

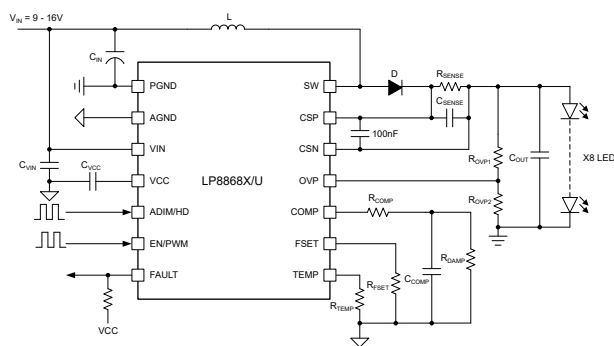
LP8868-Q1 Automotive Multi-Topology LED Driver with Inductive Fast Dimming

1 Features

- AEC-Q100 Qualified for automotive applications:
 - Temperature grade 1: -40°C to $+125^{\circ}\text{C}$, T_A
- Integrated MOSFET for buck, buck-boost and boost topology
 - Wide input voltage: 4.5 V to 65 V
 - Integrated 5.2-A and 150-m Ω MOSFET
 - Switching frequency 100 kHz to 2.2 MHz
 - Spread spectrum for lower EMI
- High precision power FET dimming
 - Up to 4-A output current in buck topology
 - Analog dimming (dimming ratio 256 :1)
 - Fast PWM dimming (150-ns pulse width)
 - Hybrid and flexible dimming
- Full protection features:
 - Fault output
 - LED open and short protection
 - Cycle-by-cycle current limit
 - Switching FET failure protection
 - Thermal shutdown
 - Configurable thermal foldback curve

2 Applications

- Automotive infotainment
- Automotive instrument clusters
- Heads-up displays(HUD)
- Automotive lighting



Typical Boost LED Driver Application Schematic

3 Description

The LP8868-Q1 family is a non-synchronous multi-topology solution with 4.5-V to 65-V wide input range. By integrating the low-side NMOS switch, the device is capable of driving LEDs with high power density and high efficiency. The family also supports common cathode connection and single layer PCB design. The switching frequency is configurable from 100 kHz to 2.2 MHz with an optional spread spectrum feature for better EMI performance.

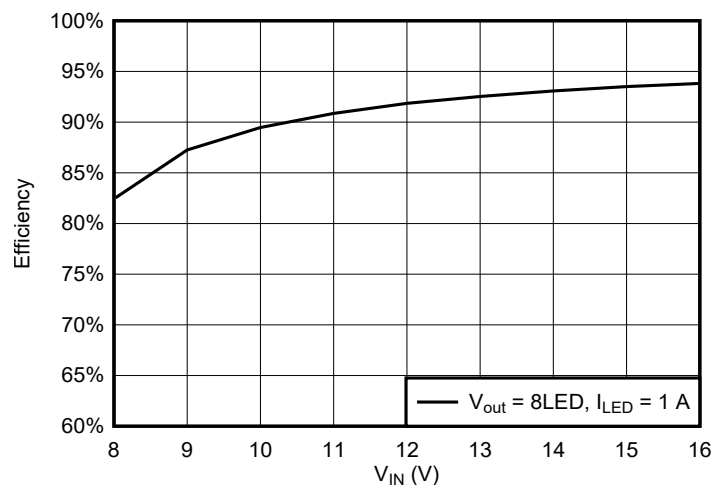
The LP8868-Q1 family supports four dimming options, including analog, PWM, hybrid and flexible dimming. Each dimming method can be configured through the PWM and ADIM input pins by means of simple high and low signals. The family adopts an adaptive off-time current mode control along with smart and accurate sampling to enable inductive fast dimming (IFD) and achieve high dimming accuracy.

The LP8868-Q1 family also provides multiple systematic protections, including LED open and short, sense resistor open and short, configurable thermal foldback and thermal shutdown. Fault output will send out acknowledge signals as soon as any fault condition is detected.

Package Information

PART NUMBER	PACKAGE ¹	BODY SIZE (NOM)
LP8868-Q1	VSON (14)	4.5 mm x 3.0 mm
	HVSSOP (12) ²	4 mm x 3.0 mm

1. For all available packages, see the orderable addendum at the end of the data sheet.
2. Product preview



Efficiency VS Input Voltage



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4 Comparison Table

Part Number	Topology	MOSFET Current Limit (Typical)	Spread Spectrum
LP8868XQDMTRQ1	Boost	6 A	Enabled
LP8868YQDMTRQ1	Buck-boost	6 A	Enabled
LP8868ZQDMTRQ1	Buck	6 A	Enabled
LP8868UQDMTRQ1	Boost	6 A	Disabled
LP8868VQDMTRQ1	Buck-boost	6 A	Disabled
LP8868WQDMTRQ1	Buck	6 A	Disabled

5 Pin Configuration and Functions

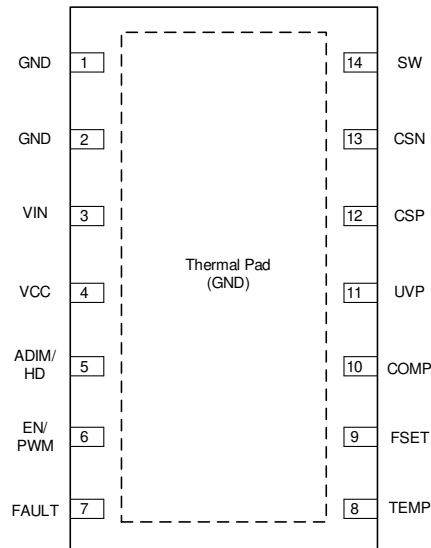
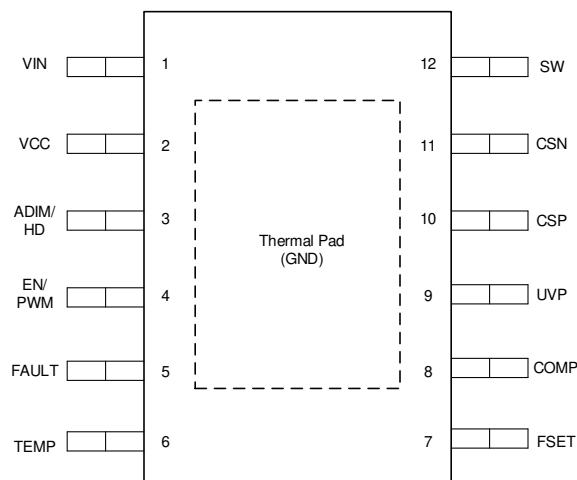


Figure 5-1. 14-Pin Buck VSON Top View



HVSSOP is product preview

Figure 5-2. 12-Pin Buck HVSSOP Top View

Table 5-1. Pin Functions

PIN		TYPE ⁽¹⁾	DESCRIPTION
No.	NAME		
1	PGND	G	Power ground pin.
2	AGND	G	Analog ground pin.
3	VIN	P	Input power pin.
4	VCC	P	Internal LDO output pin. Connect with a 10-V, 1-uF capacitor to GND.
5	ADIM/HD	I	Analog dimming or hybrid dimming pin. Pull high for PWM dimming only, pull low for hybrid dimming, input PWM signal for analog dimming.
6	PWM/EN	I	PWM dimming or EN pin. Pull high for always on, pull low for disabling the device, input PWM signal for PWM dimming.

Table 5-1. Pin Functions (continued)

PIN		TYPE ⁽¹⁾	DESCRIPTION
No.	NAME		
7	FAULT	O	Open drain output. Pull low when fault is detected.
8	TEMP	I/O	Thermal foldback pin. Put different resistor values to GND to set different thermal foldback behavior curves.
9	FSET	I/O	Switching frequency set pin, with range of 100 kHz ~ 2.2 MHz. Put different resistor values to GND for different switching frequencies.
10	COMP	I/O	Error-amplifier output. Connect capacitors to GND. Different capacitor values determine different softstart times and bandwidths.
11	UVP	I	Undervoltage detection pin. Put different resistor dividers to set the LED open detection thresholds.
12	CSP	I	LED current sense positive pin.
13	CSN	I	LED current sense negative pin.
14	SW	P	Switching node pin. Internally connected to the low-side MOSFET. Connect with the power inductor and the schottky diode.
Pad	Thermal Pad	G	Power ground pin.

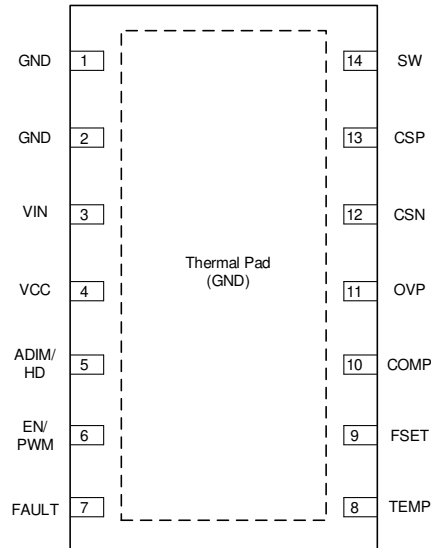


Figure 5-3. 14-Pin Boost/Buck-Boost VSON Top View

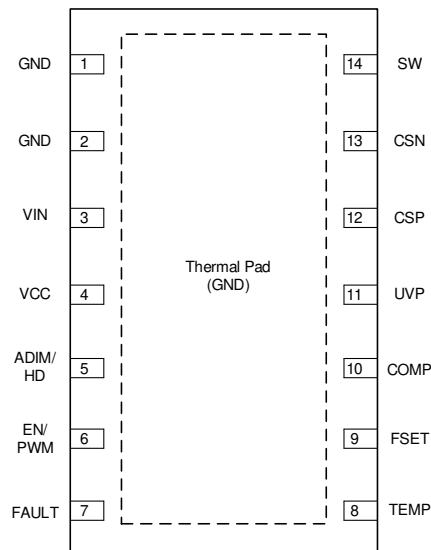


Figure 5-4. 12-Pin Boost/Buck-Boost HVSSOP Top View

HVSSOP is product preview

Table 5-2. Pin Functions for boost/buck-boost topology

PIN		TYPE ⁽¹⁾	DESCRIPTION
VSON Package	NAME		
1	PGND	G	Power ground pin.
2	AGND	G	Analog ground pin.
3	VIN	P	Input power pin.
4	VCC	P	Internal LDO output pin. Connect with a 10-V, 1- μ F capacitor to GND.
5	ADIM/HD	I	Analog dimming or hybrid dimming pin. Pull high for PWM dimming only, pull low for hybrid dimming, input PWM signal for analog dimming.

