Click here for production status of specific part numbers.

MAX25600

Synchronous High-Voltage Four-Switch Buck- Boost LED Controller

General Description

The MAX25600 is a synchronous 4-switch buck-boost LED driver controller. The controller regulates the LED current for LED string voltages from 0V to 60V. The MAX25600 can be used as a seamless buck-boost LED driver for applications that require an efficient buck-boost LED driver with synchronous rectification. The MAX25600 is ideal for high-power applications that require a current source with PWM dimming capability.

The device provides seamless transition between buck, boost, and buck-boost modes depending on the ratio of input to output voltage. The MAX25600 is ideal for LED driver applications in automotive, industrial, and other LED lighting applications. A fault flag indicates open LED, shorted LED, or thermal shutdown conditions. The device uses Maxim's proprietary average-current-mode control scheme and allows adjustable 200kHz to 700kHz fixed-frequency operation. In addition, ±6% triangular spread spectrum is added internally to the oscillator to improve EMI performance. The MAX25600 provides both analog and digital PWM dimming, and has built-in analog PWM dimming at a dimming frequency of 200Hz. The adjustable soft-start feature limits the current peaks and voltage overshoots at startup. The MAX25600 integrates a high-side p-channel dimming MOSFET driver for PWM dimming applications that require fast rising and falling edges of the LED current. It also features robust output open and short protection, is AEC qualified, and is suitable for automotive applications.

Applications

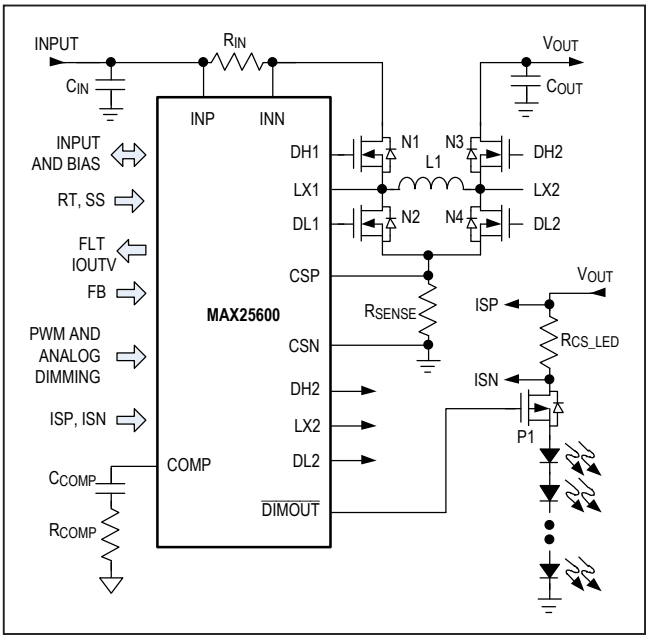
- Automotive Exterior Lighting and General, Commercial, and Industrial Lighting

Ordering Information appears at end of data sheet.

Benefits and Features

- Automotive Ready: AEC-Q100 Qualified
- Integration Minimizes BOM for High-Brightness LED Driver, Saving Space and Reducing Cost
  - Wide Input Voltage Range from 5V to 60V
  - H-Bridge Single Inductor Buck-Boost Architecture
  - Constant-Current and Constant-Voltage Regulation
  - 28-pin TSSOP with EP Pad and 28-pin (5mm x 5mm) TQFN with EP Pad Packages
- Wide Dimming Ratio allows High Contrast Ratio
  - Analog and PWM Dimming
  - Flicker-Free PWM Dimming with Spread Spectrum
  - Integrated pMOS Dimming FET Gate Driver
  - 200Hz On-Board Ramp Simplifies PWM Dimming
- Protection Features and Wide Temperature Range Increase System Reliability
  - Short Circuit,Overvoltage, and Thermal Protection
  - LED Current Monitor and Input Current Limiter
  - -40ºC to +125ºC Operating Junction-Temperature Range

Simplified Application Circuit



Absolute Maximum Ratings

IN, UVEN to SGND .....-0.3V to +52V
INN, INP, ISP, ISN, and $\overline{DIMOUT}$ to PGND .....-0.3V to +65V
ISP to ISN.....-0.3V to +0.6V
LX1, LX2 to PGND .....-1.0V to +65V
BST_ to LX_ .....-0.3V to +6V
DH_ to LX_..... -0.3V to V_{BST}+0.3V
DL1, DL2 to PGND..... -0.3V to (V_{VCC}+0.3)V
CSP, CSN to SGND .....-2.5V to +6V
CSP to CSN .....-0.5V to +0.5V
IOUTV, COMP, RT, SS ..... -0.3 to V_{VCC}+0.3
V_{CC} to SGND .....-0.3V to +6V
ICTRL, PWM, FB, and $\overline{FLT}$ to SGND.....-0.3V to +6V
PGND to SGND.....-0.3V to +0.3V
V_{CC} Short-Circuit Duration.....Continuous
Continuous Power Dissipation (TQFN)
(T_A = +70°C, derate, 28.6mW/°C above +70°C).....2285mW
Continuous Power Dissipation (TSSOP)
(T_A = +70°C, derate 27mW/°C above +70°C).....2162mW
Operating Junction-Temperature Range
(Note 1, 2)..... -40°C to +125°C
Storage-Temperature Range..... -40°C to +150°C
Soldering Temperature (reflow).....+260°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.

Package Information

28-TQFN

PACKAGE CODE	T2855Y+5C
Outline Number	21-100130
Land Pattern Number	90-0027
Thermal Resistance, Single-Layer Board:	
Junction to Ambient (θ_{JA})	68°C/W
Junction to Case (θ_{JC})	11°C/W
Thermal Resistance, Four-Layer Board:	
Junction to Ambient (θ_{JA})	60°C/W
Junction to Case (θ_{JC})	11°C/W

28-TSSOP

PACKAGE CODE	U28E+1C
Outline Number	21-100182
Land Pattern Number	90-100069
Thermal Resistance, Single-Layer Board:	
Junction to Ambient (θ_{JA})	78°C/W
Junction to Case (θ_{JC})	13°C/W
Thermal Resistance, Four-Layer Board:	
Junction to Ambient (θ_{JA})	71.6°C/W
Junction to Case (θ_{JC})	13°C/W

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 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 **www.maximintegrated.com/thermal-tutorial**.

Electrical Characteristics

(V_{IN} = 12V, V_{UVEN} = 12V, limits are 100% tested at T_A = +25°C and T_A = +125°C. Limits over the operating temperature range and relevant supply-voltage range are guaranteed by design and characterization.)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
SUPPLY VOLTAGE						
INP Input Voltage Range	V_{INP}		5.0		60	V
Supply Current	I_{INQ}	No switching		3	6	mA
UNDERVOLTAGE LOCKOUT						
Undervoltage Lockout Rising	V_{UVEN_THR}	V_{UVEN} rising	1.12	1.24	1.37	V
Hysteresis				106		mV
Shutdown Current	I_{SHTDN}	V_{UVEN} = 0, V_{IN} = 12V		6	20	µA
V_{CC} REGULATOR						
Output Voltage	V_{CC}	5.5V < V_{IN} < 40V; I_{VCC} = 1mA	4.9	5.0	5.1	V
		I_{VCC} = 30mA, 5.5V < V_{IN} < 36V	4.9	5.0	5.1	
		I_{VCC} = 1mA to 60mA, 6V < V_{IN} < 25V	4.9	5.0	5.1	
Dropout Voltage	V_{CC_DROP}	V_{IN} = 4.5V, I_{VCC} = 5mA		60	110	mV
V_{CC} UVLO Rising	V_{CC_UVLOR}	Rising		4.0		V
V_{CC} UVLO Falling	V_{CC_UVLOF}	Falling		3.75		V
Short-Circuit Current Limit	I_{VCC_SC}			70		mA
INPUT CURRENT-SENSE AMPLIFIER						
Input Current-Sense Common-Mode Range			5		60	V
Input Current-Sense Threshold		3V < V_{INP} < 60V	88	100	112	mV
INP Bias Current		V_{INP} - V_{INN} = 100mV, V_{INP} = 60V		50		µA
INN Bias Current		V_{INP} - V_{INN} = 100mV, V_{INP} = 60V		10		µA
CSP, CSN CURRENT-SENSE AMPLIFIER						
Voltage Gain (Boost, Buck modes)		V_{CSP} = 100mV, V_{CSN} = 0	9.5	10	10.5	V/V
ANALOG DIMMING						
ICTRL Control Input Voltage Range	$ICTRL_{RNG}$		0.2		1.2	V
ICTRL Zero Current Threshold	I_{CTRLZC_VTH}	(V_{ISP} - V_{ISN}) < 5mV	0.16	0.18	0.2	V
ICTRL Clamp Voltage	$ICTRL_{CLMP}$	ICTRL sink = 1µA	1.25	1.30	1.35	V
ICTRL Input Bias Current	$ICTRLI_{IN}$	V_{ICTRL} = 0.2 to 5.5V		20	500	nA
LED CURRENT-SENSE AMPLIFIER						
Common-Mode Input Range			-0.3		+60	V
Differential Signal Range			0		200	mV
ISN Input Bias Current	IB_{ISN}	V_{ISP} - V_{ISN} = 200mV, V_{ISP} = 60V		22	60	µA

Electrical Characteristics (continued)

