESET (MAX15036)/PGOOD (MAX15037)</b>						
EN Threshold	V <sub>IH</sub>		2.0			V
	V <sub>IL</sub>			0.8		
EN Input Bias Current	I <sub>EN</sub>			250		nA
RESET Threshold	V <sub>TH</sub>	$V_{FB} = V_{OUT}$ (Note 7)	90	92.5	95	% $V_{OUT}$
PGOOD Threshold	V <sub>TH</sub>	$V_{FB} = V_{OUT}$ (Note 7)	90	92.5	95	% $V_{OUT}$
FB to RESET or FB to PGOOD Propagation Delay	t <sub>FD</sub>		3			μs
RESET Active Timeout Period	t <sub>RP</sub>		140	200	254	ms
RESET, PGOOD Output Voltage Low	V <sub>OL</sub>	I <sub>SINK</sub> = 3mA			0.4	V
RESET, PGOOD Output Leakage Current	I <sub>LEAK</sub>	$V_+ = V_L = 5.2V$ , $V_{RESET} = 6V$ , $V_{FB} = 0.8V$			2	μA
<b>THERMAL SHUTDOWN</b>						
Thermal Shutdown	T <sub>SHDN</sub>	Temperature rising		+170		°C
Thermal-Shutdown Hysteresis				25		°C

**Note 2:** 100% tested at  $T_A = +25^\circ C$  and  $T_A = +125^\circ C$ . Limits from  $T_A = -40^\circ C$  to  $+25^\circ C$  are guaranteed by design.

**Note 3:** Output current may be limited by the power dissipation of the package. See the *Power Dissipation* section in the *Applications Information* section.

**Note 4:** From the rising edge of the SOURCE waveform to the rising edge of the CKO waveform.

**Note 5:** SYNC input frequency is equal to the switching frequency.

**Note 6:** From the SYNC rising edge to SOURCE rising edge.

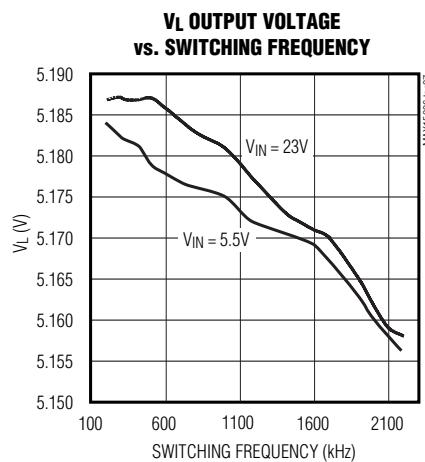
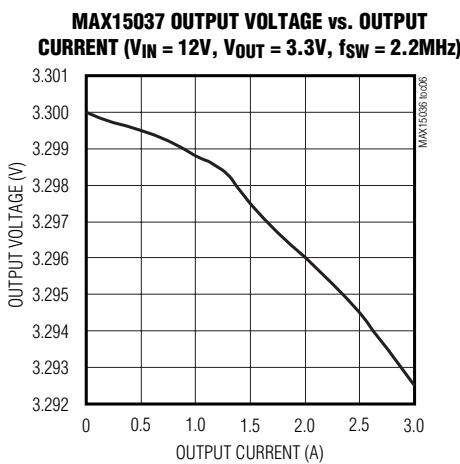
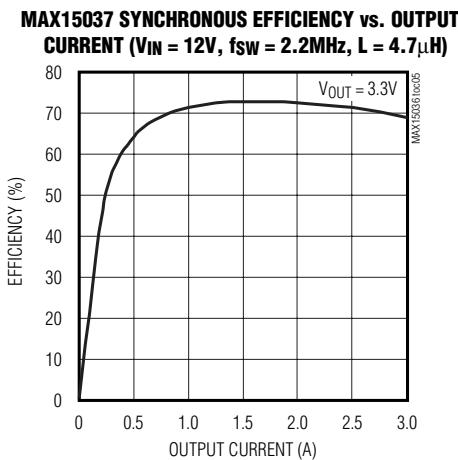
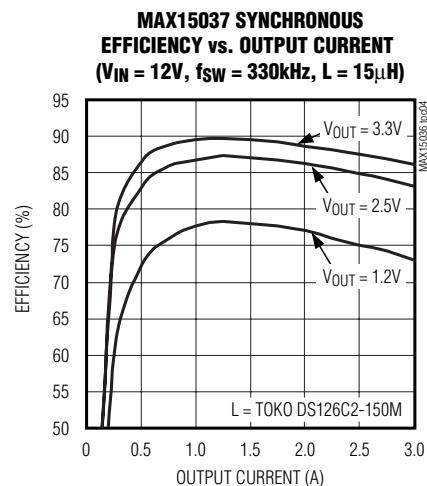
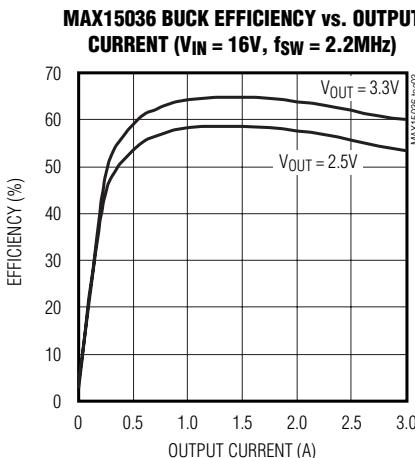
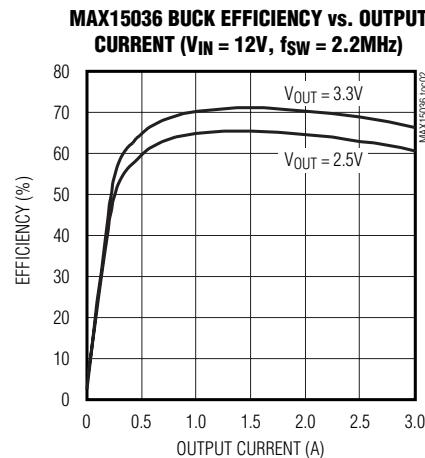
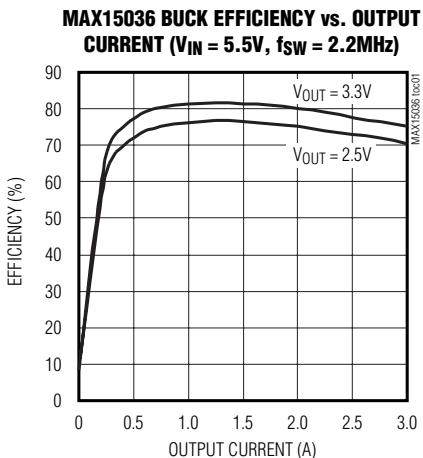
**Note 7:** RESET goes high 200ms after  $V_{OUT}$  crosses this threshold, PGOOD goes high after  $V_{OUT}$  crosses this threshold.

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MAX15036/MAX15037

### Typical Operating Characteristics

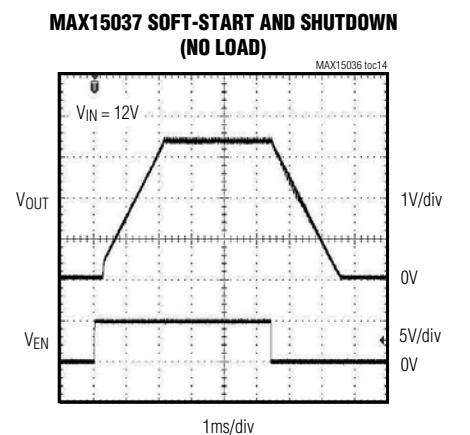
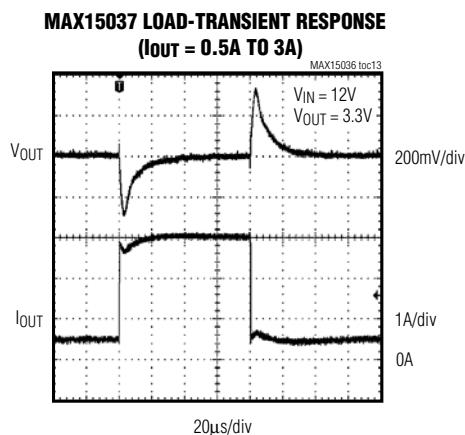
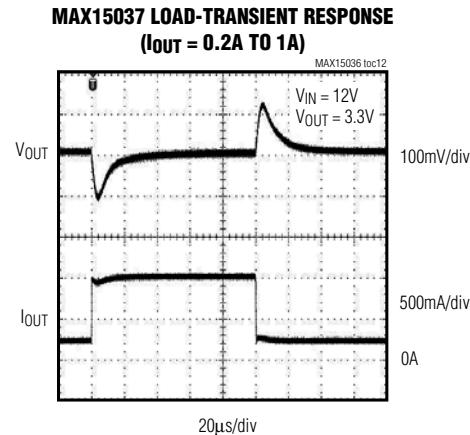
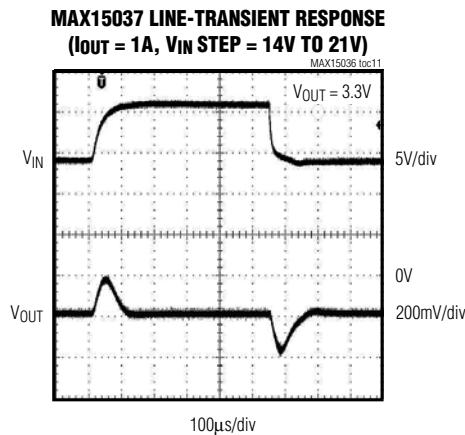
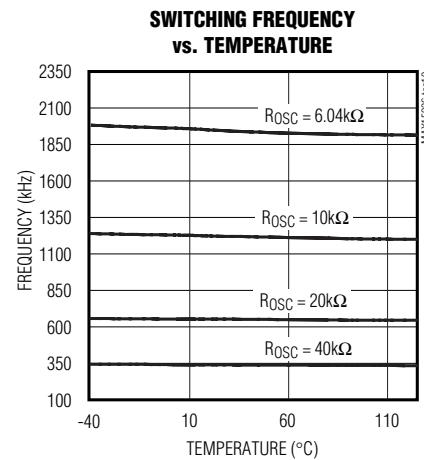
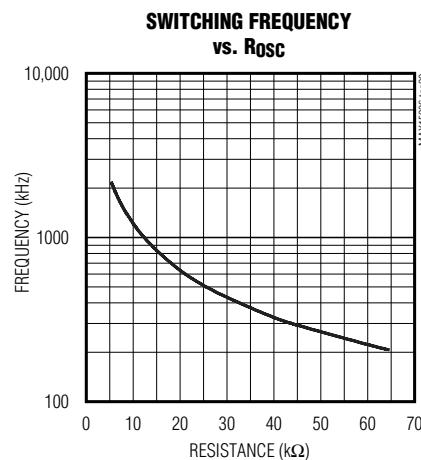
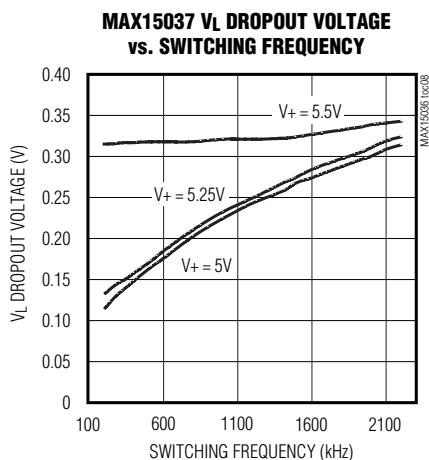
( $V_+ = V_L = 5.2V$ ,  $T_A = +25^\circ C$ , Figures 5 and 6, unless otherwise noted.)



## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### Typical Operating Characteristics (continued)

( $V_+ = V_L = 5.2V$ ,  $T_A = +25^\circ C$ , Figures 5 and 6, unless otherwise noted.)

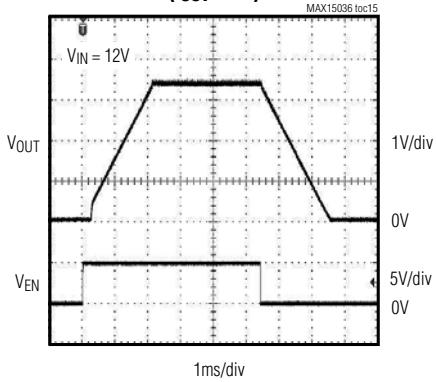


## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

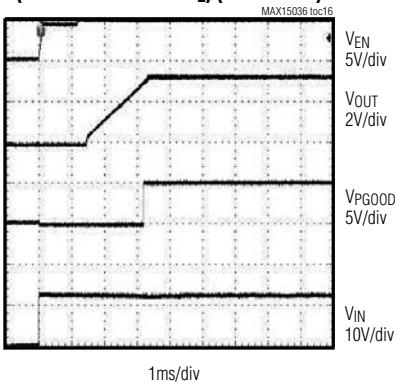
### Typical Operating Characteristics (continued)

( $V_+ = V_L = 5.2V$ ,  $T_A = +25^\circ C$ , Figures 5 and 6, unless otherwise noted.)

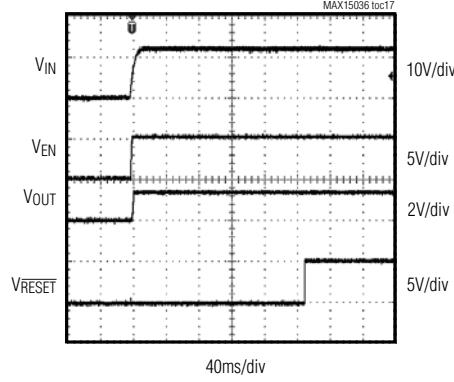
**MAX15037 SOFT-START AND SHUTDOWN ( $I_{OUT} = 2A$ )**



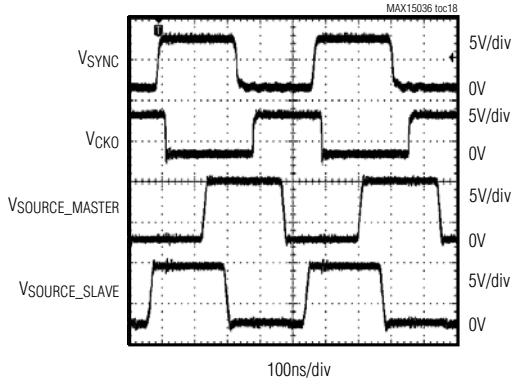
**$V_{IN}$  STARTUP WAVEFORM (EN CONNECTED TO  $V_L$ ) (MAX15037)**



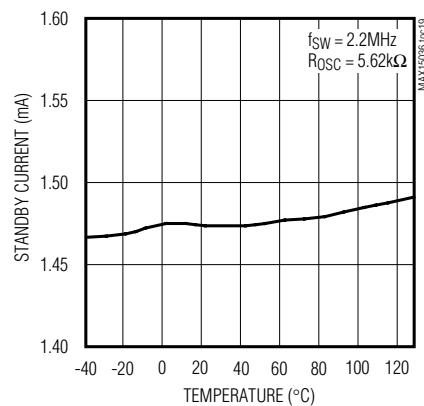
**RESET TIMEOUT (MAX15036)**



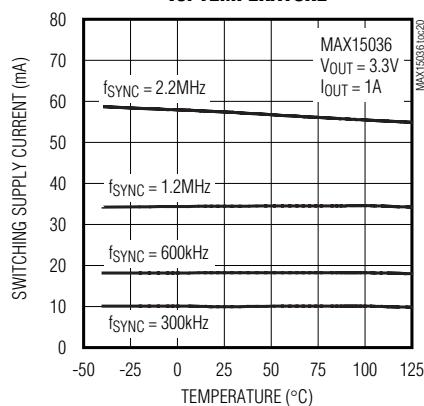
**MAX15036 EXTERNALLY SYNCHRONIZED SWITCHING WAVEFORM**



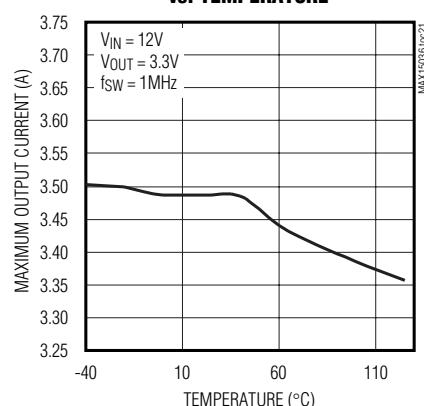
**STANDBY CURRENT vs. TEMPERATURE**



**SWITCHING SUPPLY CURRENT ( $I_{SW}$ ) vs. TEMPERATURE**



**MAXIMUM OUTPUT CURRENT vs. TEMPERATURE**



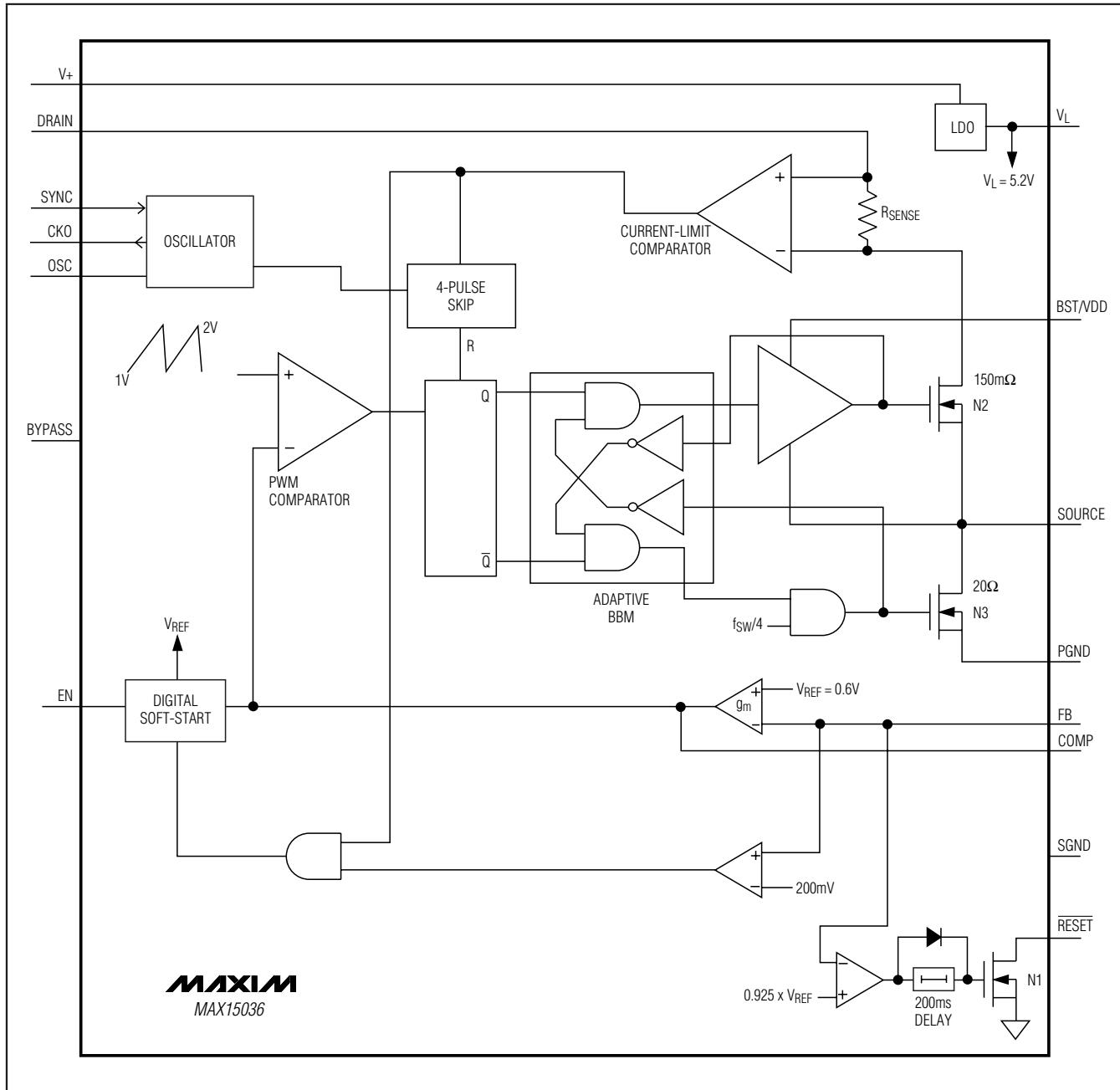
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### Pin Description