Table 5-2. Pin Functions for boost/buck-boost topology (continued)

PIN		TYPE ⁽¹⁾	DESCRIPTION
VSON Package	NAME		
6	PWM/EN	I	PWM dimming or EN pin. Pull high for always on, pull low for disabling the device, input PWM signal for PWM dimming.
7	FAULT	O	Open drain output. Pull low when fault is detected.
8	TEMP	I/O	Thermal foldback pin. Put different resistor values to GND to set different thermal foldback behavior curves.
9	FSET	I/O	Switching frequency set pin, with range of 100 kHz ~ 2.2 MHz. Put different resistor values to GND for different switching frequencies.
10	COMP	I/O	Error-amplifier output. Connect capacitors to GND. Different capacitor values determine different softstart times and bandwidths.
11	OVP	I	Overvoltage detection pin. Put different resistor dividers to set the LED open detection thresholds.
12	CSN	I	LED current sense negative pin.
13	CSP	I	LED current sense positive pin.
14	SW	P	Switching node pin. Internally connected to the low-side MOSFET. Connect with the power inductor and the schottky diode.
Pad	Thermal Pad	G	Power ground pin.

(1) I = Input, O = Output, P = Supply, G = Ground

6 Specifications

6.1 Absolute Maximum Ratings

over operating ambient temperature range (unless otherwise noted)⁽¹⁾

		MIN	MAX	UNIT
Voltage on pins	VIN, UVP, OVP, CSP, CSN, SW,	-0.3	65	V
Voltage on pins	VCC, ADIM/HD, EN/PWM, FAULT, TEMP, FSET, COMP	-0.3	5.5	V
Operation junction temperature	T _J	-40	125	°C
Storage temperature	T _{stg}	-65	150	°C

- (1) Stresses beyond those listed under *Absolute Maximum Ratings* may cause permanent damage to the device. These are stress ratings only, which do not imply functional operation of the device at these or any other conditions beyond those indicated under *Recommended Operating Conditions*. Exposure to absolute-maximum-rated conditions for extended periods may affect device reliability.

6.2 ESD Ratings

			VALUE	UNIT
V _(ESD)	Electrostatic discharge	Human-body model (HBM), per ANSI/ESDA/JEDEC JS-001 ⁽¹⁾	±2000	V
		Charged-device model (CDM), per JEDEC specification JESD22-C101 ⁽²⁾	±500	

- (1) JEDEC document JEP155 states that 500-V HBM allows safe manufacturing with a standard ESD control process.
 (2) JEDEC document JEP157 states that 250-V CDM allows safe manufacturing with a standard ESD control process.

6.3 Recommended Operating Conditions

over operating ambient temperature range (unless otherwise noted)

		MIN	MAX	UNIT
Input voltage range	VIN	4.5	63	V
Input voltage range	UVP, OVP, CSP, CSN	0	63	V
Input voltage range	VCC, ADIM/HD, EN/PWM, TEMP, FSET	0	5	V
Output voltage range	SW	0	63	V
	FAULT, COMP	0	5	V
Operating junction temperature, T _J		-40	125	°C

6.4 Thermal Information

THERMAL METRIC ⁽¹⁾		Device	UNIT
		SON	
		14 PINS	
R _{θJA}	Junction-to-ambient thermal resistance	39.1	°C/W
R _{θJC(top)}	Junction-to-case (top) thermal resistance	39.5	°C/W
R _{θJB}	Junction-to-board thermal resistance	14.7	°C/W
ψ _{JT}	Junction-to-top characterization parameter	0.9	°C/W
ψ _{JB}	Junction-to-board characterization parameter	14.7	°C/W

- (1) For more information about traditional and new thermal metrics, see the *Semiconductor and IC Package Thermal Metrics* application report, [SPRA953](#).

6.5 Electrical Characteristics

The electrical ratings specified in this section apply to all specifications in this document, unless otherwise noted. These specifications are interpreted as conditions that do not degrade the device parametric or functional specifications for the life of the product containing it. $T_J = -40^{\circ}\text{C}$ to $+125^{\circ}\text{C}$, $V_{IN} = 4.5\text{ V}$ to 60 V , (unless otherwise noted).

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
INPUT SUPPLY						
V_{VIN_UVLO}	V_{IN} undervoltage lockout	Rising V_{IN}	3.0	3.2	3.4	V
		Falling V_{IN}	2.8	3.0	3.2	V
	Hysteresis			0.2		V
I_{SD}	Shut down current from V_{IN}	$V_{IN} = 12\text{ V}$, $V_{EN/PWM} = 0\text{ V}$		0.8	2.3	μA
I_{OFF}	PWM off current from V_{IN}	$V_{IN} = 12\text{ V}$, $V_{EN/PWM} = 0\text{ V}$		2.5		mA
I_{OP}	Normal operating current	400-kHz switching frequency		4.6		mA
I_{OP}	Normal operating current	2.2-MHz switching frequency		10.0		mA
V_{VCC}	Internal LDO output voltage	$I_{VCC} = 10\text{ mA}$	5.0	5.15	5.3	V
I_{VCC_LIM}	Internal LDO output current limit		38	47	56	mA
DIMMING						
V_{PWM_L}	Low-level input voltage				0.4	V
V_{PWM_H}	High-level input voltage		1.2			V
V_{ADIM_L}	Low-level input voltage				0.4	V
V_{ADIM_H}	High-level input voltage		1.2			V
$t_{PWM_OUT_ON}$	PWM output minimum on time				150	ns
$t_{PWM_IN_ON}$	PWM input minimum on time				150	ns
$t_{PWM_IN_OFF}$	PWM input minimum off time to disable device		57		77	ms
f_{ADIM}	Analog Dimming input frequency	6-bit ADIM resolution	0.1		156	kHz
f_{ADIM}	Analog Dimming input frequency	8-bit ADIM resolution	0.1		39	kHz
FAULT						
V_{OL}	Output level low	$I = 3\text{ mA}$			0.1	V
$I_{LEAKAGE}$	Output leakage current	$V = 5\text{ V}$			1	μA
FEEDBACK AND ERROR AMPLIFIER						
$g_{M(ea)}$	Transconductance gain	ADIM 100% duty cycle, $V_{CSP-CSN} = 200\text{ mV}$, $V_{COMP} = 1.5\text{ V}$	205	265	325	$\mu\text{A/V}$
I_{COMP}	Source/sink current	ADIM 100% duty cycle, $V_{CSP-CSN} = 200\text{ mV} \pm 200\text{ mV}$, $V_{COMP} = 1.5\text{ V}$	± 24	± 40	± 56	μA
$V_{CSP-CSN}$	Current sense threshold	ADIM 100% duty cycle	194	200	206	mV
$V_{CSP-CSN}$	Current sense threshold	ADIM 12.5% duty cycle, compared with 100% duty cycle	11.875	12.5	13.125	%
$V_{CSP-CSN}$	Current sense threshold	ADIM 1.17% duty cycle, compared with 100% duty cycle	0.82	1.17	1.52	%
$I_{LEAK_CSP/N}$	CSP+CSN pin leakage current	$V_{IN} = 60\text{ V}$, $V_{EN/PWM} = 5\text{ V}$	22		31	μA
$I_{LEAK_CSP/N}$	CSP+CSN pin leakage current	$V_{IN} = 60\text{ V}$, $V_{EN/PWM} = 0\text{ V}$	10		15	μA
POWER STAGE						
R_{DSON}	Switching FET on resistance	$V_{IN} \geq 5\text{ V}$		150		m Ω
t_{min_ON}	Switching FET minimum on time			100		ns
t_{min_OFF}	Switching FET minimum off time			100		ns
f_{SW}	Switching FET frequency		0.1		2.2	MHz
CURRENT LIMIT						
I_{LIM}	Switching FET cycle-by-cycle current limit (LP8868X/LP8868Y/LP8868U/LP8868V)		5.8	6.5	7.6	A
I_{LIM}	Switching FET cycle-by-cycle current limit (LP8868Z/LP8868W)		5.2	6	7	A
THERMAL PROTECTION						
T_{th}	Thermal foldback starting temperature threshold	$R_{TEMP} = 20\text{ k}\Omega$		130		$^{\circ}\text{C}$

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PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
T_{TSD}	Thermal shutdown temperature			165		$^{\circ}\text{C}$
	Hysteresis			15		$^{\circ}\text{C}$

6.6 Typical Characteristics

$V_{IN} = 12\text{ V}$, $L = 22\ \mu\text{H}$, $F_{SW} = 400\text{ kHz}$, unless otherwise specified