(V_{IN} = 12V, V_{UVEN} = 12V, limits are 100% tested at T_A = +25°C and T_A = +125°C. Limits over the operating temperature range and relevant supply-voltage range are guaranteed by design and characterization.)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
ISP Input Bias Current	IB_{ISP}	V_{ISP} - V_{ISN} = 200mV, V_{ISP} = 60V		350	550	µA
Voltage Gain		V_{ISP} - V_{ISN}) = 200mV, 3V < V_{ISP}, V_{ISN} < 60V	4.9	5.00	5.1	V/V
LED Current-Sense Regulation Voltage		V_{ICTRL} = 1.3V, 3V < (V_{ISP} - V_{ISN}) < 60V	213.8	220	226.2	mV
		V_{ICTRL} = 1.2V, 3V < (V_{ISP} - V_{ISN}) < 60V	194	200	206	
		V_{ICTRL} = 0.4V, 3V < (V_{ISP} - V_{ISN}) < 60V	36	40	44	
LED Current-Sense Regulation Voltage (Low Range)		V_{ICTRL} = 1.2V, 0V < (V_{ISP}- V_{ISN}) < 3V	192	200	208	mV
		V_{ICTRL} = 0.4V, 0V < (V_{ISP} - V_{ISN}) < 3V	35	40	45	
Common-Mode Input Range Selector	RNG_{SEL}	V_{ISP} rising	2.75	2.85	2.95	V
		V_{ISP} falling	2.5	2.6	2.7	
OSCILLATOR						
Switching-Frequency Range			200		700	kHz
Bias Voltage at RT	V_{RT}			1.25		V
Oscillator Frequency Accuracy		Dither disabled	-10		+10	%
Frequency Dither	f_{DITH}	Dither enabled, f_{SW} = 200kHz to 700kHz		±6		%
Switching Frequency	f_{SW}	R_{RT} = 50KΩ	370	400	430	kHz
		R_{RT}=105KΩ	180	200	220	
		R_{RT}=25.5KΩ	630	700	770	
SLOPE COMPENSATION ON CSP						
Slope Compensation Current-Ramp Height	I_{SLOPE_PK}	Peak-current ramp added to CSP input per switching cycle at 100% duty cycle	42.5	50	57.5	µA
ERROR AMPLIFIER						
Transconductance	g_M	V_{ISP} - V_{ISN} = 200mV	1170	1800	2430	µS
COMP Sink Current	$COMP_{ISINK}$	V_{COMP} = 5V		300		µA
COMP Source Current	$COMP_{ISRC}$	V_{COMP} = 0V		300		µA
COMP Clamp Boost Mode	$COMP_{CLBST}$			1.93		V
COMP Clamp Buck Mode	$COMP_{CLBK}$			1.74		V
nMOS GATE DRIVERS						
DH1 Sourcing Resistance	R_{DH1_SRC}			2.2		Ω
DH2 Sourcing Resistance	R_{DH2_SRC}			1.3		Ω
DH_ Sinking Resistance	R_{DH_SINK}			0.9		Ω
DL_ Sourcing Resistance	R_{DL_SRC}	DL_ = high		1.3		Ω
DL_ Sinking Resistance	R_{DL_SINK}	DL_ = low		0.9		Ω
DH_ to DL_ Dead Time		DH_ fall to DL_ rise		20		ns
DL_ to DH_ Dead Time		DL_ fall to DH_ rise		20		ns

Electrical Characteristics (continued)

(V_{IN} = 12V, V_{UVEN} = 12V, limits are 100% tested at T_A = +25°C and T_A = +125°C. Limits over the operating temperature range and relevant supply-voltage range are guaranteed by design and characterization.)

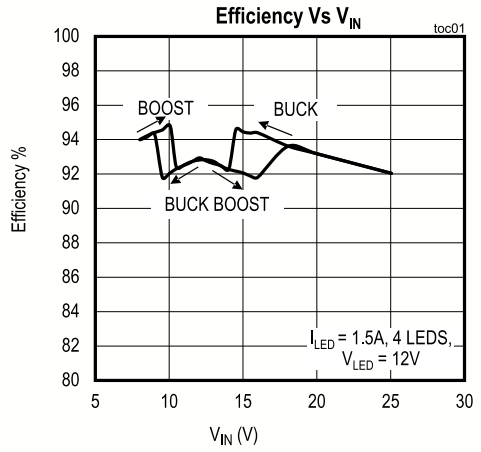
PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
PWM DIMMING						
Internal Ramp Frequency	f_{RAMP}		180	200	220	Hz
External Sync Frequency Range			30		2000	Hz
External Sync Low Level Voltage	V_{LTH}				0.4	V
External Sync High Level Voltage	V_{HTH}		2.2			V
DIM Comparator Offset Voltage			170	200	230	mV
Internal PWM Duty Cycle	V_{PWM}	V_{PWM} = 1.0V		28.6		%
		V_{PWM} = 1.7V		53.6		
DIM Voltage for 100% Duty Cycle			3.2			V
DIMMING MOSFET GATE DRIVER						
Peak Pullup Current	$I_{DIMOUTPU}$	PWM = 0V, (V_{ISP} - $V_{\overline{DIMOUT}}$) = 5V	30	50	85	mA
Peak Pulldown Current	$I_{DIMOUTPD}$	(V_{ISP} - $V_{\overline{DIMOUT}}$) = 0V	10	25	45	mA
$\overline{DIMOUT}$ Low Voltage with Respect to ISP			-5.4	-5.0	-4.6	V
DIM Turn-On Comparator		ISP rising		5.7		V
DIM Turn-Off Comparator		ISP falling		5.3		V
FAULT						
Minimum On Time (Buck)	t_{ON_MIN}			150	240	ns
FB Overvoltage Threshold	V_{TH_OVP}	FB rising	1.22	1.24	1.28	V
FB Overvoltage Hysteresis				0.1		V
Short Fault Threshold		FB falling		150		mV
Soft-Start Pullup Current				15		µA
$\overline{FLT}$ Output Voltage		I_{SINK} is 1mA after fault		0.05	0.3	V
$\overline{FLT}$ Leakage Current		$V_{\overline{FLT}}$ = 5.5V			1	µA
Hiccup Activation Threshold				400		mV
OFF TIME CONTROL						
Linear Range of Pulse Doubler				5		µs
THERMAL SHUTDOWN						
Thermal Shutdown Threshold	$T_{SHUTDOWN}$	Temperature Rising		165		°C
Thermal Shutdown Hysteresis	T_{HYS}			15		°C

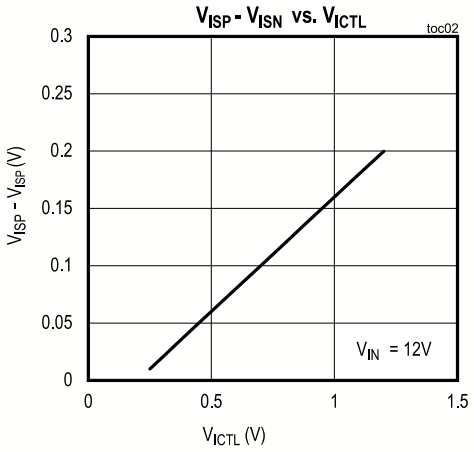
Note 1: The MAX25600 is guaranteed to meet performance specifications over the -40°C to +125°C operating junction-temperature range. High junction temperatures degrade operating lifetimes. Operating lifetime is derated for junction temperatures greater than +125°C.

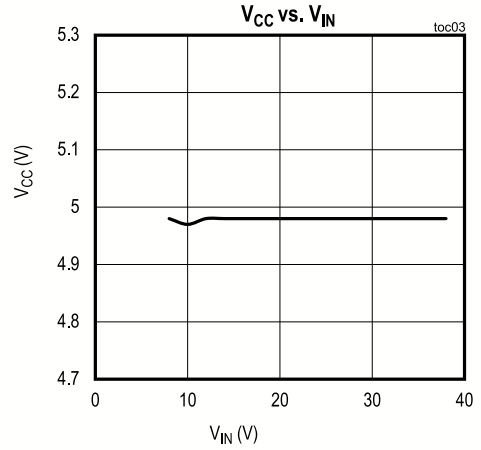
Note 2: The MAX25600 includes overtemperature protection that is intended to protect the device during momentary overload conditions. Junction temperature will exceed the maximum operating junction temperature when overtemperature protection is active. Continuous operation above the specified absolute maximum operating junction temperature may impair device reliability.

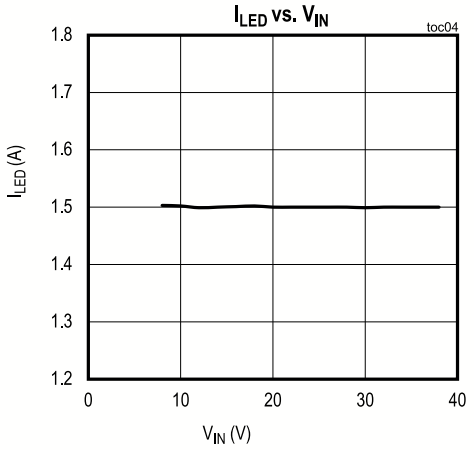
Typical Operating Characteristics

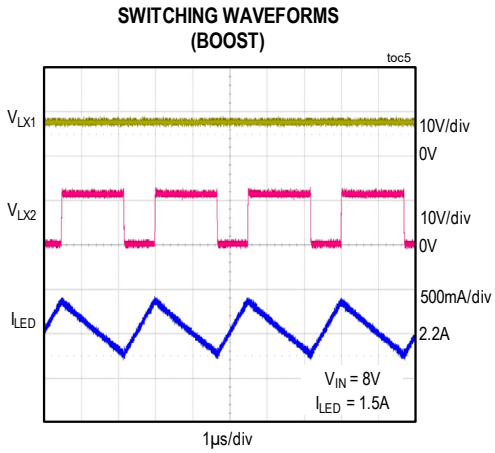
(T_A = +25°C, unless otherwise noted.)

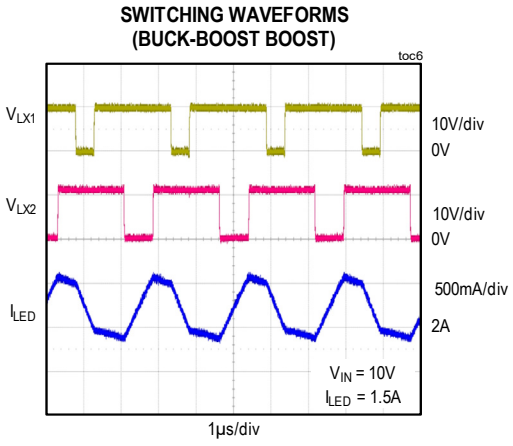








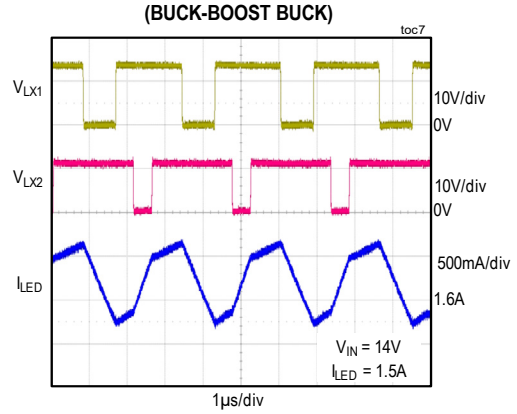




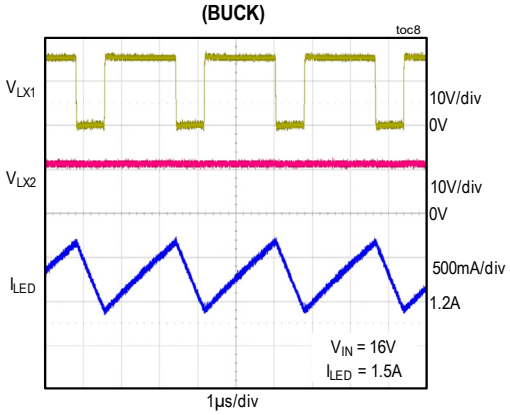
Typical Operating Characteristics (continued)

(T_A = +25°C, unless otherwise noted.)