PIN	NAME	FUNCTION
1, 2	DRAIN	Internal Power MOSFET Drain Connection. Buck converter operation—use the MOSFET as a high-side switch and connect DRAIN to the input supply. Boost converter operation (MAX15036 only)—use the MOSFET as a low-side switch and connect DRAIN to the inductor and diode junction.
3	COMP	Transconductance Error Amplifier Output. Connect a compensation network from COMP to SGND or from COMP to FB to SGND (see the <i>Compensation</i> section).
4	FB	Feedback Input. Connect a resistive divider from the output to FB to SGND to set the output voltage.
5	OSC	Switching Frequency Set Input. Connect a resistor Rosc from OSC to SGND to set the switching frequency. When using external synchronization, program Rosc so that $(0.8 \times f_{SYNC}) \leq f_{SW} \leq (1.2 \times f_{SYNC})$ . Rosc is still required when external synchronization is used.
6	BYPASS	Reference Bypass Connection. Bypass to SGND with a $0.22\mu F$ or greater ceramic capacitor.
7	V+	Input Supply Voltage. V+ can range from 5.5V to 23V. Connect V+ and VL together for 4.5V to 5.5V input operation. Bypass V+ to SGND with a minimum of $0.1\mu F$ ceramic capacitor.
8	VL	Internal Regulator Output. Bypass VL to SGND with a $4.7\mu F$ ceramic capacitor and to PGND with a $0.1\mu F$ ceramic capacitor. Connect V+ to VL for 4.5V to 5.5V operation.
9	CKO	Clock Output (MAX15036 Only). CKO is an output with the same frequency as the converter's switching frequency and $115^\circ$ out-of-phase. CKO is used to synchronize the MAX15036 to other MAX15036/MAX15037s.
	DL	Low-Side Synchronous Rectifier Driver (MAX15037 Only). DL sources 0.7A and sinks 1A to quickly turn on and off the external synchronous rectifier MOSFET.
10	SGND	Signal Ground
11	PGND	Power Ground. Connect the rectifier diode's anode, the input capacitor negative terminal, the output capacitor negative terminal, and VL bypass capacitor negative terminal to PGND.
12	SOURCE	Internal Power MOSFET Source Connection. Buck converter operation—connect SOURCE to the switched side of the inductor as shown in Figure 5. Boost converter operation (MAX15036 only)—connect SOURCE to PGND.
13	SYNC	External Synchronization Input. Connect SYNC to an external logic-level clock to synchronize the MAX15036/MAX15037. Connect SYNC to SGND when not used.
14	RESET	Open-Drain Active-Low Reset Output (MAX15036 Only). RESET remains low while the converter's output is below 92.5% of VOUT's nominal set point. When VOUT rises above 92.5% of its nominal set point, RESET goes high after the reset timeout period of 200ms (typ).
	PGOOD	Open-Drain Power-Good Output (MAX15037 Only). PGOOD remains low while the output is below 92.5% of its nominal set point.
15	BST/VDD	Internal MOSFET Driver Supply Input. Buck converter operation—bootstrap flying capacitor connection. Connect BST/VDD to an external ceramic capacitor and diode (see Figure 5). Boost converter operation (MAX15036 only)—driver bypass capacitor connection. Connect a low-ESR $0.1\mu F$ ceramic capacitor from BST/VDD to PGND.
16	EN	Enable Input. A logic-low turns off the converter. A logic-high turns on the device. Connect EN to VL for an always-on application.
—	EP	Exposed Pad. Connect to SGND. Solder EP to SGND to enhance thermal dissipation.

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MAX15036/MAX15037



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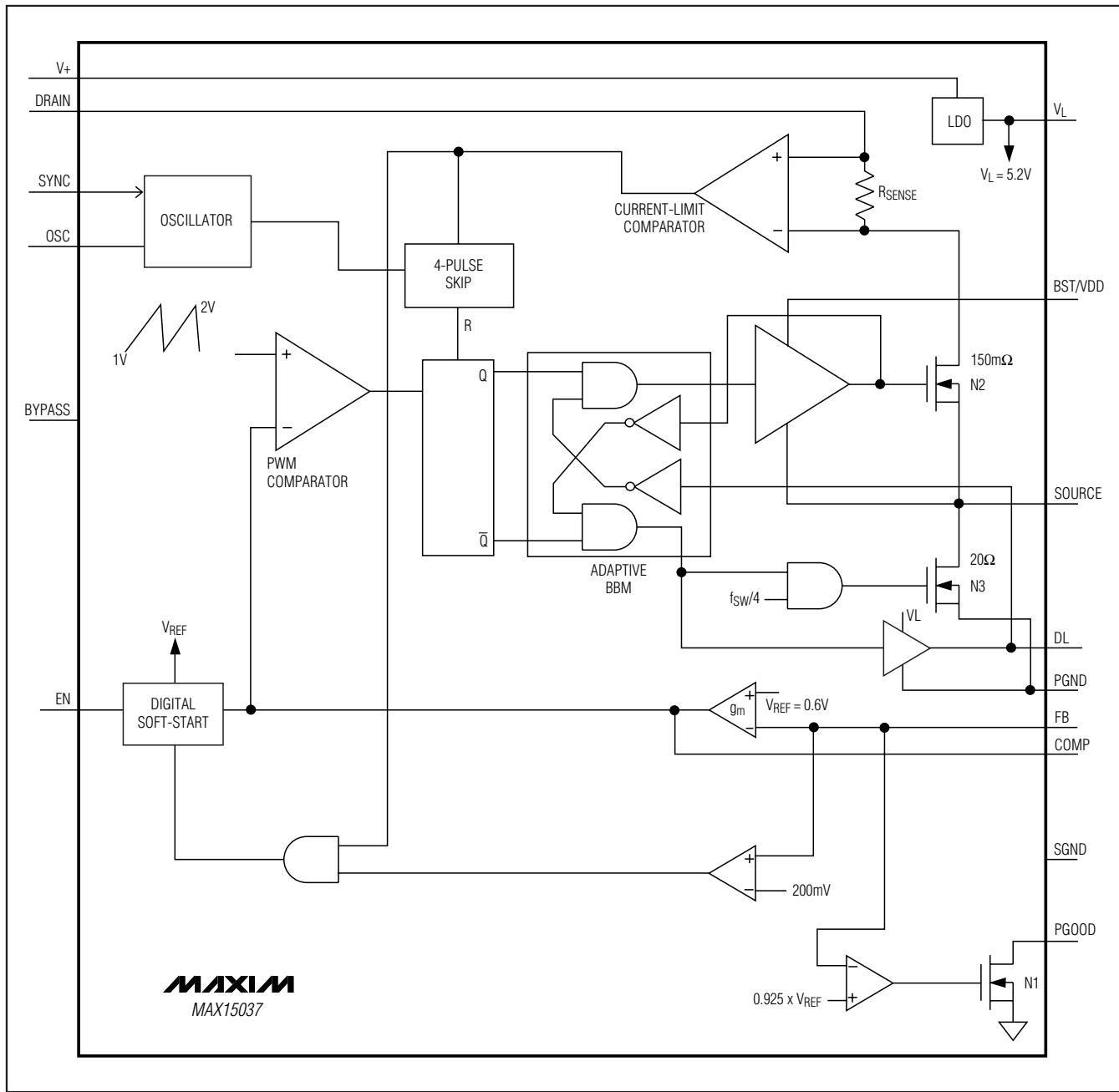


Figure 2. MAX15037 Block Diagram

# 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

## Detailed Description

### PWM Controller

The MAX15036/MAX15037 use a pulse-width modulation (PWM) voltage-mode control scheme. The MAX15036 is a nonsynchronous converter and uses an external low-forward-drop Schottky diode for rectification. The MAX15037 is a synchronous converter and drives a low-side, low-gate-charge MOSFET for higher efficiency. The controller generates the clock signal from an internal oscillator or the SYNC input when driven by an external clock. An internal transconductance error amplifier produces an integrated error voltage at COMP, providing high DC accuracy. The voltage at COMP sets the duty cycle using a PWM comparator and an internal 1Vp-p voltage ramp. At each rising edge of the clock, the converter's high-side n-channel MOSFET turns on and remains on until either the appropriate or maximum duty cycle is reached or the maximum current limit for the switch is detected.

In the case of the MAX15036 boost operation, the MOSFET is a low-side switch. During each on-time, the inductor current ramps up. During the second half of the switching cycle, the low-side switch turns off and forward biases the Schottky diode. During this time, the DRAIN voltage is clamped to 0.4V ( $V_D$ ) above  $V_{OUT}$  and the inductor provides energy to the output as well as replenishes the output capacitor charge.

### MAX15036

During each high-side MOSFET on-time (Figure 5), the inductor current ramps up. During the second half of the switching cycle, the high-side MOSFET turns off and forward biases the Schottky rectifier (D2 in Figure 5). During this time, the SOURCE voltage is clamped to 0.5V below ground. The inductor releases the stored energy as its current ramps down, and provides current to the output. During the MOSFET off-time, when the Schottky rectifier is conducting, the bootstrap capacitor (C10 in Figure 5) is recharged from the  $V_L$  output. At light loads, the MAX15036 goes into discontinuous conduction mode operation when the inductor current completely discharges before the next switching cycle commences. When the MAX15036 operates in discontinuous conduction, the bootstrap capacitor can become undercharged. To prevent this, an internal low-side 20Ω switch (see N3 in Figure 1) turns on, during the off-time, once every 4 clock cycles. This ensures that the negative terminal of the bootstrap capacitor is pulled to PGND often enough to allow it to fully charge to  $V_L$ , ensuring the internal power switch properly turns on. The operation of the bootstrap capacitor wake-up switch causes a small increase in the output voltage ripple at light loads. Under overload conditions,

when the inductor current exceeds the peak current limit of the internal switch, the high-side MOSFET turns off quickly and waits until the next clock cycle.

### MAX15037

The MAX15037 is intended for synchronous buck operation only. During the high-side MOSFET on-time, the inductor current ramps up. When the MOSFET turns off, the inductor reverses polarity and forward biases the Schottky rectifier in parallel with the low-side synchronous MOSFET. The SOURCE voltage is clamped to 0.5V below ground until the break-before-make time (tBBM) of 25ns is over. After tBBM, the synchronous rectifier MOSFET turns on. The inductor releases the stored energy as its current ramps down, and continues providing current to the output. The bootstrap capacitor is also recharged from the  $V_L$  output when the MOSFET turns off. The synchronous rectifier keeps the circuit in continuous conduction mode operation even at light load. Under overload conditions, when the inductor current exceeds the peak current limit of the internal switch, the high-side MOSFET turns off and waits until the next clock cycle.

The MAX15037, with the synchronous rectifier driver output (DL), has an adaptive break-before-make circuit to avoid cross conduction between the internal power MOSFET and the external synchronous rectifier MOSFET. When the synchronous rectifier MOSFET is turning off, the internal high-side power MOSFET is kept off until  $V_{DL}$  falls below 0.97V. Similarly, DL does not go high until the internal power MOSFET gate voltage falls below 1.24V.

### Input Voltage ( $V_+$ )/Internal Linear Regulator ( $V_L$ )

All internal control circuitry operates from an internally regulated nominal voltage of 5.2V ( $V_L$ ). At higher input voltages ( $V_+$ ) of 5.5V to 23V,  $V_L$  is regulated to 5.2V. At 5.5V or below, the internal linear regulator operates in dropout mode, where  $V_L$  follows  $V_+$ . Depending on the load on  $V_L$ , the dropout voltage can be high enough to reduce  $V_L$  to below the undervoltage lockout (UVLO) threshold.