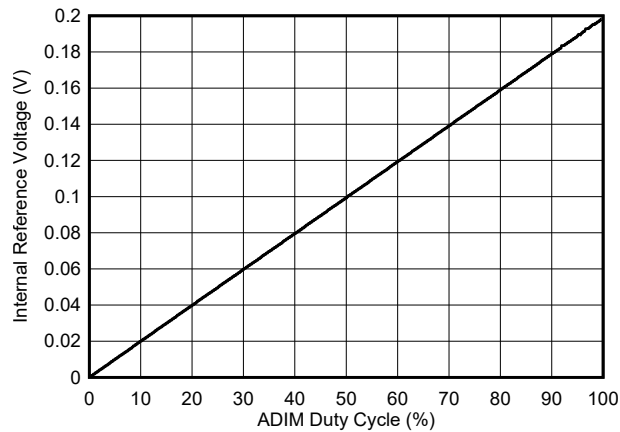


Figure 6-1. ADIM Duty Cycle vs. Reference Voltage in Analog Dimming

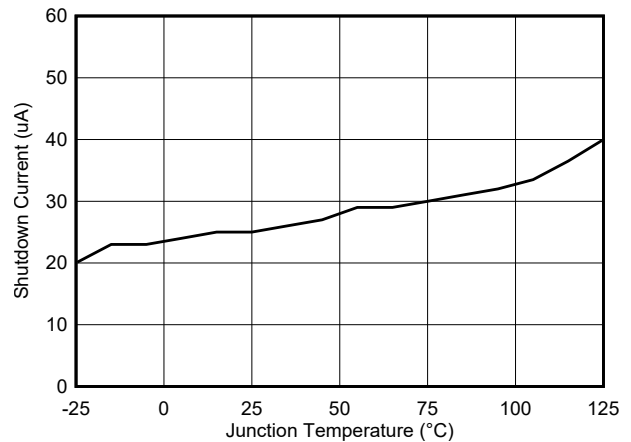


Figure 6-2. Shutdown Current vs. Junction Temperature

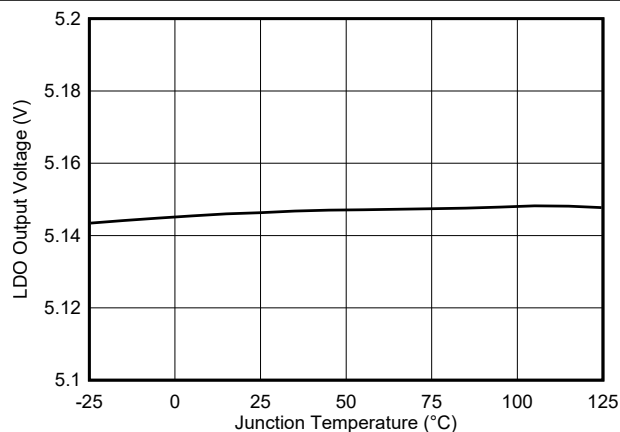


Figure 6-3. Internal LDO Output vs. Junction Temperature

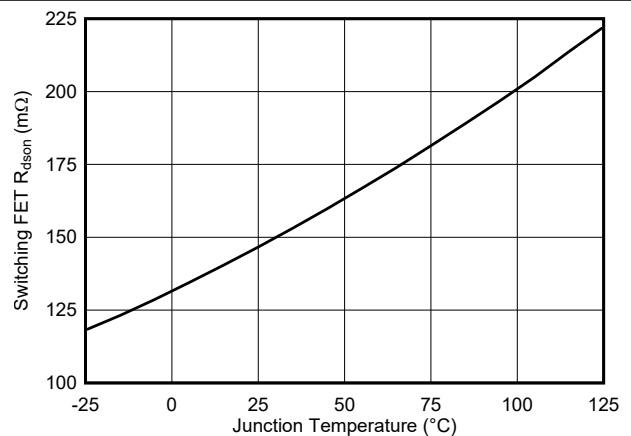


Figure 6-4. Switching FET R_{dson} vs. Junction Temperature

6.6 Typical Characteristics (continued)

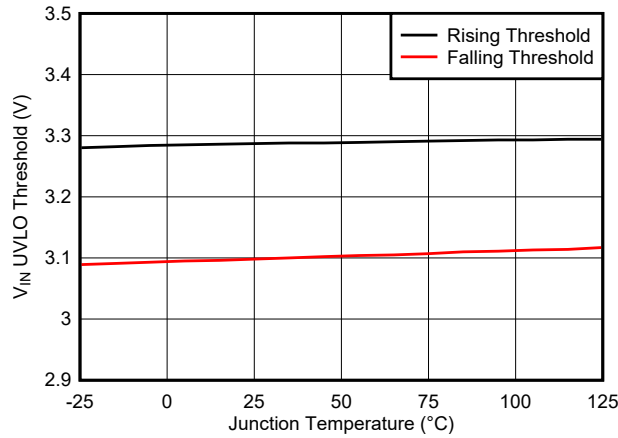


Figure 6-5. VIN UVLO Threshold vs. Junction Temperature

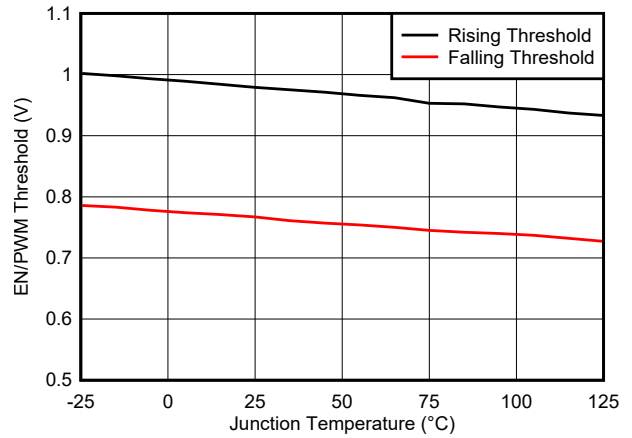


Figure 6-6. EN/PWM Threshold vs. Junction Temperature

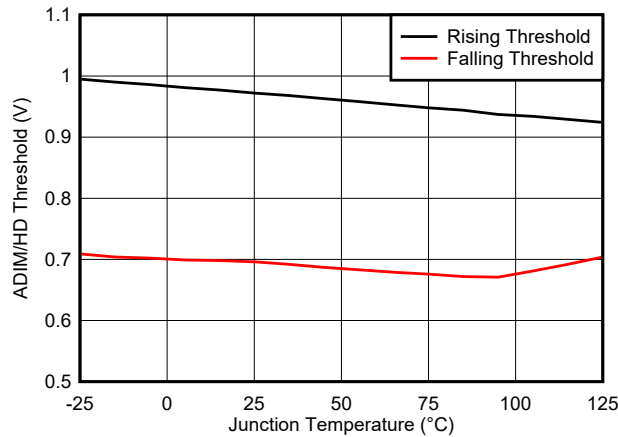


Figure 6-7. ADIM/HD Threshold vs. Junction Temperature

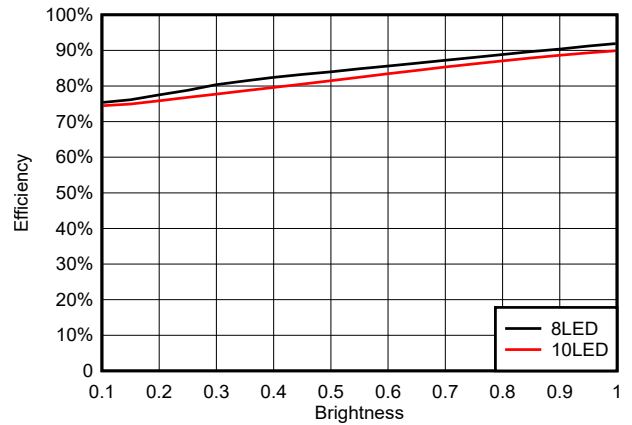


Figure 6-8. Efficiency at 1-A Max Output Current with PWM Dimming, 22uH Inductor, Boost Topology

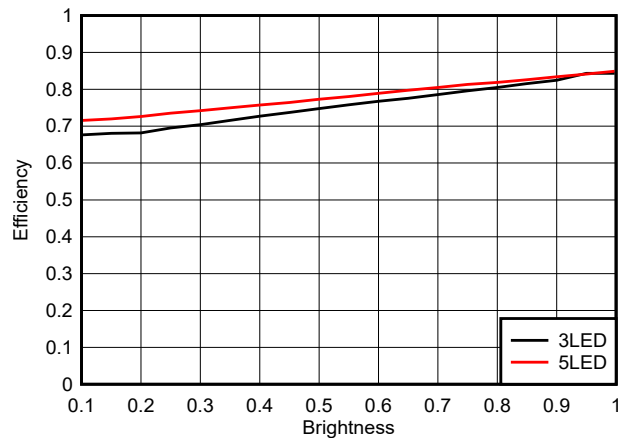


Figure 6-9. Efficiency at 1-A Max Output Current with PWM Dimming, 22uH Inductor, Buck-Boost Topology

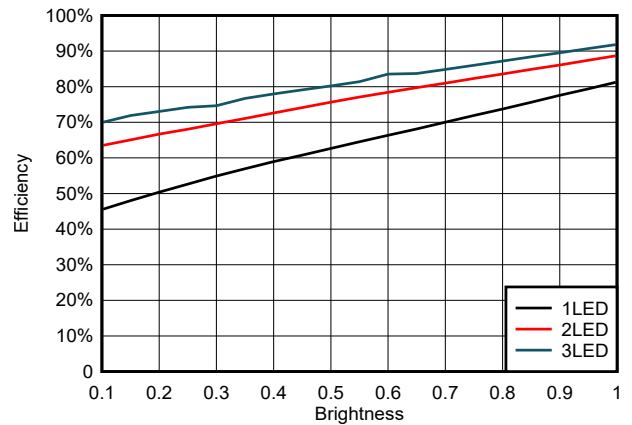


Figure 6-10. Efficiency at 3-A Max Output Current with PWM Dimming, 22uH Inductor, Buck Topology

6.6 Typical Characteristics (continued)

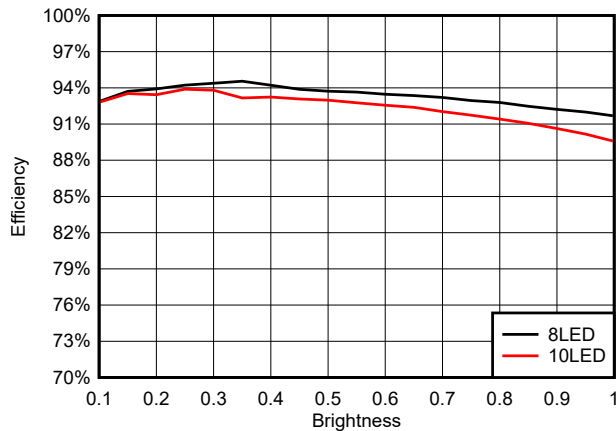


Figure 6-11. Efficiency at 1-A Max Output Current with Analogy Dimming, 22uH Inductor, Boost Topology

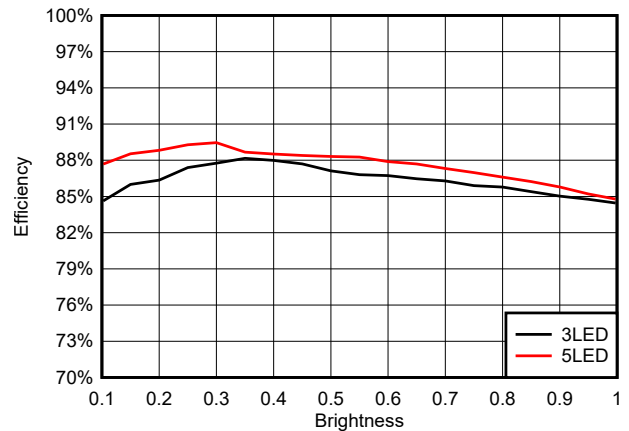


Figure 6-12. Efficiency at 1-A Max Output Current with Analogy Dimming, 22uH Inductor, Buck-Boost Topology

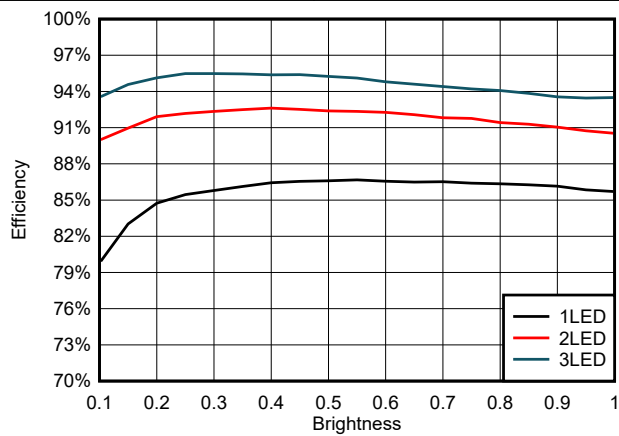


Figure 6-13. Efficiency at 3-A Max Output Current with Analogy Dimming, 22uH Inductor, Buck Topology

7 Detailed Description

7.1 Overview

The LP8868-Q1 family is a 4-A non-synchronous Buck / Boost / Buck-Boost LED driver with 4.5-V to 65-V wide input range. By integrating the low-side NMOS switch with constant current and constant voltage controls, the device is capable of not only driving LEDs but also charging batteries with high power density and high efficiency.

