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Maxim Integrated
MAX5951ETJ+

For any questions, you can email us directly:
sales@integrated-circuit.com

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Datasheet of MAX5951ETJ+ - IC REG CTRLR BUCK PWM VM 32-TQFN
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19-0546; Rev 0; 5/06

12V/5V Input Buck PWM Controller

General Description

The MAX5951 is a 12V pulse-width modulated (PWM), step-down, DC-DC controller. The device operates over the input-voltage range of 8V to 16V or 5V ±10%, and provides an adjustable output from 0.8V to 5.5V. The device delivers up to 10A of load current with excellent load-and-line regulation.

The MAX5951 PWM section utilizes a voltage-mode control scheme for good noise immunity and offers external compensation, allowing for maximum flexibility with a wide selection of inductor values and capacitor types. The device operates at a fixed switching frequency that is programmable from 100kHz to 1MHz and can be synchronized to an external clock signal though the SYNCIN input. The device includes undervoltage lockout (UVLO) and digital soft-start. Protection features include lossless valley-mode current limit, hiccup-mode output short-circuit protection, and thermal shutdown.

The MAX5951 is available in a space-saving 5mm x 5mm, 32-pin thin QFN package and is specified for operation over the -40°C to +85°C extended temperature range. Refer to the MAX5950 data sheet for a pin-compatible PWM controller with hot swap.

Applications

PCI-e Express Modules™

General 12V or 5V Input PWM Controllers

Blade Servers

RAID

Base Stations

Workstations

PCI-e Express Modules is a trademark of PCI-SIG.

Features

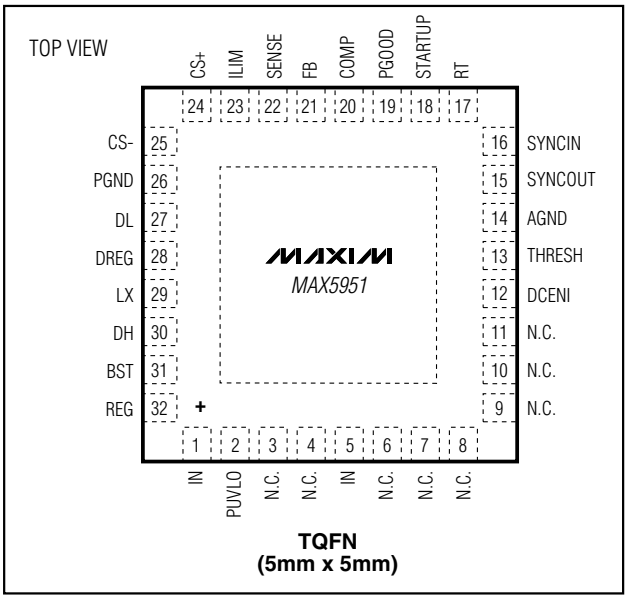
- ♦ **8V to 16V or 5V ±10% Input-Voltage Range**
- ♦ **Lossless Valley-Mode Current Sensing**
- ♦ **Output Voltage Adjustable from 0.8V to 5.5V**
- ♦ **Voltage-Mode Control**
- ♦ **External Compensation for Maximum Flexibility**
- ♦ **Digital Soft-Start**
- ♦ **Sequencing or Ratiometric Tracking**
- ♦ **Startup Synchronization**
- ♦ **Programmable PGOOD Output**
- ♦ **Programmable Switching Frequency from 100kHz to 1MHz**
- ♦ **External Frequency Synchronization**
- ♦ **SYNCIN and SYNCOUT Enable 180° Out-of-Phase Operation**
- ♦ **Thermal Shutdown and Short-Circuit Protection**
- ♦ **Space-Saving, 5mm x 5mm, 32-Pin TQFN Package**

Ordering Information

PART	TEMP RANGE	PIN-PACKAGE	PKG CODE
MAX5951ETJ+	-40°C to +85°C	32 TQFN	T3255-4

+Denotes lead-free package.

Pin Configuration



TQFN (5mm x 5mm)

For pricing, delivery, and ordering information, please contact Maxim/Dallas Direct! at 1-888-629-4642, or visit Maxim's website at www.maxim-ic.com.

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ABSOLUTE MAXIMUM RATINGS

IN to AGND .....-0.3V to +24V
BST to AGND .....-0.3V to +30V
BST to LX .....-0.3V to +6V
CS- to AGND .....-0.3V to (V_{IN} + 0.3V)
REG, DREG, PUVLO, DCENI, SYNCIN, THRESH, SENSE to AGND .....-0.3V to +6V
RT, ILIM, STARTUP, PGOOD, FB, CS+ to AGND ....-0.3V to +6V
SYNCOUT, COMP to AGND .....-0.3V to (V_{REG} + 0.3V)
DL to PGND .....-0.3V to (V_{DREG} + 6V)
DH to LX .....-0.3V to (V_{BST} + 0.3V)
PGND to AGND .....-0.3V to +0.3V
Input Current (any pin) .....±50mA

Continuous Power Dissipation (T_A = +70°C)
32-Pin TQFN (derate 34.5 mW/°C above +70°C) .2758.6mW
32-Pin TQFN (θ_{JA}) .....+29°C/W
32-Pin TQFN (θ_{JC}) .....+2.1°C/W
Operating Ambient Temperature Range .....-40°C to +85°C
Maximum Junction Temperature .....+150°C
Storage Temperature Range .....-60°C to +150°C
Lead Temperature (soldering, 10s) .....+300°C

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} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 100kΩ, R_{ILIM} = 60kΩ, C_{REG} = 2.2µF, T_A = T_J = -40°C to +85°C, unless otherwise noted. Typical values are at T_A = T_J = +25°C.) (Note 1)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
Input-Voltage Range	V_{IN}		8		16	V
		V_{IN} = V_{REG} = V_{DREG} (Note 2)	4.5		5.5	
Standby Supply Current		V_{IN} = 16V, V_{PUVLO} = 0V		0.3	0.5	mA
Quiescent Supply Current		V_{IN} = 16V, V_{FB} = 0.9V		1.6	2.6	mA
Switching Supply Current		V_{IN} = 16V, V_{FB} = 0V		5.0	8.0	mA
PWM UVLO						
Default PWM Undervoltage Lockout Threshold		V_{IN} rising	6.7		7.3	V
PWM Undervoltage Lockout Hysteresis				0.7		V
PUVLO Threshold	V_{PUVLO}	V_{PUVLO} rising	1.202	1.220	1.238	V
PUVLO Hysteresis				122		mV
PUVLO Input Impedance			180	310	500	kΩ
PWM DCENI CONTROL						
DCENI Comparator Input Common-Mode Range			0		3	V
DCENI Comparator Offset		V_{DCENI} - V_{THRESH}	-10		+10	mV
DCENI Comparator Hysteresis				100		mV
DCENI Input Current			-1		+1	µA
THRESH Operating Voltage Range			0.6		2.5	V
THRESH Input Current		V_{THRESH} > 0.6V	-1.5		+1	µA
		V_{THRESH} < 0.3V	-5		+1	
Default DCENI Threshold		V_{THRESH} < 0.3V	1.202	1.22	1.238	V

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ELECTRICAL CHARACTERISTICS (continued)

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PARAMETER	SYMBOL	CONDITIONS		MIN	TYP	MAX	UNITS
PWM PGOOD OUTPUT							
SENSE Threshold		V_{SENSE} rising		788	800	812	mV
SENSE Hysteresis					100		mV
SENSE Input Bias Current				-1		+1	µA
PGOOD Internal Pullup Current					10		µA
PGOOD Output Voltage Low		I_{PGOOD} = -2.4mA				50	mV
INTERNAL VOLTAGE REGULATOR							
Output Voltage Set Point	V_{REG}			4.7		5.3	V
Line Regulation		V_{IN} = 8V to 16V				1	mV/V
Load Regulation		I_{REG} = 0 to 50mA				150	mV
PWM OSCILLATOR							
Oscillator Frequency Range	f_{SW}	V_{SYNCIN} = 0V, f_{SW} = 5 x 10^{10} / R_{RT} Hz		100		1000	kHz
Oscillator Accuracy		T_A = T_J = +25°C	f_{SW} < 500kHz	-2.5		+2.5	%
			f_{SW} > 500kHz	-4		+4	
		T_A = T_J = -40°C to +85°C	f_{SW} < 500kHz	-3.5		+3.5	%
			f_{SW} > 500kHz	-5		+5	
RT Voltage	V_{RT}	50kΩ < R_{RT} < 500kΩ			2		V
Maximum Duty Cycle		V_{SYNCIN} = 0V, V_{IN} = 12V		82	88		%
SYNCIN High-Level Voltage				2.1			V
SYNCIN Low-Level Voltage						0.8	V
SYNCIN Pulldown Resistor				50	100	150	kΩ
SYNCIN Rising to SYNCOUT Falling Delay					10		ns
SYNCIN Falling to SYNCOUT Rising Delay					30		ns
Maximum SYNCIN Frequency				1			MHz
SYNCOUT Voltage High	$VH_{SYNCOUT}$	$I_{SYNCOUT}$ = +1.2mA		V_{REG} - 0.1			V
SYNCOUT Voltage Low	$VL_{SYNCOUT}$	$I_{SYNCOUT}$ = -2.4mA				50	mV
PWM ERROR AMPLIFIER							
FB Input Range				0		V_{REF}	V
FB Input Current				-250		+250	nA
COMP Output Voltage Range		I_{COMP} = -500µA to +500µA		0.25		V_{REG} - 0.5	V
Open-Loop Gain					80		dB
Unity-Gain Bandwidth	f_{GBW}				2.5		MHz
Reference Voltage	V_{REF}	I_{COMP} = -500µA to +500µA		792	800	808	mV

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ELECTRICAL CHARACTERISTICS (continued)

(V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 100kΩ, R_{ILIM} = 60kΩ, C_{REG} = 2.2µF, T_A = T_J = -40°C to +85°C, unless otherwise noted. Typical values are at T_A = T_J = +25°C.) (Note 1)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
PWM COMPARATOR						
Comparator Offset Voltage				0.3		V
Comparator Propagation Delay				40		ns
PWM DIGITAL SOFT-START						
Soft-Start Duration				1024		Clocks
Reference Voltage Steps				128		Steps
				6.3		mV
PWM RAMP GENERATOR						
Ramp Amplitude				1.8		V
PWM CURRENT-LIMIT COMPARATOR AND HICCUP MODE						
Cycle-by-Cycle Valley Current-Limit Threshold Adjustment Range		Limit = V_{ILIM} / 10	50		350	mV
Cycle-by-Cycle Valley Current-Limit Threshold Tolerance		V_{ILIM} = 0.5V	44.5		55.5	mV
		V_{ILIM} = 3.5V	330		366	
ILIM Reference Current		V_{ILIM} = 0 to 3.5V, T_A = T_J = +25°C	19	20	21	µA
ILIM Reference Current Tempco				3333		ppm/°C
CS+, CS- Input Bias Current		V_{CS+} = 0V, V_{CS-} = -0.3V, current out of the CS_	-1		+20	µA
PWM HICCUP MODE						
Number of Cumulative Current-Limit Events to Hiccup	N_{CL}			8		Clocks
Number of Consecutive Noncurrent-Limit Cycles to Clear N_{CL}	N_{CLR}			3		Clocks
Hiccup Timeout	N_{HT}			512		Clocks
PWM STARTUP INPUT						
STARTUP Threshold	V_{SUT}		1.1		1.9	V
STARTUP Threshold Hysteresis				250		mV
Internal Pullup Current	I_{START}			10		µA
STARTUP Output Voltage Low		$I_{STARTUP}$ = -2.4mA			0.1	V
PWM DH DRIVER						
Peak Source Current		$V_{DH,LX}$ = 0V, pulse width < 100ns, $V_{BST,LX}$ = 5V		2		A
Peak Sink Current		$V_{DH,LX}$ = 5V, pulse width < 100ns, $V_{BST,LX}$ = 5V		2		A
DH Resistance Sourcing		I_{DH} = 50mA, $V_{BST,LX}$ = 5V		1	3	Ω
DH Resistance Sinking		I_{DH} = -50mA, $V_{BST,LX}$ = 5V		1	3	Ω