For input voltages of lower than 5.5V, connect  $V_+$  and  $V_L$  together. The load on  $V_L$  is proportional to the switching frequency of the converter. See the  $V_L$  Output Voltage vs. Switching Frequency graph in the *Typical Operating Characteristics*. For an input voltage higher than 5.5V, use the internal regulator.

Bypass  $V_+$  to SGND with a low-ESR 0.1μF or greater ceramic capacitor placed as close as possible to the MAX15036/MAX15037. Current spikes from  $V_L$  disturb the internal circuitry powered by  $V_L$ . Bypass  $V_L$  with a low-ESR 0.1μF ceramic capacitor to PGND and a low-ESR 4.7μF ceramic capacitor to SGND.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### Enable

EN is an active-high input that turns the MAX15036/MAX15037 on and off. EN is a TTL-logic input with 2.0V and 0.8V logic-high and low levels, respectively. When EN is asserted high, the internal digital soft-start cycle slowly ramps up the internal reference and provides some soft-start at the output. Hysteresis provides immunity to the glitches during logic turn-on of the converter. Large voltage variations at EN can interrupt the soft-start sequence and can cause a latch-up. Ensure that EN remains high for at least 5ms once it is asserted. Force EN low to turn off the internal power MOSFET and cause RESET to go low (MAX15036) or cause PGOOD to go low (MAX15037). Connect EN to VL when not used.

### Soft-Start/Soft-Stop

The MAX15036/MAX15037 include UVLO with hysteresis to prevent chattering during startup. The UVLO circuit holds the MAX15036/MAX15037 off until V<sub>+</sub> reaches 4.5V and turns the devices off when V<sub>+</sub> falls below 4.3V. The MAX15036/MAX15037 also offer a soft-start feature that reduces surge currents and glitches on the input during turn-on. During turn-on when the UVLO threshold is reached or EN goes from low to high, the digital soft-start ramps up the reference (VBYPASS) in 64 steps. During a turn-off (by driving EN or V<sub>+</sub> low), the reference is reduced to zero slowly. The soft-start and soft-stop periods (t<sub>SS</sub>) are 4096 cycles of the internal oscillator. To calculate the soft-start/soft-stop period use the following equation:

$$t_{SS} = \frac{4096}{f_{SW}}$$

f<sub>SW</sub> is the switching frequency of the converter.

### Oscillator/Synchronization (SYNC)/Clock Output (CLKOUT)

The clock frequency (or switching frequency) is generated internally and is adjustable through an external resistor connected from OSC to SGND. The relationship between ROSC and f<sub>SW</sub> is:

$$ROSC = \frac{125 \times 10^8 \Omega/s}{f_{SW}}$$

The adjustment range for f<sub>SW</sub> is from 200kHz to 2.2MHz.

Connect a logic-level clock between 200kHz to 2.2MHz at SYNC to externally synchronize the MAX15036/MAX15037's oscillator (see Figure 8). The MAX15036/MAX15037 synchronize to the rising edge of the SYNC clock. The rising edge of the SYNC clock corresponds to

the turn-on edge of the internal n-channel power MOSFET with a fixed propagation delay. When operating the MAX15036/MAX15037 with an external SYNC clock, ROSC must be installed. Program the internal switching frequency so that  $(0.8 \times f_{SYNC}) \leq f_{SW} \leq (1.2 \times f_{SYNC})$ . The minimum pulse width for f<sub>SYNC</sub> is 100ns. Connect SYNC to SGND if synchronization is not used.

The CKO output (MAX15036 only) is a logic-level clock with the same frequency as f<sub>SW</sub> and with 115° phase shift with respect to SYNC clock. Two MAX15036s can be connected in a master/slave configuration for two-phase (180°) interleaved operation. The CKO output of the master drives the SYNC input of the slave to form a dual-phase converter. To achieve the 180° out-of-phase operation, program the internal switching frequency of both converters close to each other by using the same ROSC value. When synchronizing the master-slave configuration using external clock, program the internal switching frequency using ROSC close to the external clock frequency (f<sub>SYNC</sub>) for 180° ripple phase operation (see Figure 8). Any difference in the internal switching frequency and f<sub>SYNC</sub> changes the phase delay. If both master and slave converters use the same power source, and share input bypass capacitors, the effective switching frequency at the input is twice the switching frequency of the individual converter. Higher ripple frequency at the input capacitor means a lower RMS ripple current into the capacitor.

### Current Limit

The MAX15036/MAX15037 protect against output over-load and short-circuit conditions when operated in a buck configuration. An internal current-sensing stage develops a voltage proportional to the instantaneous switch current. When the switch current reaches 4.6A (typ), the power MOSFET turns off and remains off until the next on cycle.

During a severe overload or short-circuit condition when the output voltage is pulled to ground, the discharging slope of the inductor is V<sub>DS</sub> (the voltage across the synchronous FET), or V<sub>F</sub> (the voltage across the rectifying diode) divided by L. The short off-time does not allow the current to properly ramp down in the inductor, causing a dangerous current runaway and possibly destruction of the device. To prevent this, the MAX15036/MAX15037 include a frequency foldback feature. When the current limit is detected the frequency is reduced to 1/4th of the programmed switching frequency. When the output voltage falls below 1/3rd of its nominal set point (V<sub>FB</sub> = 0.2V), the converter is turned off and soft-start cycle is initiated. This reduces the RMS current sourced by the converter during the fault condition.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

At high input-to-output differential, and high switching frequency, the on-time drops to the order of 100ns. Even though the MAX15036/MAX15037 can control the on-time as low as 100ns, the internal current-limit circuit may not detect the overcurrent within this time. In that case, the output current during the fault may exceed the current limit specified in the *Electrical Characteristics* table. The MAX15036/MAX15037 may still be protected against the output short-circuit fault through the overtemperature shutdown. However, the output switch current may be as high as 5.6A. If the minimum on-time for a given frequency and duty cycle is less than 200ns, choose the inductor with a saturation current of greater than 5.6A.

### Power-On Reset (RESET) (MAX15036 Only)

RESET is an active-low open-drain output that goes low when V<sub>OUT</sub> falls below 92.5% of its nominal set point. RESET goes high impedance when V<sub>OUT</sub> rises above 92.5% of its nominal set point, the soft-start period is complete, and the 200ms (typ) timeout period has elapsed. Connect a pullup resistor from RESET to a logic voltage or to V<sub>L</sub>. The internal open-drain MOSFET at RESET can sink 3mA while providing a TTL-compatible logic-low signal. Connect RESET to SGND or leave unconnected when not used.

### Power-Good (PGOOD) (MAX15037 Only)

PGOOD is an open-drain, active-high output that goes low when V<sub>OUT</sub> is below 92.5% of its nominal set point and goes high impedance when V<sub>OUT</sub> goes above 92.5% its nominal set point. Connect a pullup resistor from PGOOD to a logic voltage or to V<sub>L</sub>. PGOOD can sink up to 3mA while still providing a TTL-compatible logic-low output. Pulling EN low forces PGOOD low. Connect PGOOD to SGND or leave unconnected when not used.

### Thermal-Overload Protection

During a continuous output short-circuit or overload condition, the die junction temperature in the MAX15036/MAX15037 can exceed its limit. The MAX15036/MAX15037 provide an internal thermal shutdown to turn off the device when the die temperature

reaches +170°C. A thermal sensor monitors the die temperature and turns the device on again when the temperature reduces by +25°C. During thermal shutdown, the internal power MOSFET shuts off, DL pulls to SGND, V<sub>L</sub> shuts down, RESET (MAX15036)/PGOOD (MAX15037) goes low, and soft-start resets.

### Applications Information

#### Setting the Switching Frequency

The controller generates the switching frequency (f<sub>sw</sub>) through the internal oscillator or the signal at SYNC (f<sub>SYNC</sub>), when driven by an external oscillator. The switching frequency is equal to f<sub>sw</sub> or f<sub>SYNC</sub>.

A resistor, Rosc, from OSC to SGND sets the internal oscillator. The relationship between f<sub>sw</sub> and Rosc is:

$$R_{OSC} = \frac{125 \times 10^8}{f_{SW}}$$

where f<sub>sw</sub> is in Hertz, and Rosc is in ohms. For example, a 1.25MHz switching frequency is set with Rosc = 10kΩ. Higher frequencies allow designs with lower inductor values and less output capacitance. Consequently, peak currents and I<sup>2</sup>R losses are lower at higher switching frequencies, but core losses, gate-charge currents, and switching losses increase.

Rising clock edges on SYNC are interpreted as a synchronization input. If the SYNC signal is lost, the internal oscillator takes control of the switching rate, returning the switching frequency to that set by Rosc. This maintains output regulation even with intermittent SYNC signals. When using an external synchronization signal, set Rosc so that  $(0.8 \times f_{SYNC}) \leq f_{sw} \leq (1.2 \times f_{SYNC})$ .

#### Buck Converter

Use the internal n-channel power MOSFET as a high-side switch to configure the MAX15036/MAX15037 as a buck converter. In this configuration, SOURCE is connected to the inductor, DRAIN is connected to the input, and BST/VDD connects to the cathode of the bootstrap diode and capacitor. Figures 5 and 6 show the typical application circuits for MAX15036/MAX15037, respectively, in a buck configuration.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### Effective Input Voltage Range

The MAX15036/MAX15037 can operate with input supplies ranging from 4.5V to 5.5V or 5.5V to 23V. The input voltage range ( $V_+$ ) can be constrained to a minimum by the duty-cycle limitations and to a maximum by the on-time limitation. The minimum input voltage is determined by:

$$V_{IN\_MIN} = \frac{V_{OUT} + V_{DROP1}}{D_{MAX}} + V_{DROP2} - V_{DROP1}$$

$D_{MAX}$  is the maximum duty cycle of 87.5% (typ).  $V_{DROP1}$  is the total drop in the inductor discharge path that includes the diode's forward voltage drop (or the drop across the synchronous rectifier MOSFET), and the drops across the series resistance of the inductor and PCB traces.  $V_{DROP2}$  is the total drop in the inductors charging path, which includes the drop across the internal power MOSFET, and the drops across the series resistance of the inductor and PCB traces.

The maximum input voltage can be determined by:

$$V_{IN\_MAX} = \frac{V_{OUT}}{t_{ON\_MIN} \times f_{SW}}$$

where  $t_{ON\_MIN} = 100\text{ns}$  and  $f_{SW}$  is the switching frequency.