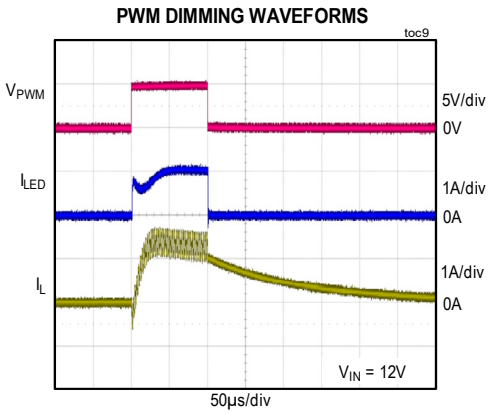
SWITCHING WAVEFORMS (BUCK-BOOST BUCK)



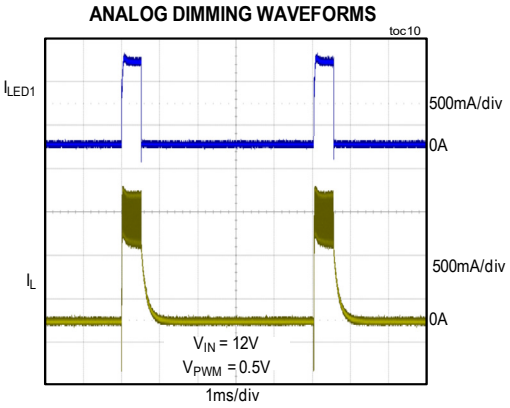
SWITCHING WAVEFORMS (BUCK)



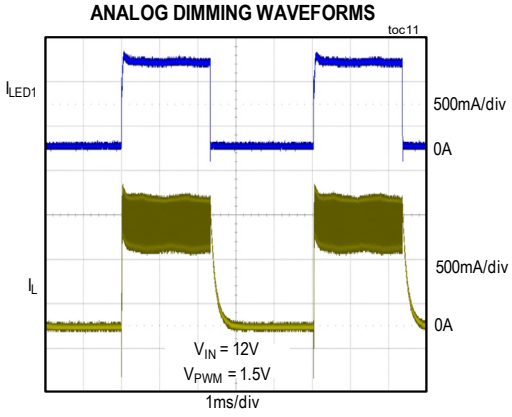
PWM DIMMING WAVEFORMS



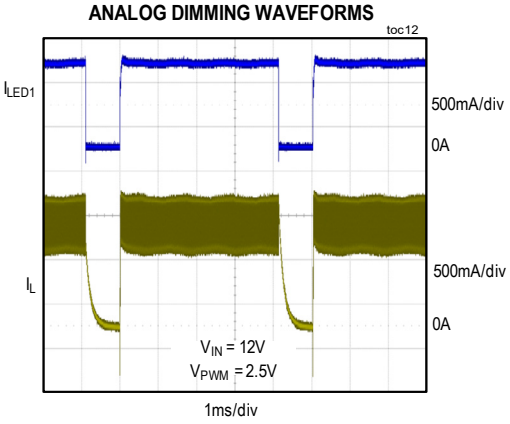
ANALOG DIMMING WAVEFORMS



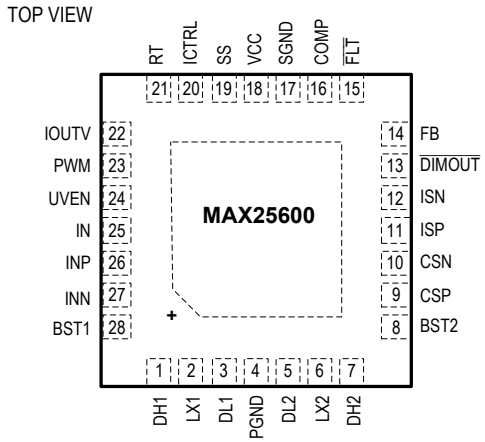
ANALOG DIMMING WAVEFORMS



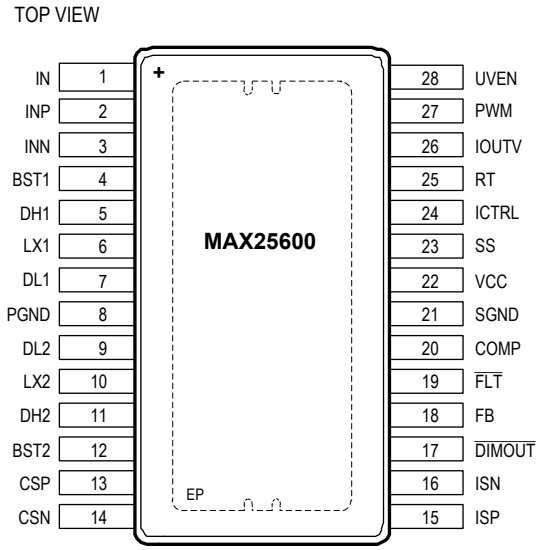
ANALOG DIMMING WAVEFORMS



Pin Configuration



TQFN
(5mm x 5mm)



