

Absolute Maximum Ratings

IN, ON/OFF to SGND.....-0.3V to +45V
 LX to SGND.....-0.3V to ($V_{IN} + 0.3V$)
 BST to SGND.....-0.3V to ($V_{IN} + 12V$)
 BST to LX.....-0.3V to +12V
 PGND to SGND.....-0.3V to +0.3V
 REG, DVREG, SYNC to SGND.....-0.3V to +12V
 FB, COMP, SS to SGND.....-0.3V to ($V_{REG} + 0.3V$)
 C+ to PGND (MAX5082 only).....($V_{DVREG} - 0.3V$) to +12V
 C- to PGND (MAX5082 only).....-0.3V to ($V_{DVREG} + 0.3V$)
 Continuous current through internal power MOSFET (pins 11/12 connected together and pins 13/14 connected together)

$T_J = +125^{\circ}C$3A
 $T_J = +150^{\circ}C$2A
 Continuous Power Dissipation* ($T_A = +70^{\circ}C$)
 16-Pin TQFN (derate 33.3mW/ $^{\circ}C$ above $+70^{\circ}C$)....2666.7mW
 16-Pin TQFN (θ_{JA}).....30 $^{\circ}C/W$
 16-Pin TQFN (θ_{JC}).....1.7 $^{\circ}C/W$
 Operating Temperature Range.....-40 $^{\circ}C$ to $+125^{\circ}C$
 Maximum Junction Temperature..... $+150^{\circ}C$
 Storage Temperature Range.....-60 $^{\circ}C$ to $+150^{\circ}C$
 Lead Temperature (soldering, 10s)..... $+300^{\circ}C$

*As per JEDEC 51 Standard.

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

($V_{IN} = V_{ON/OFF} = 12V$, $V_{REG} = V_{DVREG}$, $V_{SYNC} = PGND = SGND$, $T_A = T_J = -40^{\circ}C$ to $+125^{\circ}C$, unless otherwise noted. Typical values are at $T_A = +25^{\circ}C$.) (Note 1)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
Input Voltage Range	V_{IN}	MAX5082	4.5		40	V
		MAX5083	7.5		40	
Undervoltage Lockout Threshold	UVLO	V_{IN} rising, MAX5082	3.9		4.2	V
		V_{IN} rising, MAX5083	6.8		7.3	
Undervoltage Lockout Hysteresis	UVLO _{HYST}	MAX5082		0.4		V
		MAX5083		0.7		
Switching Supply Current (PWM Operation)	I_{SW}	$V_{FB} = 0V$, MAX5082		10.5		mA
		$V_{FB} = 0V$, MAX5083		9.5		
Efficiency		$V_{IN} = 12V$, $V_{OUT} = 3.3V$, $I_{OUT} = 1.5A$		84		%
		$V_{IN} = 4.5V$, $V_{OUT} = 3.3V$, $I_{OUT} = 1.5A$ (MAX5082)		88		
No-Load Supply Current (PFM Operation)		MAX5082		1.4	2.5	mA
		MAX5083		1.3	2.3	
Shutdown Current	I_{SHDN}	$V_{ON/OFF} = 0V$, $V_{IN} = 40V$		200	300	μA
ON/OFF CONTROL						
Input Voltage Threshold	$V_{ON/OFF}$	$V_{ON/OFF}$ rising	1.20	1.23	1.25	V
Input Voltage Hysteresis				0.12		V
Input Bias Current		$V_{ON/OFF} = 0$ to 40V	-250		+250	nA
ERROR AMPLIFIER/SOFT-START						
Soft-Start Current	I_{SS}		8	15	24	μA
Reference Voltage (Soft-Start)	V_{SS}		1.215	1.228	1.240	V
FB Regulation Voltage	V_{FB}	$I_{COMP} = -500\mu A$ to $+500\mu A$	1.215	1.228	1.240	V
FB Input Range			0		1.5	V
FB Input Current			-250		+250	nA
COMP Voltage Range		$I_{COMP} = -500\mu A$ to $+500\mu A$	0.25		4.50	V
Open-Loop Gain				80		dB
Unity-Gain Bandwidth				1.8		MHz
FB Offset Voltage		$I_{COMP} = -500\mu A$ to $+500\mu A$	-5		+5	mV

Electrical Characteristics (continued)

($V_{IN} = V_{ON/OFF} = 12V$, $V_{REG} = V_{DVREG}$, $V_{SYNC} = PGND = SGND$, $T_A = T_J = -40^{\circ}C$ to $+125^{\circ}C$, unless otherwise noted. Typical values are at $T_A = +25^{\circ}C$.) (Note 1)

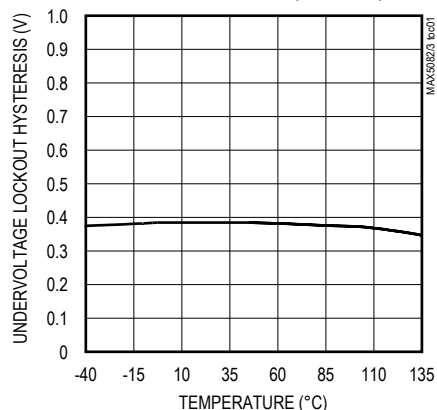
PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
OSCILLATOR						
Frequency	f _{SW}	V _{SYNC} = 0V	225	250	275	kHz
Maximum Duty Cycle	D _{MAX}	V _{SYNC} = 0V, V _{IN} = 4.5V, MAX5082	87			%
		V _{SYNC} = 0V, V _{IN} = 7.5V, MAX5083	87			
		V _{SYNC} = 0V, V _{IN} ≤ 40V	87			
SYNC High-Level Voltage			2.2			V
SYNC Low-Level Voltage					0.8	V
SYNC Frequency Range	f _{SYNC}		150		350	kHz
PWM Modulator Gain		f _{SYNC} = 150kHz to 350kHz		10		V/V
Ramp Level Shift (Valley)				0.3		V
POWER SWITCH						
Switch On-Resistance		V _{BST} - V _{LX} = 6V		0.3	0.6	Ω
Switch Gate Charge		V _{BST} - V _{LX} = 6V		6		nC
Switch Leakage Current		V _{IN} = 40V, V _{LX} = V _{BST} = 0V			10	μA
BST Leakage Current		V _{BST} = V _{LX} = V _{IN} = 40V			10	μA
CHARGE PUMP						
C- Output Voltage Low		MAX5082 only, sinking 10mA			0.1	V
C- Output Voltage High		MAX5082 only, relative to DVREG, sourcing 10mA			0.1	V
DVREG to C+ On-Resistance		MAX5082 only, sourcing 10mA			10	Ω
LX to PGND On-Resistance		Sinking 10mA			12	Ω
CURRENT-LIMIT COMPARATOR						
Pulse-Skip Threshold	I _{PFM}		100	200	300	mA
Cycle-by-Cycle Current Limit	I _{ILIM}		1.9	2.7	3.5	A
Number of Consecutive ILIM Events to Hiccup				4		
Hiccup Timeout				512		Clock periods
INTERNAL VOLTAGE REGULATOR						
Output Voltage	V _{REG}	MAX5082	4.75	5	5.25	V
		MAX5083	7.6	8	8.4	
Line Regulation		V _{IN} = 5.5V to 40V, MAX5082			1	mV/V
		V _{IN} = 9.0V to 40V, MAX5083			1	
Load Regulation		I _{REG} = 0 to 20mA			0.25	V
Dropout Voltage		V _{IN} = 4.5V, I _{REG} = 20mA, MAX5082			0.5	V
		V _{IN} = 4.5V, I _{REG} = 20mA, MAX5083			0.5	
THERMAL SHUTDOWN						
Thermal Shutdown Temperature		Temperature rising		+160		°C
Thermal Shutdown Hysteresis				20		°C

Note 1: 100% production tested at $T_A = +25^{\circ}C$ and $T_A = +125^{\circ}C$. Limits at $-40^{\circ}C$ are guaranteed by design.