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ELECTRICAL CHARACTERISTICS (continued)

(V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 100kΩ, R_{ILIM} = 60kΩ, C_{REG} = 2.2µF, T_A = T_J = -40°C to +85°C, unless otherwise noted. Typical values are at T_A = T_J = +25°C.) (Note 1)

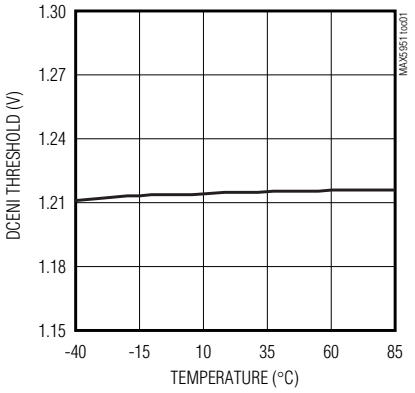
PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
PWM DL DRIVER						
Peak Source Current		V_{DL} = 0V, pulse width < 100ns, V_{DREG} = 5V		2		A
Peak Sink Current		V_{DL} = 5V, pulse width < 100ns, V_{DREG} = 5V		2		A
DL Resistance Sourcing		I_{DL} = 50mA, V_{DREG} = 5V		1	3	Ω
DL Resistance Sinking		I_{DL} = -50mA, V_{DREG} = 5V		1	3	Ω
Break-Before-Make Time				25		ns
THERMAL SHUTDOWN						
Thermal Shutdown Temperature		T_J rising		135		°C
Thermal Shutdown Hysteresis				15		°C

Note 1: Limits at -40°C are guaranteed by design and are not production tested.
Note 2: For 5V applications, connect REG directly to IN.

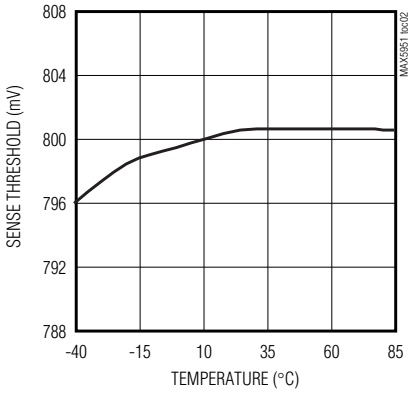
Typical Operating Characteristics

(*Typical Operating Circuits*. V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 49.9kΩ, R_{ILIM} = 48.7kΩ, C_{REG} = 2.2µF, T_A = +25°C.)

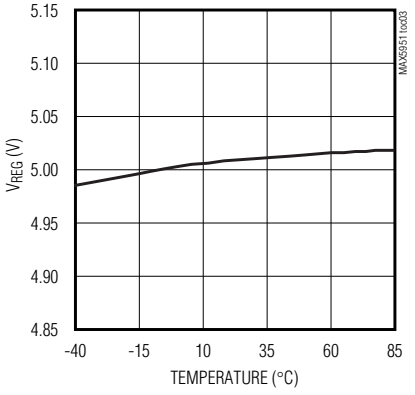
DEFAULT DCENI THRESHOLD vs. TEMPERATURE



SENSE THRESHOLD vs. TEMPERATURE



REG OUTPUT VOLTAGE vs. TEMPERATURE



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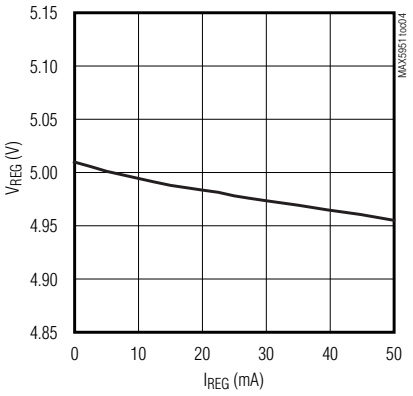
12V/5V Input Buck PWM Controller

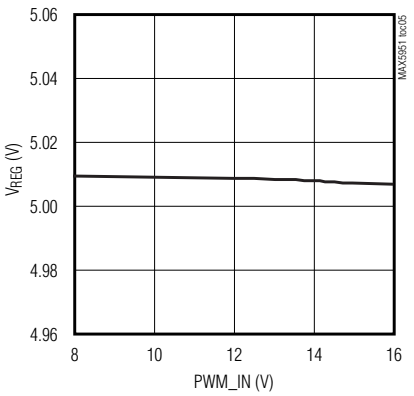
Typical Operating Characteristics (continued)

(*Typical Operating Circuits.* V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 49.9kΩ, R_{ILIM} = 48.7kΩ, C_{REG} = 2.2µF, T_A = +25°C.)

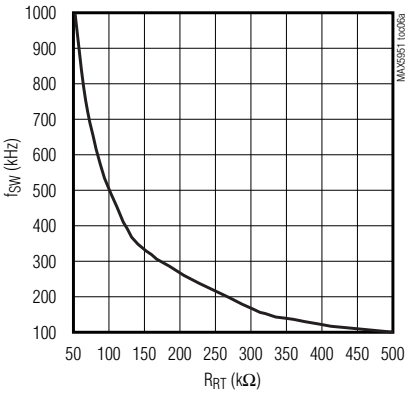
REG OUTPUT VOLTAGE vs. LOAD CURRENT



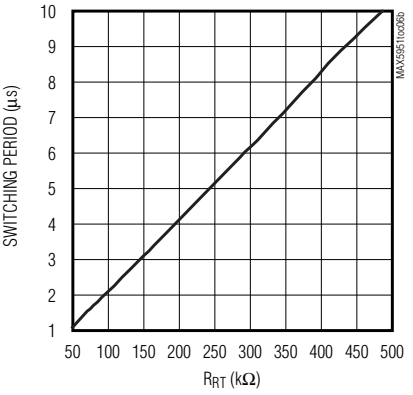
REG OUTPUT VOLTAGE vs. INPUT VOLTAGE



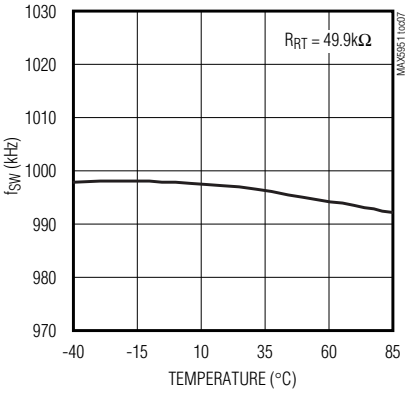
SWITCHING FREQUENCY vs. RRT



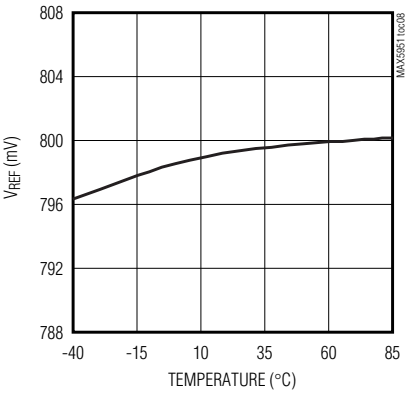
SWITCHING PERIOD vs. RRT



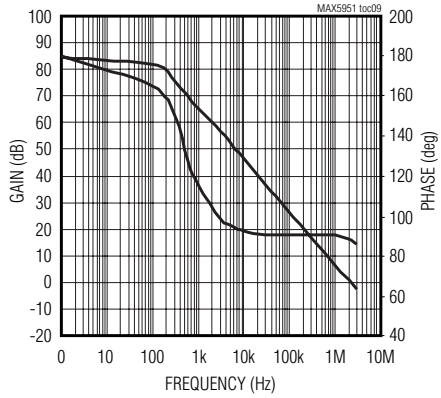
SWITCHING FREQUENCY vs. TEMPERATURE



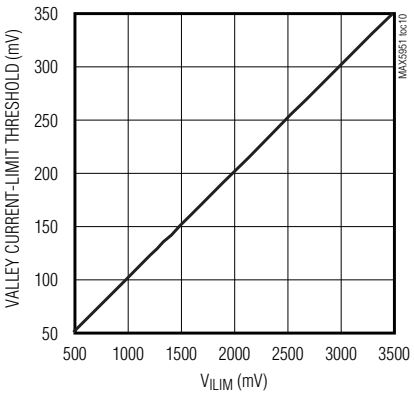
FB REFERENCE VOLTAGE vs. TEMPERATURE



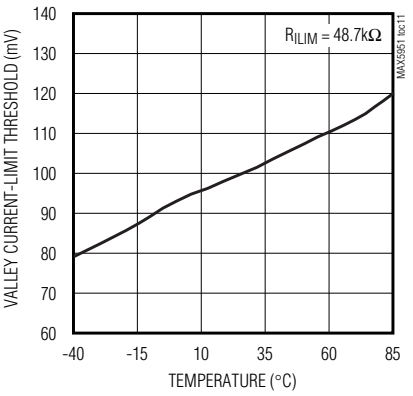
OPEN-LOOP GAIN/PHASE vs. FREQUENCY



VALLEY CURRENT-LIMIT THRESHOLD vs. VILIM



VALLEY CURRENT-LIMIT THRESHOLD vs. TEMPERATURE



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Typical Operating Characteristics (continued)

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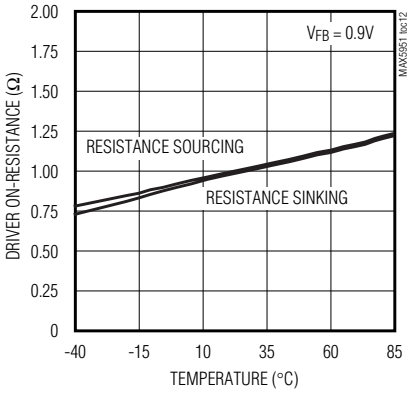
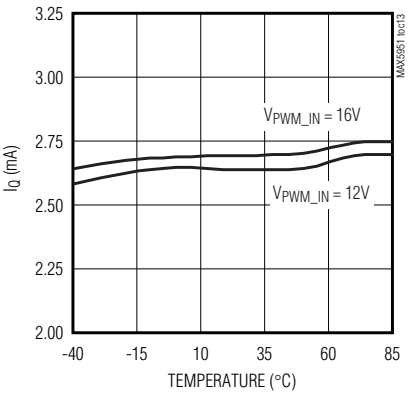
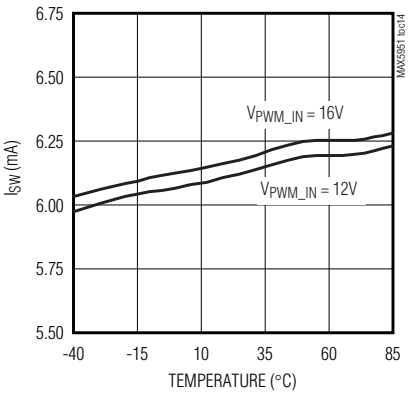
DL, DH DRIVER ON-RESISTANCE vs. TEMPERATURE



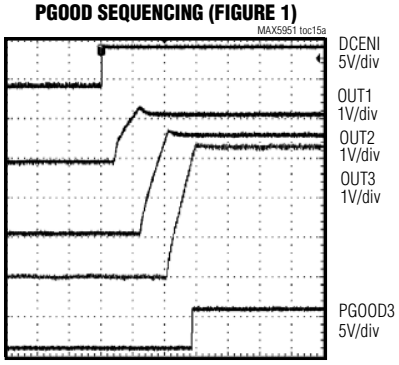
QUIESCENT SUPPLY CURRENT vs. TEMPERATURE