TSSOP

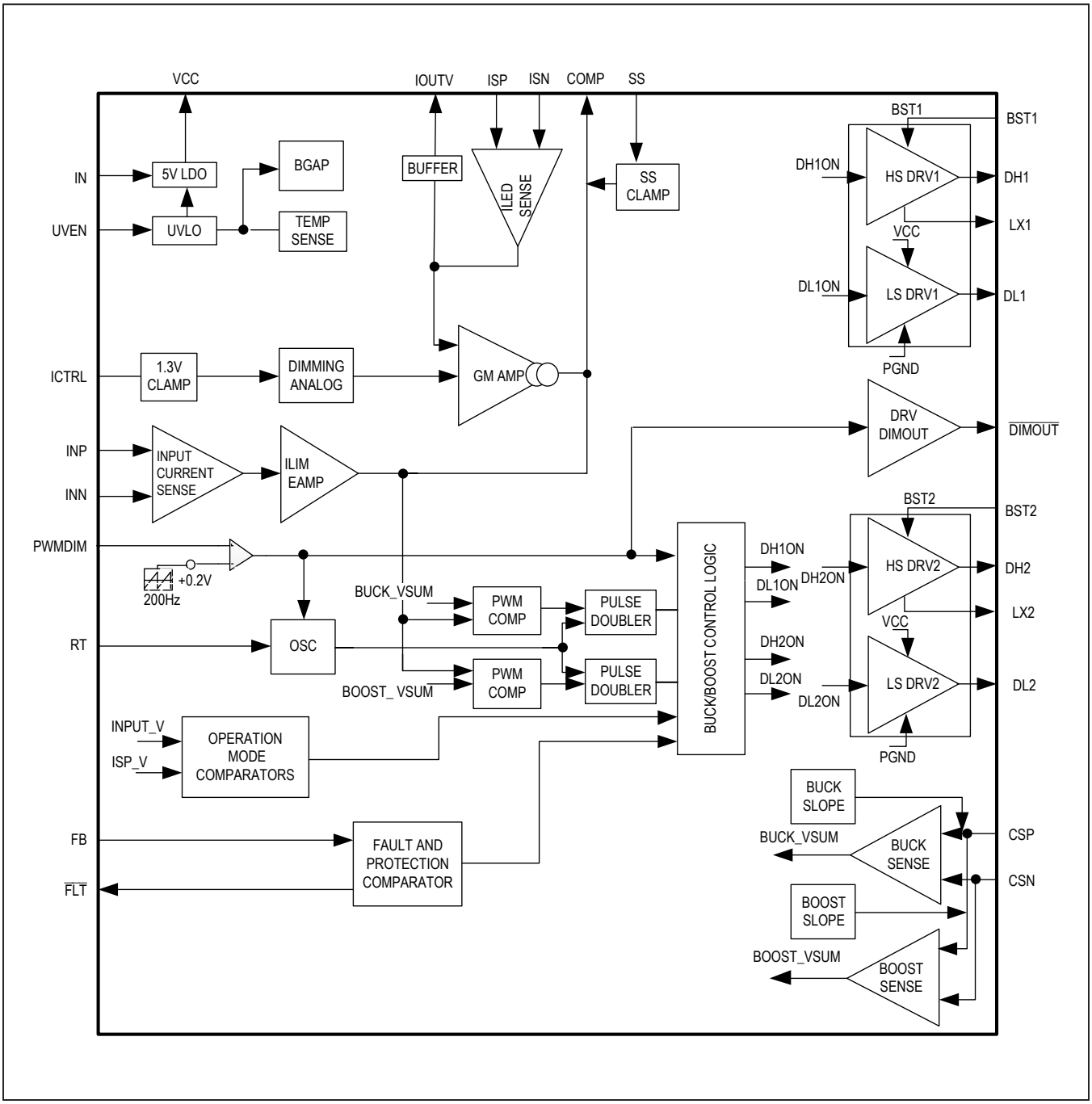
Pin Description

PIN		NAME	FUNCTION
TQFN	TSSOP		
1	5	DH1	Top Gate Drive for *Buck* Section of the MAX25600. Drives the gate of the top n-channel MOSFET.
2	6	LX1	Buck-Side Switching Node. LX1 pin swings from a diode voltage drop below ground up to V_{IN}.
3	7	DL1	Bottom Gate Drive for *Buck* Section of the MAX25600. Drives the gate of the bottom n-channel MOSFET.
4	8	PGND	Power Ground Connection
5	9	DL2	Bottom Gate Drive for the *Boost* Section of the MAX25600. Drives the gate of the bottom n-channel MOSFET.
6	10	LX2	Switching Node of *Boost* Section of the MAX25600. LX2 pin swings from a diode voltage drop below ground up to V_{OUT}.
7	11	DH2	Top Gate Drive for *Boost* Section of the MAX25600. Drives the gate of the top n-channel MOSFET.
8	12	BST2	High-Side Power Supply for High-Side Gate Drive for *Boost* Section. Connect a 0.1µF ceramic capacitor from BST2 to LX2.
9	13	CSP	Positive Input to the Current-Sense Comparator for the Average-Current-Mode Controller
10	14	CSN	Negative Input to the Current-Sense Comparator for the Average-Current-Mode Controller
11	15	ISP	Positive LED Current-Sense Input. The voltage between ISP and ISN is proportionally regulated to 1.3V or (V_{ICTRL} - 0.2)/(5 x RCS_LED), whichever is less.
12	16	ISN	Negative LED Current-Sense Input
13	17	$\overline{DIMOUT}$	External Dimming p-Channel MOSFET Gate Driver
14	18	FB	Overvoltage-Protection Input for the LED String. Connect a resistive divider between the output, FB, and GND. When the voltage on FB exceeds 1.23V, a fast-acting comparator immediately stops PWM switching and pulls $\overline{DIMOUT}$ high to disconnect the LED string from the output. $V_{OVP} = 1.23\frac{(R_{FB1} + R_{FB2})}{R_{FB2}}$
15	19	$\overline{FLT}$	Active-Low, Open-Drain Fault Indicator Output. See the *Fault Indicator ($\overline{FLT}$)* section.
16	20	COMP	Compensation-Network Connection. For proper compensation, connect a suitable RC network from COMP to SGND.
17	21	SGND	Signal Ground Connection
18	22	V_{CC}	5V Regulator Output. Connect a minimum 2.2µF ceramic capacitor from V_{CC} to SGND for stable operation.
19	23	SS	Soft-Start Pin. A minimum 10nF capacitor is recommended on this pin.

Pin Description (continued)

PIN		NAME	FUNCTION
TQFN	TSSOP		
20	24	ICTRL	Analog Dimming-Control Input. Connect an analog voltage from 0 to 1.3V for analog dimming of LED current. $I_{LED} = \frac{(V_{ICTRL} - 0.2V)}{5 \times R_{CS\_LED}}$ Bypass ICTRL to GND with at least a 10nF ceramic capacitor for noise filtering. Not needed if $V_{REFI} > 1.3V$.
21	25	RT	Oscillator Switching-Frequency Programming. Connect a resistor (R_{RT}) from RT to SGND to set the internal clock frequency. $f_{OSC}(kHz) = 20000/R_{RT}$ (kΩ).
22	26	IOUTV	Analog Voltage Indication of LED Current. Bypass to SGND with a 0.1µF ceramic capacitor.
23	27	PWM	Dimming Control Input. Connect PWM to an external 3.3V or 5V PWM signal for PWM dimming. For analog-voltage-controlled PWM dimming, connect PWM to V_{CC} through a resistive voltage-divider with voltage between 0.2V and 3V. The dimming frequency is 200Hz under these conditions, and the duty cycle is (V_{PWM} - 0.2)/2.8. Connect PWM to SGND to turn off the LEDs. Connect PWM to V_{CC} for 100% duty cycle. Bypass PWM to SGND with a 0.1µF ceramic capacitor when using analog PWM.
24	28	UVEN	Undervoltage-Lockout (UVEN) Threshold/Enable Input. UVEN is a dual-function adjustable UVLO threshold input with an enable feature. Connect UVEN to V_{IN} through a resistive voltage-divider to program the UVLO threshold. The UVLO threshold is given by $V_{UVLO} = 1.23\frac{(R_{UVEN1} + R_{UVEN2})}{R_{UVEN2}}$
25	1	IN	Positive Power-Supply Input. Bypass IN to PGND with at least a 1µF ceramic capacitor.
26	2	INP	Positive Input for the Input Current Limit
27	3	INN	Negative Input for the Input Current Limit. Add an RC low-pass filter from INN to INP to provide a filtered DC voltage from INP to INN. The resistor should be 100Ω and the capacitor should be 0.1µF.
28	4	BST1	High-Side Power Supply for High-Side Gate Drive for *Buck* Section. Connect a 0.1µF ceramic capacitor from BST1 to LX1.
—	—	EP	Exposed Pad. Connect EP to a large-area contiguous copper ground plane for effective power dissipation. Do not use as the main IC ground connection. EP must be connected to SGND.

Functional Block Diagram



Detailed Description

The MAX25600 is a synchronous 4-switch buck-boost LED driver controller. The controller regulates the LED current for LED string voltages from 0V to 60V. The MAX25600 can be used as a seamless buck-boost LED driver for applications that require an efficient buck-boost LED driver with synchronous rectification. The MAX25600 is ideal for high-power applications that require a current source with PWM dimming capability.

The device provides seamless transition between buck, boost, and buck-boost modes depending on the ratio of input to output voltage. The MAX25600 is ideal for LED driver applications in automotive, industrial, and other LED lighting applications. A fault flag indicates open LED, shorted LED, or thermal-shutdown conditions. The device uses Maxim's proprietary average-current-mode control scheme and allows adjustable 200kHz to 700kHz fixed-frequency operation. In addition, ±6% triangular spread spectrum is added internally to the oscillator to improve EMI performance. The MAX25600 provides both analog and digital PWM dimming, and has built-in analog PWM dimming at a dimming frequency of 200Hz. Adjustable soft-start limits the current peaks and voltage overshoots at startup. The MAX25600 integrates a high-side p-channel dimming MOSFET driver for PWM dimming applications that require fast rising and falling edges of the LED current. It also features robust output open and short protection, is AEC qualified, and is suitable for automotive applications.

V_{CC} Regulator

The V_{CC} supply is the low-voltage analog supply for the chip and derives power from the input voltage from IN to PGND. An internal power-on-reset (POR) monitors the V_{CC} voltage and the IN voltage. A POR is generated when V_{CC} drops below its UVLO threshold, causing the IC to reset. The chip exits reset state once the input voltage goes back up and the V_{CC} linear regulator output is back in regulation.

Undervoltage Lockout

The MAX25600 features an adjustable UVLO using the enable input (UVEN). Connect UVEN to V_{IN} through a resistive divider to set the UVLO threshold. The MAX25600 is enabled when V_{UVEN} exceeds the 1.24V (typ) threshold. UVEN also functions as an enable/disable input to the device. Drive UVEN low to disable the output, and high to enable the output.

H-Bridge Operation

The H-bridge configuration using the MAX25600 is shown in Figure 1. The H-bridge consists of the four switches N1, N2, N3, and N4. Switches N1 and N2 are in series with the input voltage, and switches N3 and N4 are connected to the output. Inductor L is connected as shown. There are four different configurations in which the circuit operates, depending on the ratio of the input and output voltage.

Table 1 shows the status of the switches in the H-Bridge in each configuration.

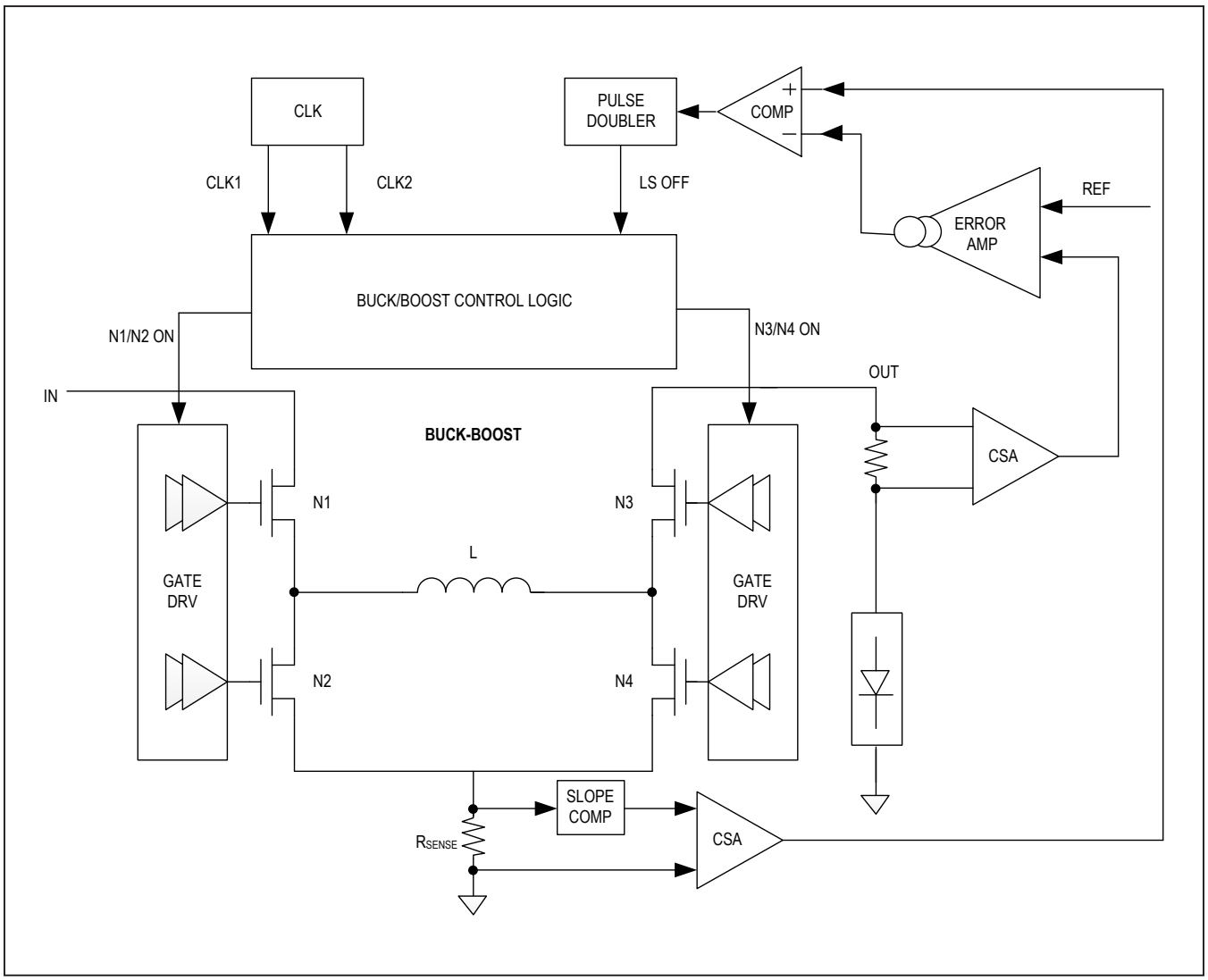


Figure 1. H-Bridge LED Driver

Table 1. Status of Switches in H-Bridge

SWITCH	BOOST MODE	BUCK-BOOST MODE (BOOST CONTROL)	BUCK-BOOST MODE (BUCK CONTROL)	BUCK MODE
N1	ON	PWM	PWM	PWM
N2	OFF	PWM	PWM	PWM
N3	PWM	PWM	PWM	ON
N4	PWM	PWM	PWM	OFF

Buck Mode

When the input voltage is much higher than the output voltage, then the MAX25600 operates in buck mode. In this configuration, switch N3 is always on and switch N4 is always off. Switch N2 is turned on at the beginning of the clock cycle (CLK1), and the inductor current ramps down. The MAX25600 uses an average-current-mode control scheme to determine the ON pulse width for switch N2. Once N2 is turned off, N1 is turned on. Switches N1 and N2 will alternate, behaving like a synchronous buck regulator.