The switching frequency is configurable through FSET pin, ranging from 100 kHz to 2.2 MHz, with optional spread spectrum feature to decrease the EMC emission and reduce the input filter size. The device supports four dimming options, including analog dimming, PWM dimming, hybrid dimming and flexible dimming. Each dimming method can be configured through the PWM and ADIM input pins by means of simple high/low sequencing signals at startup.

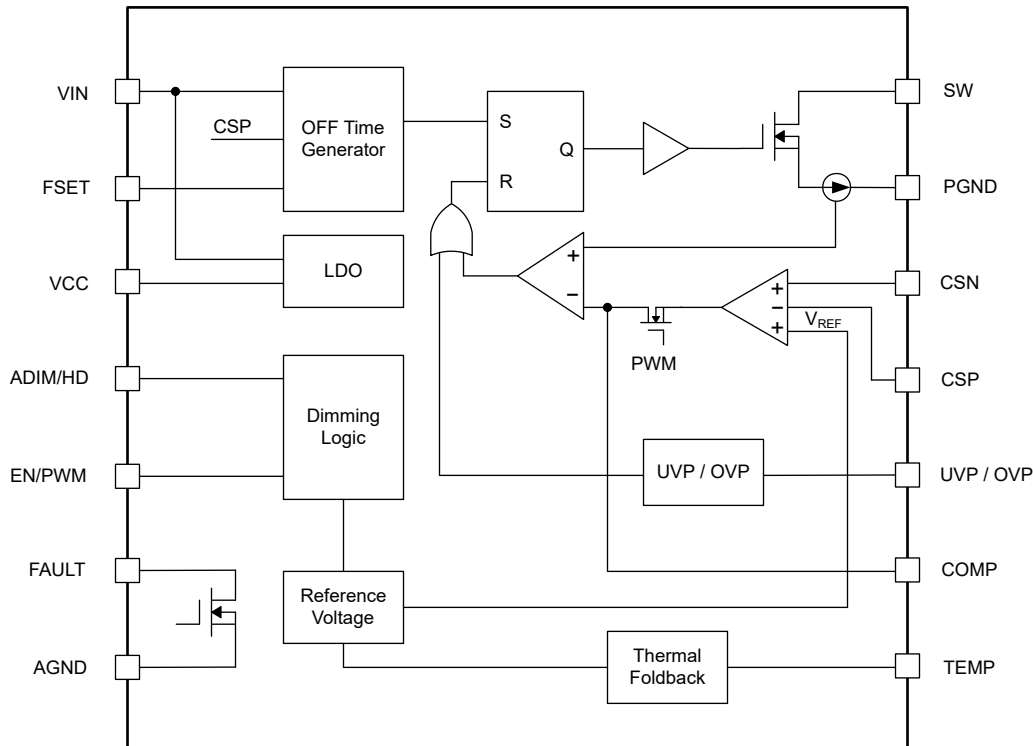
The device adopts an adaptive off-time current mode control along with smart and accurate sampling to enable inductive fast dimming (IFD) and achieve high dimming ratio. The compensation bandwidth can be adjusted through an external capacitor on COMP pin based on system requirement.

The LP8868-Q1 family has extensive fault detection feature:

- LED open and short detection
- Sense resistor open and short detection
- Configurable thermal foldback and thermal shutdown protection

Fault condition is indicated through the FAULT output pin.

7.2 Functional Block Diagram



7.3 Feature Description

7.3.1 Adaptive Off-Time Current Mode Control

The LP8868-Q1 family adopts an adaptive off-time current mode control to support fast transient response over a wide range of operation. The switching frequency is configurable through FSET pin, ranging from 100 kHz to 2.2 MHz.

For average output current regulation, the sensed voltage across the sensing resistor between the CSP and CSN pins is compared with the internal voltage reference, V_{REF} , through the error amplifier. The output of the error amplifier, V_{COMP} , passes through an external compensation network and is then compared with the peak current feedback at the PWM comparator

During each switching cycle, when the internal N-MOSFET is turned on, the peak current is sensed through the internal FET. When the sensed value of peak current reaches V_{COMP} at the input of PWM comparator, the N-MOSFET is turned off and the adaptive off-time counter starts counting. Once the adaptive off-time counter stops counting, the counter keeps reset until when the N-MOSFET turns off. The counting off time is determined by the external resistor connected to the FSET pin and the input/output feedforward. Thus, the device is able to maintain a nearly constant switching frequency at steady state and regulate the output average current at a desired value.

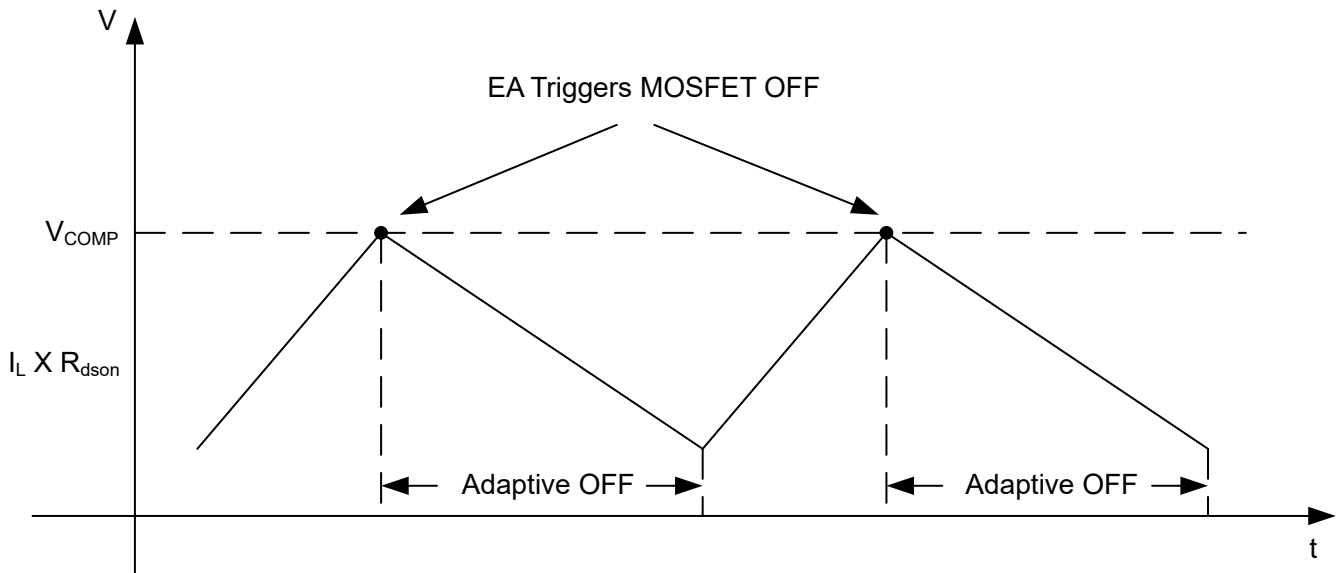


Figure 7-1. Adaptive off-time current mode control method

7.3.1.1 Switching Frequency Settings

The switching frequency of LP8868 family is adjustable from 100 kHz to 2.2 MHz by means of changing R_{FSET} connected between FSET pin and AGND. The default switching frequency is 100 kHz when the FSET pin is connected to nothing.

The resistor value and the corresponding switching frequency are listed in the below table:

Table 7-1. Switching Frequency vs. R_{FSET} resistor value

Switching Frequency	Resistor Value (k Ω)
100 kHz	232
200 kHz	138
300 kHz	83
400 kHz	59
600 kHz	38
800 kHz	28
1 MHz	23
1.2 MHz	18
1.5 MHz	13
1.8 MHz	11
2.2 MHz	9

For example, if R_{FSET} is set to 59 k Ω , the corresponding switching frequency is 400 kHz.

In most cases, lower switching frequency has higher system efficiency and better thermal performance.

7.3.1.2 Spread Spectrum

The LP8868X/Y/Z-Q1 enables the spread spectrum feature ($\pm 7\%$ from central frequency, 2-kHz modulation frequency) which reduces EMI noise at the switching frequency and its harmonic frequencies.

On the other hand, the LP8868U/V/W-Q1 disables the spread spectrum feature toward better brightness performance in low brightness scenario.

7.3.2 Setting LED Current

The output current of the LED is controlled with external resistor R_{sense} between CSP and CSN pins. R_{sense} value for the target current can be calculated using equation [Equation 1](#). Noted that, to relised IFD function and improve the accuracy of the output current in low duty cycle, the capacitor in parallel with sense resistor is required for boost and buck-boost topology. And it is optional for buck topology.

$$R_{SENSE} = \frac{V_{REF}}{I_{LED_FS}} \quad (1)$$

where

- $V_{REF} = 200$ mV
- R_{SENSE} is current setting resistor, m Ω
- I_{LED} is output current, mA

For example, if R_{sense} is set to 100 m Ω . I_{LED} will be 2 A.

7.3.3 Internal Soft Start

The LP8868-Q1 family implements the internal soft-start function. Once V_{IN} rises above V_{VIN_MIN} , the internal LDO starts to charge V_{CC} capacitor. It takes approximately 800 μs for V_{CC} to rise above V_{VIN_UVLO} if a 1- μF capacitor is connected to V_{CC} pin. If EN/PWM pin is pulled high before V_{CC} rises above V_{VIN_UVLO} , the POR is enabled right after V_{CC} above V_{VIN_UVLO} and waits for 100 μs to start dimming mode. EN/PWM pin has to stay high for more than 5 μs after V_{CC} rises above V_{VIN_UVLO} . In this case, if using 1- μF V_{CC} capacitor, it is recommended to wait for 1 ms to start dimming mode after V_{IN} rises above V_{VIN_MIN} .

If EN/PWM pin has the first PWM pulse appearing after V_{CC} rises above V_{VIN_UVLO} , the device waits for 200 μs to enable POR and another 100 μs to start dimming mode. Hence, without triggering V_{IN} UVLO, the device can be renabled after disabled and waits for 300 μs to start dimming mode. Note that the initial enable PWM pulse lasting more than 5 μs is required at EN/PWM input pin to enable the device. After dimming mode is started, the device enters four different dimming modes based on the configuration of ADIM/HD pin and EN/PWM pin.