SWITCHING SUPPLY CURRENT vs. TEMPERATURE

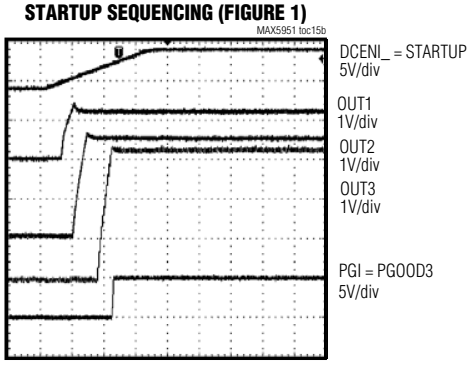


PGOOD SEQUENCING (FIGURE 1)



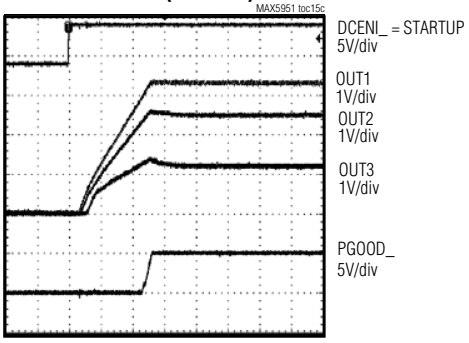
PWREN SWITCHING FROM HIGH TO LOW

STARTUP SEQUENCING (FIGURE 1)



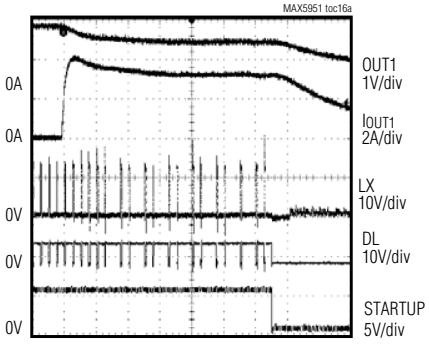
PWREN SWITCHING FROM HIGH TO LOW

TRACKING (FIGURE 1)



PWREN SWITCHING FROM HIGH TO LOW

OVERLOAD RESPONSE



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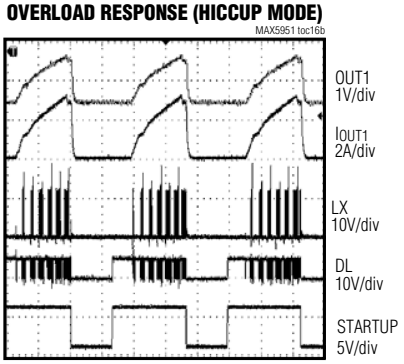
12V/5V Input Buck PWM Controller

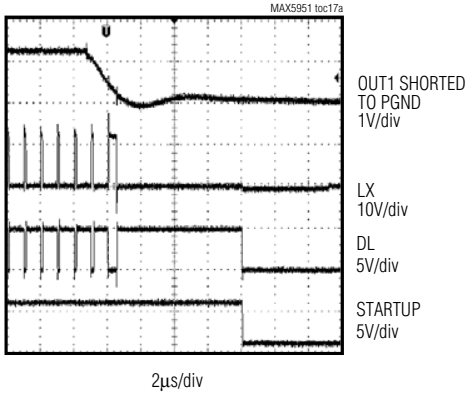
Typical Operating Characteristics (continued)

(*Typical Operating Circuits*. V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 49.9kΩ, R_{ILIM} = 48.7kΩ, C_{REG} = 2.2µF, T_A = +25°C.)

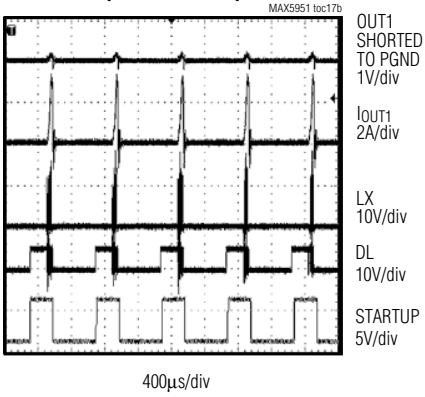
OVERLOAD RESPONSE (HICCUP MODE)



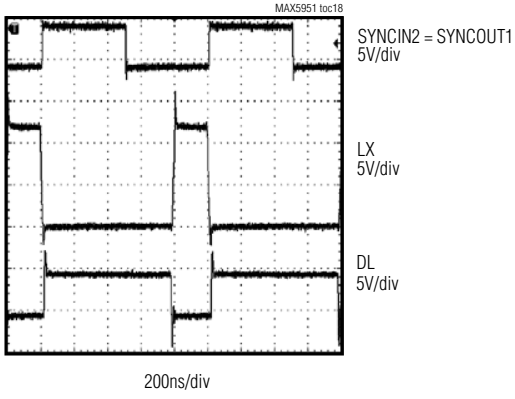
SHORT-CIRCUIT RESPONSE



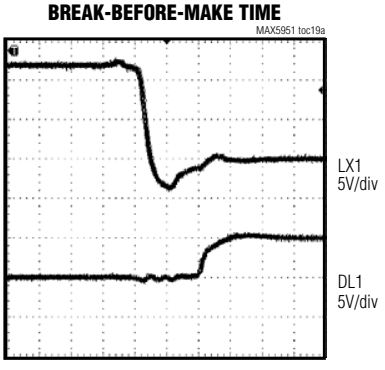
SHORT-CIRCUIT RESPONSE (HICCUP MODE)



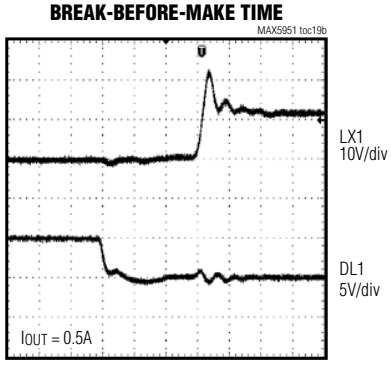
SYNCHRONIZATION



BREAK-BEFORE-MAKE TIME



BREAK-BEFORE-MAKE TIME



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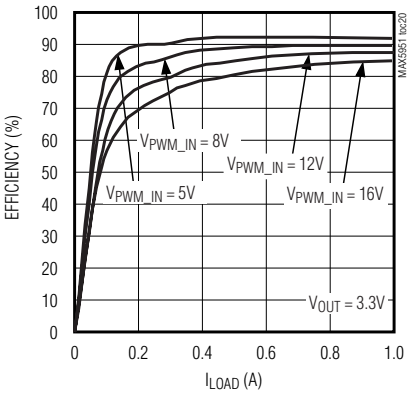
12V/5V Input Buck PWM Controller

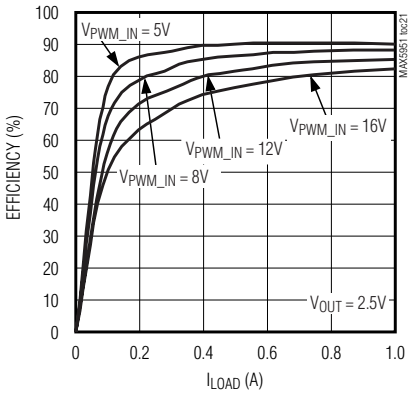
Typical Operating Characteristics (continued)

(*Typical Operating Circuits*. V_{IN} = 12V or V_{IN} = V_{REG} = 5V, V_{DREG} = V_{REG}, V_{PGND} = 0V, V_{SYNCIN} = 0V, R_{RT} = 49.9kΩ, R_{ILIM} = 48.7kΩ, C_{REG} = 2.2µF, T_A = +25°C.)

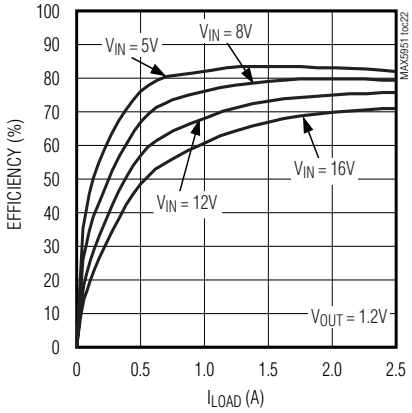
EFFICIENCY vs. LOAD CURRENT



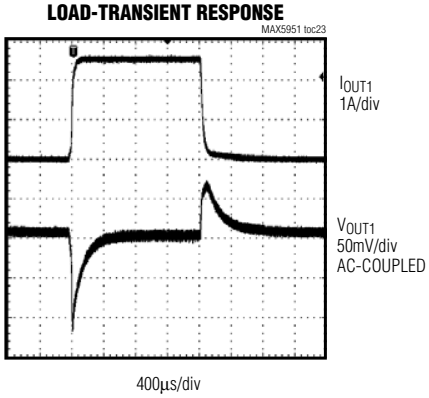
EFFICIENCY vs. LOAD CURRENT



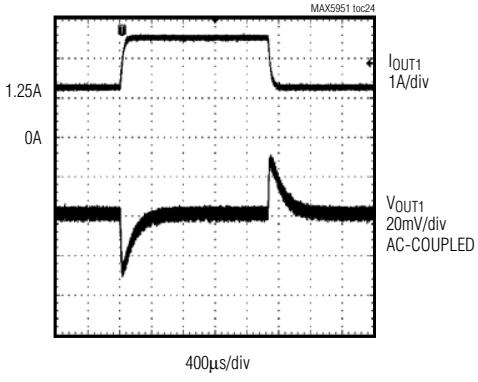
EFFICIENCY vs. LOAD CURRENT



LOAD-TRANSIENT RESPONSE



LOAD-TRANSIENT RESPONSE



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12V/5V Input Buck PWM Controller

Pin Description

PIN	NAME	FUNCTION
1, 5	IN	Supply Input Connection. Connect to an external voltage source from 8V to 16V. For 5V input application, connect IN = REG to a 5V ±10% source. Connect an external divider from IN to PUVLO to AGND to lower the startup voltage.
2	PUVLO	PWM UVLO Divider Center Point. Use an external divider to override the internal PWM UVLO divider. The rising threshold is set to 1.220V with 122mV hysteresis. Leave PUVLO unconnected for the default PWM UVLO.
3, 4, 6–11	N.C.	No Connection. Do not connect.
12	DCENI	DC-DC Enable Input. DCENI must be above V_{THRESH} for the PWM controller to start. Connect to REG if not used.
13	THRESH	DC-DC Enable Input Threshold Set. Connect a resistive divider from REG to THRESH to AGND to set the DCENI threshold. Connect to ground for a default threshold of 1.220V.
14	AGND	Analog Ground Connection. Solder the exposed pad to a large AGND plane. Connect AGND and PGND together at one point near the input bypass capacitor return terminal.
15	SYNCOUT	Synchronization Output. SYNCOUT is a synchronization signal to drive the SYNCIN of a second MAX5950 or MAX5951, if used. Leave SYNCOUT unconnected when not used.
16	SYNCIN	Synchronization Input. SYNCIN accepts the SYNCOUT from another MAX5950 or MAX5951 and shifts switching by 180°, allowing the reduction of the input bypass capacitors. When used, drive with a frequency at least 20% higher than the frequency programmed through the RT pin. If phase staggering is desired, use 50% duty cycle. Connect SYNCIN to AGND when not used.
17	RT	Oscillator Timing Resistor Connection. Connect a 500kΩ to 50kΩ resistor from RT to AGND to program the switching frequency from 100kHz to 1MHz.
18	STARTUP	Startup Input. STARTUP coordinates simultaneous soft-start for multiple converters. See the *Tracking (STARTUP)* section.
19	PGOOD	Power-Good Output. PGOOD output goes high when SENSE is above V_{REF} and STARTUP is high.
20	COMP	Error-Amplifier Output. Connect COMP to the compensation feedback network.
21	FB	Feedback Regulation Point. Connect to the center tap of a resistive divider from the converter output to AGND to set the output voltage. The FB voltage regulates to the reference voltage.
22	SENSE	Output Voltage Sense. Connect a resistive divider from the converter output to SENSE to AGND to monitor the programmed output voltage. SENSE is compared to the internal reference, V_{REF}.
23	ILIM	Valley Current-Limit Set Output. Connect a 25kΩ to 175kΩ resistor, R_{ILIM}, from ILIM to AGND to program the valley current-limit threshold from 50mV to 350mV. ILIM sources 20µA out to R_{ILIM}. The resulting voltage divided by 10 is the valley current limit. Alternatively, a resistive divider from REG to ILIM to AGND can be used to set the valley current limit.