### Setting the Output Voltage

For 0.6V or greater output voltages, connect a resistive divider from  $V_{OUT}$  to FB to SGND. Select the FB to SGND resistor ( $R_2$ ) from  $1\text{k}\Omega$  to  $10\text{k}\Omega$  and calculate the resistor from OUT to FB ( $R_1$ ) by the following equation:

$$R_1 = R_2 \times \left[ \frac{V_{OUT}}{V_{FB}} - 1 \right]$$

where  $V_{FB} = 0.6\text{V}$ , see Figure 3.

For designs that use a Type III compensation scheme, first calculate  $R_1$  for stability requirements (see the Compensation section) then choose  $R_2$  so that:

$$R_2 = \frac{R_1 \times V_{FB}}{V_{OUT} - V_{FB}}$$

See Figure 4.

### Inductor Selection

Three key inductor parameters must be specified for operation with the MAX15036/MAX15037: inductance value ( $L$ ), peak inductor current ( $I_{PEAK}$ ), and inductor saturation current ( $I_{SAT}$ ). The minimum required inductance is a function of operating frequency, input-to-output voltage differential, and the peak-to-peak inductor current ( $\Delta I_{P-P}$ ). Higher  $\Delta I_{P-P}$  allows for a lower inductor value, while a lower  $\Delta I_{P-P}$  requires a higher inductor value. A lower inductor value minimizes size and cost, improves large-signal and transient response, but reduces efficiency due to higher peak currents and higher peak-to-peak output voltage ripple for the same output capacitor. On the other hand, higher inductance increases efficiency by reducing the ripple current. Resistive losses due to extra wire turns can exceed the benefit gained from lower ripple current levels especially when the inductance is increased without also allowing for larger inductor dimensions. A good compromise is to choose  $\Delta I_{P-P}$  equal to 30% of the full load current. Use the following equation to calculate the inductance:

$$L = \frac{V_{OUT}(V_{IN} - V_{OUT})}{V_{IN} \times f_{SW} \times \Delta I_{P-P}}$$

$V_{IN}$  and  $V_{OUT}$  are typical values so that efficiency is optimum for typical conditions. The switching frequency is set by  $f_{SW}$  (see the Setting the Switching Frequency section). The peak-to-peak inductor current, which reflects the peak-to-peak output ripple, is worse at the maximum input voltage. See the Output Capacitor Selection section to verify that the worst-case output ripple is acceptable. The inductor saturation current is also important to avoid runaway current during continuous output short-circuit. At high input-to-output differential, and high switching frequency, the on-time drops to the order of 100ns. Though the MAX15036/MAX15037 can control the on-time as low as 100ns, the internal current-limit circuit may not detect the overcurrent within this time. In that case, the output current during the fault may exceed the current limit specified in the Electrical Characteristics table. The overtemperature shutdown protects the MAX15036/MAX15037 against the output short-circuit fault. However, the output current may reach 5.6A. Choose an inductor with a saturation current of greater than 5.6A when the minimum on-time for a given frequency and duty cycle is less than 200ns.

### Input Capacitors

The discontinuous input current of the buck converter causes large input ripple current. The switching frequency, peak inductor current, and the allowable peak-to-peak input voltage ripple dictate the input capacitance requirement. Increasing the switching frequency or the inductor value lowers the peak-to-average current ratio yielding a lower input capacitance requirement.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

The input ripple comprises mainly of  $\Delta V_Q$  (caused by the capacitor discharge) and  $\Delta V_{ESR}$  (caused by the ESR of the input capacitor). The total voltage ripple is the sum of  $\Delta V_Q$  and  $\Delta V_{ESR}$ . Assume the input voltage ripple from the ESR and the capacitor discharge is equal to 50% each. The following equations show the ESR and capacitor requirement for a target voltage ripple at the input:

$$ESR = \frac{\Delta V_{ESR}}{\left(I_{OUT} + \frac{\Delta I_{P-P}}{2}\right)}$$

$$C_{IN} = \frac{I_{OUT} \times D(1-D)}{\Delta V_Q \times f_{SW}}$$

where

$$\Delta I_{P-P} = \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{V_{IN} \times f_{SW} \times L} \text{ and}$$

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

where  $I_{OUT}$  is the output current,  $D$  is the duty cycle, and  $f_{SW}$  is the switching frequency. Use additional input capacitance at lower input voltages to avoid possible undershoot below the UVLO threshold during transient loading.

### Output Capacitor Selection

The allowable output voltage ripple and the maximum deviation of the output voltage during step load currents determine the output capacitance and its ESR.

The output ripple comprises of  $\Delta V_Q$  (caused by the capacitor discharge) and  $\Delta V_{ESR}$  (caused by the ESR of the output capacitor). Use low-ESR ceramic or aluminum electrolytic capacitors at the output. For aluminum electrolytic capacitors, the entire output ripple is contributed by  $\Delta V_{ESR}$ . Use the  $ESR_{OUT}$  equation to calculate the ESR requirement and choose the capacitor accordingly. If using ceramic capacitors, assume the contribution to the output ripple voltage from the ESR and the capacitor discharge to be equal. The following equations show the output capacitance and ESR requirement for a specified output voltage ripple.

$$ESR = \frac{\Delta V_{ESR}}{\Delta I_{P-P}}$$

$$C_{OUT} = \frac{\Delta I_{P-P}}{8 \times \Delta V_Q \times f_{SW}}$$

where:

$$\Delta I_{P-P} = \frac{(V_{IN} - V_{OUT}) \times V_{OUT}}{V_{IN} \times f_{SW} \times L}$$

$$V_{OUT\_RIPPLE} \approx \Delta V_{ESR} + \Delta V_Q$$

$\Delta I_{P-P}$  is the peak-to-peak inductor current as calculated above and  $f_{SW}$  is the individual converter's switching frequency.

The allowable deviation of the output voltage during fast transient loads also determines the output capacitance and its ESR. The output capacitor supplies the step load current until the controller responds with a greater duty cycle. The response time ( $t_{RESPONSE}$ ) depends on the closed-loop bandwidth of the converter. The high switching frequency of the MAX15036/MAX15037 allows for a higher closed-loop bandwidth, thus reducing  $t_{RESPONSE}$  and the output capacitance requirement. The resistive drop across the output capacitor's ESR and the capacitor discharge causes a voltage droop during a step load. Use a combination of low-ESR tantalum and ceramic capacitors for better transient load and ripple/noise performance. Keep the maximum output voltage deviation below the tolerable limits of the electronics being powered. When using a ceramic capacitor, assume an 80% and 20% contribution from the output capacitance discharge and the ESR drop, respectively. Use the following equations to calculate the required ESR and capacitance value:

$$ESR_{OUT} = \frac{\Delta V_{ESR}}{I_{STEP}}$$

$$C_{OUT} = \frac{I_{STEP} \times t_{RESPONSE}}{\Delta V_Q}$$

where  $I_{STEP}$  is the load step and  $t_{RESPONSE}$  is the response time of the controller. The controller response time depends on the control-loop bandwidth.

### Boost Converter

The MAX15036 can be configured for step-up conversion since the internal MOSFET can be used as a low-side switch. Use the following equations to calculate the inductor ( $L_{MIN}$ ), input capacitor ( $C_{IN}$ ), and output capacitor ( $C_{OUT}$ ) when using the converter in boost operation.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### Inductor

Choose the minimum inductor value so the converter remains in continuous mode operation at minimum output current ( $I_{OUTMIN}$ ).

$$L_{MIN} = \frac{V_{IN}^2 \times D \times n}{2 \times f_{SW} \times V_{OUT} \times I_{OUTMIN}}$$

where

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

and  $I_{OUTMIN} = 0.25 \times I_{OUT}$ .

The  $V_D$  is the forward voltage drop of the external Schottky diode,  $D$  is the duty cycle, and  $V_{DS}$  is the voltage drop across the internal switch. Select the inductor with low DC resistance and with a saturation current ( $I_{SAT}$ ) rating higher than the peak switch current limit of 5.6A.

### Input Capacitor

The input current for the boost converter is continuous and the RMS ripple current at the input is low. Calculate the capacitor value and ESR of the input capacitor using the following equations.

$$C_{IN} = \frac{\Delta I_{P-P} \times D}{4 \times f_{SW} \times \Delta V_Q}$$

$$ESR = \frac{\Delta V_{ESR}}{\Delta I_{P-P}}$$

where

$$\Delta I_{P-P} = \frac{(V_{IN} - V_{DROP}) \times D}{L \times f_{SW}}$$

where  $V_{DROP}$  is the total voltage drop across the internal MOSFET plus the voltage drop across the inductor ESR.  $\Delta I_{P-P}$  is the peak-to-peak inductor ripple current as calculated above.  $\Delta V_Q$  is the portion of input ripple due to the capacitor discharge and  $\Delta V_{ESR}$  is the contribution due to ESR of the capacitor.

### Output Capacitor

For the boost converter, the output capacitor supplies the load current when the main switch is on. The required output capacitance is high, especially at higher duty cycles. Also, the output capacitor ESR needs to be low enough to minimize the voltage drop due to the ESR while supporting the load current. Use the following equation to calculate the output capacitor for a specified output ripple tolerance.

$$ESR = \frac{\Delta V_{ESR}}{I_{OUT}}$$

$$C_{OUT} = \frac{I_{OUT} \times D_{MAX}}{\Delta V_Q \times f_{SW}}$$

$I_{OUT}$  is the load current,  $\Delta V_Q$  is the portion of the ripple due to the capacitor discharge, and  $\Delta V_{ESR}$  is the contribution due to the ESR of the capacitor.  $D_{MAX}$  is the maximum duty cycle at minimum input voltage.

### Power Dissipation

The MAX15036/MAX15037 are available in thermally enhanced 16-pin, 5mm x 5mm TQFN packages that dissipate up to 2.7W at  $T_A = +70^\circ\text{C}$ . When the die temperature reaches  $+170^\circ\text{C}$ , the MAX15036/MAX15037 shut down (see the *Thermal-Overload Protection* section). The power dissipated in the device is the sum of the power dissipated from supply current ( $P_Q$ ), power dissipated due to switching the internal power MOSFET ( $P_{SW}$ ), and the power dissipated due to the RMS current through the internal power MOSFET (PMOSFET). The total power dissipated in the package must be limited so the junction temperature does not exceed its absolute maximum rating of  $+150^\circ\text{C}$  at maximum ambient temperature.