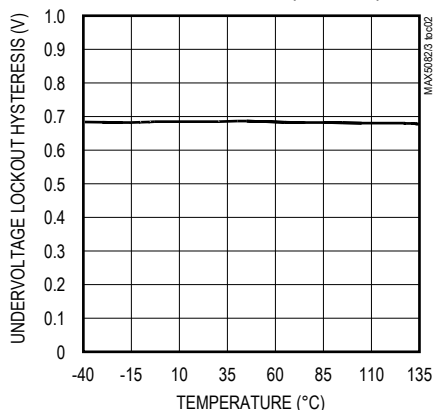
Typical Operating Characteristics

($V_{IN} = 12V$, see Figure 5 (MAX5082) and Figure 6 (MAX5083), $T_A = +25^\circ C$, unless otherwise noted.)

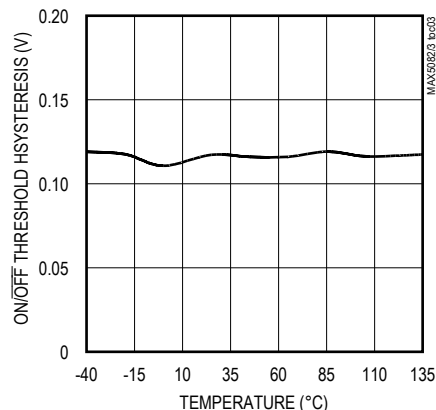
**UNDERVOLTAGE LOCKOUT HYSTERESIS
vs. TEMPERATURE (MAX5082)**



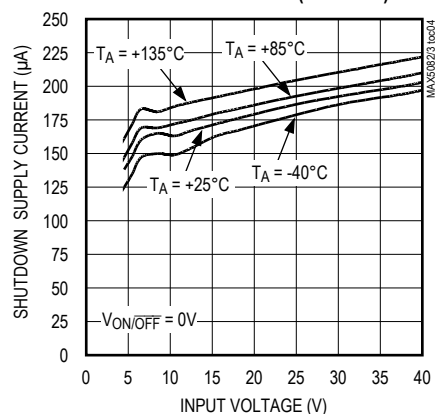
**UNDERVOLTAGE LOCKOUT HYSTERESIS
vs. TEMPERATURE (MAX5083)**



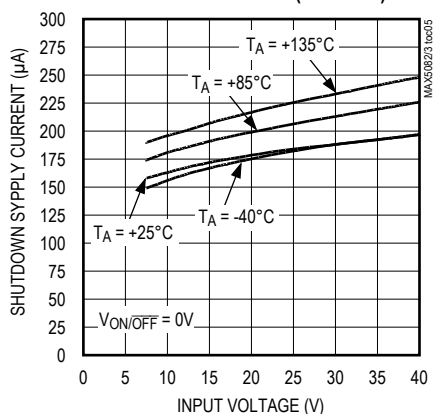
**ON/OFF THRESHOLD HYSTERESIS
vs. TEMPERATURE**



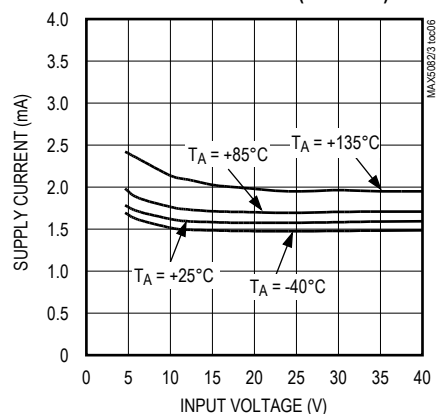
**SHUTDOWN SUPPLY CURRENT
vs. INPUT VOLTAGE (MAX5082)**



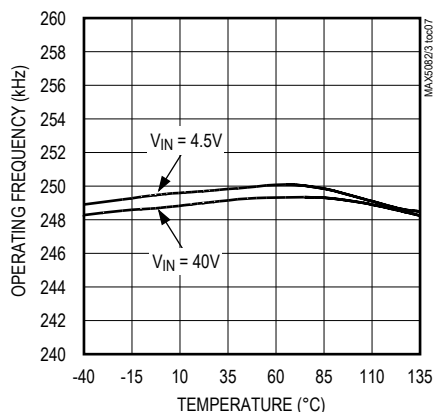
**SHUTDOWN SUPPLY CURRENT
vs. INPUT VOLTAGE (MAX5083)**



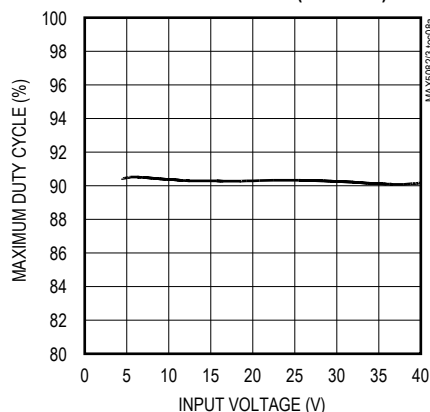
**NO-LOAD SUPPLY CURRENT
vs. INPUT VOLTAGE (MAX5082)**