Boost Mode

When the input voltage is much lower than the output voltage, then the MAX25600 operates in boost mode. In this configuration, switch N1 is always on and switch N2 is always off. Switch N4 is turned on at the beginning of the clock cycle (CLK2), and the inductor current ramps up. The MAX25600 uses an average-current-mode control scheme to determine the ON pulse width for switch N4. Once N4 is turned off, N3 is turned on. Switches N3 and N4 will alternate, behaving like a synchronous boost regulator.

Buck-Boost Mode

When V_{IN} is close to V_{OUT}, the MAX25600 operates in buck-boost configuration. In this configuration, all four switches have PWM voltages on the gates, and all four switches are switching at the switching frequency. There are two different configurations in the buck-boost mode.

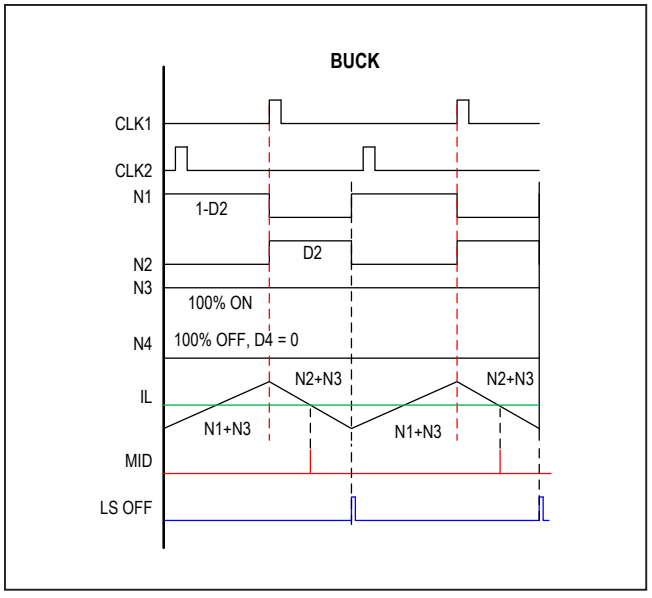


Figure 2. Buck-Mode Waveforms

When V_{IN} is slightly higher than V_{OUT}, the MAX25600 operates in the buck-boost region, where switch N2 is controlled by the PWM. Switch N4 is turned on for the beginning 16.7% cycle triggered by clock CLK2, and switch N3 is turned on for the remaining 83.3% cycle. Control of switch N2 is initiated by clock CLK1. N2 is turned on and N1 is turned off when CLK1 goes high. The MAX25600 uses average-current-mode control to determine the ON pulse width of N2. When N2 is turned off, N1 is turned on immediately

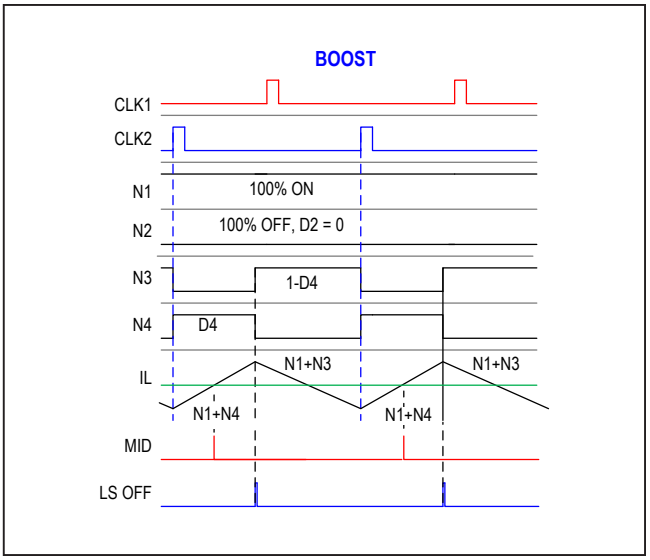


Figure 3. Boost-Mode Waveforms

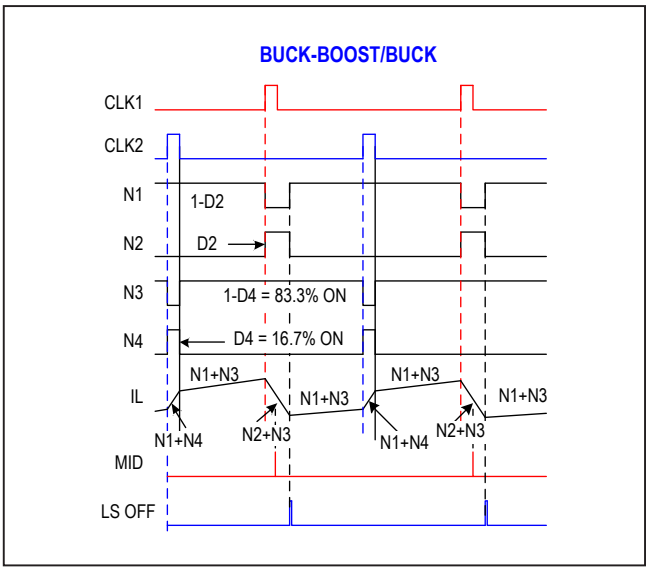


Figure 4. Buck-Boost Buck-Mode Waveforms

When V_{IN} is slightly lower than V_{OUT}, the MAX25600 operates in the buck-boost region where switch N4 is controlled by the PWM. Switch N2 is turned on for the beginning 16.7% cycle triggered by clock CLK1 and switch N1 is turned on for the remaining 83.3% cycle. Control of switch N4 is initiated by clock CLK2. N4 is turned on and N3 is turned off when CLK2 goes high. The MAX25600 uses average-current-mode control to determine the ON pulse width of N4. When N4 is turned off, N3 is turned on immediately.

Maximum Proprietary Average-Current-Mode Control

A novel average-current-mode control scheme is used in this current-mode buck-boost H-bridge converter. Instead of regulating the peak/valley current in buck/boost mode, average inductor current is regulated regardless of operating mode. As long as the inductor current is not changed abruptly during the mode transitions, the command signal remains at a nearly constant value unrelated to operating modes. As a result, seamless mode transition can be achieved. Since the converter is operating at a fixed switching frequency, additional slope compensation must be added to the inductor current-sense signal, which may require slight changes to the command signal to compensate for the error introduced by the slope compensation signal.

Average-Current-Mode Buck

When operating in buck mode, the pulse doubler controls the duty cycle of switch N2. The pulse width of switch N2 is 2x tpw.

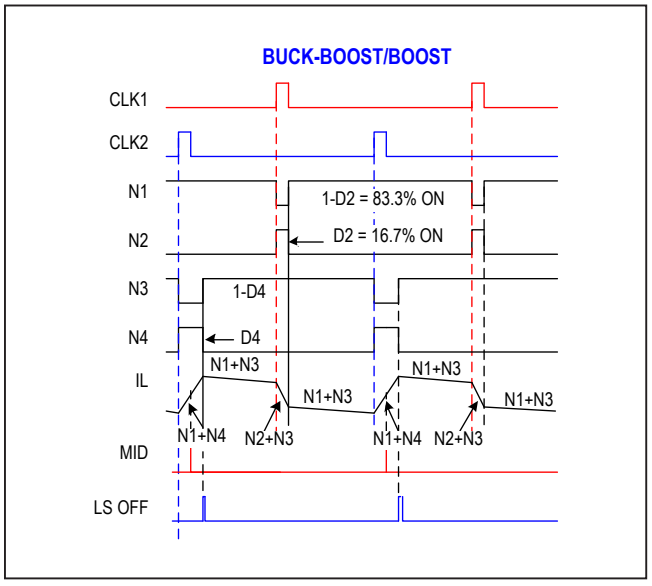


Figure 5. Buck-Boost Boost-Mode Waveforms

Average Current Mode Boost

When operating in boost mode, the pulse doubler controls the duty cycle of switch N4. The pulse width of switch N4 is 2x tpw.

Soft-Start

The SS pin can be used to program soft-start by connecting an external capacitor from SS pin to ground. An internal 15µA pullup current charges the capacitor on the SS pin, creating a voltage ramp. An internal diode from COMP to the SS pin clamps the voltage on the COMP pin. A ceramic capacitor of 0.1µF or higher is recommended on the SS pin.