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12V/5V Input Buck PWM Controller

Pin Description (continued)

PIN	NAME	FUNCTION
24	CS+	Positive Current-Sense Input. Connect CS+ to the synchronous MOSFET source (connected to PGND).
25	CS-	Negative Current-Sense Input. Connect CS- to the synchronous MOSFET drain (connected to LX).
26	PGND	Power-Ground Connection. Connect the input filter capacitor's negative terminal, the source of the synchronous MOSFET, and the output filter capacitor's return to PGND. Connect externally to AGND at a single point near the input capacitor return terminal.
27	DL	Low-Side Gate-Driver Output. DL is the gate-driver output for the synchronous MOSFET.
28	DREG	Gate-Drive Supply for the Low-Side MOSFET Driver. Connect externally to REG and the anode of the boost diode.
29	LX	Source Connection of the High-Side MOSFET and Drain Connection of the Synchronous MOSFET. Connect the inductor and the negative side of the boost capacitor to LX.
30	DH	High-Side Gate-Driver Output. DH drives the gate of the high-side MOSFET.
31	BST	High-Side Gate-Driver Supply. Connect BST to the cathode of the boost diode and to the positive terminal of the boost capacitor.
32	REG	5V Regulator Output. Bypass with a 2.2µF ceramic capacitor to AGND.
EP	EP	Exposed Pad. Connect the exposed pad to AGND.

Detailed Description

The MAX5951 is a PWM, step-down, DC-DC controller. The device operates over the input-voltage range of 8V to 16V or 5V ±10% (V_{IN} = V_{REG}) and provides an adjustable output from 0.8V to 5.5V. The device delivers up to 10A of load current with excellent load-and-line regulation.

The MAX5951 PWM controller utilizes a voltage-mode control scheme for good noise immunity and offers external compensation allowing for maximum flexibility with a wide selection of inductor values and capacitor types. The device operates at a fixed switching frequency that is programmable from 100kHz to 1MHz and can be synchronized to an external clock signal through the SYNC input. The device includes UVLO and digital soft-start. Protection features include valley-mode current limit, hiccup-mode output short-circuit protection, and thermal shutdown.

PWM Controller

PWM UVLO

V_{IN} must exceed the default PWM UVLO threshold (7V typ) before any PWM operation can commence. The UVLO circuitry keeps the MOSFET drivers, oscillator, and all the internal circuitry shut down to reduce current consumption.

Override the internal PWM UVLO divider by connecting an external resistive divider from IN to PUVLO to AGND. The PUVLO threshold is 1.220V with 122mV hysteresis.

Digital Soft-Start

The MAX5951 soft-start feature allows the load voltage to ramp up in a controlled manner, eliminating output voltage overshoot. Soft-start begins after V_{IN} exceeds the UVLO threshold. The soft-start circuitry gradually ramps up the reference voltage. This controls the rate of rise of the output voltage and reduces input surge currents during startup. The soft-start duration is 1024 clock cycles. The output voltage is incremented through 128 equal steps. The output reaches regulation when soft-start is completed, regardless of output capacitance and load.

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12V/5V Input Buck PWM Controller

Internal Linear Regulator (REG)

REG is the output terminal of a 5V LDO, which is powered from IN and provides power to the IC. Bypass REG to AGND with a 2.2µF ceramic capacitor. Place the capacitor physically close to the MAX5951 to provide good bypassing. REG is intended for powering only the internal circuitry and should not be used to supply power to external loads.

Low-Side MOSFET Driver Supply (DREG)

DREG is the supply input for the low-side MOSFET driver. Connect DREG to REG externally. Adding an RC filter (5Ω resistor and 2.2µF ceramic capacitor) from REG to DREG filters out the high peak currents of the MOSFET drivers.

High-Side MOSFET Driver Supply (BST)

BST supplies the power for the high-side MOSFET drivers. Connect the bootstrap diode from BST to DREG (anode at DREG and cathode at BST). Connect a bootstrap 1µF ceramic capacitor between BST and LX.

MOSFET Gate Drivers (DH, DL)

The high-side (DH) and low-side (DL) drivers drive the gates of the external n-channel MOSFETs. The drivers' 2A peak source-and-sink current capability provides ample drive to ensure fast rise and fall times of the switching MOSFETs. Short rise and fall times minimize switching losses. For low output-voltage applications where the duty cycle is less than 50%, choose a high-side MOSFET (Q2) with a moderate $R_{DS(ON)}$. Choose a low-side MOSFET (Q1) with a very low $R_{DS(ON)}$.

The gate driver circuitry also provides a break-before-make time (25ns, typ) to prevent shoot-through currents during transition.

Oscillator/Synchronization Input (SYNCIN)/ Synchronization Output (SYNCOUT)

Use an external resistor at RT to program the MAX5951 switching frequency from 100kHz to 1MHz.

Choose the appropriate resistor at RT to calculate the desired output switching frequency (f_{SW}):

$$f_{SW}\ (Hz) = (5 \times 10^{10}) / R_{RT}\ (\Omega)$$

Connect an external clock (SYNCOUT from another MAX5950 or MAX5951) at SYNCIN for external clock synchronization. For proper synchronization, the external frequency must be at least 20% higher than the frequency programmed through the RT input. If SYNCIN is 50% duty cycle, SYNCOUT is shifted by 180°, allowing the reduction of the DC-DC converter input bypass capacitor.

SYNCOUT is a synchronization signal that is used to drive the SYNCIN of a second MAX5950 or MAX5951.

Tracking (STARTUP)

The STARTUP input in conjunction with digital soft-start provides simple ratiometric tracking. When using multiple MAX5950s or MAX5951s, in addition to connecting SYNCIN and SYNCOUT signals appropriately, connect the STARTUP of all the devices together. STARTUP synchonizes the soft-start of all the devices' references, and hence their respective output voltages track ratiometrically. See Figure 1 and the *Typical Operating Circuits*.

The STARTUP input has an internal 10µA pullup current, but can be driven by external logic. When using multiple converters, connect the STARTUP of all the devices together.

Startup Sequencing (DCENI, THRESH)

The DCENI input must be above V_{THRESH} for the PWM controller to start. By connecting the DCENI inputs of multiple devices together and having different start thresholds ($V_{THRESH_}$), the startup of the PWM controllers can be staggered to provide power sequencing. Connect a resistive divider from REG to THRESH to AGND to set the start thresholds of each device. Connect THRESH to AGND to produce a default 1.220V threshold for DCENI. See Figure 1 and the *Typical Operating Circuits.*

Power-Good Sequencing (PGOOD, SENSE)

The PGOOD outputs and DCENI inputs can be daisy-chained to generate power sequencing. The PGOOD output is pulled high when the voltage at SENSE is above V_{REF} (800mV, typ). Connect a resistive divider from the power-supply output voltage to SENSE to AGND to set the power-good threshold. See Figure 1 and the *Typical Operating Circuits.*

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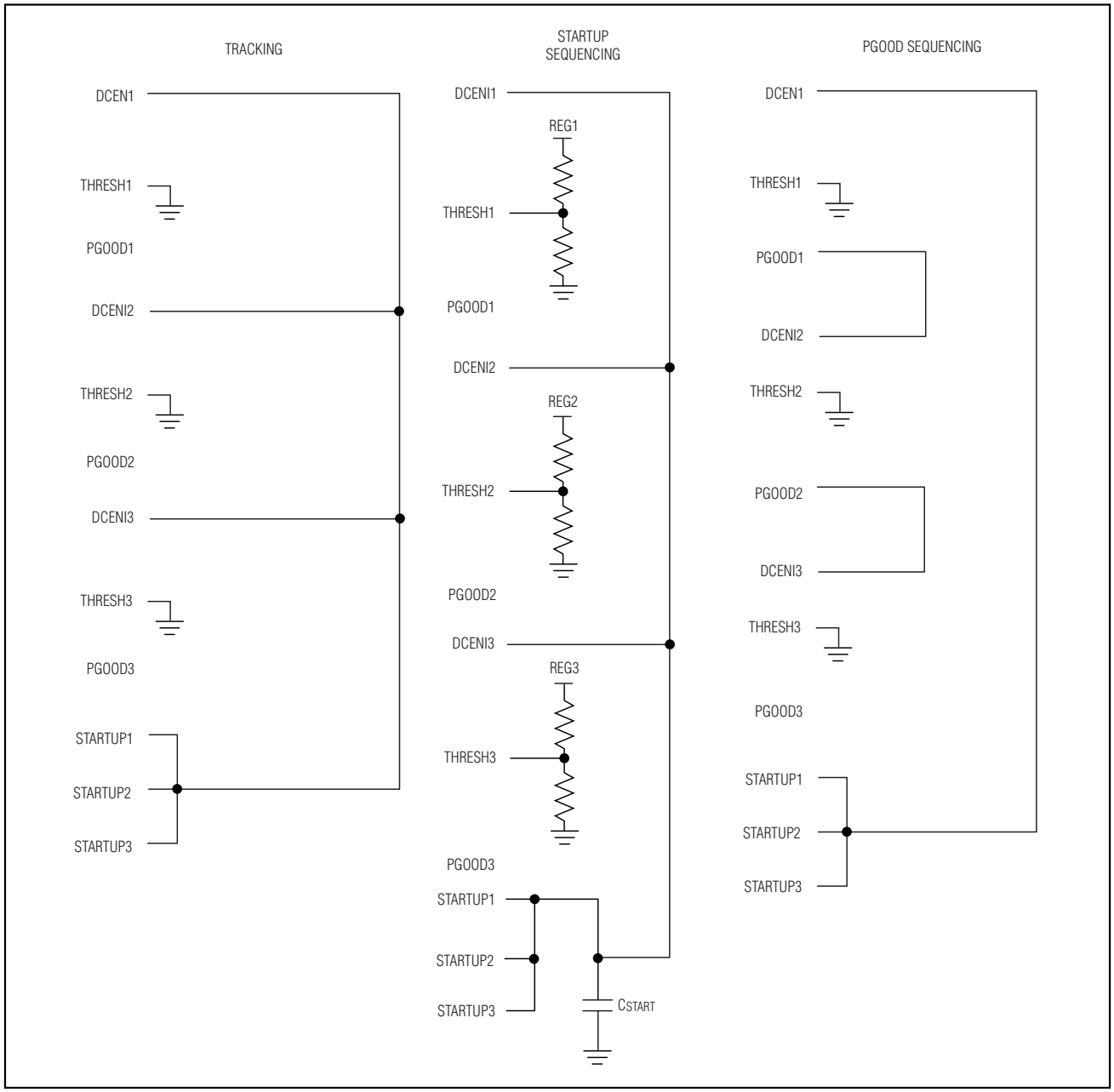

Figure 1. Tracking, STARTUP Sequencing, and PGOOD Sequencing Configurations

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12V/5V Input Buck PWM Controller

Error Amplifier

The output of the internal error amplifier (COMP) is available for frequency compensation (see the *Compensation Design Guidelines* section). The inverting input is FB; the output is COMP. The error amplifier has an 80dB open-loop gain and a 2.5MHz GBW product. See the *Typical Operating Characteristics* for the Open-Loop Gain and Phase vs. Frequency graph.