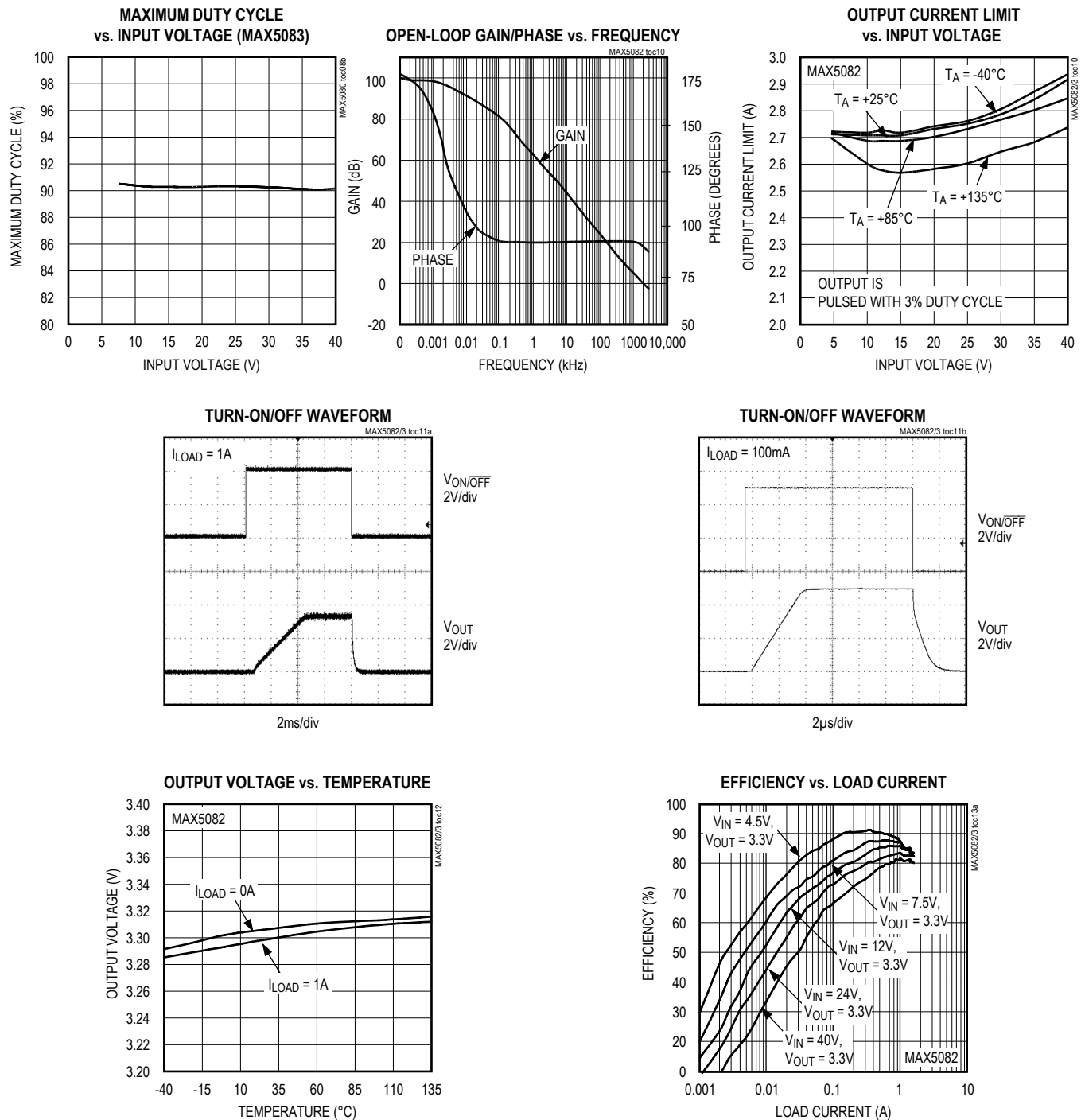
**OPERATING FREQUENCY
vs. TEMPERATURE**



**MAXIMUM DUTY CYCLE
vs. INPUT VOLTAGE (MAX5082)**

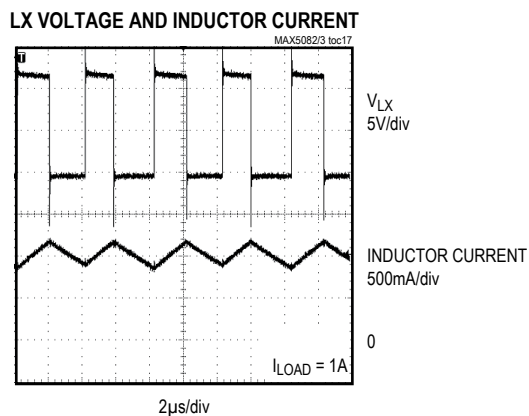
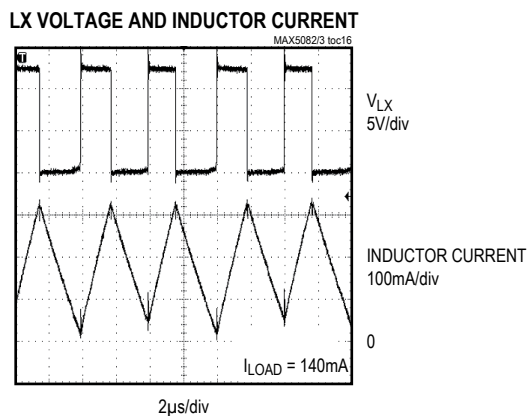
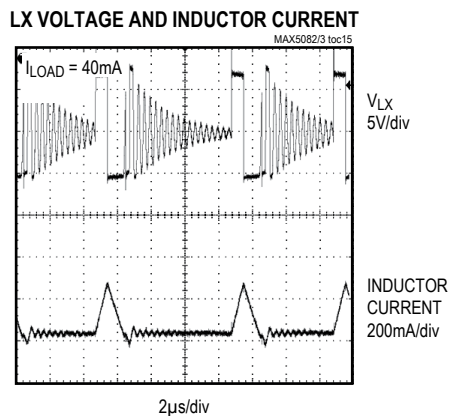
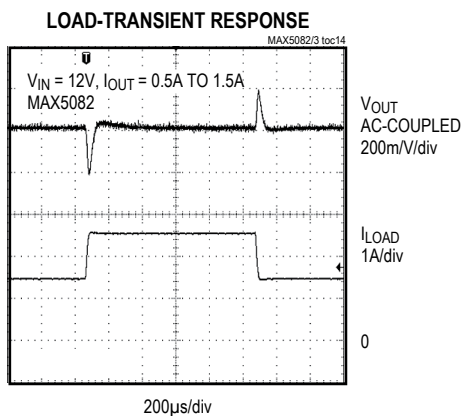
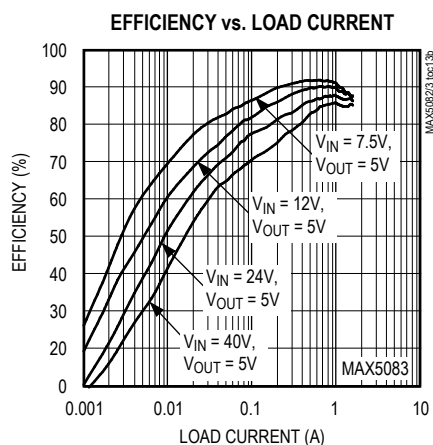


Typical Operating Characteristics (continued)

(V_{IN} = 12V, see Figure 5 (MAX5082) and Figure 6 (MAX5083), T_A = +25°C, unless otherwise noted.)

Typical Operating Characteristics (continued)

($V_{IN} = 12V$, see Figure 5 (MAX5082) and Figure 6 (MAX5083), $T_A = +25^\circ C$, unless otherwise noted.)