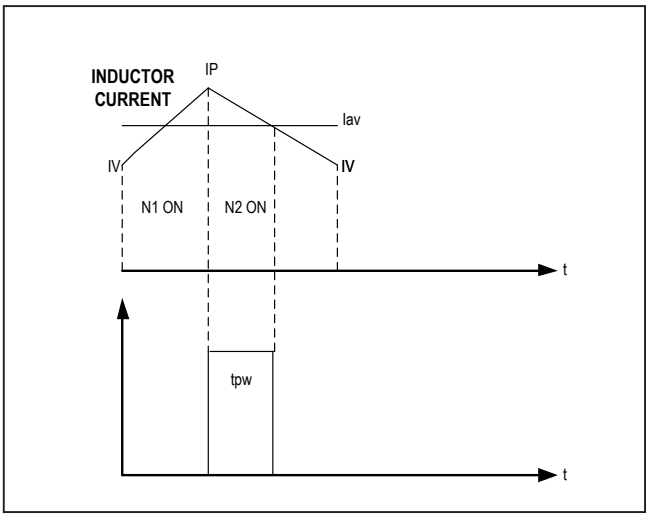


Figure 6. Pulse Doubler Buck

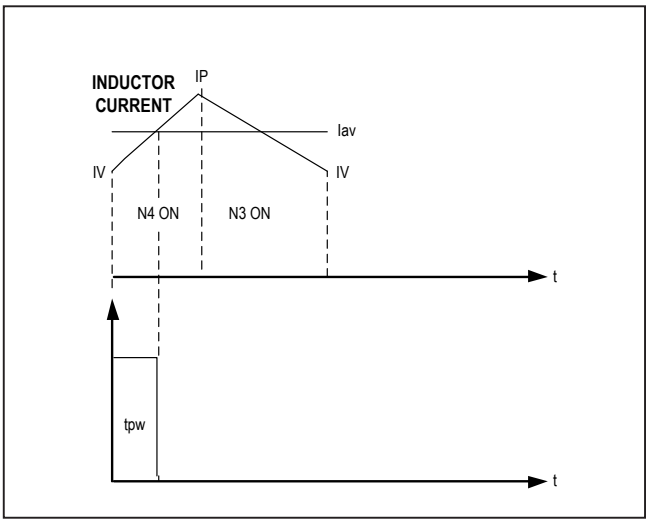


Figure 7. Pulse Doubler Boost

Switching Frequency

The internal oscillators of the MAX25600 are programmable from 200 kHz to 700 kHz using a single resistor at RT. Use the following formula to calculate the switching frequency:

$$f_{OSC}(kHz) = 20000/R_{RT}(k\Omega)$$

where R_{RT} is the resistor from RT to SGND.

The switching frequency oscillator in MAX25600 is synchronized to the leading edge of the PWM dimming pulse on input PWMDIM. The MAX25600 has built-in frequency dithering of ±6% of the programmed frequency to alleviate EMI problems.

Analog Dimming (ICTRL)

The MAX25600 offers an analog dimming-control input (ICTRL). The voltage at ICTRL sets the LED current level when VICTRL < 1.2V. The LED current can be linearly adjusted from zero with the voltage on ICTRL. For VICTRL > 1.3V, an internal reference sets the LED current. The maximum withstand voltage of this input is 5.5V. The LED current is guaranteed to be at zero when the ICTRL voltage is at or below 0.18V. The LED current can be linearly adjusted from zero to full scale for the ICTRL voltage in the range of 0.2V to 1.2V.

PWM Dimming (PWM)

In the MAX25600, the PWM functions with either analog or PWM control signals. Once the internal pulse detector detects three successive edges of a PWM signal with a frequency between 30Hz and 2kHz, the MAX25600 synchronizes to the external signal and pulse-width modulates the LED current at the external DIM input frequency with the same duty cycle as the PWM input. If an analog control signal is applied to the PWM pin, the MAX25600 compares the DC input to an internally generated 200Hz ramp to pulse-width-modulate the LED current (f_{DIM} = 200Hz). The output-current duty cycle is linearly adjustable from 0% to 100% (0.2V < V_{PWM} <3.0V). Use the following formula to calculate the voltage V_{PWM} necessary for a given output-current duty cycle D:

$$V_{PWM} = (D \times 2.8) + 0.2V$$

where V_{PWM} is the voltage applied to PWM pin, in volts.

Input-Current Limit

The MAX25600 features circuitry that limits the input current during line dropouts. If desired, this circuitry can be disabled by shorting INN and INP pins together. If DC input-current limiting is desired during low input voltages, then a current-sense resistor R_{IN} should be used. Use the circuit shown in Figure 8 to limit the input current.

An RC filter and a series resistor to INN should be used as shown. The input current is limited to IN_{MAX} where IN_{MAX} is given by the following equation:

$$IN_{MAX} = 0.1/R_{IN}$$

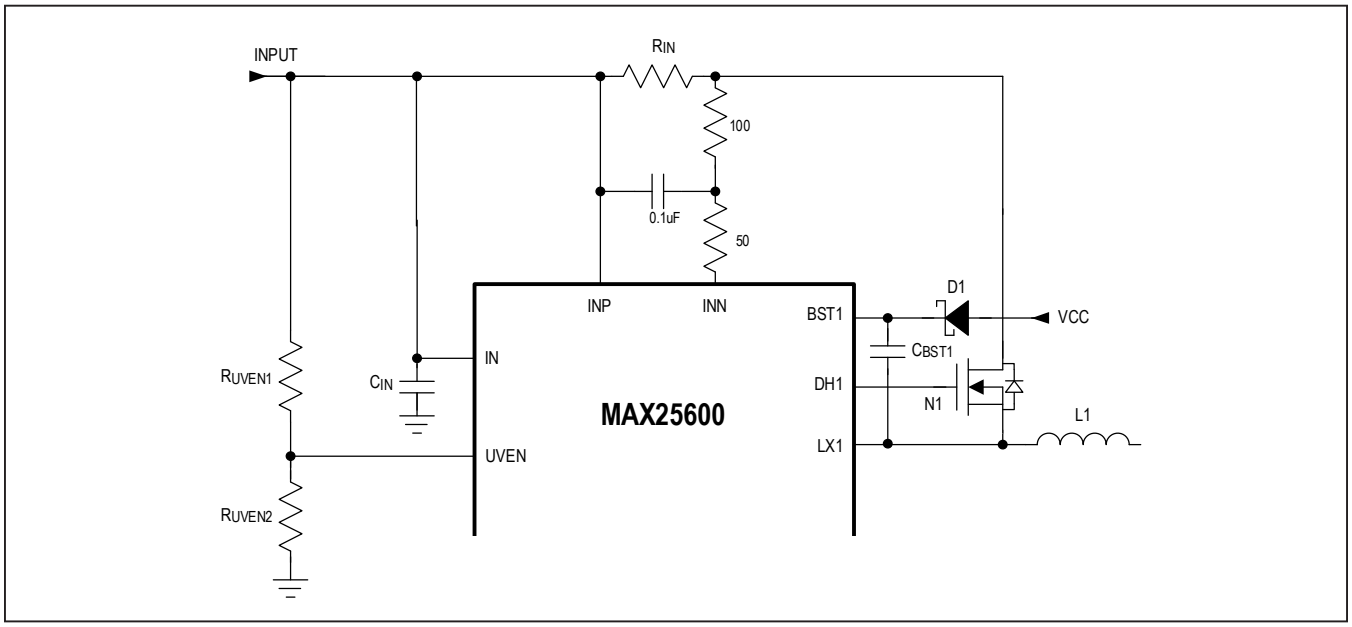


Figure 8. Input-Current Limit

Output-Current Monitor (IOUTV)

The MAX25600 includes a current monitor on the IOUTV pin. The IOUTV voltage is an analog voltage indication of the LED current when DIM is high. The voltage on the IOUTV pin is given by the following equation:

$$V_{IOUTV} = I_{LED} \times R_{CS\_LED} \times 5 + 0.2V$$

Control Loop and Error Amplifier

The sensed inductor current is controlled by the voltage on the COMP pin, which is the output of the error amplifier. The error amplifier has three inputs.

The control input ICTRL sets the LED current.

Overvoltage Protection (FB)

Pin FB sets the overvoltage-threshold limit across the LEDs. Use a resistive divider between ISP to FB and SGND to set the overvoltage-threshold limit. An internal overvoltage-protection comparator senses the differential voltage across FB and SGND. If the differential voltage is greater than 1.24V, the switching is turned off, $\overline{DIMOUT}$ goes high, and $\overline{FLT}$ asserts. When the differential voltage drops by 70mV, switching is enabled if PWM is high and $\overline{DIMOUT}$ goes low. $\overline{FLT}$ deasserts only if PWM is high and $V_{(ISP - ISN)}$ is > 20mV.

Fault Indicator $\overline{FLT}$

The MAX25600 features an active-low, open-drain fault indicator ($\overline{FLT}$). $\overline{FLT}$ asserts when one of the following conditions occur:

- Overvoltage or open across the LED string
- Short-circuit condition across the LED string
- Overtemperature condition

For overvoltage or open across the LED string, the $\overline{FLT}$ asserts only when an overvoltage occurs with the PWM in the high state. Once asserted, the $\overline{FLT}$ remains low and will only change state if PWM is high, the overvoltage condition is removed, and the voltage across the LED current-sense resistor is greater than 20mV. The $\overline{FLT}$ signal never changes state when PWM is low.

DC-DC Converter Application

The MAX25600 can also be used as a voltage regulator. Simplified typical circuit is shown in Figure 9.

In a typical application circuit, the PWM pin would be connected to V_{CC} and the LED string would be replaced by a resistor (R_{BOT}). The output voltage is available from ISP to ground. The programmed output voltage can be adjusted by controlling the voltage on ICTRL, or by adjusting the resistors R_{LED} and R_{BOT}. The MAX25600 controls the output voltage by regulating the voltage across R_{LED}.