PWM Comparator

An internal ramp is compared against the output of the error amplifier to generate the PWM signal. The amplitude of the ramp, V_{RAMP}, is 1.8V.

Output Short-Circuit Protection (Hiccup Mode)

The current-limit circuit employs a lossless valley current-limiting algorithm that uses the MOSFET's on-resistance as the current-sensing element. Once the high-side MOSFET turns off, the voltage across the low-side MOSFET is monitored. If the voltage across the low-side MOSFET ($R_{DS(ON)}$ x $I_{INDUCTOR}$) does not exceed the current-limit threshold, the high-side MOSFET turns on normally at the start of the next cycle. If the voltage across the low-side MOSFET exceeds the current-limit threshold just before the beginning of a new PWM cycle, the controller skips that cycle. During severe overload or short-circuit conditions, the switching frequency of the device appears to decrease because the on-time of the low-side MOSFET extends beyond a clock cycle.

If the current-limit threshold is exceeded for eight cumulative clock cycles (N_{CL}), the device shuts down (both DH and DL are pulled low) for 512 clock cycles (hiccup timeout) and restarts with a soft-start sequence. If three consecutive cycles pass without a current-limit event, the count of N_{CL} is cleared (see Figure 2). Hiccup mode protects against continuous output short circuit.

Thermal-Overload Protection

The MAX5951 features an integrated thermal-overload protection with temperature hysteresis. 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 +135°C, an internal thermal sensor shuts down the device, turning off the power MOSFETs and allowing the die to cool. After the die temperature falls by +15°C, the part restarts with a soft-start sequence.

PWM Controller Design Procedures

Setting the Undervoltage Lockout

Connect an external resistive divider from IN to PUVLO to AGND to override the internal PWM UVLO divider. The rising threshold at PUVLO is set to 1.220V with 120mV hysteresis. First select the PUVLO to AGND resistor (R2), then calculate the resistor from IN to PUVLO (R1) using the following equation:

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

where V_{IN} is the input voltage at which the converter needs to turn on, V_{PUVLO} = 1.220V, and R2 is chosen to be less than 20kΩ. See Figure 3.

Leave PUVLO unconnected for the default PWM UVLO threshold. In this case, an internal voltage-divider monitors the supply voltage at IN and allows startup when IN rises above 7V (typ).

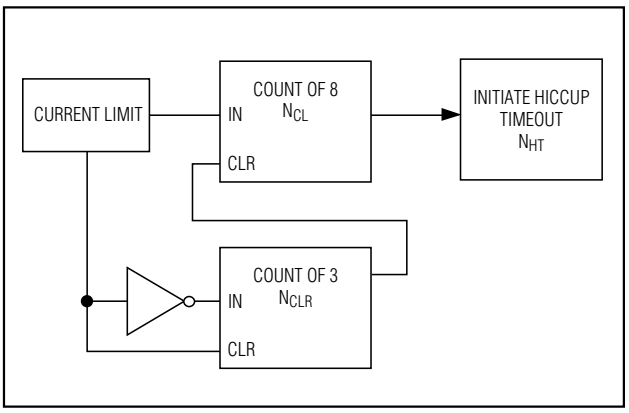

Figure 2. Hiccup Block Diagram

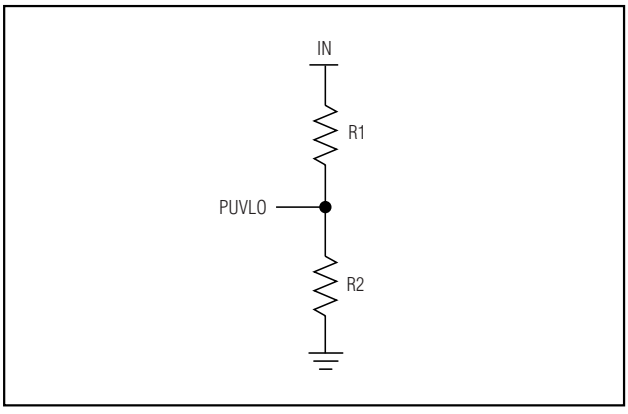

Figure 3. External PWM UVLO Divider

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12V/5V Input Buck PWM Controller

Setting the Output Voltage

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

$$R4 = \frac{R3}{\left[\frac{V_{OUT}}{V_{FB}} - 1\right]}$$

where V_{FB} = 0.8V.

Inductor Selection

Three key inductor parameters must be specified for operation with the MAX5951: 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. A lower inductance 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. A higher inductance increases efficiency by reducing the ripple current; however, 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 rule of thumb is to choose ΔI_{P-P} equal to 30% 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 is programmable between 100kHz and 1000kHz (see the *Oscillator/Synchronization Input (SYNCIN)/ Synchronization Output (SYNCOUT)* 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 current ripple is acceptable. The inductor saturation current (I_{SAT}) is also important to avoid runaway current during continuous output short-circuit conditions. Select an inductor with an I_{SAT} specification higher than the maximum peak current.

Input Capacitor Selection

The discontinuous input current of the buck converter causes large input ripple currents; therefore, the input capacitor must be carefully chosen to withstand the input ripple current and maintain the input voltage ripple within design requirements. The total voltage ripple is the sum of ΔV_Q (caused by the capacitor discharge) and ΔV_{ESR} (caused by the ESR of the input capacitor), which peaks at the end of the on cycle. Calculate the input capacitance and ESR required for a specified ripple using the following equations:

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

$$C_{IN} = \frac{I_{LOAD(MAX)} \times \left(\frac{V_{OUT}}{V_{IN}}\right)}{\Delta V_Q \times f_{SW}}$$

where

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

$I_{LOAD(MAX)}$ is the maximum output current, ΔI_{P-P} is the peak-to-peak inductor current, and f_{SW} is the switching frequency.

The MAX5951 includes UVLO hysteresis to avoid possible unintentional chattering during turn-on. Use additional bulk capacitance if the input source impedance is high. When the input voltage is near the UVLO, additional input capacitance helps avoid possible undershoot below the UVLO threshold during transient loading.

Output Capacitor Selection

The allowed output-voltage ripple and the maximum deviation of the output voltage during load steps determine the required 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 ESR of the output capacitor). The equations for calculating the peak-to-peak output-voltage ripple are:

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

$$\Delta V_{ESR} = ESR \times \frac{\Delta I_{P-P}}{2}$$

ΔV_{ESR} and ΔV_Q are not directly additive since they are out of phase from each other. If using ceramic capacitors, which generally have low ESR, ΔV_Q dominates. If using electrolytic capacitors, ΔV_{ESR} dominates.

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The allowable deviation of the output voltage during load transients also affects the choice of 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 Guidelines* section). The resistive drop across the output capacitor's ESR, the drop across the capacitor's ESL, and the capacitor discharge cause a voltage droop during the load-step. Use a combination of low-ESR tantalum/aluminum electrolyte and ceramic capacitors for better load transient and voltage ripple performance. Surface-mount capacitors and capacitors in parallel help reduce the ESL. Keep 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:

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

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

$$ESL = \frac{\Delta V_{ESL} \times t_{STEP}}{I_{STEP}}$$

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.

Setting the Current Limit

Connect a 25kΩ to 175kΩ resistor, R_{ILIM}, from ILIM to AGND to program the valley current-limit threshold between 50mV and 350mV. ILIM sources 20µA out to R_{ILIM}. The resulting voltage divided by 10 is the valley current-limit threshold.

The MAX5951 uses a valley current-sense method for current limiting. The voltage drop across the low-side MOSFET due to its on-resistance is used to sense the inductor current. The voltage drop across the low-side MOSFET at the valley point and at $I_{LOAD(MAX)}$ is:

$$V_{VALLEY} = R_{DS(ON)}(T) \times \left(I_{LOAD(MAX)} - \frac{\Delta I_{P-P}}{2} \right)$$

$R_{DS(ON)}$ is the on-resistance of the low-side MOSFET, which is temperature dependent, $I_{LOAD(MAX)}$ is the maximum DC load current, and ΔI_{P-P} is the peak-to-peak inductor current.

The 20µA current source, ILIM reference current, has a temperature coefficient of 3333ppm/°C. This allows the valley current-limit threshold:

$$\frac{R_{ILIM} \times 20\mu A(T)}{10}$$

to track and compensate for the increase in the synchronous MOSFET's $R_{DS(ON)}$ with increasing temperature. MOSFETs typically have a temperature coefficient range within 3000ppm/°C to 7000ppm/°C. Refer to the MOSFET data sheet for a device-specific temperature coefficient.

At a given temperature, the calculated V_{VALLEY} must be less than the minimum valley current-limit threshold specified.

Figure 4 illustrates the effect of MAX5951 ILIM reference current temperature coefficient to compensate for the variation of the MOSFET $R_{DS(ON)}$ over the operating junction temperature range.

Power MOSFET Selection

When selecting the MOSFETs, consider the total gate charge, $R_{DS(ON)}$, power dissipation, the maximum drain-to-source voltage, package thermal impedance, and desired current limit. The product of the MOSFET gate charge and on-resistance is a figure of merit, with a lower number signifying better performance. Choose MOSFETs optimized for high-frequency switching applications. The average gate-drive current from the MAX5951's output is proportional to the frequency and gate charge required to drive the MOSFET. The power dissipated in the MAX5951 is proportional to the input voltage and the average drive current (see the *Power Dissipation* section).

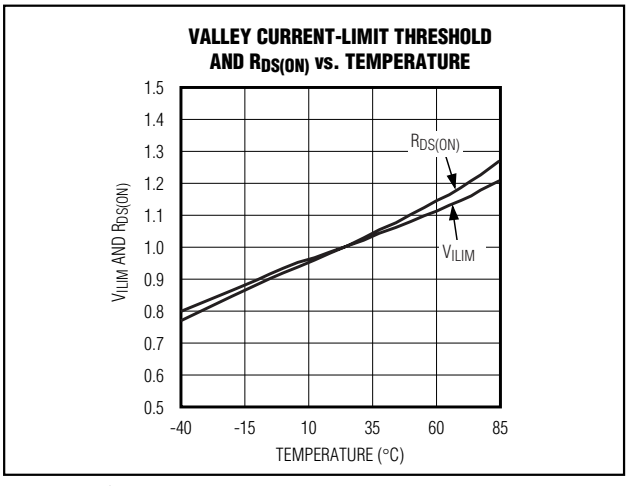

Figure 4. Current-Limit Threshold and $R_{DS(ON)}$ vs. Temperature

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Compensation Design Guidelines

The MAX5951 uses 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 compensation network must compensate for this gain drop and phase shift in order 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_{IN}/V_{RAMP}, with a double pole and a single zero set by the output inductance (L), the output capacitance (C_{OUT}), and its equivalent series resistance (ESR). Below are equations that define the power modulator:

$$G_{MOD(DC)} = \frac{V_{IN}}{V_{RAMP}}$$

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

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

The switching frequency is programmable between 100kHz and 1000kHz by an external resistor at RT. The crossover frequency (f_C), which is the frequency when the closed-loop gain is equal to unity, should be set to f_{SW} / 10 or f_{GBW} / 25, whichever is lower.