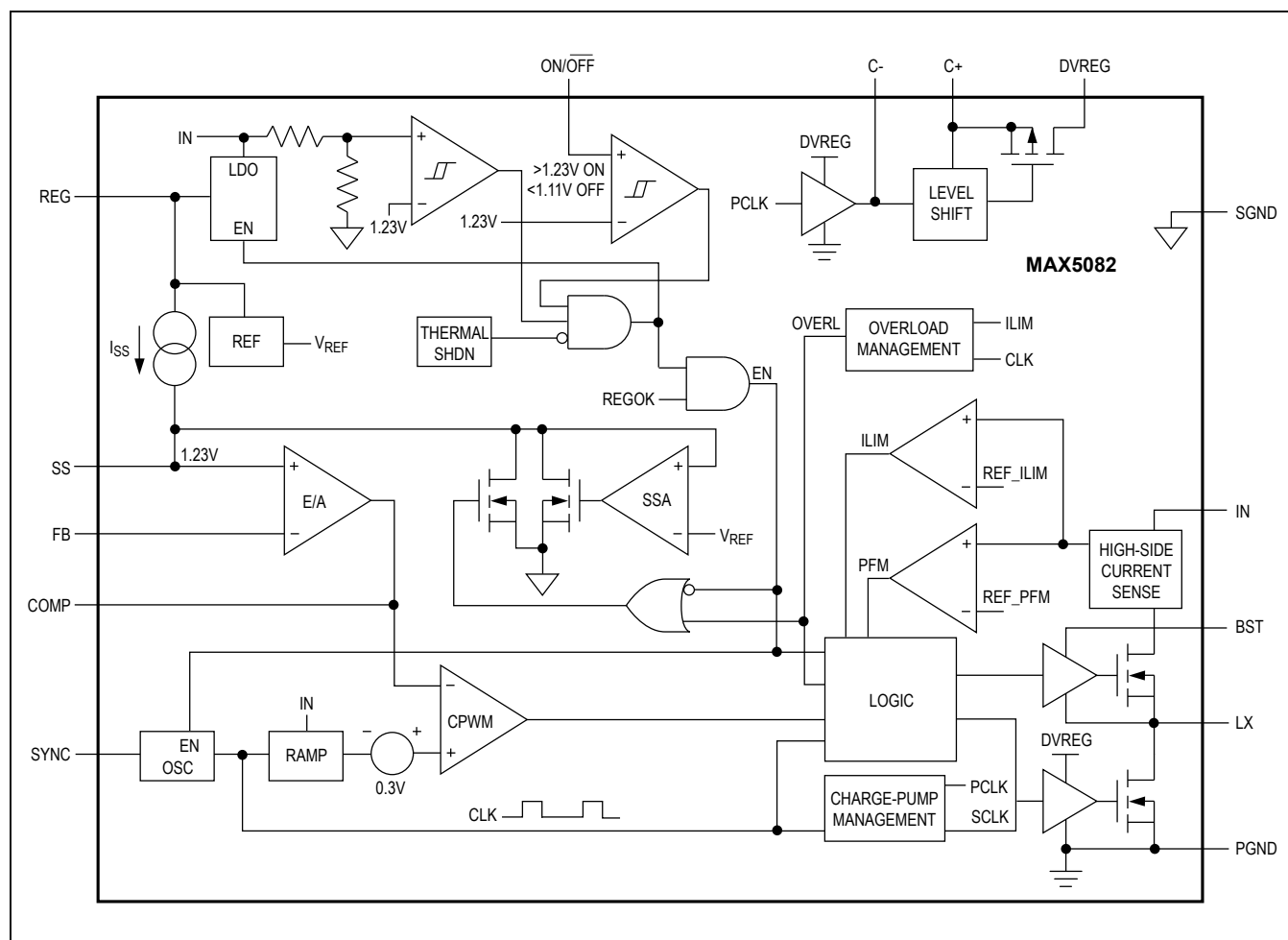
Pin Description

PIN		NAME	FUNCTION
MAX5082	MAX5083		
1	1	COMP	Error Amplifier Output. Connect COMP to the compensation feedback network.
2	2	FB	Feedback Regulation Point. Connect to the center tap of a resistive divider from converter output to SGND to set the output voltage. The FB voltage regulates to the voltage present at SS (1.23V).
3	3	ON/ $\overline{\text{OFF}}$	ON/ $\overline{\text{OFF}}$ and External UVLO Control. The ON/ $\overline{\text{OFF}}$ rising threshold is set to approximately 1.23V. Connect to the center tap of a resistive divider from IN to SGND to set the UVLO (rising) threshold. Pull ON/ $\overline{\text{OFF}}$ to SGND to shut down the device. ON/ $\overline{\text{OFF}}$ can be used for power-supply sequencing. Connect to IN for always-on operation.
4	4	SS	Soft-Start and Reference Output. Connect a capacitor from SS to SGND to set the soft-start time. See the <i>Applications Information</i> section to calculate the value of the SS capacitor.
5	5	SYNC	Oscillator Synchronization Input. SYNC can be driven by an external 150kHz to 350kHz clock to synchronize the MAX5082/MAX5083's switching frequency. Connect SYNC to SGND when not used.
6	6	DVREG	Gate Drive Supply for High-Side MOSFET Driver. Connect externally to REG for MAX5082. Connect to REG and the anode of the boost diode for MAX5083.
7	—	C+	Charge-Pump Flying Capacitor Positive Connection
8	—	C-	Charge-Pump Flying Capacitor Negative Connection
—	7, 8	N.C.	No Connection. Not internally connected. Can be left floating or connected to SGND.
9	9	PGND	Power Ground Connection. Connect the input filter capacitor's negative terminal, the anode of the freewheeling diode, and the output filter capacitor's return to PGND. Connect externally to SGND at a single point near the input capacitor's return terminal.
10	10	BST	High-Side Gate Driver Supply. Connect BST to the cathode of the boost diode and to the positive terminal of the boost capacitor.
11, 12	11, 12	LX	Source Connection of Internal High-Side Switch. Connect the inductor and rectifier diode's anode to LX.
13, 14	13, 14	IN	Supply Input Connection. Connect to an external voltage source from 4.5V to 40V (MAX5082) or a 7.5V to 40V (MAX5083).
15	15	REG	Internal Regulator Output. 5V output for the MAX5082 and 8V output for the MAX5083. Bypass to SGND with at least a 1 μ F ceramic capacitor.
16	16	SGND	Signal Ground Connection. Solder the exposed pad to a large SGND plane. Connect SGND and PGND together at one point near the input bypass capacitor return terminal.
EP	EP	—	Exposed Pad. Connect exposed pad to SGND.

Detailed Description

The MAX5082/MAX5083 are voltage-mode buck converters with internal 0.3 Ω power MOSFET switches. The MAX5082 has a wide input voltage range of 4.5V to 40V. The MAX5083's input voltage range is 7.5V to 40V. The internal low R_{DS_ON} switch allows for up to 1.5A of output current. The 250kHz fixed switching frequency, external compensation, and voltage feed-forward simplify loop compensation design and allow for a variety of L and C filter

components. Both devices offer an automatic switchover to pulse-skipping (PFM) mode, providing low quiescent current and high efficiency at light loads. Under no load, a PFM mode operation reduces the current consumption to only 1.4mA. In shutdown, the supply current falls to 200 μ A. Additional features include an externally programmable undervoltage lockout through the ON/ $\overline{\text{OFF}}$ pin, a programmable soft-start, cycle-by-cycle current limit, hiccup mode output short-circuit protection, and thermal shutdown.



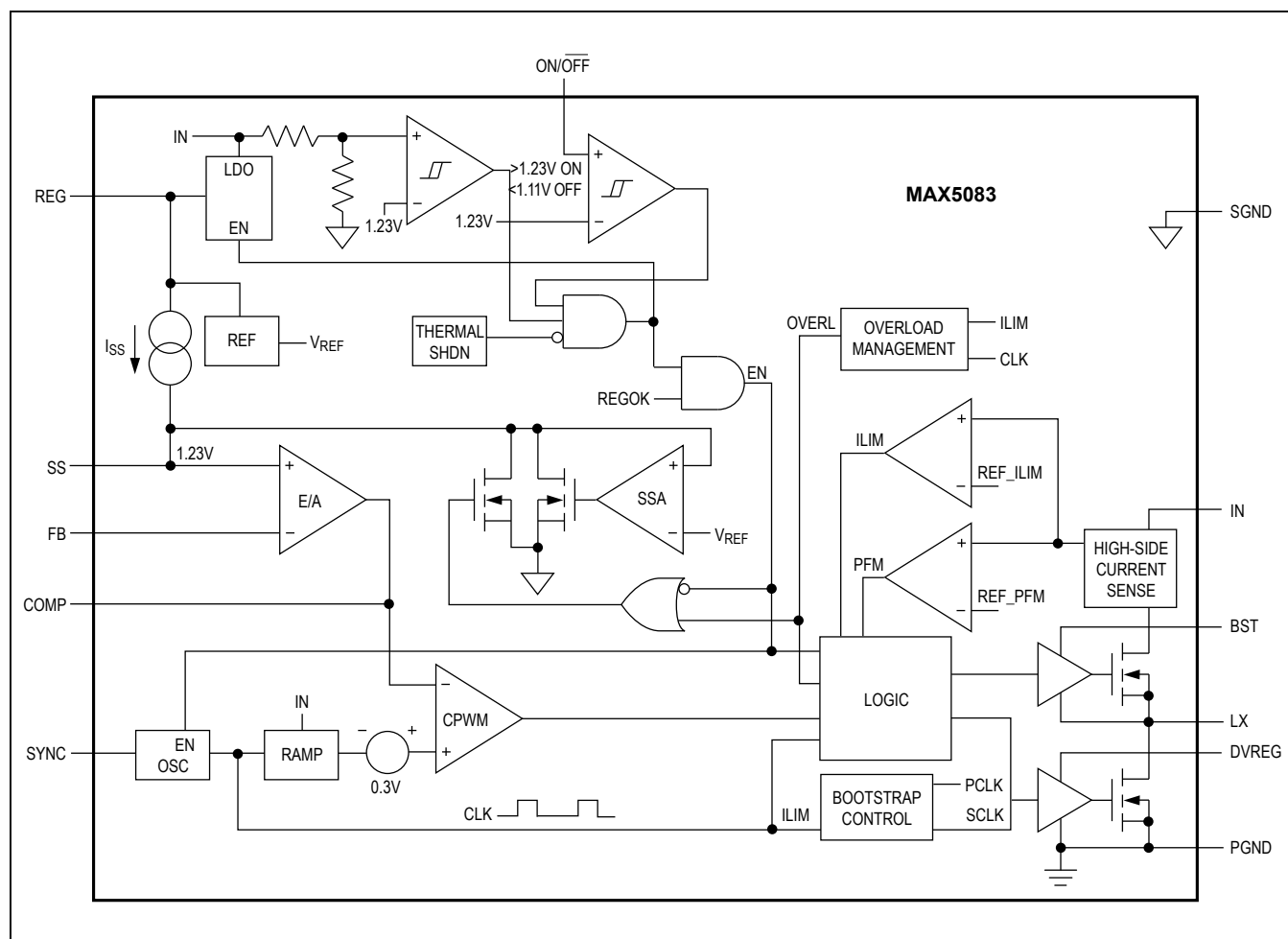


Figure 2. MAX5083 Simplified Block Diagram

Soft-Start and Reference (SS)

SS is the 1.23V reference bypass connection for the MAX5082/MAX5083 and also controls the soft-start period. At startup, after V_{IN} is applied and the internal and external UVLO thresholds are reached, the device enters soft-start. During soft-start, $15\mu A$ is sourced into the capacitor (C_{SS}) connected from SS to SGND causing the reference voltage to ramp up slowly. When V_{SS} reaches 1.23V the output becomes fully active. Set the soft-start time (t_{SS}) using the following equation:

$$t_{SS} = \frac{1.23V \times C_{SS}}{15\mu A}$$

where t_{SS} is in seconds and C_{SS} is in Farads.