The power dissipated in the switch is:

$$PMOSFET = I_{RMS\_MOSFET} \times R_{ON}$$

For the buck converter:

$$I_{RMS\_MOSFET} = \sqrt{(I_{OUT}^2 \times D) + \left( \frac{\Delta I_{P-P}^2 \times D}{12} \right)}$$

$\Delta I_{P-P}$  is the peak-to-peak inductor current ripple.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

For the boost converter:

$$I_{RMS\_MOSFET} = \sqrt{(I_{DC}^2 + I_{PK}^2 + (I_{DC} \times I_{PK})) \times \frac{D_{MAX}}{3}}$$

$$I_N = \frac{V_{OUT} \times I_{OUT}}{V_{IN} \times \eta}$$

$$\Delta I_{P-P} = \frac{(V_{IN} - V_{DROP}) \times D}{L \times f_{SW}}$$

$$I_{DC} = I_N - \frac{\Delta I_{P-P}}{2}$$

$$I_{PK} = I_N + \frac{\Delta I_{P-P}}{2}$$

The power lost due to switching the internal power MOSFET is:

$$P_{SW} = \frac{V_{IN} \times I_{OUT} \times (t_R + t_F) \times f_{SW}}{4}$$

$t_R$  and  $t_F$  are the rise and fall times of the internal power MOSFET measured at SOURCE.

The power lost due to the switching quiescent current of the device is:

$$P_Q = V_{IN} \times I_{SW} \quad (\text{MAX15036})$$

The switching quiescent current ( $I_{SW}$ ) of the MAX15036/MAX15037 is dependent on switching frequency. See the *Typical Operating Characteristics* section for the value of  $I_{SW}$  at a given frequency.

In the case of the MAX15037, the switching current includes the synchronous rectifier MOSFET gate-drive current ( $I_{SW-DL}$ ). The  $I_{SW-DL}$  depends on the total gate charge ( $Q_{g-DL}$ ) of the synchronous rectifier MOSFET and the switching frequency.

$$P_Q = V_{IN} \times (I_{SW} + I_{SW-DL}) \quad (\text{MAX15037})$$

$$I_{SW-DL} = Q_{g-DL} \times f_{SW}$$

where the  $Q_{g-DL}$  is the total gate charge of the synchronous rectifier MOSFET at  $V_{GS} = 5V$ .

The total power dissipated in the device is:

$$P_{TOTAL} = P_{MOSFET} + P_{SW} + P_Q$$

Calculate the temperature rise of the die using the following equation:

$$T_J = T_C + (P_{TOTAL} \times \theta_{JC})$$

$\theta_{JC}$  is the junction-to-case thermal resistance equal to  $1.7^{\circ}\text{C}/\text{W}$ .  $T_C$  is the temperature of the case and  $T_J$  is the junction temperature, or die temperature. The case-to-ambient thermal resistance is dependent on how well heat can be transferred from the PCB to the air. Solder the underside exposed pad to a large copper GND plane. If the die temperature reaches  $+170^{\circ}\text{C}$  the MAX15036/MAX15037 shut down and do not restart again until the die temperature cools by  $25^{\circ}\text{C}$ .

### Compensation

The MAX15036/MAX15037 have an internal transconductance error amplifier with an inverting input (FB) and output (COMP) available for external frequency compensation. The flexibility of external compensation and high switching frequencies for the MAX15036/MAX15037 allow a wide selection of output filtering components, especially the output capacitor. For cost-sensitive applications, use high-ESR aluminum electrolytic capacitors. For size-sensitive applications, use low-ESR tantalum or ceramic capacitors at the output.

Before designing the compensation components, first choose all the passive power components that meet the output ripple, component size, and component cost requirements. Secondly, choose the compensation components to achieve the desired closed-loop bandwidth and phase margin. Use a simple 1-zero, 2-pole pair (Type II) compensation if the output capacitor ESR zero frequency ( $f_{ZESR}$ ) is below the unity-gain crossover frequency ( $f_C$ ). Use a 2-zero, 2-pole (Type III) compensation when the  $f_{ZESR}$  is higher than  $f_C$ .

### Buck Converter Compensation

Use procedure 1 to calculate the compensation network components when  $f_{ZESR} < f_C$ .

#### Procedure 1 (see Figure 3)

Calculate the  $f_{ZESR}$  and  $f_{LC}$  double pole:

$$f_{ZESR} = \frac{1}{2\pi \times ESR \times C_{OUT}}$$

$$f_{LC} = \frac{1}{2\pi \times \sqrt{L \times C_{OUT}}}$$

Calculate the unity-gain crossover frequency as:

$$f_C = \frac{f_{SW}}{20}$$

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

If  $f_{ZESR}$  is lower than  $f_C$  and close to  $f_{LC}$ , use a Type II compensation network where  $R_{FCF}$  provides a midband zero ( $f_{mid,zero}$ ) and  $R_{CCF}$  provides a high-frequency pole. Calculate the modulator gain ( $G_M$ ) at the crossover frequency.

$$G_M = \frac{V_{IN}}{V_{OSC}} \times \frac{ESR}{ESR + (2\pi \times f_C \times L)} \times \frac{V_{FB}}{V_{OUT}}$$

where  $V_{OSC}$  is the 1Vp-p ramp amplitude and  $V_{FB} = 0.6V$ . The transconductance error amplifier gain at  $f_C$  is:

$$G_E/A = g_m \times R_F$$

The total loop gain at  $f_C$  should be equal to 1:

$$G_M = G_E/A = 1$$

or

$$R_F = \frac{V_{OSC} (ESR + 2\pi \times f_C \times L) V_{OUT}}{V_{FB} \times V_{IN} \times g_m \times ESR}$$

Place a zero at or below the LC double pole:

$$C_F = \frac{1}{2\pi \times R_F \times f_{LC}}$$

Place a high-frequency pole at  $f_P = 0.5 \times f_{SW}$ . Therefore  $C_{CF}$  is:

$$C_{CF} = \frac{1}{\pi \times R_F \times f_{SW}}$$

### Procedure 2 (see Figure 4)

When using a low-ESR ceramic-type capacitor as the output capacitor, the ESR frequency is much higher than the targeted unity-gain crossover frequency ( $f_C$ ). In this case, Type III compensation is recommended. Type III compensation provides a low-frequency pole ( $\approx DC$ ) and two pole-zero pairs. The locations of the zero and poles should be such that the phase margin peaks at  $f_C$ .

$$\frac{f_C}{f_Z} = \frac{f_P}{f_C} = 5$$

The  $\frac{f_C}{f_Z} = \frac{f_P}{f_C} = 5$  is a good number to get approximately 60° of phase margin at  $f_C$ . However, it is important to place the two zeros at or below the double pole to avoid conditional stability.

First, select the crossover frequency so that:

$$f_C \leq \frac{f_{SW}}{20}$$

Calculate the LC double-pole frequency,  $f_{LC}$ :

$$f_{LC} = \frac{1}{2\pi \times \sqrt{L \times C_{OUT}}}$$

$$\text{Place a zero } f_Z = \frac{1}{2\pi \times R_F \times C_F} \text{ at } 0.75 \times f_{LC}$$

where:

$$C_F = \frac{1}{2\pi \times 0.75 \times f_{LC} \times R_F}$$

with  $R_F \geq 10k\Omega$ .

Calculate  $C_A$  for a target unity crossover frequency,  $f_C$ :

$$C_A = \frac{2\pi \times f_C \times L \times C_{OUT} \times V_{OSC}}{V_{IN} \times R_F}$$

$$\text{Place a pole } (f_{P1} = \frac{1}{2\pi \times R_A \times C_A}) \text{ at } f_{ZESR}.$$

$$R_A = \frac{1}{2\pi \times f_{ZESR} \times C_A}$$

Place a second zero,  $f_{Z2}$ , at  $0.2 \times f_C$  or at  $f_{LC}$ , whichever is lower.

$$R_1 = \frac{1}{2\pi \times f_{Z2} \times C_A} - R_A$$

$$\text{Place a second pole } (f_{P2} = \frac{1}{2\pi \times R_F \times C_{CF}})$$

at 1/2 the switching frequency.

$$C_{CF} = \frac{C_F}{(2\pi \times 0.5 \times f_{SW} \times R_F \times C_F) - 1}$$

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

MAX15036/MAX15037

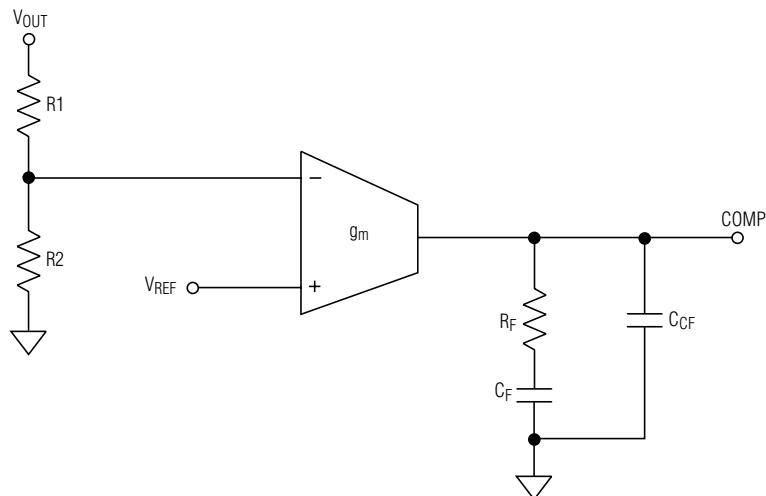


Figure 3. Type II Compensation Network

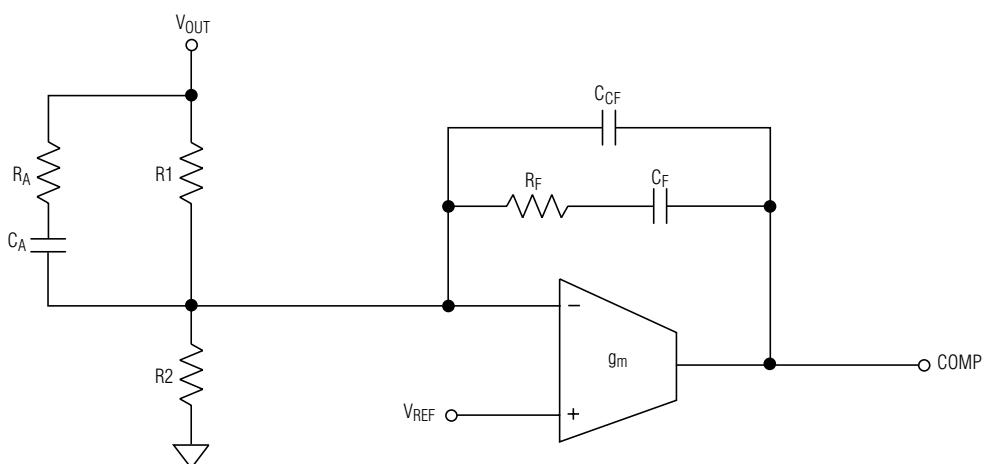


Figure 4. Type III Compensation Network

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

### Boost Converter Compensation

The boost converter compensation gets complicated due to the presence of a right-half-plane zero  $f_{ZERO,RHP}$ . The right-half-plane zero causes a drop in-phase while adding positive (+1) slope to the gain curve. It is important to drop the gain significantly below unity before the RHP frequency. Use the following procedure to calculate the compensation components. (See Figure 4.)