The error amplifier must provide a gain-and-phase boost to compensate for the rapid gain-and-phase loss from the LC double pole. This is accomplished by utilizing type 3 compensation (see Figures 5 and 6) that introduces two zeros and three poles into the control loop. The error amplifier has a low-frequency pole (f_{P1}) at the origin; two zeros at:

$$f_{Z1} = \frac{1}{2\pi \times R5 \times C7}$$

and:

$$f_{Z2} = \frac{1}{2\pi \times R3 \times C6}$$

and higher frequency poles at:

$$f_{P2} = \frac{1}{2\pi \times R6 \times C6}$$

and:

$$f_{P3} = \frac{1}{2\pi \times R5 \times C8}$$

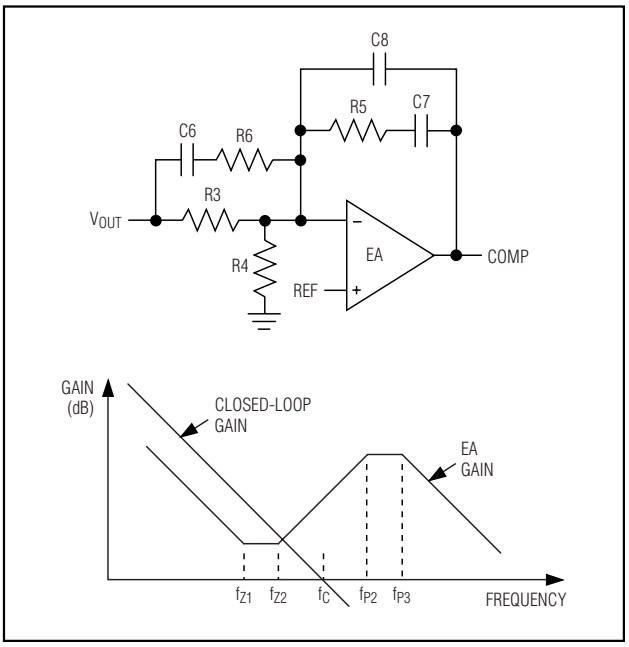

Figure 5. Error-Amplifier Compensation Circuit (Closed-Loop and Error-Amplifier Gain Plot) for Ceramic Capacitors

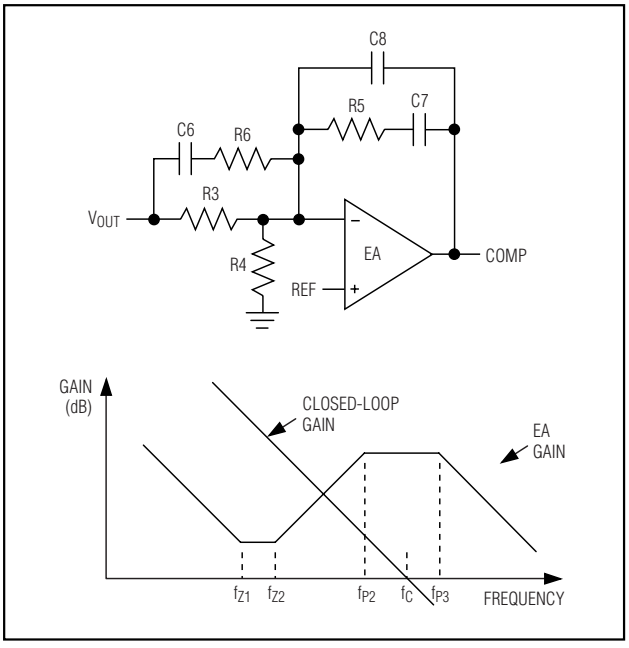

Figure 6. Error-Amplifier Compensation Circuit (Closed-Loop and Error-Amplifier Gain Plot) for Higher ESR Output Capacitors

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Compensation when $f_C < f_{ZESR}$

Figure 5 shows the error-amplifier feedback, as well as its gain response for circuits that use low-ESR output capacitors (ceramic). In this case, f_C occurs before f_{ZESR}.

f_{Z1} is set to 0.5 x f_{LC} and f_{Z2} is set to f_{LC} in order 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 Selection* and *Output Capacitor Selection* sections.

Pick a value for the feedback resistor R5 in Figure 5 (values between 1kΩ and 10kΩ are adequate). C7 is then calculated as:

$$C7 = \frac{1}{2\pi \times 0.5 \times f_{LC} \times R5}$$

f_C occurs between f_{Z2} and f_{P2}. The circuit is implemented with C7 >> C8 and R3 >> R6, in which case, the error amplifier gain (G_{EA}) at f_C is due primarily to C6 and R5. Therefore:

$$G_{EA(fc)} = 2\pi \times f_C \times C6 \times R5$$

The modulator gain at f_C is:

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

Since $G_{EA(fC)}$ x $G_{MOD(fC)}$ = 1, C6 is calculated by:

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

R3 is then calculated as:

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

f_{P2} is set at 1/2 the switching frequency (f_{SW}). R6 is then calculated by:

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

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

$$C8 = \frac{1}{2\pi \times R5 \times 5 \times f_C}$$

Compensation when $f_C > f_{ZESR}$

For larger ESR capacitors such as tantalum and aluminum electrolytic, f_{ZESR} can occur before f_C. If f_C > f_{ZESR}, 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 zeros (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 5 shows the error-amplifier feedback, as well as its gain response for circuits that use higher ESR output capacitors (tantalum, aluminum electrolytic, etc.).

Pick a value for feedback resistor R5 in Figure 5 (values between 1kΩ and 10kΩ are adequate). C7 is then calculated as:

$$C7 = \frac{1}{2\pi \times 0.5 \times f_{LC} \times R5}$$

The circuit is implemented with C7 >> C8 and R3 >> R6, in which case the error-amplifier gain between f_{P2} and f_{P3} is approximately equal to:

$$\frac{R5}{R6}$$

The modulator gain at f_C is:

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

Since $G_{EA(fC)}$ x $G_{MOD(fC)}$ = 1, R6 can then be calculated as:

$$R6 \approx \frac{R5 \times G_{MOD(DC)}}{(2\pi)^2 \times L \times C_{OUT} \times f_C^2}$$

f_{P2} is set to f_{ZESR}. C6 is then calculated as:

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

R3 is then calculated as:

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

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

$$C8 = \frac{1}{2\pi \times R5 \times 5 \times f_C}$$

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PWM Controller Applications Information

Power Dissipation

The 32-pin TQFN thermally enhanced package can dissipate 2.7W. Calculate power dissipation in the MAX5951 as a product of the input voltage and the total REG output current (I_{REG}). I_{REG} includes quiescent current (I_Q) and gate drive current (I_{DREG}):

$$P_D = V_{IN} \times I_{REG}$$

$$I_{REG} = I_Q + [f_{SW} \times (Q_{G1} + Q_{G2})]$$

where Q_{G1} and Q_{G2} are the total gate charge of the low-side and high-side external MOSFETs. f_{SW} is the switching frequency of the converter, and I_Q is the quiescent current of the device at the switching frequency.

Use the following equation to calculate the maximum power dissipation (P_{DMAX}) in the chip at a given ambient temperature (T_A):

$$P_{DMAX} = 34.5 \times (150 - T_A)$$

PC Board Layout Guidelines

Use the following guidelines to lay out the switching voltage regulator:

1) Place the IN and DREG bypass capacitors close to the MAX5951 PGND pin. Place the REG bypass capacitor close to the AGND pin.
2) Minimize the area and length of the high-current loops from the input capacitor, upper switching MOSFET, inductor, and output capacitor back to the input capacitor negative terminal.
3) Keep short the current loop formed by the synchronous switching MOSFET, inductor, and output capacitor.
4) Keep AGND and PGND isolated and connect them at one single point close to the negative terminal of the input filter capacitor.
5) Run the current-sense lines CS+ and CS- close to each other to minimize the loop area.
6) Avoid long traces between the REG/DREG bypass capacitors, driver output of the MAX5951, MOSFET gates, and PGND. Minimize the loop formed by the REG bypass capacitors, bootstrap diode, bootstrap capacitor, the MAX5951, and the upper MOSFET gate.
7) Place the bank of output capacitors close to the load.
8) Distribute the power components evenly across the board for proper heat dissipation.
9) Provide enough copper area at and around the switching MOSFETs and inductor to aid in thermal dissipation.
10) Use 2oz copper to keep the trace inductance and resistance to a minimum. Thin copper PC boards can compromise efficiency since high currents are involved in the application. Also, thicker copper conducts heat more effectively, thereby reducing thermal impedance.

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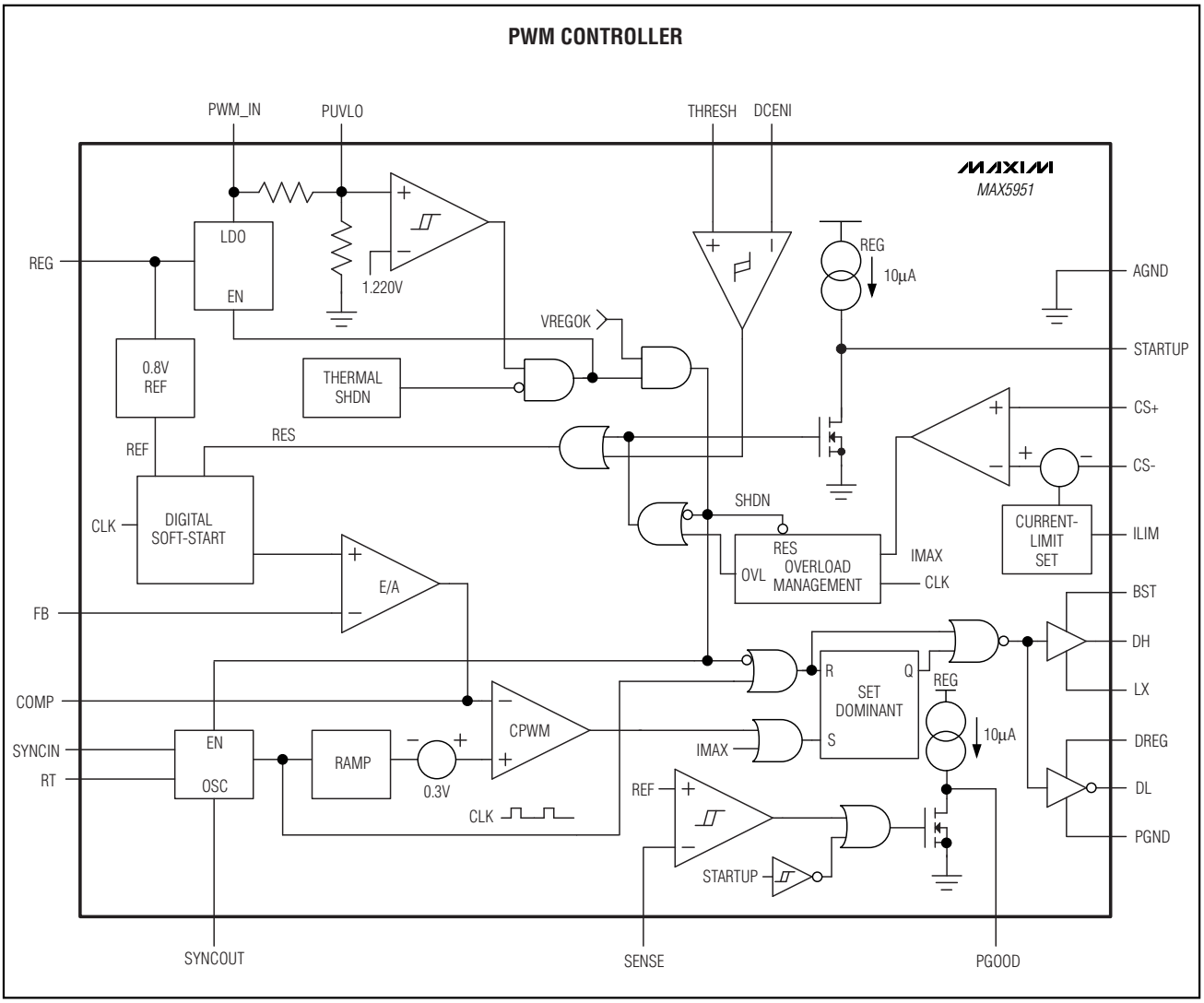
Simplified Block Diagram