Internal Charge Pump (MAX5082)

The MAX5082 features an internal charge pump to enhance the turn-on of the internal MOSFET, allowing for operation with input voltages down to 4.5V. Connect a flying capacitor (C_F) between C+ and C-, a boost diode from C+ to BST, as well as a bootstrap capacitor (C_{BST}) between BST and LX to provide the gate-drive voltage for the high-side n-channel DMOS switch. During the on-time, the flying capacitor is charged to V_{DVREG} . During the off-time, the positive terminal of the flying capacitor (C+) is pumped to two times V_{DVREG} and charge is dumped onto C_{BST} to provide twice the regulator voltage across the high-side DMOS driver. Use a ceramic capacitor of at least $0.1\mu F$ for C_{BST} and C_F located as close to the device as possible.

For applications that do not require a 4.5V minimum input, use the MAX5083. In this device, the charge pump is

omitted and the input voltage range is from 7.5V to 40V. In this situation, the boost diode and the boost capacitor are still required (see the *MAX5083 Typical Operating Circuit*).

Gate Drive Supply (DVREG)

DVREG is the supply input for the internal high-side MOSFET driver. The power for DVREG is derived from the output of the internal regulator (REG). Connect DVREG to REG externally. We recommend the use of an RC (1Ω and 0.47μF) filter from REG to DVREG to filter the noise generated by the switching of the charge pump. In the MAX5082, the high-side drive supply is generated using the internal charge pump along with the bootstrap diode and capacitor. In the MAX5083, the high-side MOSFET driver supply is generated using only the bootstrap diode and capacitor.

Error Amplifier

The output of the internal error amplifier (COMP) is available for frequency compensation (see the *Compensation Design* section). The inverting input is FB, the noninverting input SS, and the output COMP. The error amplifier has an 80dB open-loop gain and a 1.8MHz GBW product. See the *Typical Operating Characteristics* for the Gain and Phase vs. Frequency graph.

Oscillator/Synchronization Input (SYNC)

With SYNC tied to SGND, the MAX5082/MAX5083 use their internal oscillator and switch at a fixed frequency of 250kHz. For external synchronization, drive SYNC with an external clock from 150kHz to 350kHz. When driven with an external clock, the device synchronizes to the rising edge of SYNC.

PWM Comparator/Voltage Feed-Forward

An internal 250kHz ramp generator is compared against the output of the error amplifier to generate the PWM signal. The maximum amplitude of the ramp (VRAMP) automatically adjusts to compensate for input voltage and oscillator frequency changes. This causes the V_{IN}/V_{RAMP} to be a constant 10V/V across the input voltage range of 4.5V to 40V (MAX5082) or 7.5V to 40V (MAX5083) and the SYNC frequency range of 150kHz to 350kHz.

Output Short-Circuit Protection (Hiccup Mode)

The MAX5082/MAX5083 protects against an output short circuit by utilizing hiccup-mode protection. In hiccup mode, a series of sequential cycle-by-cycle current-limit events will cause the part to shut down and restart with a soft-start sequence. This allows the device to operate with a continuous output short circuit.

During normal operation, the current is monitored at the drain of the internal power MOSFET. When the current limit is exceeded, the internal power MOSFET turns off until the next on-cycle and a counter increments. If the counter counts four consecutive current-limit events, the device discharges the soft-start capacitor and shuts down for 512 clock periods before restarting with a soft-start sequence. Each time the power MOSFET turns on and the device does not exceed the current limit, the counter is reset.

Thermal-Overload Protection

The MAX5082/MAX5083 feature an integrated thermal-overload protection. Thermal-overload protection limits the total power dissipation in the device and protects it in the event of an extended thermal fault condition. When the die temperature exceeds +160°C, an internal thermal sensor shuts down the part, turning off the power MOSFET and allowing the IC to cool. After the temperature falls by 20°C, the part will restart with a soft-start sequence.

Applications Information

Setting the Undervoltage Lockout

When the voltage at ON/OFF rises above 1.23V, the MAX5082/MAX5083 turns on. Connect a resistive divider from IN to ON/OFF to SGND to set the UVLO threshold (see Figure 5). First select the ON/OFF to the SGND resistor (R2) then calculate the resistor from IN to ON/OFF (R1) using the following equation:

$$R1 = R2 \times \left[\frac{V_{IN}}{V_{ON/OFF}} - 1 \right]$$

where V_{IN} is the input voltage at which the converter turns on, $V_{ON/OFF} = 1.23V$ and R2 is chosen to be less than 600kΩ.

If the external UVLO divider is not used, connect ON/OFF to IN directly. In this case, an internal undervoltage lockout feature monitors the supply voltage at IN and allows operation to start when IN rises above 4.1V (MAX5082) and 7.1V (MAX5083).

Setting the Output Voltage

Connect a resistive divider from OUT to FB to SGND to set the output voltage (see Figure 5). First calculate the resistor from OUT to FB using the guidelines in the *Compensation Design* section. Once R3 is known, calculate R4 using the following equation:

$$R4 = \frac{R3}{\left[\frac{V_{OUT_1}}{V_{FB}} \right]}$$

where $V_{FB} = 1.23V$.

Inductor Selection

Three key inductor parameters must be specified for operation with the MAX5082/MAX5083: 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 (ΔI_{P-P}). Higher ΔI_{P-P} allows for a lower inductor value while a lower ΔI_{P-P} requires a higher inductor value. A lower inductor value minimizes size and cost and 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 ΔI_{P-P} equal to 40% of the full load current. Calculate the inductor using the following equation:

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

V_{IN} and V_{OUT} are typical values so that efficiency is optimum for typical conditions. The switching frequency (f_{SW}) is fixed at 250kHz or can vary between 150kHz and 350kHz when synchronized to an external clock (see the *Oscillator/Synchronization Input (SYNC)* section). The peak-to-peak inductor current, which reflects the peak-to-peak output ripple, is worst at the maximum input voltage. See the *Output Capacitor Selection* section to verify that the worst-case output ripple is acceptable. The inductor saturating current (I_{SAT}) is also important to avoid runaway current during continuous output short circuit. Select an inductor with an I_{SAT} specification higher than the maximum peak current limit of 3.5A.

Input Capacitor Selection

The discontinuous input current of the buck converter causes large input ripple currents and therefore the input capacitor must be carefully chosen to keep the input voltage ripple within design requirements. The input volt-

age ripple is comprised of ΔV_Q (caused by the capacitor discharge) and ΔV_{ESR} (caused by the ESR of the input capacitor). The total voltage ripple is the sum of ΔV_Q and ΔV_{ESR} . Calculate the input capacitance and ESR required for a specified ripple using the following equations:

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

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

where

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

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

I_{OUT_MAX} is the maximum output current, D is the duty cycle, and f_{SW} is the switching frequency.

The MAX5082/MAX5083 includes internal and external UVLO hysteresis and soft-start to avoid possible unintentional chattering during turn-on. However, use a bulk capacitor if the input source impedance is high. Use enough input capacitance at lower input voltages to avoid possible undershoot below the undervoltage lockout threshold during transient loading.

Output Capacitor Selection

The allowable output voltage ripple and the maximum deviation of the output voltage during load steps determine the output capacitance and its ESR. The output ripple is mainly composed of ΔV_Q (caused by the capacitor discharge) and ΔV_{ESR} (caused by the voltage drop across the equivalent series resistance of the output capacitor). The equations for calculating the peak-to-peak output voltage ripple are:

$$\Delta V_Q = \frac{\Delta I_{P-P}}{16 \times C_{OUT} \times f_{SW}}$$

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

Normally, a good approximation of the output voltage ripple is $\Delta V_{RIPPLE} \approx \Delta V_{ESR} + \Delta V_Q$. If using ceramic capacitors, assume the contribution to the output voltage ripple from ESR and the capacitor discharge to be equal to 20% and 80%, respectively. ΔI_{P-P} is the peak-to-peak inductor current (see the *Input Capacitors Selection* section) and f_{SW} is the converter's switching frequency.