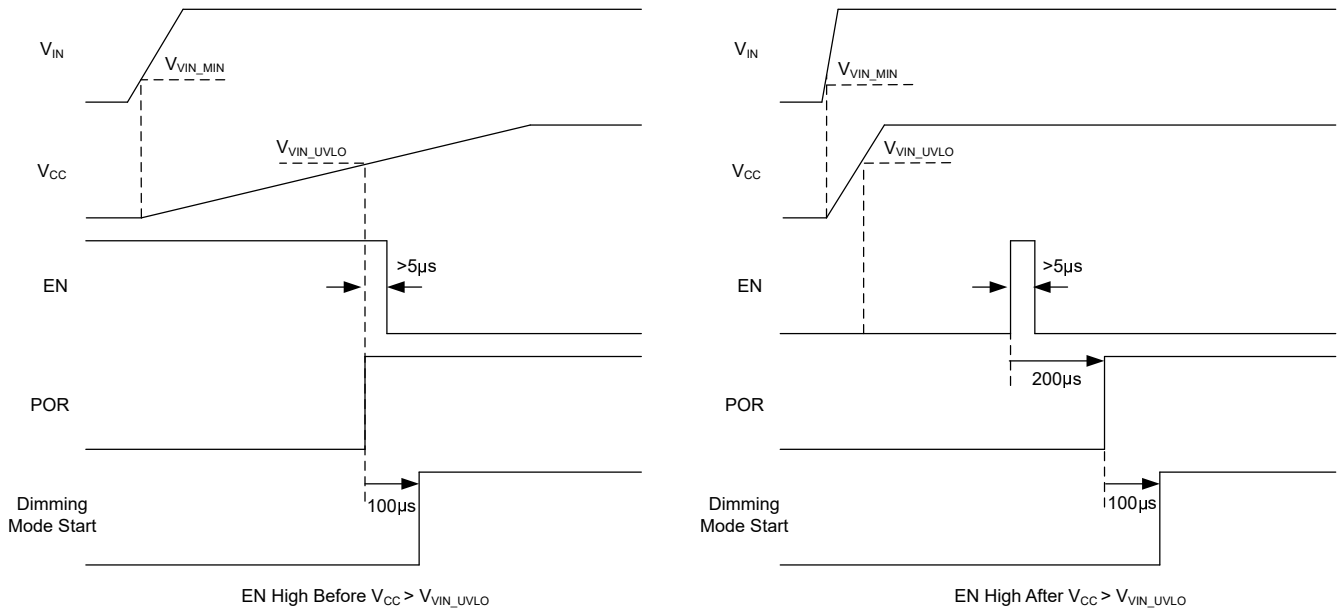


Figure 7-2. Startup Sequence

7.3.4 Dimming Mode

The LP8868 family has four optional dimming modes determined by the waveform in PWM and ADIM pins. The dimming mode is started either 1 ms after VIN exits UVLO or 300 μs after renable by EN/PWM pin. The configuration to one of the four dimming modes are shown as below.

Table 7-2. Dimming Mode Configuration

Dimming Mode	EN/PWM Pin	ADIM/HD Pin
PWM Dimming	PWM signal	High
Analog Dimming	High	PWM signal
Hybrid Dimming	PWM signal	Low
Flexible Dimming	PWM signal	PWM signal

7.3.4.1 PWM dimming

The PWM dimming mode is enabled when the ADIM/HD input pin is always high and the PWM/EN input pin is configured by a PWM input signal. Device supports PWM input signals with ultra-narrow pulse width down to 200 ns in PWM dimming mode. The PWM output duty cycle can be changed in the opposite direction only when PWM input duty cycle changes by more than 0.38%.

In PWM dimming mode, when the PWM input signal at the PWM pin turns from low to high, the internal NMOS FET starts switching and the inductor current rises to the determined value. The LED current is then regulated at the determined value as long as the PWM input signal stays high. When the PWM input signal turns from high to low, the internal FET is turned off causing the inductor current falling to zero. The internal FET maintains off and the LED current stays zero if the PWM input signal stays low.

7.3.4.2 Analog dimming

The LP8868 family supports analog dimming which regulates the LED current through the ADIM/HD pin. The analog dimming mode is enabled when the PWM/EN pin is always high and the ADIM/HD pin is configured by a PWM input signal. And the internal digital circuits is able to respond to the duty cycle change of the PWM input signal with tens of micro-seconds delay

The internal voltage reference, V_{REF} , changes in proportion to the duty cycle of the PWM input signal at the ADIM/HD pin. V_{REF} is 200 mV when the PWM input signal at the ADIM/HD pin has a 100% duty cycle, for instance, and V_{REF} is 20 mV when the PWM input signal has a 10% duty cycle.

7.3.4.3 Hybrid Dimming

The LP8868 family supports a unique hybrid dimming function to maximize the dimming performance, especially when both high dimming frequency and high dimming ratio are needed. The hybrid dimming mode is enabled when the ADIM/HD pin is always low and the PWM/EN pin is configured by a PWM input signal.

When the hybrid dimming is enabled, the LED current is regulated by the analog dimming at high brightness level (12.5% - 100%) and by the PWM dimming at low brightness level (0% - 12.5%), respectively. At high brightness level, the internal voltage reference, V_{REF} , changes in proportion to the duty cycle of the PWM input signal at the PWM/EN pin. At low brightness level, V_{REF} stays unchanged and an internal PWM generator is enabled. Thus, the LED is turned on and off corresponding to the on and off of the internal PWM signal of which the frequency and the duty cycle are configured by the PWM input signal at the PWM/EN pin. The detailed hybrid dimming behavior is illustrated in the below figure.

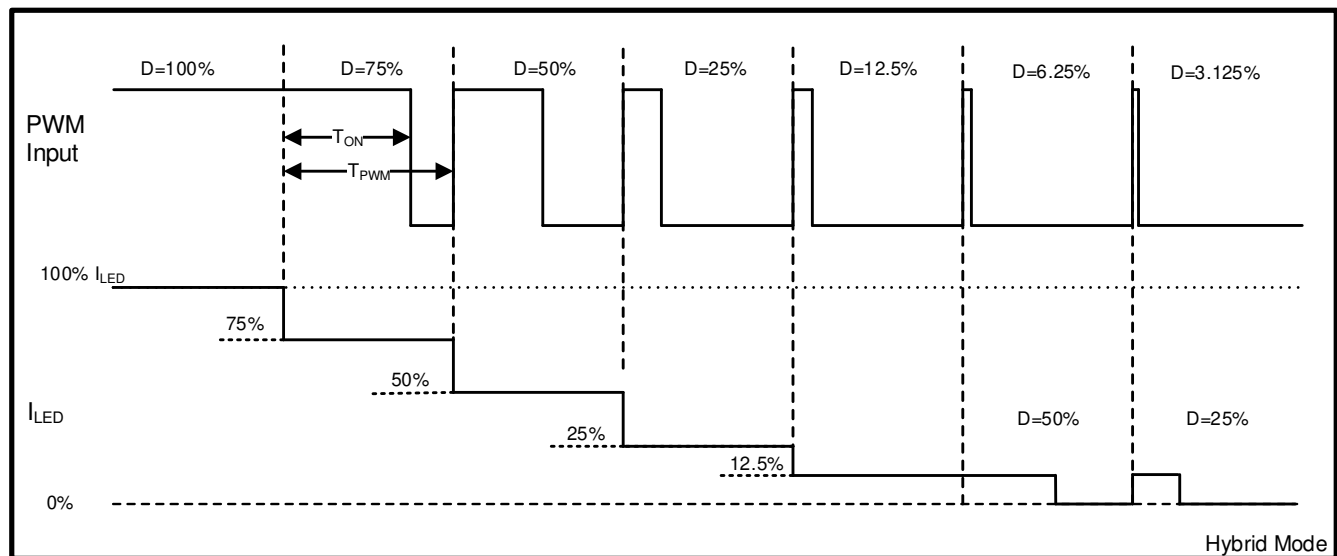


Figure 7-3. Hybrid Dimming

7.3.4.4 Flexible Dimming

The LP8868 family also supports flexible dimming to maximize the flexibility of dimming control, in which the LED current value and the on/off behavior can be controlled independently. The flexible dimming mode is enabled when both the ADIM/HD pin and the PWM/EN pin are configured by PWM input signals at the same time. Therefore, in flexible dimming mode, the LED is turned on and off corresponding to the on and off of the PWM input signal at the PWM/EN pin while the reference voltage changes in proportion to the duty cycle of the PWM input signal at the ADIM/HD pin.

7.3.5 Undervoltage Lockout

The LP8868 family implements an internal undervoltage-lockout (UVLO) circuitry connecting to the VCC pin. The UVLO is triggered and then device is disabled when the VCC pin voltage falls below the internal UVLO threshold voltage, V_{VCC_UVLO} typically 3.0 V, with a typical 0.2-V hysteresis. The VCC pin is the output of an internal regulator of which the input is supplied by the VIN pin. Therefore, if VIN pin voltage falls close to above the V_{VCC_UVLO} (around 500 mV above), the UVLO will be triggered.

7.3.6 Fault Protection

The LP8868 family is able to provide fault protections and send fault report signals in many fault conditions, including LED open, LED short to GND, sense resistor open and short, internal switching FET fault and thermal shutdown. The fault criterion for different topology is shown in below.

Table 7-3. Protections in Buck topology

TYPE	CRITERION	BEHAVIOR
LED open load	$V_{UVP} < 1.2 \text{ V}$ for 100us	Fault pin pulls low. The device stops switching when $V_{UVP} < 1.2 \text{ V}$.
LED+ and LED- short circuit	$V_{IN} - V_{CSP} < 300 \text{ mV}$ for 30ms	Fault pin pulls low. The keeps normal behavior.
LED- short to GND	$V_{UVP} < 1.2 \text{ V}$ for 100us	Fault pin pulls low. The device stops switching and recovers when fault is removed.
Sense-resistor open load	$V_{CSP} - V_{CSN} > 300\text{mV}$ for 20us	Fault pin pulls low. The device stops switching and recovers when fault is removed.
Sense-resistor short circuit	Error Amplifier output high for 100us	Fault pin pulls low. The device keeps switching under the cycle-by-cycle current limit.
Switching FET open	Error Amplifier output high for 100us	Fault pin pulls low. The device keeps maximum duty cycle turn-on switching.
Switching FET short	$V_{CSP} - V_{CSN} > 300 \text{ mV}$ for 20us	Fault pin pulls low. The device stops switching and recovers when fault is removed.
Thermal shutdown	$T_J > T_{TSD}$ for 100us	Fault pin pulls low. The device stops switching when $T_J > T_{TSD}$, and is re-activated when T_J falls below the hysteresis level.
VIN UVLO	$VCC < 3 \text{ V}$	Fault pin pulls low. The device stops switching and recovers when fault is removed.