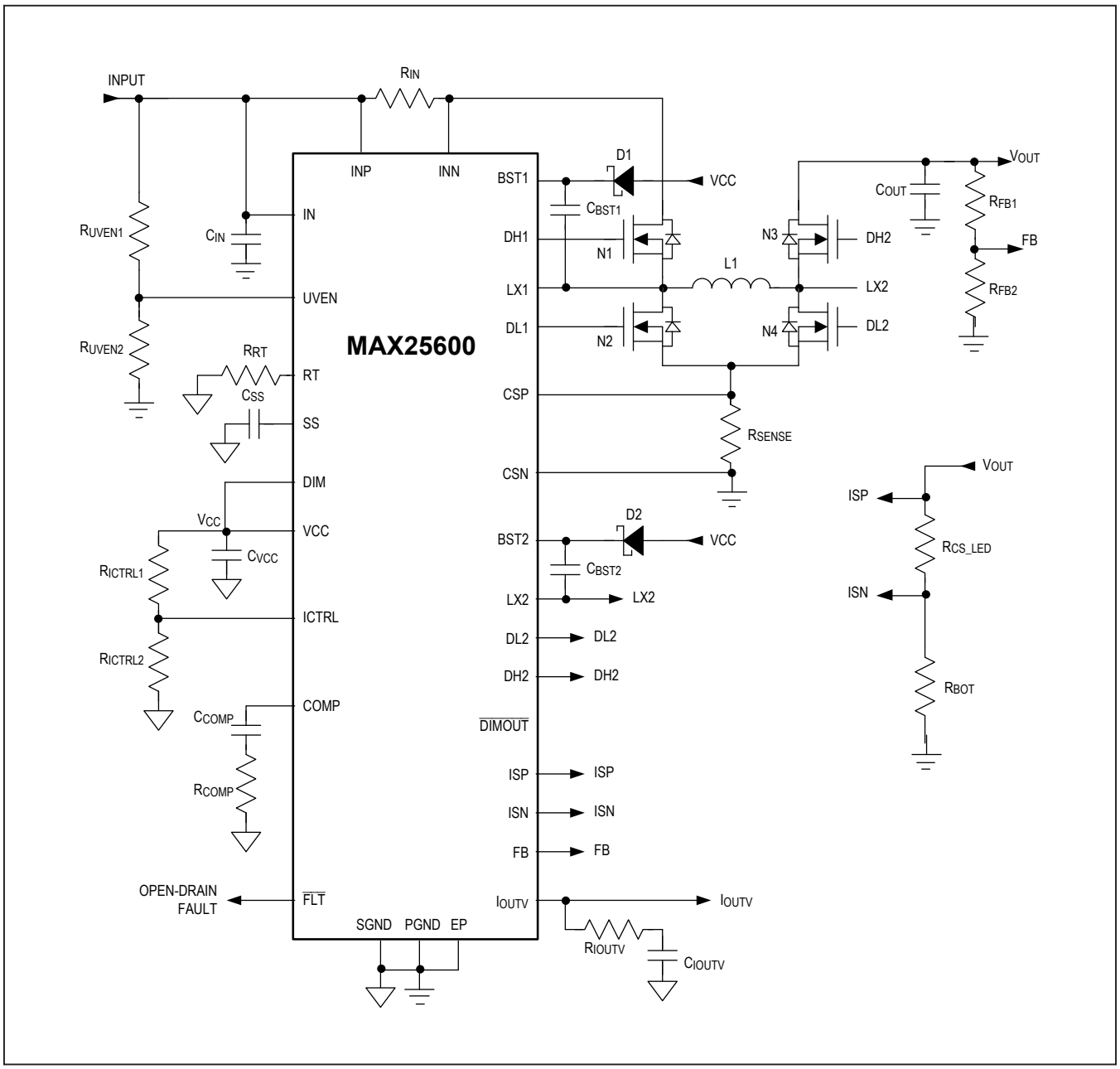


Figure 9. DC-DC Converter Application

Applications Information

Figure 10 shows a functional MAX25600 application circuit. External-component selection is driven by the input voltage range and LED string voltage and current requirements.

V_{CC} Regulator

The internal 5V regulator is used to power the internal control circuitry inside the MAX25600. This regulator can provide a load of 50mA to internal and external circuitry, and requires an external ceramic capacitor for stable operation. A 2.2µF ceramic capacitor is adequate for most applications. Place the ceramic capacitor close to the IC to minimize trace length to the internal V_{CC} pin and also to the IC ground. Choose a low-ESR, X7R ceramic capacitor for optimal performance. The IC powers up once the voltage on V_{CC} crosses the undervoltage lockout (V_{CC} UVLO) rising threshold and shuts down when V_{CC} falls below the (V_{CC} UVLO) falling threshold.

Programming Input UVLO Threshold

The input UVLO threshold is set by resistors R_{UVEN1} and R_{UVEN2} (see the Simplified Application Circuit). The MAX25600 turns on when the voltage across R_{UVEN2} exceeds 1.24V, the UVLO threshold. Use the following equation to set the desired UVLO threshold:

$$V_{UVEN} = 1.24 \times (R_{UVEN1} + R_{UVEN2})/R_{UVEN2}, \text{ in volts}$$

The UVEN pin can also be used as a separate enable pin where an external logic signal can switch the MAX25600 on and off.

Programming the LED Current

Normal sensing of the LED current should be done on the high side, where the LED current-sense resistor is connected to the anode of the LED string. The LED current is programmed using the resistor R_{LED} (see the *Simplified Application Circuit*). The LED current can also be programmed by adjusting the voltage on ICTRL when $V_{ICTRL} \leq 1.2V$ (analog dimming). The current is given by the following equation:

$$I_{LED} = (V_{ICTRL} - 0.2)/(5 \times R_{CS\_LED})$$

For voltages greater than 1.3V on the ICTRL pin, the LED current is clamped to the current given by the following equation:

$$I_{LED} = (1.3 - 0.2)/(5 \times R_{CS\_LED})$$

LED current can also be sensed on the ground side, if needed. In some applications, the LED current can be sensed by a current-sense resistor R_{LED} to ground.

Programming the Switching Frequency

The internal oscillator of the MAX25600 is programmable from 200kHz to 700kHz using a single resistor at RT. Use the following formula to calculate the value of the resistor

$$R_{RT}: R_{RT}(k\Omega) = 20000/f_{SW}(kHz)$$

where f_{SW} is the desired switching frequency, in kHz.

An additional ±6% spread spectrum is added internally to the oscillator to improve EMI performance.

Programming Input-Current Limit

The MAX25600 has an input current-sense amplifier that can be used to limit the input, as calculated by the following equation:

$$I_{IN} = 0.1/R_{IN}$$

A low-pass RC filter is needed for loop stability. For most applications, a 100Ω resistor R_F and a 100nF capacitor C_F is sufficient. An added 50Ω resistor R_{INN} in series with the INN pin should be added, as shown in Figure 10.

Setting the Overvoltage Threshold

The overvoltage threshold is set by resistors R_{OVP1} and R_{OVP2} (see the Simplified Application Circuit). The overvoltage circuit in the MAX25600 is activated when the voltage on FB with respect to GND exceeds 1.24V. Use the following equation to set the desired overvoltage threshold:

$$V_{OVP} = 1.24 \times (R_{OVP1} + R_{OVP2})/R_{OVP2}$$

Inductor Selection

In the boost converter, the average inductor current varies with the line voltage. The maximum average current occurs at the lowest line voltage. When operating the boost converter, the average inductor current is equal to the input current. Calculate maximum duty cycle using the following equation:

$$D_{MAX} = (V_{LED} - V_{INMIN})/V_{LED}$$

where V_{LED} is the forward voltage of the LED string, in volts, and V_{INMIN} is the minimum input supply voltage, in volts.

Use the following equations to calculate the maximum average inductor current (IL_{AVG}), peak-to-peak inductor-current ripple (ΔIL), and peak inductor current (I_{LP}):

Maximum average inductor current is given by

$$IL_{AVG} = I_{LED}/(1 - D_{MAX})$$

Allowing the peak-to-peak inductor ripple to be ΔIL, the peak inductor current is given by

$$IL_P = IL_{AVG} + 0.5 \times \Delta IL$$

The inductance value (L) of inductor L1, in Henrys (H), is calculated as

$$L = V_{INMIN} \times D_{MAX}/(f_{SW} \times \Delta IL)$$

where f_{SW} is the switching frequency in Hertz, V_{INMIN} is in volts, and ΔIL is in amperes.

Choose an inductor that has a minimum inductance greater than the calculated value. The current rating of the inductor should be higher than I_{LP} at the operating temperature.

When operating in buck mode, the average inductor current is the same as the LED current. The peak inductor current occurs at the maximum input line voltage where the duty cycle is at the minimum:

$$D_{MIN} = V_{LED}/V_{INMAX}$$

where V_{LED} is the forward voltage of the LED string, in volts, and V_{INMAX} is the maximum input supply voltage, in volts.

The peak inductor current is given by

$$I_{LP} = I_{LED} + 0.5 \times \Delta IL$$

The inductance value (L) of inductor L1, in Henrys, is calculated as

$$L = (V_{INMAX} - V_{LED}) \times D_{MIN}/(f_{SW} \times \Delta IL)$$

where f_{SW} is the switching frequency in Hertz, V_{INMAX} is in volts, and ΔIL is in amperes.

Choose an inductor that has a minimum inductance greater than the calculated value. The chosen inductor for the application should have an inductance that is the larger of the two calculated values from the boost and the buck configurations.

Input Capacitor Selection

The discontinuous input-current waveform of the buck converter causes large ripple currents in the input capacitor. The switching frequency, peak inductor current, and the allowable peak-to-peak voltage ripple reflected back to the source dictate the capacitance requirement. The input ripple consists of ΔVQ (caused by the capacitor discharge) and ΔV_{ESR} (caused by the ESR of the capacitor). Use low-ESR ceramic capacitors with high ripple-current capability at the input. A good starting point for selection of C_{IN} is to use an input-voltage ripple of 2% to 10% of V_{IN}. C_{IN_MIN} can be selected as follows:

$$C_{IN\_MIN} = 2(I_{LED} \times t_{ON})/\Delta V_{IN}$$

where t_{ON} is the on-time pulse width per switching cycle.

When selecting a ceramic capacitor, pay special attention to the operating conditions of the application. Ceramic capacitors can lose more than half of their capacitance at their rated DC-voltage bias and also lose capacitance with extremes in temperature. In applications with PWM dimming where the input connections to the source have wiring inductance, additional electrolytic capacitors may need to be added at the input to prevent large sags and surges in the input voltages during the PWM rising and falling edges. These line sags and surges can also cause the H-bridge to switch between boost, buck-boost, and buck configurations in one dimming cycle, which is undesirable and can cause flicker problems.

Output Capacitor Selection

The function of the output capacitor is to reduce the output ripple to acceptable levels. The ESR, ESL, and bulk capacitance of the output capacitor contribute to the output ripple. In most applications, the output ESR and ESL effects can be dramatically reduced by using low-ESR ceramic capacitors. To reduce the ESL and ESR effects, connect multiple ceramic capacitors in parallel to achieve the required bulk capacitance. To minimize audible noise generated by the ceramic capacitors during PWM dimming, it may be necessary to minimize the number of ceramic capacitors on the output. In these cases, an additional electrolytic or tantalum capacitor provides most of the bulk capacitance.