The allowable deviation of the output voltage during fast load transients also determines the output capacitance, its ESR, and its equivalent series inductance (ESL). The output capacitor supplies the load current during a load step until the controller responds with a greater duty cycle. The response time (t_{RESPONSE}) depends on the closed-loop bandwidth of the converter (see the *Compensation Design* section). The resistive drop across the output capacitor's ESR, the drop across the capacitor's ESL (ΔV_{ESL}), and the capacitor discharge causes a voltage droop during the loadstep. Use a combination of low-ESR tantalum/aluminum electrolyte and ceramic capacitors for better transient load and voltage ripple performance. Nonleaded capacitors and capacitors in parallel help reduce the ESL. Keep the maximum output voltage deviation below the tolerable limits of the electronics being powered. Use the following equations to calculate the required ESR, ESL, and capacitance value during a load step:

$$\begin{aligned} \text{ESR} &= \frac{\Delta V_{\text{ESR}}}{I_{\text{STEP}}} \\ C_{\text{OUT}} &= \frac{I_{\text{STEP}} \times t_{\text{RESPONSE}}}{\Delta V_{\text{Q}}} \\ \text{ESL} &= \frac{\Delta V_{\text{ESL}} \times t_{\text{STEP}}}{I_{\text{STEP}}} \end{aligned}$$

where I_{STEP} is the load step, t_{STEP} is the rise time of the load step, and t_{RESPONSE} is the response time of the controller.

Compensation Design

The MAX5082/MAX5083 use a voltage-mode control scheme that regulates the output voltage by comparing the error amplifier output (COMP) with an internal ramp to produce the required duty cycle. The output lowpass LC filter creates a double pole at the resonant frequency, which has a gain drop of -40dB/decade. The error amplifier must compensate for this gain drop and phase shift to achieve a stable closed-loop system.

The basic regulator loop consists of a power modulator, an output feedback divider, and a voltage-error amplifier. The power modulator has a DC gain set by $V_{\text{IN}}/V_{\text{RAMP}}$, with a double pole and a single zero set by the output inductance (L), the output capacitance (C_{OUT}) (C_5 in

the *Typical Application Circuit*) and its equivalent series resistance (ESR). The power modulator incorporates a voltage feed-forward feature, which automatically adjusts for variations in the input voltage resulting in a DC gain of 10. The following equations define the power modulator:

$$\begin{aligned} G_{\text{MOD(DC)}} &= \frac{V_{\text{IN}}}{V_{\text{RAMP}}} = 10 \\ f_{\text{LC}} &= \frac{1}{2\pi\sqrt{L \times C_{\text{OUT}}}} \\ f_{\text{ZESR}} &= \frac{1}{2\pi \times C_{\text{OUT}} \times \text{ESR}} \end{aligned}$$

The switching frequency is internally set at 250kHz or can vary from 150kHz to 350kHz when driven with an external SYNC signal. The crossover frequency (f_{C}), which is the frequency when the closed-loop gain is equal to unity, should be set at 15kHz or below therefore:

$$f_{\text{C}} \leq 15\text{kHz}$$

The error amplifier must provide a gain and phase bump to compensate for the rapid gain and phase loss from the LC double pole. This is accomplished by utilizing a type 3 compensator that introduces two zeroes and 3 poles into the control loop. The error amplifier has a low-frequency pole (f_{P1}) near the origin.

The two zeros are at:

$$f_{\text{Z1}} = \frac{1}{2\pi \times R_5 \times C_7} \text{ and } f_{\text{Z2}} = \frac{1}{2\pi \times (R_6 + R_3) \times C_6}$$

and the higher frequency poles are at:

$$f_{\text{P2}} = \frac{1}{2\pi \times R_6 \times C_6} \text{ and } f_{\text{P3}} = \frac{1}{2\pi \times R_5 \times \left(\frac{C_7 \times C_8}{C_7 + C_8} \right)}$$

Compensation When $f_{\text{C}} < f_{\text{ZESR}}$

Figure 3 shows the error amplifier feedback as well as its gain response for circuits that use low-ESR output capacitors (ceramic). In this case f_{ZESR} occurs after f_{C} .

f_{Z1} is set to $0.8 \times f_{\text{LC(MOD)}}$ and f_{Z2} is set to f_{LC} to compensate for the gain and phase loss due to the double pole. Choose the inductor (L) and output capacitor (C_{OUT}) as described in the *Inductor and Output Capacitor Selection* section.

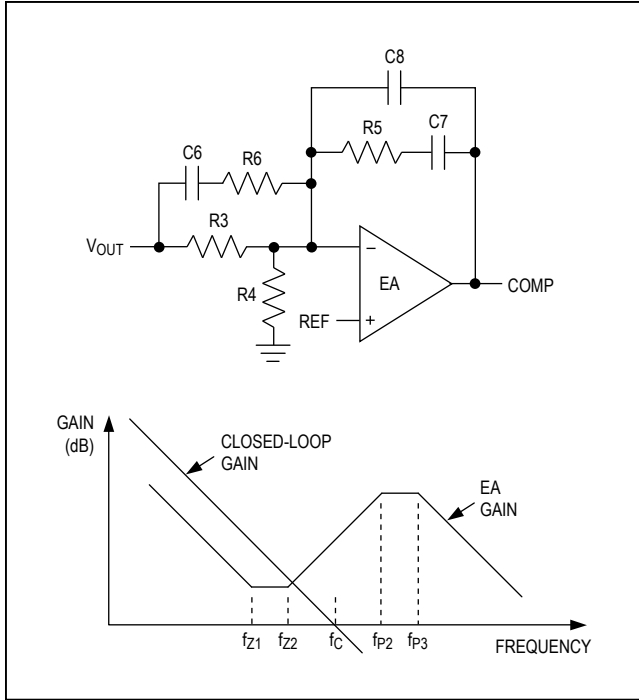


Figure 3. Error Amplifier Compensation Circuit (Closed-Loop and Error-Amplifier Gain Plot) for Ceramic Capacitors

Pick a value for the feedback resistor R5 in Figure 3 (values between 1kΩ and 10kΩ are adequate).

C7 is then calculated as:

$$C7 = \frac{1}{2\pi \times 0.8 \times f_{LC} \times R5}$$

f_C occurs between f_{Z2} and f_{P2} . The error-amplifier gain (GEA) at f_C is due primarily to C6 and R5. Therefore, $G_{EA}(f_C) = 2\pi \times f_C \times C6 \times R5$ and the modulator gain at f_C is:

$$G_{MOD}(f_C) = \frac{G_{MOD}(DC)}{(2\pi)^2 \times L \times C_{OUT} \times f_C^2}$$

Since $G_{EA}(f_C) \times G_{MOD}(f_C) = 1$, C6 is calculated by:

$$C6 = \frac{f_C \times L \times C_{OUT} \times 2\pi}{R5 \times G_{MOD}(DC)}$$

f_{P2} is set at one-half the switching frequency (f_{SW}). R6 is then calculated by:

$$R6 = \frac{1}{2\pi \times C6 \times 0.5 \times f_{SW}}$$

Since $R3 \gg R6$, $R3 + R6$ can be approximated as R3. R3 is then calculated as:

$$R3 \approx \frac{1}{2\pi \times f_{LC} \times C6}$$

f_{P3} is set at $5 \times f_C$. Therefore, C8 is calculated as:

$$C8 = \frac{C7}{(2\pi \times C7 \times R5 \times f_{P3} - 1)}$$

Compensation When $f_C > f_{ZESR}$

For larger ESR capacitors such as tantalum and aluminum electrolytic ones, f_{ZESR} can occur before f_C . If $f_{ZESR} < f_C$, then f_C occurs between f_{P2} and f_{P3} . f_{Z1} and f_{Z2} remain the same as before however, f_{P2} is now set equal to f_{ZESR} . The output capacitor's ESR zero frequency is higher than f_{LC} but lower than the closed-loop crossover frequency. The equations that define the error amplifier's poles and zeroes (f_{Z1} , f_{Z2} , f_{P1} , f_{P2} , and f_{P3}) are the same as before. However, f_{P2} is now lower than the closed-loop crossover frequency. Figure 4 shows the error amplifier feedback as well as its gain response for circuits that use higher-ESR output capacitors (tantalum or aluminum electrolytic).