Table 7-4. Protections in Boost / Buck-Boost topology

TYPE	CRITERION	BEHAVIOR
LED open load	$V_{OVP} > 1.2 \text{ V}$ for 100us	Fault pin pulls low. The device stops switching when $V_{OVP} > 1.2 \text{ V}$.
LED+ and LED- short circuit (Buck-Boost)	$V_{CSP} - V_{IN} < 300\text{mV}$ for 30ms	Fault pin pulls low. The device keeps normal behavior.
LED+ short to GND	$CSP - V_{CSN} > 300\text{mV}$ for 20us	Fault pin pulls low. The device stops switching and recovers when fault is removed.
Sense-resistor open load	$V_{CSP} - V_{CSN} > 300\text{mV}$ for 20us	Fault pin pulls low. The device stops switching and recovers when fault is removed.
Sense-resistor short circuit	Error Amplifier output high for 100us	Fault pin pulls low. The device keeps switching under the cycle-by-cycle current limit.
Switching FET open	Error Amplifier output high for 100us	Fault pin pulls low. The device keeps maximum duty cycle turn-on switching.
Switching FET short	Error Amplifier output high for 100us	Fault pin pulls low. The device tries to keep switching.
Thermal shutdown	$T_J > T_{TSD}$ for 100us	Fault pin pulls low. The device stops switching when $T_J > T_{TSD}$, and is re-activated when T_J falls below the hysteresis level.
VIN UVLO	$VCC < 3 \text{ V}$	Fault pin pulls low. The device stops switching and recovers when fault is removed.

7.3.7 Thermal Foldback

The LP8868 family integrates thermal shutdown protection to prevent the device from overheating. In order to provide design margin of system thermal performance, the device enables a programmable thermal foldback function which automatically reduces the full-scale output current, I_{FS} , at high junction temperature. When the device along with the LEDs are mounted on the same thermal substrate, the thermal performance is effectively improved due to the reduction of dissipation need for both and LED.

As the junction temperature rises above the thermal foldback threshold temperature, T_{th} , the full-scale current starts to reduce following the current-temperature curve shown in the below figure. The current starts to reduce from the 100% level at typically rate of 2% of I_{FS} per $^{\circ}C$ until it drops to 50% of the full scale. Once the junction temperature rises $25^{\circ}C$ above the T_{th} , the current continues to decrease at a lower rate until the temperature reaches above the overtemperature shutdown threshold temperature, T_{TSD} .

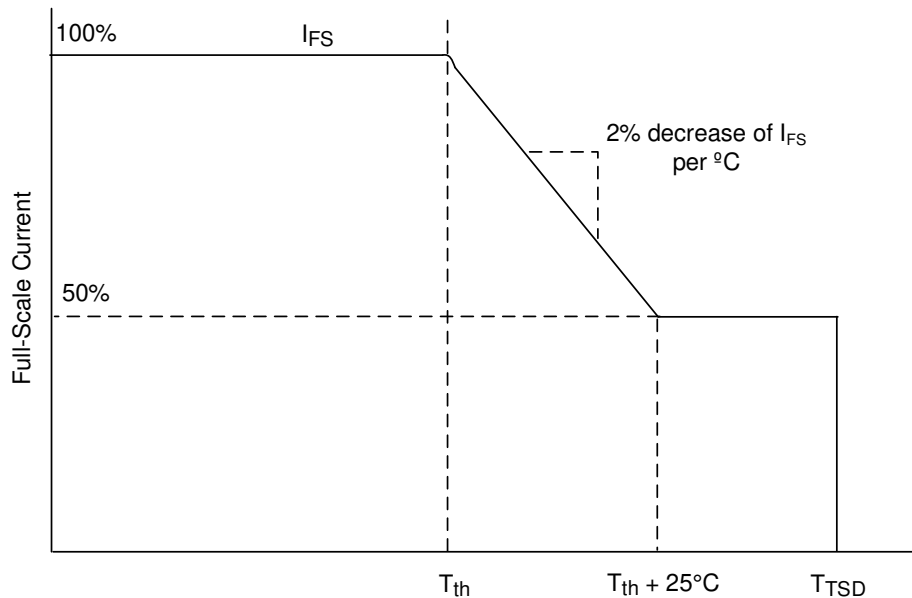


Figure 7-4. Thermal Foldback

The T_{th} can be adjusted by changing the resistor R_{TEMP} connected between the TEMP and GND pin. The T_{th} and the corresponding R_{TEMP} value are listed in below table.

Table 7-5. T_{th} vs. R_{TEMP} resistor value

T_{th} ($^{\circ}C$)	Resistor Value (k Ω)
80	200
90	100
100	60
110	40
120	28
130	20
140	15
150	10

8.2.1.2 Detailed Design Procedure

8.2.1.2.1 Inductor Selection

For this design, the input voltage is 9 V to 16 V. The output is 8 white LEDs in series and the inductor current ripple by requirement is less than 40% of maximum LED current. To choose a proper peak-to-peak inductor current ripple, the low-side FET current limit should not be violated when the converter works in no-load condition. This requires half of the peak-to-peak inductor current ripple to be lower than that limit. Another consideration is to ensure reasonable inductor core loss and copper loss caused by the peak-to-peak current ripple. Once this peak-to-peak inductor current ripple is chosen, use [Equation 2](#) to calculate the recommended value of the inductor L.

$$L = \frac{V_{IN(max)} \times (V_{OUT} - V_{IN(max)})}{V_{OUT} \times K_{IND} \times I_{L(max)} \times f_{SW}} \quad (2)$$

where

- K_{IND} is a coefficient that represents the amount of inductor ripple current relative to the maximum LED current.
- $I_{L(max)}$ is the maximum inductor current.
- f_{SW} is the switching frequency.
- $V_{IN(max)}$ is the maximum input voltage.
- V_{OUT} is the sum of the voltage across LED load and the voltage across sense resistor.

With the chosen inductor value, the user can calculate the actual inductor current ripple using [Equation 3](#).

$$I_{L(ripple)} = \frac{V_{IN(max)} \times (V_{OUT} - V_{IN(max)})}{V_{OUT} \times L \times f_{SW}} \quad (3)$$

The design ratings of inductor RMS current and saturation current must be greater than those seen in the system requirement. This is to ensure no inductor overheat or saturation occurring. During power up, transient conditions or fault conditions, the inductor current may exceed its normal operating current and reach the current limit. Therefore, it is preferred to select a saturation current rating equal to or greater than the converter current limit. The peak-inductor-current and RMS current equations are shown in [Equation 4](#) and [Equation 5](#).

$$I_{L(peak)} = I_{L(max)} + \frac{I_{L(ripple)}}{2} \quad (4)$$

$$I_{L(rms)} = \sqrt{I_{L(max)}^2 + \frac{I_{L(ripple)}^2}{2}} \quad (5)$$

In this design, $V_{IN(max)} = 16V$, $V_{OUT} = 24 V$, $I_{LED} = 1 A$, considering the efficiency as 0.9, $I_{L(max)} = 2.96 A$, $f_{SW} = 400 \text{ kHz}$, choose $K_{IND} = 0.4$, the calculated inductance is $11.25 \mu\text{H}$. A $10\text{-}\mu\text{H}$ inductor is chosen. With this inductor, the ripple, peak, and rms currents of the inductor are $1.212 A$, $3.569 A$, $3.084 A$, respectively.

8.2.1.2.2 Input Capacitor Selection

An input capacitor is required to reduce the surge current drawn from the input supply and the switching noise coming from the device. Electrolytic capacitors are recommended for energy storage. Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. For most applications, it is recommended to place a $10\text{-}\mu\text{F}$ ceramic capacitor along with a $0.1\text{-}\mu\text{F}$ capacitor from VIN to PGND/AGND to provide high-frequency filtering. The input capacitor voltage rating must be greater than the maximum input voltage. Use [Equation 6](#) to calculate the input ripple voltage, where ESR_{CIN} is the ESR of input capacitor, and K_{DR} is the derating coefficient of ceramic capacitance at the applied DC voltage.

$$V_{IN(ripple)} = \frac{I_{L(ripple)}}{8 \times C_{IN} \times f_{SW}} \quad (6)$$

In this design, a 33- μ F, 100V electrolytic capacitor, a 1- μ F, 100V X7R ceramic capacitor and a 0.1- μ F, 100V X7R ceramic capacitor are chosen, yielding around 100-mV input ripple voltage.

8.2.1.2.3 Output Capacitor Selection

The output capacitor reduces the high-frequency current ripple through the LED string. Excessive current ripple increases the RMS current in the LED string, therefore increasing the LED temperature.

1. Calculate the total dynamic resistance of the LED string (R_{LED}) using the LED manufacturer's datasheet.
2. Calculate the required impedance of the output capacitor (Z_{COUT}) given the acceptable peak-to-peak ripple current through the LED string, $I_{LED(ripple)}$. $I_{L(ripple)}$ is the peak-to-peak inductor ripple current as calculated with the selected inductor.
3. Calculate the minimum effective output capacitance required.
4. Increase the output capacitance appropriately due to the derating effect of applied DC voltage.

See [Equation 7](#), [Equation 8](#), and [Equation 9](#).

$$R_{LED} = \frac{\Delta V_F}{\Delta I_F} \times \# \text{ of LEDs} \quad (7)$$

$$Z_{COUT} = \frac{R_{LED} \times I_{LED(ripple)}}{I_{L(max)} - I_{LED(ripple)}} \quad (8)$$

$$C_{COUT} = \frac{1}{2\pi \times f_{SW} \times Z_{COUT}} \quad (9)$$

Once the output capacitor is chosen, [Equation 10](#) can be used to estimate the peak-to-peak ripple current through the LED string.

$$I_{LED(ripple)} = \frac{Z_{COUT} \times I_{L(ripple)}}{Z_{COUT} + R_{LED}} \quad (10)$$

Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. In this design, a 10- μ F, 50-V X7R ceramic capacitor is chosen.

8.2.1.2.4 Sense Resistor Selection

The maximum LED current is 1 A at 100% PWM duty and the corresponding V_{REF} is 200 mV. By using [Equation 11](#), the sense resistance is calculated as 200 m Ω . Note that the power consumption of the sense resistor is 200 mW, requiring enough margin of the resistor's power rating in selection.

$$R_{SENSE} = \frac{V_{REF}}{I_{LED_FS}} \quad (11)$$

In boost topology, C_{SENSE} is required to achieve the IFD control. Using [Equation 12](#), a 10- μ F, 50-V X7R ceramic capacitor is chosen for C_{SENSE} to suppress the ac magnitude of sense feedback less than 200 mV.