For simplicity, assume that the contributions from ESR and the bulk capacitance are equal, allowing 50% of the ripple for the bulk capacitance. The capacitance for boost configuration is given by

$$C_{BOOSTOUT} > (I_{LED} \times 2 \times D_{MAX})/(V_{OUTRIPPLE} \times f_{SW})$$

where I_{LED} is in amperes, C_{OUT} is in farads, f_{SW} is in Hertz, and $V_{OUTRIPPLE}$ is in volts.

The remaining 50% of allowable ripple is for the ESR of the output capacitor. Based on this, the ESR of the output capacitor is given by

$$ESR_{COUT} < (V_{OUTRIPPLE} \times V_{INMIN})/(2 \times V_{LED} \times I_{LED})$$

When operating in buck configuration, the required capacitance is given by

$$C_{BUCKOUT} > (V_{INMIN} - V_{LED}) \times V_{LED})/(V_{OUTRIPPLE} \times 2 \times L \times V_{INMAX} \times f_{SW} \times 2)$$

Based on this, the ESR of the output capacitor is given by

$$ESR_{COUT} < V_{OUTRIPPLE} \times L \times f_{SW})/(2 \times V_{LED} \times (1 - D_{MIN}))$$

H-Bridge Control Loop Current-Sense Selection (R_{SENSE})

The current-sense resistor on the low side of the H-bridge is chosen based on maximum LED current and the total output power. The control loop uses average-current-mode control in both boost and buck mode to control the H-bridge switches. When operating in boost mode, the maximum average load current at minimum input voltage

V_{INMIN} is given by $I_{OUT(BOOST\_MAX)}$. $I_{OUT(BOOST\_MAX)}$ at minimum input voltage is given by

$$I_{OUT(BOOST\_MAX)} = \frac{50\ mV}{R_{SENSE}} \times \frac{V_{INMIN}}{V_{OUT}}$$

When operating in buck mode, the maximum average inductor current is given by $I_{OUT(BUCK\_MAX)}$. The maximum current occurs at the maximum LED current in buck mode.

$$I_{OUT(BUCK\_MAX)} = \frac{50\ mV}{R_{SENSE}}$$

Slope Compensation

Slope compensation should be added to fixed-frequency converters operating in continuous-conduction mode with more than 50% duty cycle to avoid current-loop instability and sub-harmonic oscillations.

In the MAX25600, the slope-compensating ramp is added to the current-sense signal before it is fed to the PWM comparator. Connect a resistor (R_{SC}) from CS to the switch current-sense resistor terminal for programming the amount of slope compensation.

The device generates a current ramp with a slope of $50\mu A/t_{OSC}$ for slope compensation. The current-ramp signal is forced into an external resistor (R_{SC}) connected between CS and the source of the external MOSFET, thereby adding a programmable slope-compensating voltage (V_{SLOPE}) at the current-sense input CS. Therefore:

$$dV_{SLOPE})/dt = (R_{SC} \times 50\mu A)/t_{OSC}$$

The slope-compensation voltage that must be added to the current signal at minimum line voltage, with margin of 1.5x, is as follows:

Boost configuration:

$$V_{SLOPE} = D_{MAX} \frac{(V_{LED} - 2V_{INMIN}) \times R_{SENSE}}{L \times f_{SW}} \times 2 \times 1.5$$

Buck configuration:

$$V_{SLOPE} = D_{MAX} \frac{V_{LED} \times R_{SENSE}}{L \times f_{SW}} \times 2 \times 1.5$$

Control Loop Compensation

The LED current-control loop comprising the switching converter, LED current amplifier, and the error amplifier should be compensated for stable control of the LED current. For most applications the design needs to be stabilized for boost mode of operation and the buck mode should be automatically stable. The switching converter small-signal transfer function has a right half-plane (RHP) zero for the boost configuration as the inductor current is in continuous-conduction mode. The RHP zero adds a 20dB/ decade gain together with a 90° phase lag, which is difficult to compensate. The easiest way to avoid this zero is to roll off the loop gain to 0dB at a frequency less than 1/5 of the RHP zero frequency with a -20dB/decade slope.

The worst-case RHP zero frequency (f_{ZRHP}) is calculated as follows:

Boost configuration:

$$f_{ZRHP} = \frac{V_{LED} \times \left(1 - D_{MAX}\right)^2}{2\pi \times L \times I_{LED}}$$

The switching converter small-signal transfer function also has an output pole for the boost configuration. The effective output impedance that determines the output pole frequency together with the output filter capacitance is calculated as:

Boost configuration:

$$R_{OUT} = \frac{\left(R_{LED} + R_{CS\_LED}\right) \times V_{LED}}{\left(R_{LED} + R_{CS\_LED}\right) \times I_{LED} + V_{LED}}$$

where R_{LED} is the dynamic impedance of the LED string at the operating current.

The output pole frequency is calculated as follows:

$$f_P = \frac{1}{2\pi R_{OUT} C_{OUT}}$$

The feedback-loop compensation is done by connecting a resistor (R_{COMP}) and capacitor (C_{COMP}) in series from COMP to GND. R_{COMP} is chosen to set the high frequency integrator gain for fast transient response, while C_{COMP} is chosen to set the integrator zero to maintain loop stability. For optimum performance, choose the components using the following equations:

$$f_C = 0.2 \times f_{ZRHP}$$

$$R_{COMP} = \frac{\left(f_{ZRHP} \times R_{SENSE}\right)}{\left(f_p \times (1 - D_{MAX})\right) \times R_{CS\_LED} \times 5 \times G_M}$$

where G_M = 1.8mS (transconductance of error ampliflier), R_{SENSE} = Current-sense resistor on low side of H-Bridge, and R_{CS_LED} = LED current-sense resistor.

$$C_{COMP} = \frac{1}{2 \times 3.14 \times R_{COMP} \times 5 \times f_p}$$

Typical Application Circuits

MAX25600 in Buck-Boost LED Driver Application

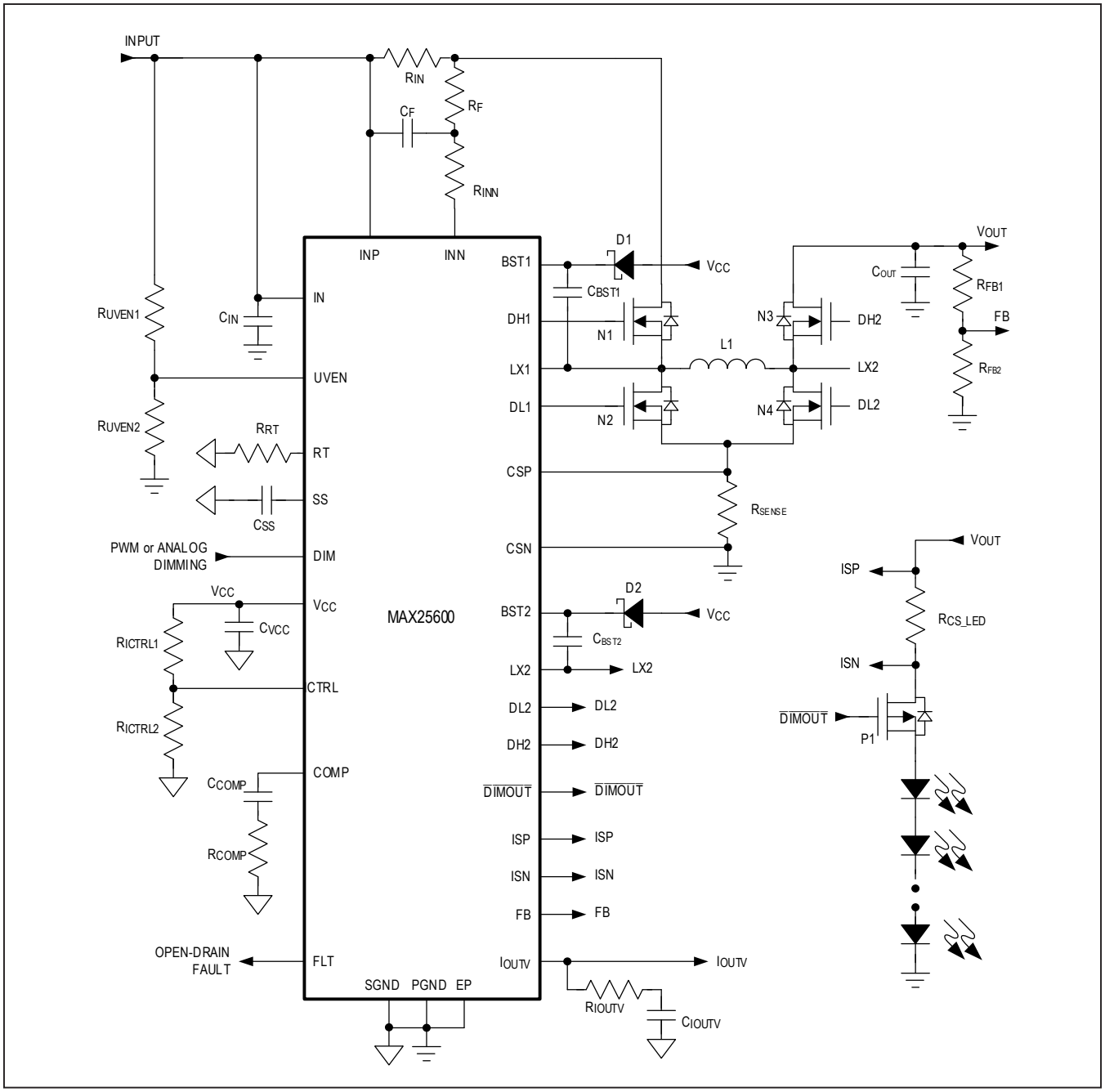


Figure 10. MAX25600 in Buck-Boost LED Driver Application

Ordering Information

PART	TEMP RANGE	PIN-PACKAGE
MAX25600ATI/VY+	-40°C to +125°C	28 TQFN-Cu
MAX25600AUI/V+	-40°C to +125°C	28 TSSOP-Cu

+ Denotes a lead(Pb)-free/RoHS-compliant package.
T Denotes tape-and-reel.
Y Denotes Side-Wettable

Revision History

REVISION NUMBER	REVISION DATE	DESCRIPTION	PAGES CHANGED
0	12/18	Initial release	—
1	1/19	Added future-product notation to MAX25600AUI/V+** in *Ordering Information* table	23
2	10/19	Updated *Electrical Characteristics*; updated *Applications Information;* removed future-product notation from MAX25600AUI/V+** in *Ordering Information* table	3–5, 21, 23

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