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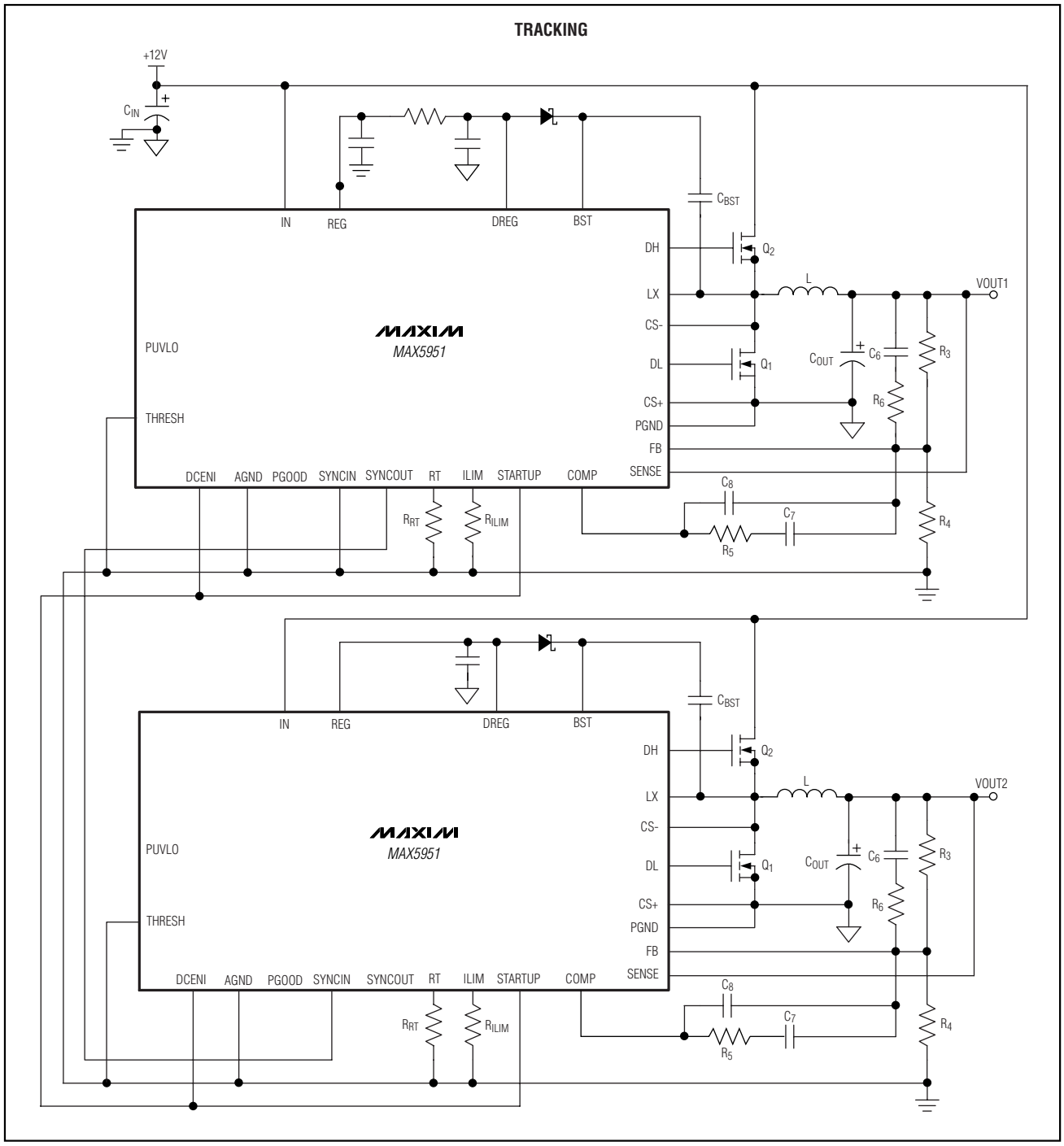
Typical Operating Circuits



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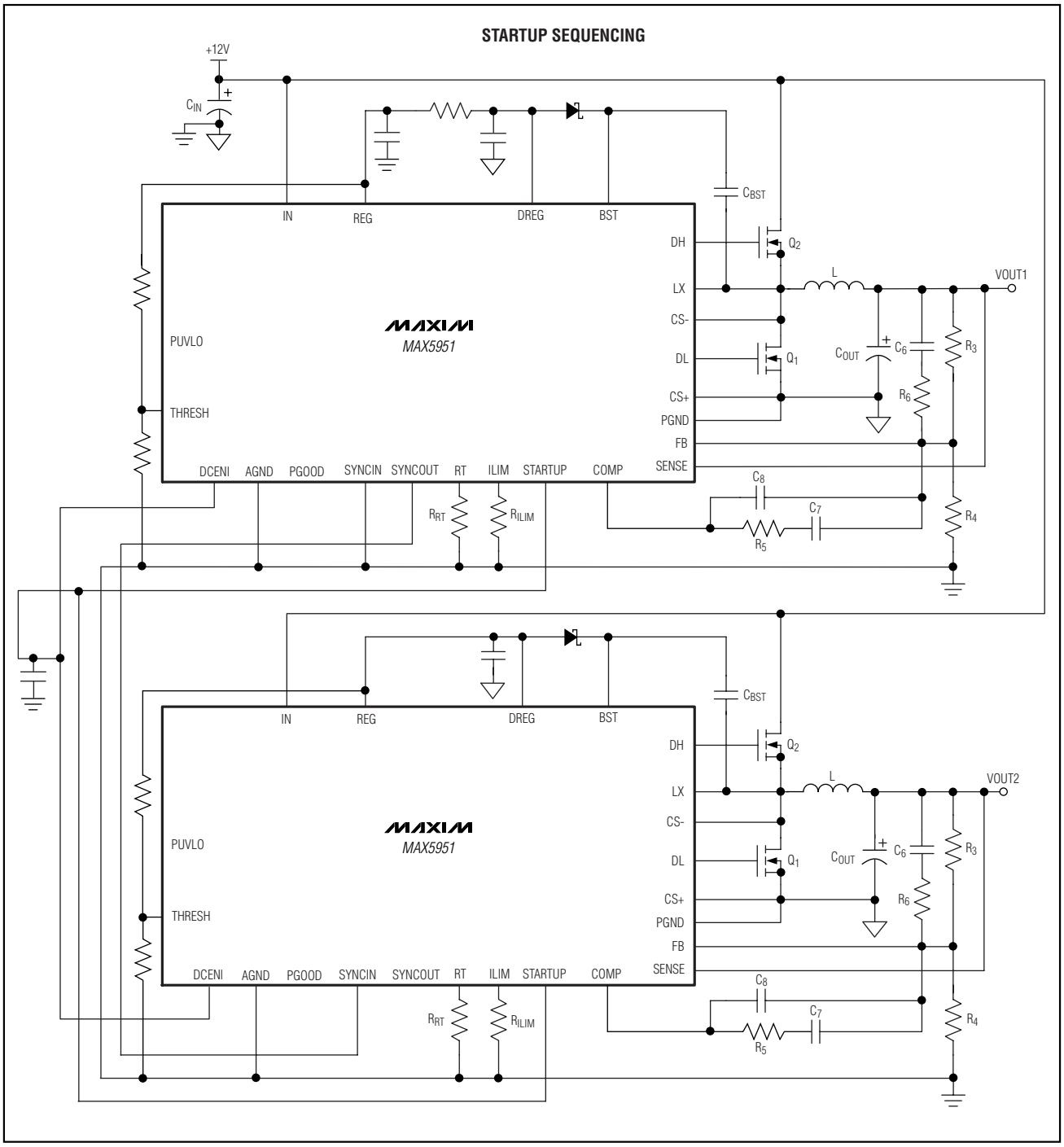
Typical Operating Circuits (continued)



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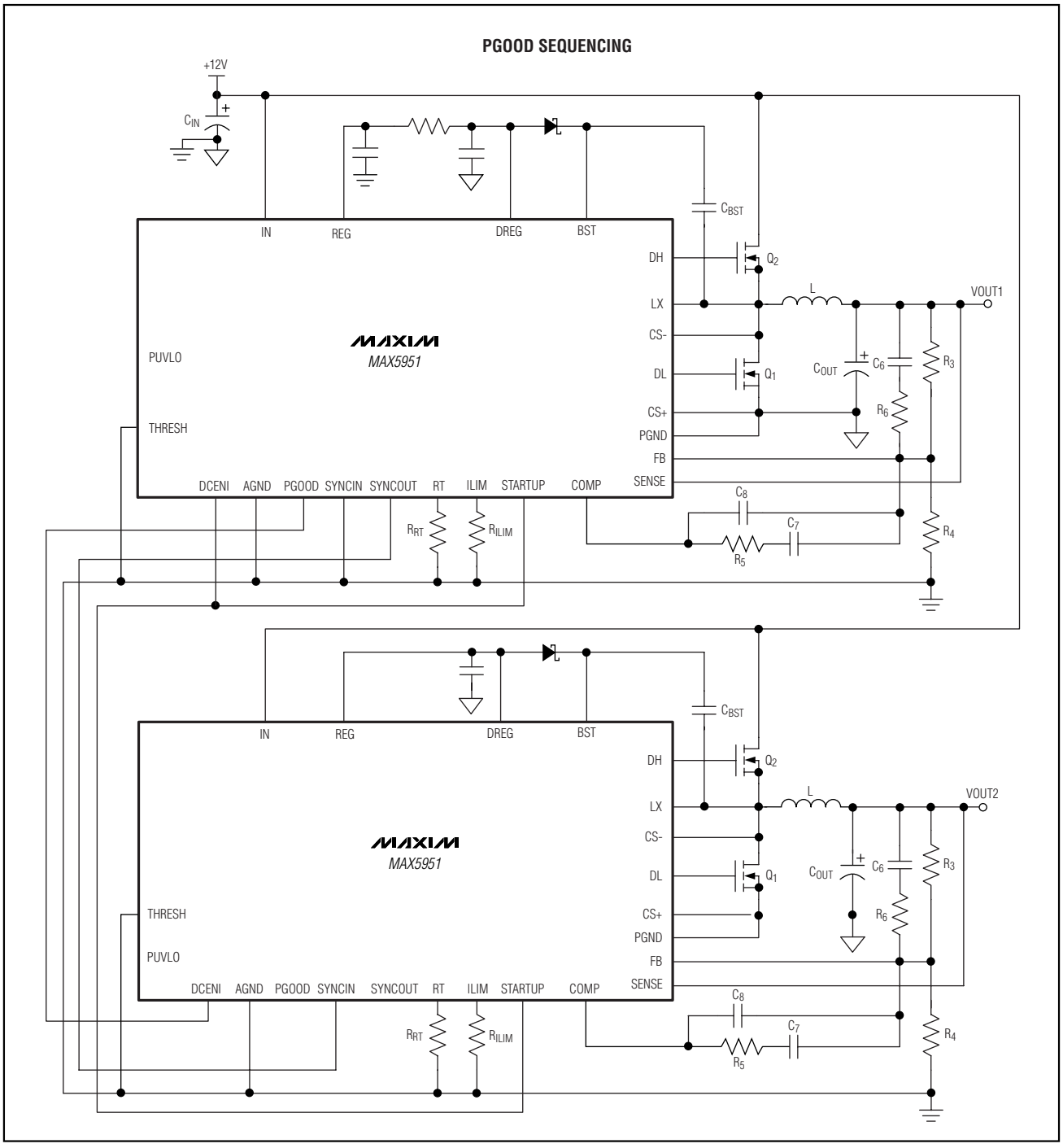
Typical Operating Circuits (continued)