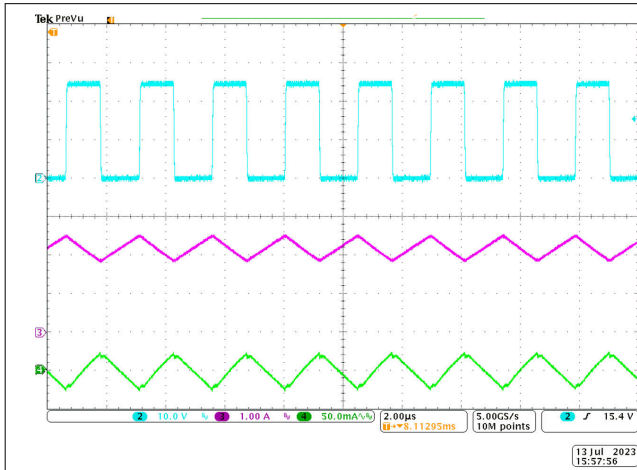
$$C_{SENSE} = \frac{0.25 \times I_{L(max)}}{200mV \times f_{SW}} \quad (12)$$

8.2.1.2.5 Other External Components Selection

In this design, a 0.1- μ F, 50-V X7R ceramic capacitor is chosen for high-frequency filtering of sense feedback.

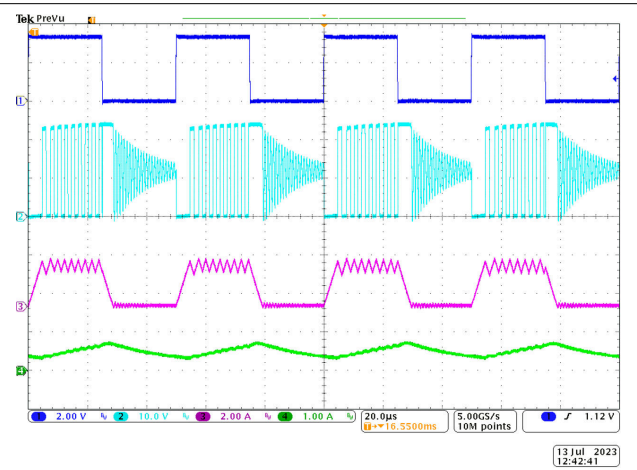
For loop stability, it is recommended to select a 1-nF, 10-V X7R ceramic capacitor for C_{COMP} and a 1-k Ω resistor for R_{COMP} . A 1-M Ω resistor is chosen for R_{DAMP} to suppress the overshoot current at rising edge of PWM on.

8.2.1.3 Application Curves



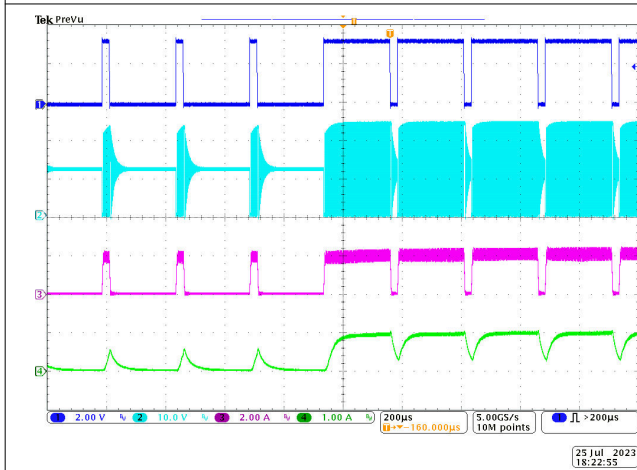
Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current
 (AC)

Figure 8-2. LED Current Ripple at ADIM = 100%, 1 kHz and $F_{sw} = 400$ kHz



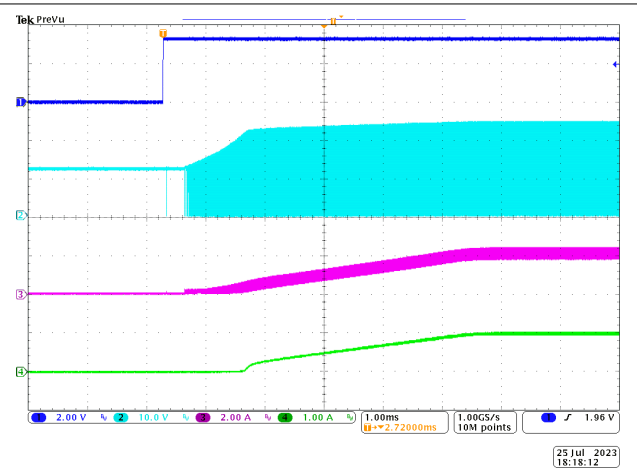
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

Figure 8-3. LED Current Ripple at PWM = 50%, 20 kHz and $F_{sw} = 400$ kHz



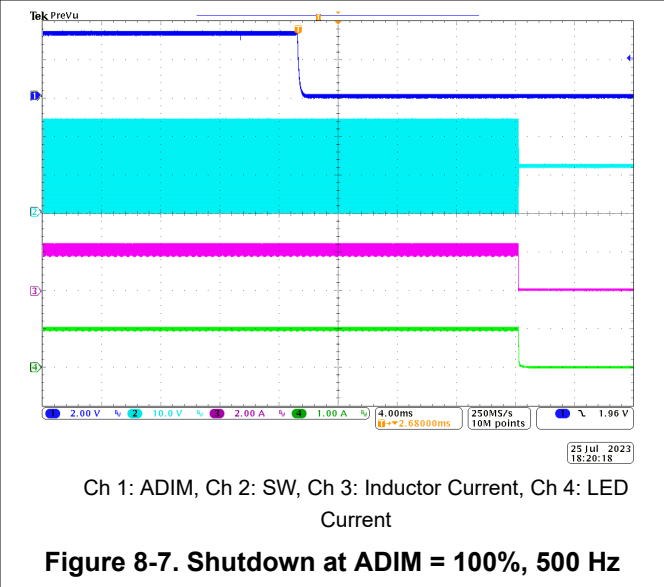
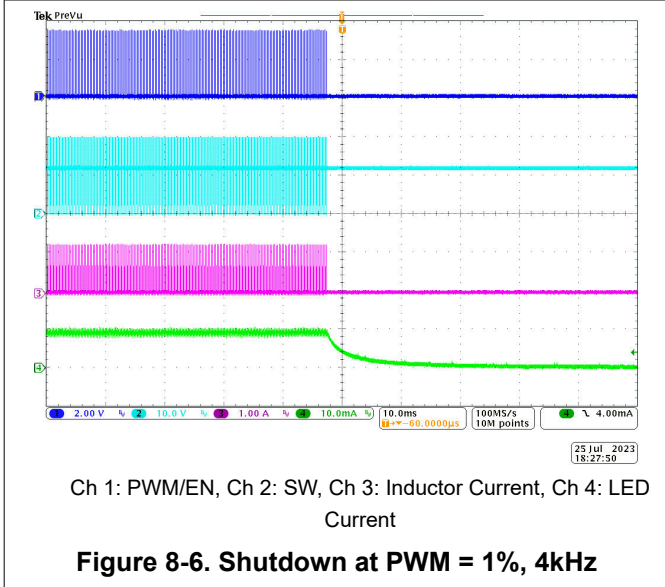
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

**Figure 8-4. LED Current Transient for a PWM
 Transition from 10% to 99%, 4 kHz**



Ch 1: ADIM, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

Figure 8-5. Startup at ADIM = 100%, 500Hz



8.2.2.2 Detailed Design Procedure

8.2.2.2.1 Inductor Selection

For this design, the input voltage is 9 V to 16 V. The output is 8 white LEDs in series and the inductor current ripple by requirement is less than 40% of maximum LED current. To choose a proper peak-to-peak inductor current ripple, the low-side FET current limit should not be violated when the converter works in no-load condition. This requires half of the peak-to-peak inductor current ripple to be lower than that limit. Another consideration is to ensure reasonable inductor core loss and copper loss caused by the peak-to-peak current ripple. Once this peak-to-peak inductor current ripple is chosen, use [Equation 13](#) to calculate the recommended value of the inductor L.

$$L = \frac{V_{IN(max)} \times V_{OUT}}{(V_{OUT} + V_{IN(max)}) \times K_{IND} \times I_{L(max)} \times f_{SW}} \quad (13)$$

where

- K_{IND} is a coefficient that represents the amount of inductor ripple current relative to the maximum LED current.
- $I_{L(max)}$ is the maximum inductor current.
- f_{SW} is the switching frequency.
- $V_{IN(max)}$ is the maximum input voltage.
- V_{OUT} is the sum of the voltage across LED load and the voltage across sense resistor.

With the chosen inductor value, the user can calculate the actual inductor current ripple using [Equation 14](#).

$$I_{L(ripple)} = \frac{V_{IN(max)} \times V_{OUT}}{(V_{OUT} + V_{IN(max)}) \times L \times f_{SW}} \quad (14)$$

The design ratings of inductor RMS current and saturation current must be greater than those seen in the system requirement. This is to ensure no inductor overheat or saturation occurring. During power up, transient conditions or fault conditions, the inductor current may exceed its normal operating current and reach the current limit. Therefore, it is preferred to select a saturation current rating equal to or greater than the converter current limit. The peak-inductor-current and RMS current equations are shown in [Equation 15](#) and [Equation 16](#).

$$I_{L(peak)} = I_{L(max)} + \frac{I_{L(ripple)}}{2} \quad (15)$$

$$I_{L(rms)} = \sqrt{I_{L(max)}^2 + \frac{I_{L(ripple)}^2}{2}} \quad (16)$$

In this design, $V_{IN(max)} = 16V$, $V_{OUT} = 15V$, $I_{LED} = 1A$, considering the efficiency as 0.8, $I_{L(max)} = 2.083A$, $f_{SW} = 400kHz$, choose $K_{IND} = 0.4$, the calculated inductance is 23.22 μH . A 22- μH inductor is chosen. With this inductor, the ripple, peak, and rms currents of the inductor are 1.09 A, 2.629 A, 2.222 A, respectively.