- 1) Calculate the LC double-pole frequency,  $f_{LC}$ , and the right half plane zero frequency.

$$f_{LC} = \frac{1 - D}{2\pi \times \sqrt{LC_{OUT}}}$$

$$f_{ZERO, RHP} = \frac{(1 - D)^2 R_{(MIN)}}{2\pi \times L}$$

where:

$$D = 1 - \frac{V_{IN}}{V_{OUT}}$$

$$R_{(MIN)} = \frac{V_{OUT}}{I_{OUT(MAX)}}$$

Target the unity-gain crossover frequency for:

$$f_C \leq \frac{f_{ZERO, RHP}}{5}$$

- 2) Place a zero  $(f_{Z1} = \frac{1}{2\pi \times R_F \times C_F})$  at  $0.75 \times f_{LC}$ .

$$C_F = \frac{1}{2\pi \times 0.75 \times f_{LC} \times R_F}$$

where  $R_F \geq 10k\Omega$ .

- 3) Calculate  $C_A$  for a target crossover frequency,  $f_C$ :

$$C_A = \frac{V_{OSC} [(1 - D)^2 + \omega_C^2 LC_{OUT}]}{\omega_C R_F V_{IN}}$$

where  $\omega_C = 2\pi f_C$ .

- 4) Place a pole  $(f_{P1} = \frac{1}{2\pi \times R_A \times C_A})$  at  $f_{ZERO,RHP}$ .

$$R_A = \frac{1}{2\pi \times f_{ZERO,RHP} \times C_A}$$

- 5) Place the second zero  $(f_{Z2} = \frac{1}{2\pi \times R_1 \times C_A})$  at  $f_{LC}$ .

$$R_1 = \frac{1}{2\pi \times f_{LC} \times C_A} - R_A$$

- 6) Place the second pole  $(f_{P2} = \frac{1}{2\pi \times R_F \times C_{CF}})$  at  $1/2$  the switching frequency.

$$C_{CF} = \frac{C_F}{(2\pi \times 0.5 \times f_{SW} \times R_F \times C_F) - 1}$$

### Improving Noise Immunity

When using the MAX15036/MAX15037 in noisy environments, adjust the controller's compensation to improve the system's noise immunity. In particular, high-frequency noise coupled into the feedback loop causes duty-cycle jitter. One solution is to lower the crossover frequency (see the Compensation section).

### PCB Layout Guidelines

Careful PCB layout is critical to achieve low-switching power losses and clean stable operation. Use a multi-layer board whenever possible for better noise immunity. Follow these guidelines for good PCB layout:

- 1) Solder the exposed pad to a large copper plane under the IC. To effectively use this copper area as a heat exchanger between the PCB and the ambient, expose this copper area on the top and bottom

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

side of the PCB. Do not make a direct connection of the exposed pad copper plane to the SGND (pin 10) underneath the IC. Connect this plane and SGND together at the return terminal of the V+ bypass capacitor

- 2) Isolate the power components and high-current paths from sensitive analog circuitry.
- 3) Keep the high-current paths short, especially at the ground terminals. This practice is essential for stable, jitter-free operation.
- 4) Connect SGND and PGND together close to the return terminals of the  $V_L$  and V+ high-frequency bypass capacitors near the IC. Do not connect them together anywhere else.
- 5) Keep the power traces and load connections short. This practice is essential for high efficiency. Use thick copper PCBs to enhance full-load efficiency and power dissipation capability.
- 6) Ensure that the feedback connection from FB to  $C_{OUT}$  is short and direct.
- 7) Route high-speed switching nodes (BST/VDD, SOURCE) away from the sensitive analog areas (BYPASS, COMP, FB, and OSC). Use internal PCB layers for SGND as EMI shields to keep radiated noise away from the IC, feedback dividers, and the analog bypass capacitors.

### Layout Procedure

- 1) Place the power components (inductor,  $C_{IN}$ , and  $C_{OUT}$ ) first, with ground terminals close to each other. Make all these connections on the top layer with wide, copper-filled areas (2oz copper recommended).
- 2) Group the gate-drive components (boost diodes and capacitors, and  $V_L$  bypass capacitor) together near the controller IC.
- 3) Make the ground connections as follows:
  - a) Create a small-signal ground plane underneath the IC.
  - b) Connect this plane to SGND and use this plane for the ground connection for BYPASS, COMP, FB, and OSC.
  - c) Connect SGND and PGND together at the return terminal of V+ and  $V_L$  bypass capacitors near the IC. Make this the only connection between SGND and PGND.