Pick a value for the feedback resistor R5 in Figure 4 (values between 1kΩ and 10kΩ are adequate).

C7 is then calculated as:

$$C7 = \frac{1}{2\pi \times 0.8 \times f_{LC} \times R5}$$

The error amplifier gain between f_{P2} and f_{P3} is approximately equal to $R5/R6$ (given that $R6 \ll R3$). R6 can then be calculated as:

$$R6 \approx \frac{R5 \times 10 \times f_{LC}^2}{f_C^2}$$

C6 is then calculated as:

$$C6 = \frac{C_{OUT} \times ESR}{R6}$$

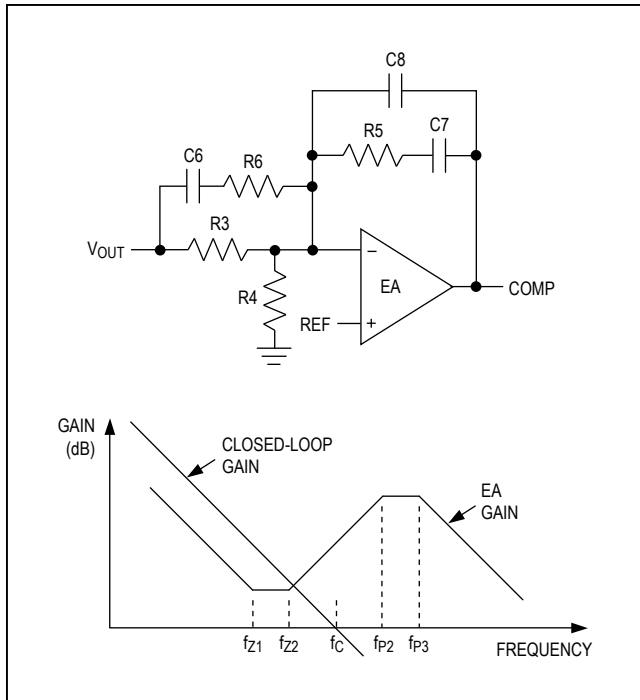


Figure 4. Error Amplifier Compensation Circuit (Closed-Loop and Error Amplifier Gain Plot) for Higher ESR Output Capacitors

Since $R3 \gg R6$, $R3 + R6$ can be approximated as $R3$. $R3$ is then calculated as:

$$R3 \approx \frac{1}{2\pi \times f_{LC} \times C6}$$

f_{p3} is set at $5 \times f_c$. Therefore, $C8$ is calculated as:

$$C8 = \frac{C7}{(2\pi \times C7 \times R5 \times f_{p3} - 1)}$$

Power Dissipation

The MAX5082/MAX5083 is available in a thermally enhanced package and can dissipate up to 2.7W at $T_A = +70^\circ\text{C}$. When the die temperature reaches $+160^\circ\text{C}$, the part shuts down and is allowed to cool. After the part cools by 20°C , the device restarts with a soft-start.

The power dissipated in the device is the sum of the power dissipated from supply current (P_Q), transition losses due to switching the internal power MOSFET (P_{SW}), and the power dissipated due to the RMS current through the internal power MOSFET (P_{MOSFET}). The total power dissipated in the package must be limited such that the junction temperature does not exceed its absolute maximum rating of $+150^\circ\text{C}$ at maximum ambient temperature. Calculate the power lost in the MAX5082/MAX5083 using the following equations:

The power loss through the switch:

$$P_{MOSFET} = I_{RMS_MOSFET}^2 \times R_{ON}$$

$$I_{RMS_MOSFET} = \sqrt{\left[I_{PK}^2 + (I_{PK} \times I_{DC}) + I_{DC}^2 \right] \times \frac{D}{3}}$$

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

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

R_{ON} is the on-resistance of the internal power MOSFET (see the *Electrical Characteristics*).

The power loss due to switching the internal MOSFET:

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

where t_R and t_F are the rise and fall times of the internal power MOSFET measured at LX.

The power loss due to the switching supply current (I_{SW}):

$$P_Q = V_{IN} \times I_{SW}$$

The total power dissipated in the device will be:

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

Chip Information

TRANSISTOR COUNT: 4300

PROCESS: BiCMOS/DMOS

Typical Application Circuits

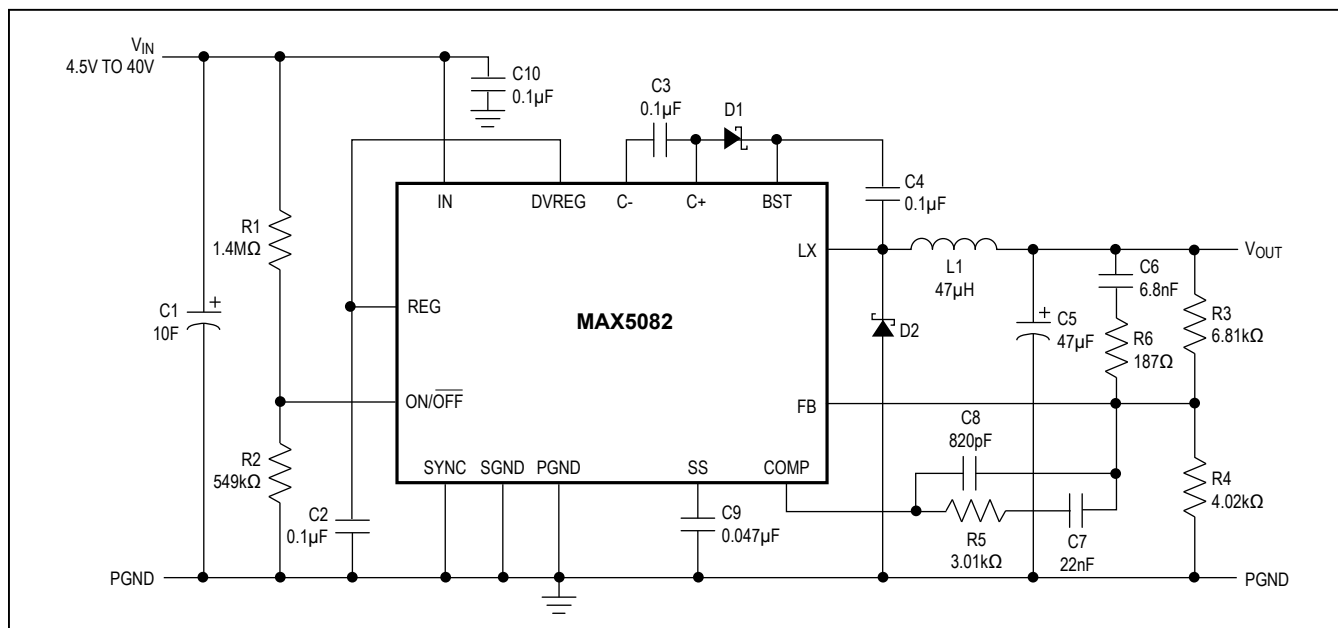


Figure 5. MAX5082 Typical Application Circuit

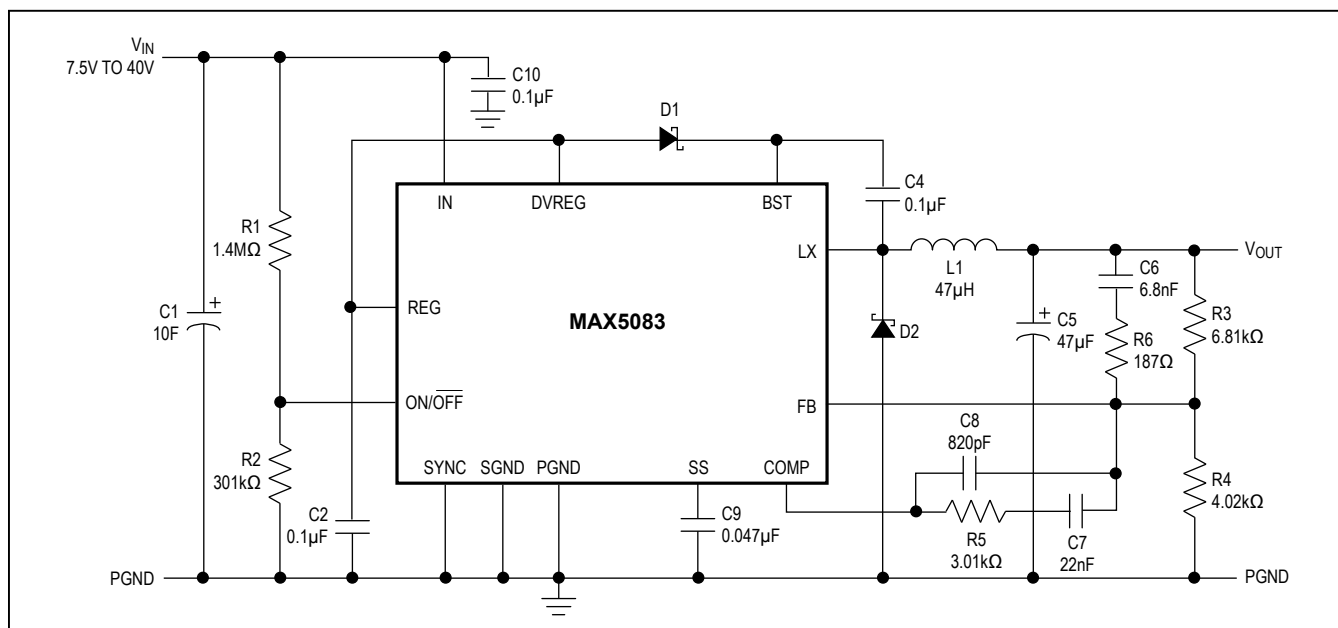
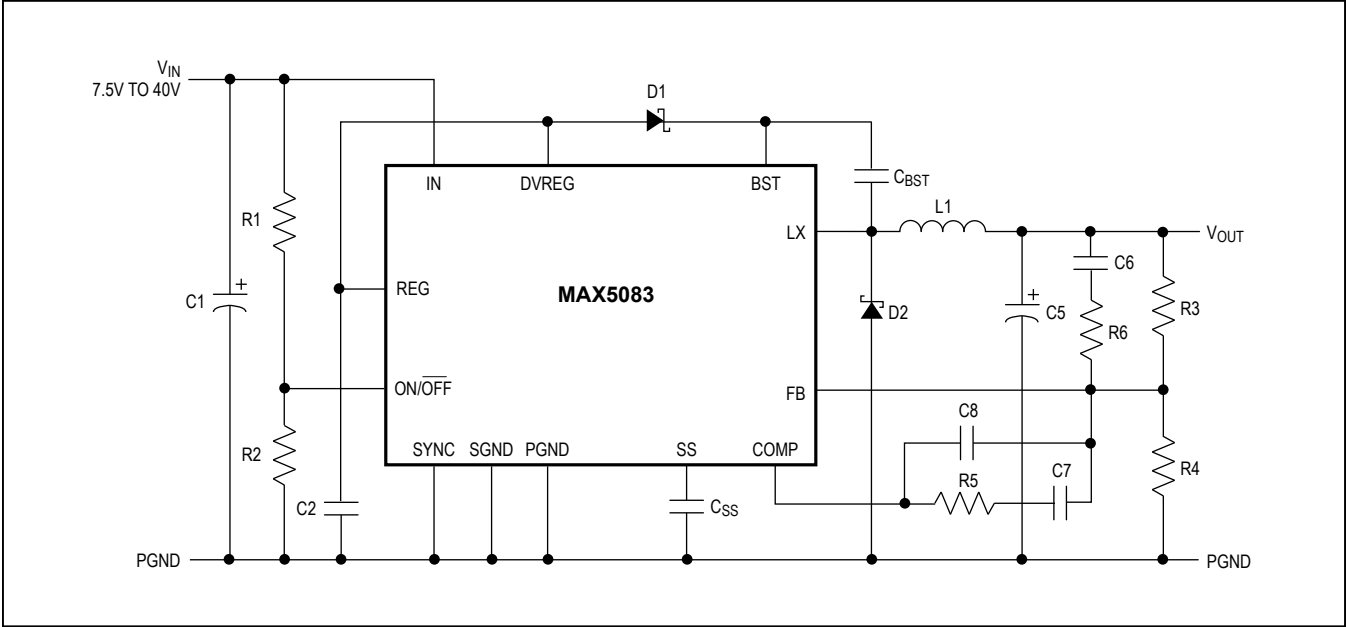
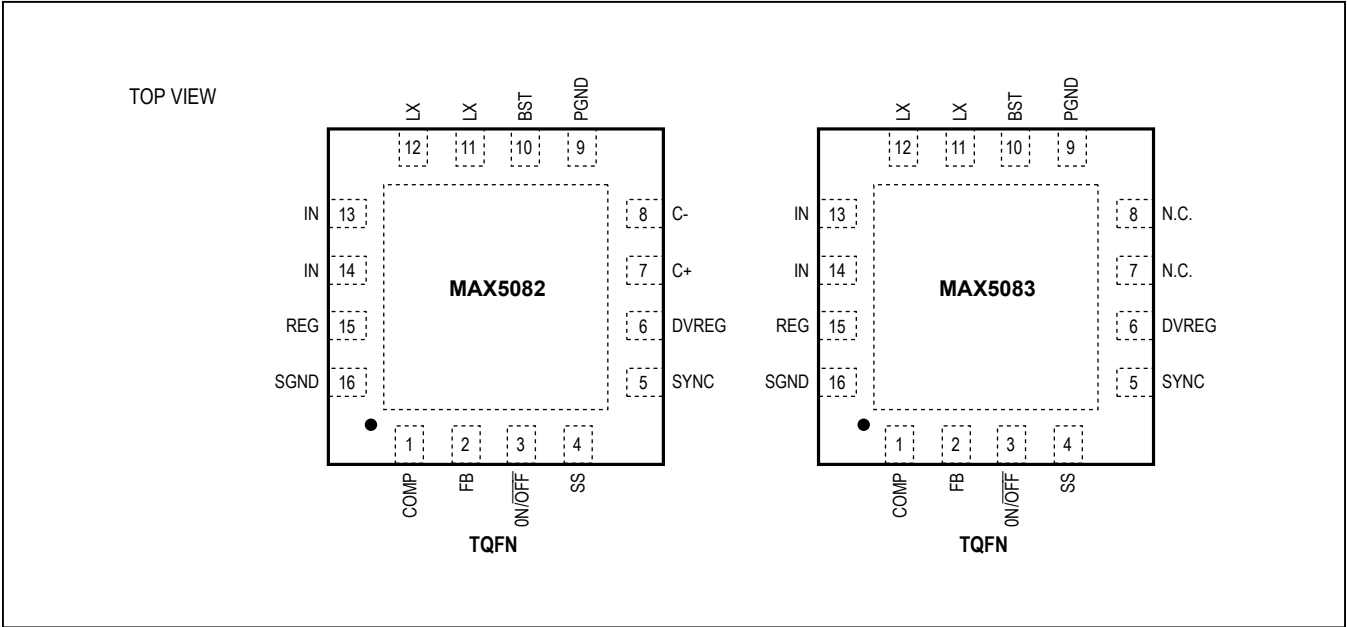


Figure 6. MAX5083 Typical Application Circuit

Typical Operating Circuits (continued)



Pin Configurations



Package Information

For the latest package outline information and land patterns (footprints), go to www.maximintegrated.com/packages. Note that a "+", "#", or "-" in the package code indicates RoHS status only. Package drawings may show a different suffix character, but the drawing pertains to the package regardless of RoHS status.

PACKAGE TYPE	PACKAGE CODE	DOCUMENT NO.	LAND PATTERN NO.
16 TQFN-EP	T1655+3	21-0140	Refer to Application Note 1891

REVISION HISTORY

REVISION NUMBER	REVISION DATE	DESCRIPTION	PAGES CHANGED
0	5/05	Initial release	—
1	5/14	Removed automotive reference under <i>Applications</i> section on Page 1	1

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