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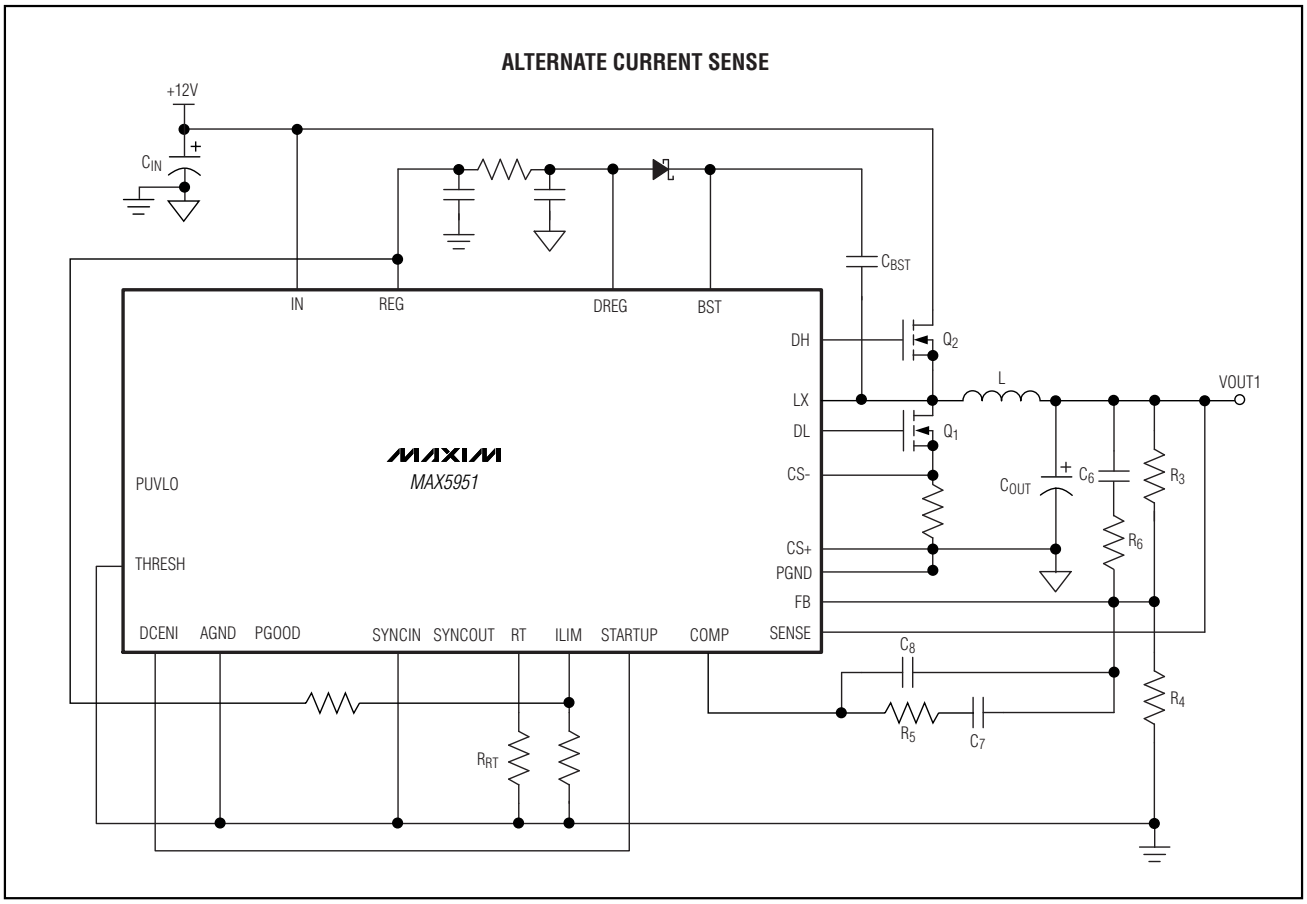
Typical Operating Circuits (continued)



Chip Information

PROCESS: BiCMOS

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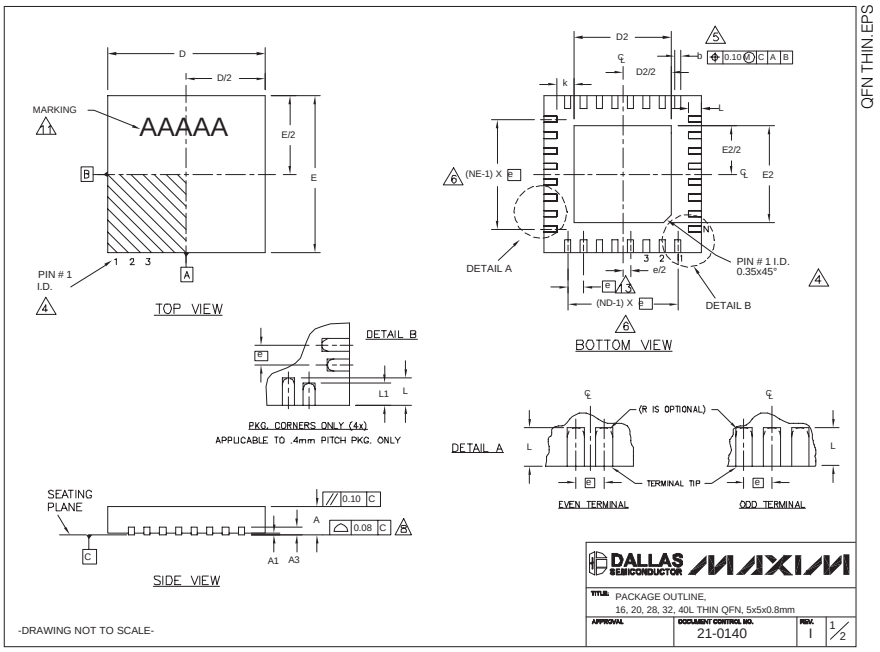
Package Information

(The package drawing(s) in this data sheet may not reflect the most current specifications. For the latest package outline information, go to **www.maxim-ic.com/packages**.)

QFN THIN.EPS



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DALLAS SEMICONDUCTOR MAXIM

TITLE: PACKAGE OUTLINE, 16, 20, 28, 32, 40L THIN QFN, 5x5x0.8mm			
APPROVAL	DOCUMENT CONTROL NO. 21-0140	REV. I	1/2

COMMON DIMENSIONS

PKG.	16L 5x5			20L 5x5			28L 5x5			32L 5x5			40L 5x5		
SYMBOL	MIN.	NOM.	MAX.	MIN.	NOM.	MAX.	MIN.	NOM.	MAX.	MIN.	NOM.	MAX.	MIN.	NOM.	MAX.
A	0.70	0.75	0.80	0.70	0.75	0.80	0.70	0.75	0.80	0.70	0.75	0.80	0.70	0.75	0.80
A1	0	0.02	0.05	0	0.02	0.05	0	0.02	0.05	0	0.02	0.05	0	0.02	0.05
A3	0.20 REF.			0.20 REF.			0.20 REF.			0.20 REF.			0.20 REF.		
b	0.25	0.30	0.35	0.25	0.30	0.35	0.20	0.25	0.30	0.20	0.25	0.30	0.15	0.20	0.25
D	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10
E	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10	4.90	5.00	5.10
e	0.80 BSC.			0.65 BSC.			0.50 BSC.			0.50 BSC.			0.40 BSC.		
k	0.25	-	-	0.25	-	-	0.25	-	-	0.25	-	-	0.25	0.35	0.45
L	0.30	0.40	0.50	0.45	0.55	0.65	0.45	0.55	0.65	0.30	0.40	0.50	0.40	0.50	0.60
L1	-	-	-	-	-	-	-	-	-	-	-	-	0.30	0.40	0.50
N	16			20			28			32			40		
ND	4			5			7			8			10		
NE	4			5			7			8			10		
JEDEC	WHHB			WHHC			WHHD-1			WHHD-2			-----		

EXPOSED PAD VARIATIONS

PKG. CODES	D2			E2			L	DOWN BONDS ALLOWED
	MIN.	NOM.	MAX.	MIN.	NOM.	MAX.	±0.15	
T1655-2	3.00	3.10	3.20	3.00	3.10	3.20	**	YES
T1655-3	3.00	3.10	3.20	3.00	3.10	3.20	**	NO
T1655N-1	3.00	3.10	3.20	3.00	3.10	3.20	**	NO
T2055-3	3.00	3.10	3.20	3.00	3.10	3.20	**	YES
T2055-4	3.00	3.10	3.20	3.00	3.10	3.20	**	NO
T2055-5	3.15	3.25	3.35	3.15	3.25	3.35	0.40	YES
T2855-3	3.15	3.25	3.35	3.15	3.25	3.35	**	YES
T2855-4	2.60	2.70	2.80	2.60	2.70	2.80	**	YES
T2855-5	2.60	2.70	2.80	2.60	2.70	2.80	**	NO
T2855-6	3.15	3.25	3.35	3.15	3.25	3.35	**	NO
T2855-7	2.60	2.70	2.80	2.60	2.70	2.80	**	YES
T2855-8	3.15	3.25	3.35	3.15	3.25	3.35	0.40	YES
T2855N-1	3.15	3.25	3.35	3.15	3.25	3.35	**	NO
T3255-3	3.00	3.10	3.20	3.00	3.10	3.20	**	YES
T3255-4	3.00	3.10	3.20	3.00	3.10	3.20	**	NO
T3255-5	3.00	3.10	3.20	3.00	3.10	3.20	**	YES
T3255N-1	3.00	3.10	3.20	3.00	3.10	3.20	**	NO
T4055-1	3.20	3.30	3.40	3.20	3.30	3.40	**	YES

**SEE COMMON DIMENSIONS TABLE

NOTES:

1. DIMENSIONING & TOLERANCING CONFORM TO ASME Y14.5M-1994.
2. ALL DIMENSIONS ARE IN MILLIMETERS. ANGLES ARE IN DEGREES.
3. N IS THE TOTAL NUMBER OF TERMINALS.
4. THE TERMINAL #1 IDENTIFIER AND TERMINAL NUMBERING CONVENTION SHALL CONFORM TO JESD 95-1 SPP-012. DETAILS OF TERMINAL #1 IDENTIFIER ARE OPTIONAL, BUT MUST BE LOCATED WITHIN THE ZONE INDICATED. THE TERMINAL #1 IDENTIFIER MAY BE EITHER A MOLD OR MARKED FEATURE.
5. DIMENSION b APPLIES TO METALLIZED TERMINAL AND IS MEASURED BETWEEN 0.25 mm AND 0.30 mm FROM TERMINAL TIP.
6. ND AND NE REFER TO THE NUMBER OF TERMINALS ON EACH D AND E SIDE RESPECTIVELY.
7. DEPOPULATION IS POSSIBLE IN A SYMMETRICAL FASHION.
8. COPLANARITY APPLIES TO THE EXPOSED HEAT SINK SLUG AS WELL AS THE TERMINALS.
9. DRAWING CONFORMS TO JEDEC MO220, EXCEPT EXPOSED PAD DIMENSION FOR T2855-3 AND T2855-6.
10. WARPAGE SHALL NOT EXCEED 0.10 mm.
11. MARKING IS FOR PACKAGE ORIENTATION REFERENCE ONLY.
12. NUMBER OF LEADS SHOWN ARE FOR REFERENCE ONLY.
13. LEAD CENTERLINES TO BE AT TRUE POSITION AS DEFINED BY BASIC DIMENSION "e", ±0.05.

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