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

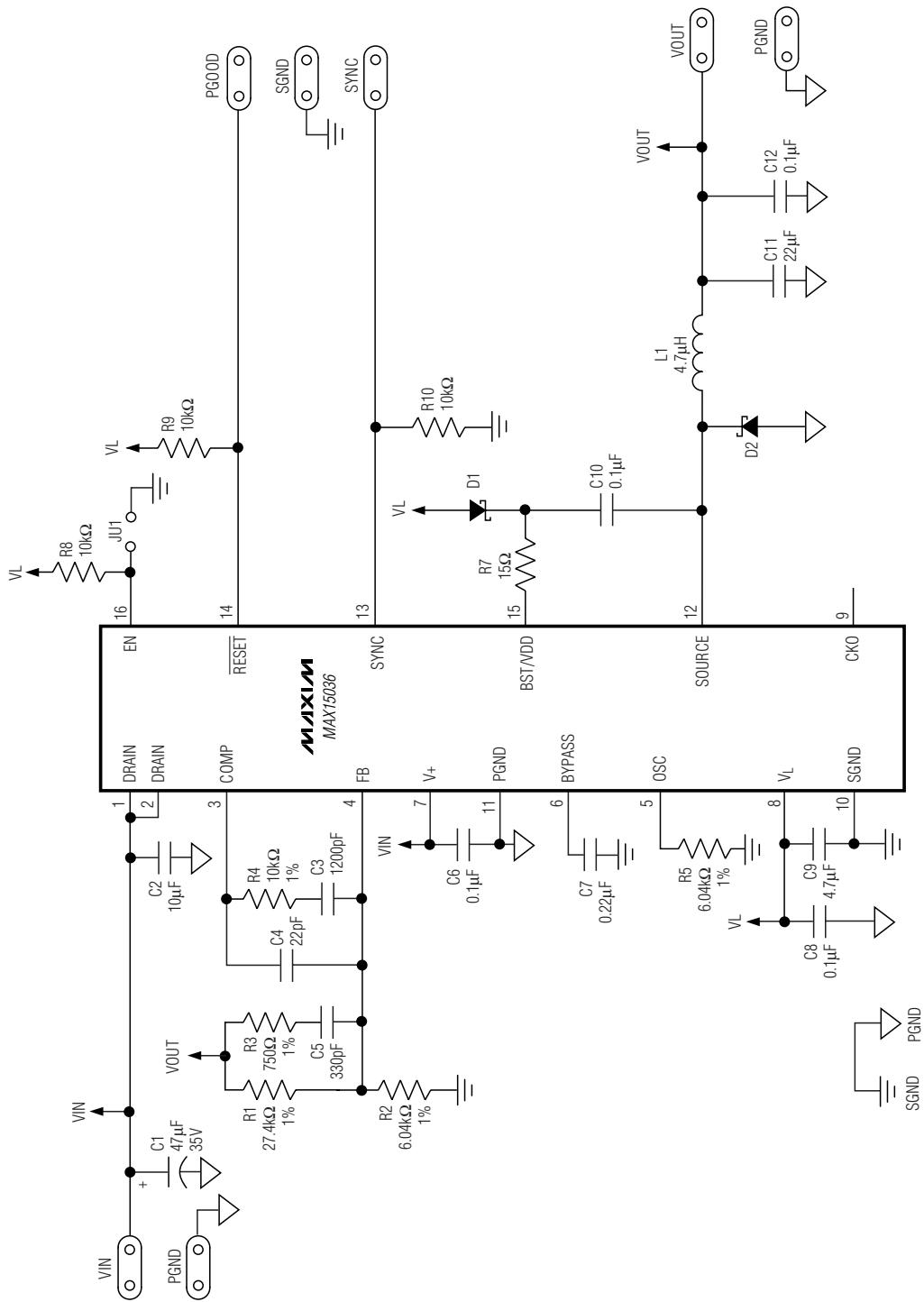


Figure 5. MAX15036 Buck Configuration

## 2.2MHz, 3A Buck or Boost Converters with an Integrated High-Side Switch

MAX15036/MAX15037

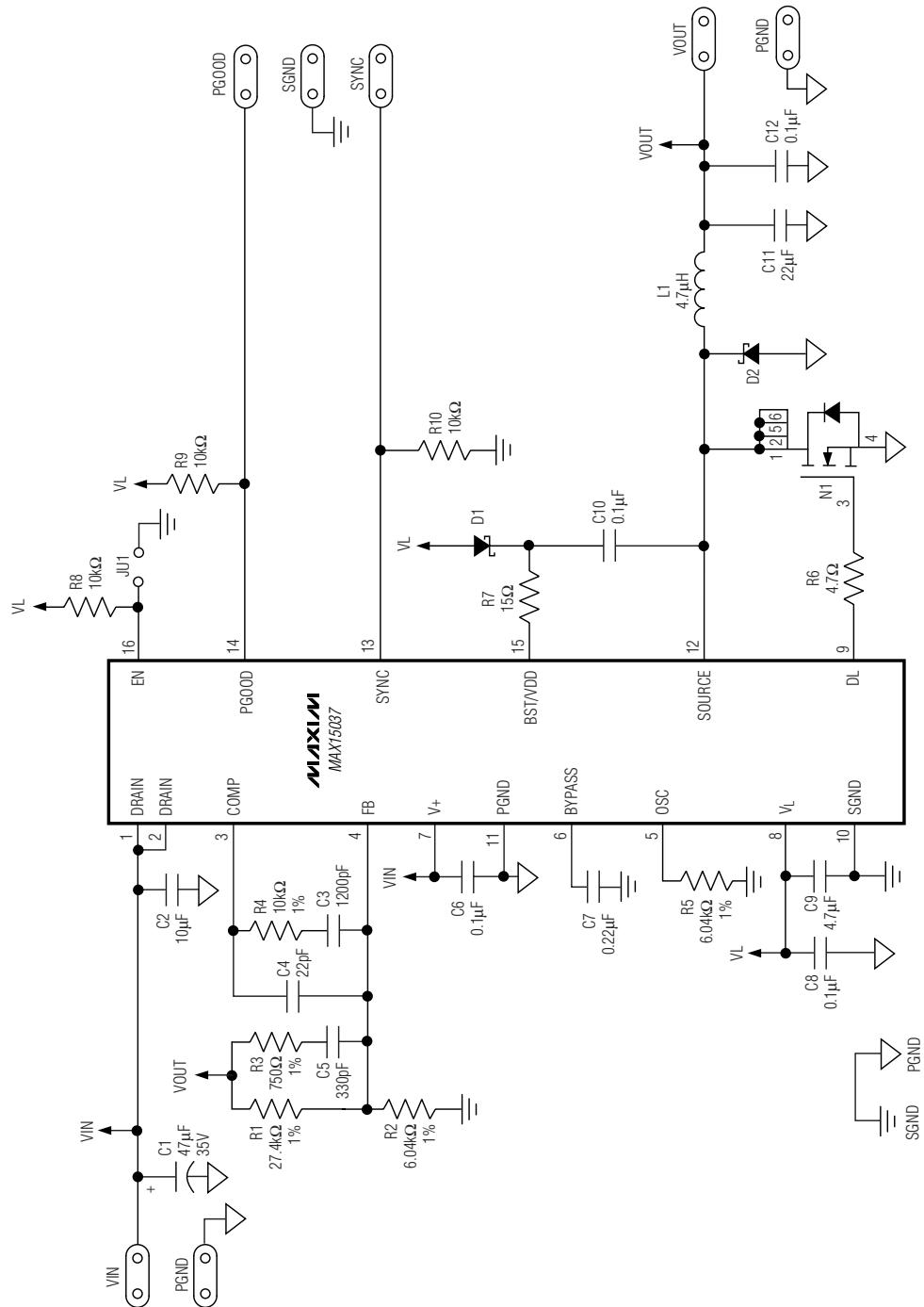


Figure 6. MAX15037 Buck Configuration

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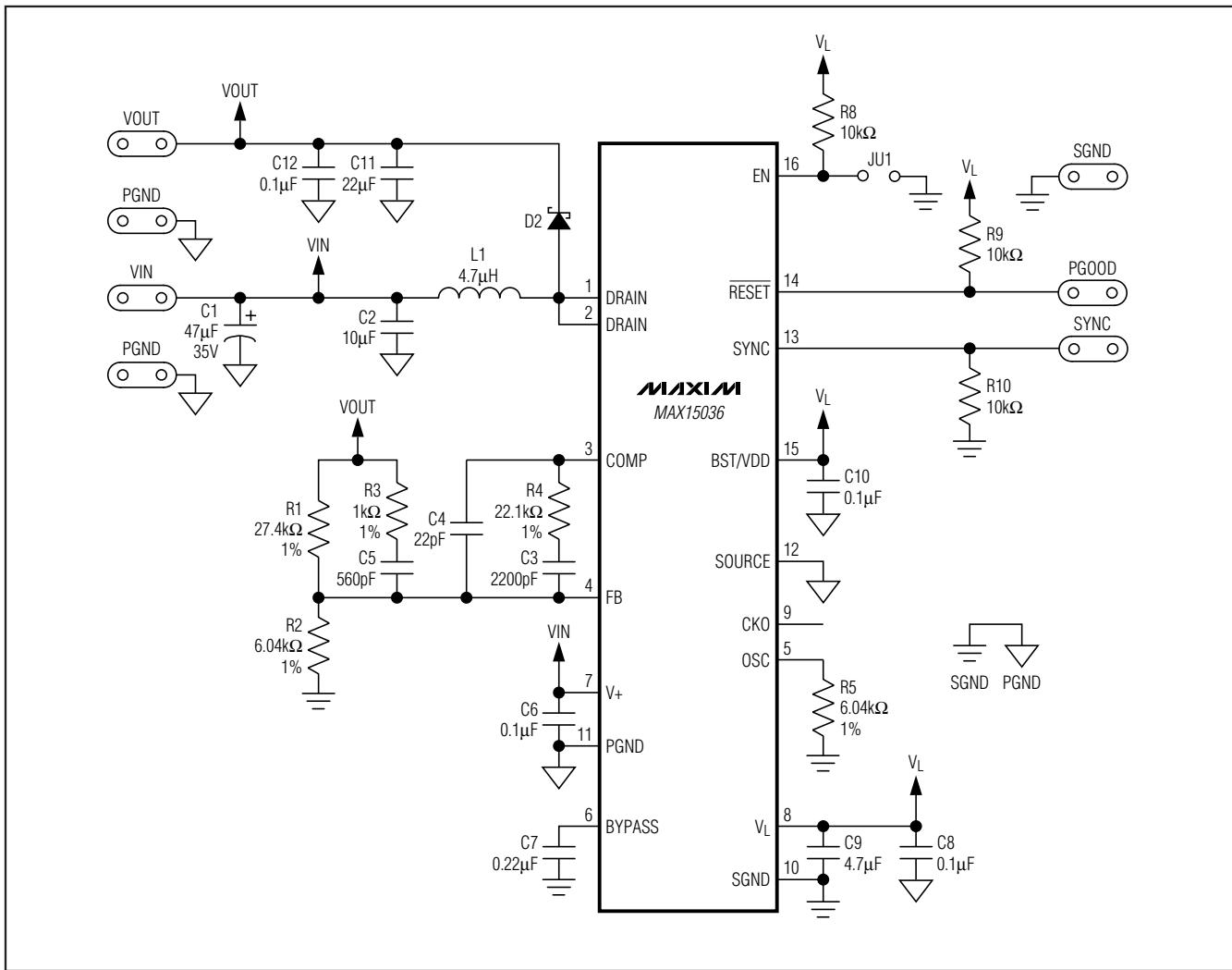


Figure 7. MAX15036 Boost Configuration

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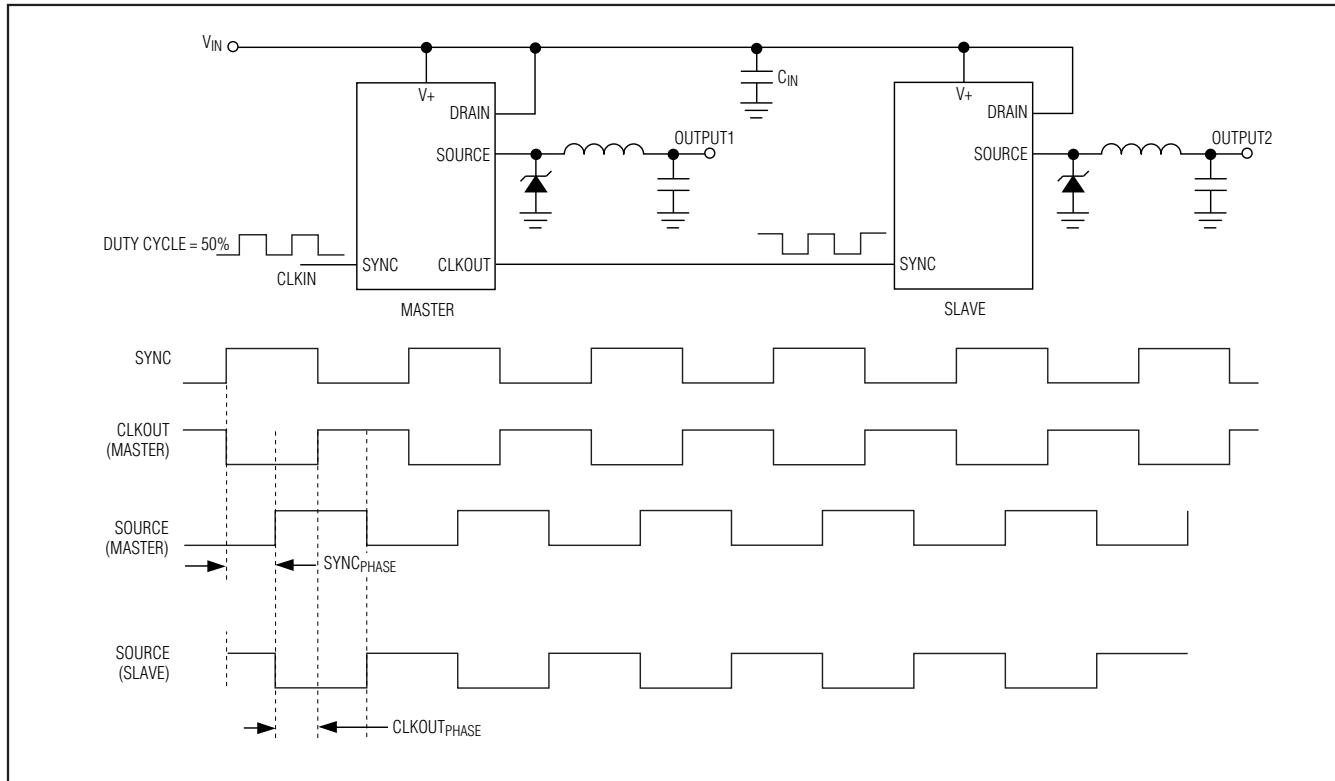
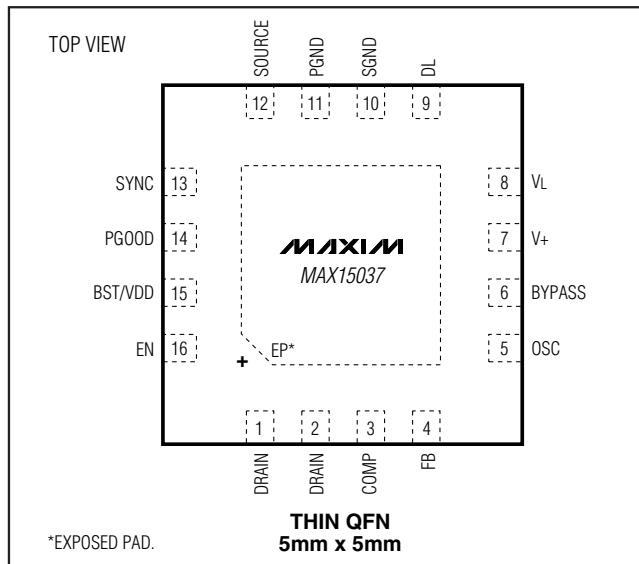


Figure 8. Synchronized Converters

### Pin Configurations (continued)



### Chip Information

PROCESS: BiCMOS

### Package Information

For the latest package outline information and land patterns, go to [www.maxim-ic.com/packages](http://www.maxim-ic.com/packages).

PACKAGE TYPE	PACKAGE CODE	DOCUMENT NO.
16 TQFN	T1655-2	<a href="#">21-0140</a>

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