8.2.2.2.2 Input Capacitor Selection

An input capacitor is required to reduce the surge current drawn from the input supply and the switching noise coming from the device. Electrolytic capacitors are recommended for energy storage. Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. For most applications, it is recommended to place a 10- μF capacitor along with a 0.1- μF capacitor from VIN to PGND/AGND to provide high-frequency filtering. The input capacitor voltage rating must be greater than the maximum input voltage. Use [Equation 17](#) to calculate the input ripple voltage, where ESR_{CIN} is the ESR of input capacitor, and K_{DR} is the derating coefficient of ceramic capacitance at the applied DC voltage.

$$V_{IN(ripple)} = I_{L(max)} \times \left(\frac{V_{OUT}}{K_{DR} \times C_{IN} \times f_{SW} \times (V_{IN(max)} + V_{OUT})} + ESR_{CIN} \right) \quad (17)$$

In this design, a 10- μ F, 100V electrolytic capacitor, a 2.2- μ F, 100V X7R ceramic capacitor and a 0.1- μ F, 100V X7R ceramic capacitor are chosen, yielding around 504-mV input ripple voltage.

8.2.2.2.3 Output Capacitor Selection

The output capacitor reduces the high-frequency current ripple through the LED string. Excessive current ripple increases the RMS current in the LED string, therefore increasing the LED temperature.

1. Calculate the total dynamic resistance of the LED string (R_{LED}) using the LED manufacturer's datasheet.
2. Calculate the required impedance of the output capacitor (Z_{COUT}) given the acceptable peak-to-peak ripple current through the LED string, $I_{LED(ripple)}$, $I_{L(ripple)}$ is the peak-to-peak inductor ripple current as calculated with the selected inductor.
3. Calculate the minimum effective output capacitance required.
4. Increase the output capacitance appropriately due to the derating effect of applied DC voltage.

See [Equation 18](#), [Equation 19](#), and [Equation 20](#).

$$R_{LED} = \frac{\Delta V_F}{\Delta I_F} \times \# \text{ of LEDs} \quad (18)$$

$$Z_{COUT} = \frac{R_{LED} \times I_{LED(ripple)}}{I_{L(max)} - I_{LED(ripple)}} \quad (19)$$

$$C_{COUT} = \frac{1}{2\pi \times f_{SW} \times Z_{COUT}} \quad (20)$$

Once the output capacitor is chosen, [Equation 21](#) can be used to estimate the peak-to-peak ripple current through the LED string.

$$I_{LED(ripple)} = \frac{Z_{COUT} \times I_{L(ripple)}}{Z_{COUT} + R_{LED}} \quad (21)$$

Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. In this design, a 10- μ F, 50-V X7R ceramic capacitor is chosen.

8.2.2.2.4 Sense Resistor Selection

The maximum LED current is 1 A at 100% PWM duty and the corresponding V_{REF} is 200 mV. By using [Equation 22](#), the sense resistance is calculated as 200 m Ω . Note that the power consumption of the sense resistor is 200 mW, requiring enough margin of the resistor's power rating in selection.

$$R_{SENSE} = \frac{V_{REF}}{I_{LED_FS}} \quad (22)$$

In buck-boost topology, C_{SENSE} is required to achieve the IFD control. Using [Equation 23](#), a 10- μ F, 50-V X7R ceramic capacitor is chosen for C_{SENSE} to suppress the ac magnitude of sense feedback less than 200 mV.

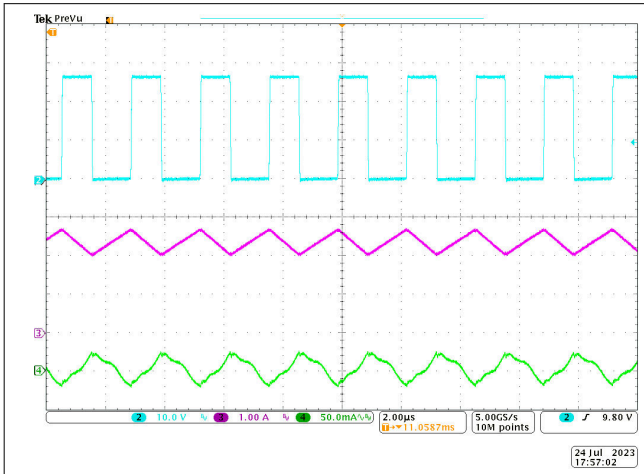
$$C_{SENSE} = \frac{0.25 \times I_{L(max)}}{200mV \times f_{SW}} \quad (23)$$

8.2.2.2.5 Other External Components Selection

In this design, a 0.1- μ F, 50-V X7R ceramic capacitor is chosen for high-frequency filtering of sense feedback.

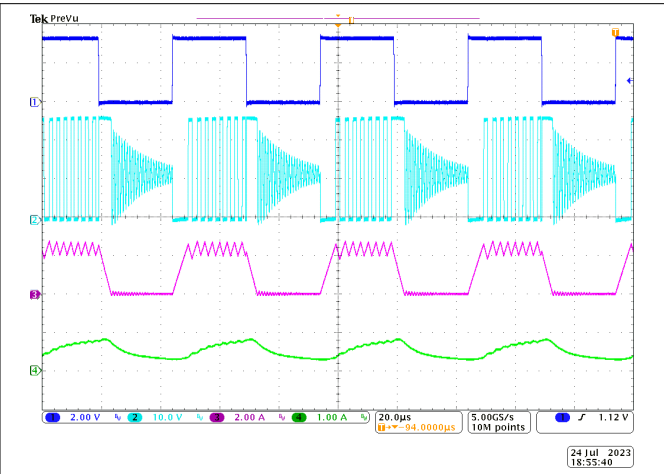
For loop stability, it is recommended to select a 1-nF, 10-V X7R ceramic capacitor for C_{COMP} and a 1-k Ω resistor for R_{COMP} . A 1-M Ω resistor is chosen for R_{DAMP} to suppress the overshoot current at rising edge of PWM on.

8.2.2.3 Application Curves



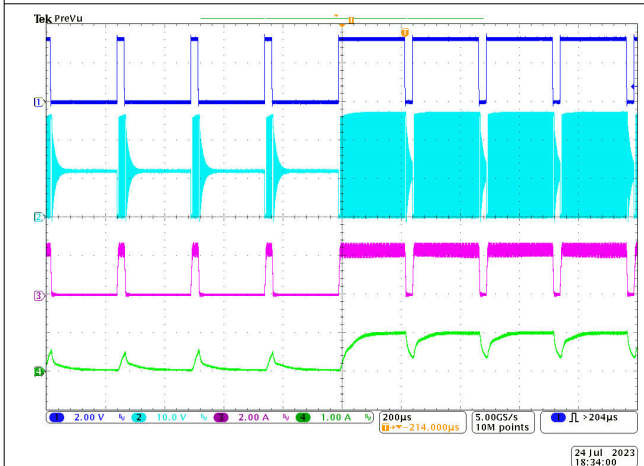
Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current Ripple (AC)

Figure 8-9. LED Current Ripple at ADIM = 100%, 1 kHz and $F_{sw} = 400$ kHz



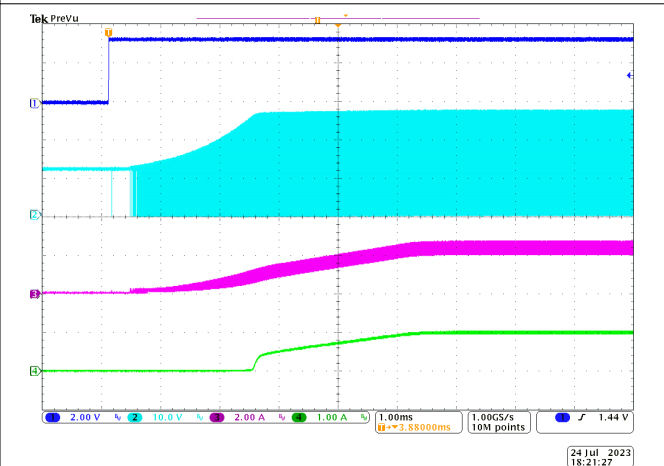
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current

Figure 8-10. LED Current Ripple at PWM = 50%, 20 kHz and $F_{sw} = 400$ kHz



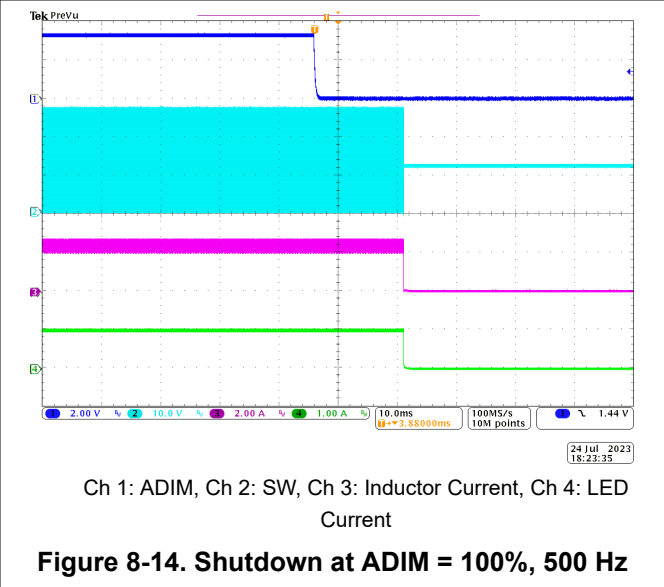
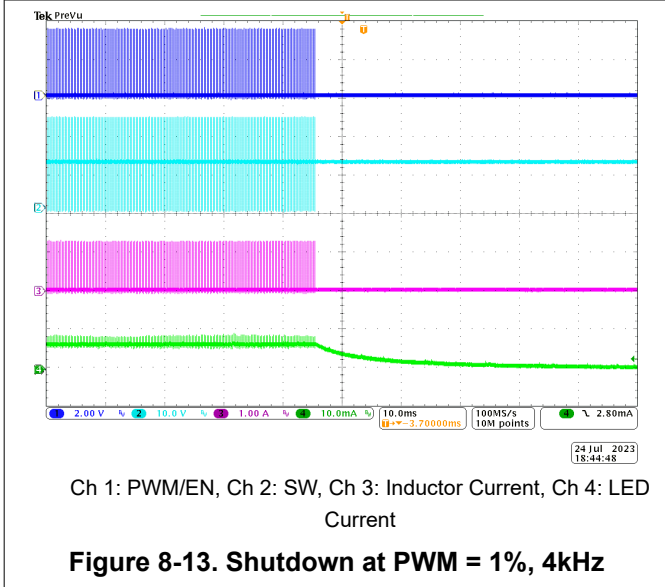
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current

Figure 8-11. LED Current Transient for a PWM Transition from 10% to 99%, 4 kHz



Ch 1: ADIM, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current

Figure 8-12. Startup at ADIM = 100%, 500Hz



8.2.3 LP8868ZQDMTRQ1 12-V Input, 2-A Output, 1-piece LED With Buck Topology

Figure 8-15 shows a typical application for the LP8868Z in a buck topology. The switching current limit is 5.2 A and the output current limit is equal to the limit of switching current limit.

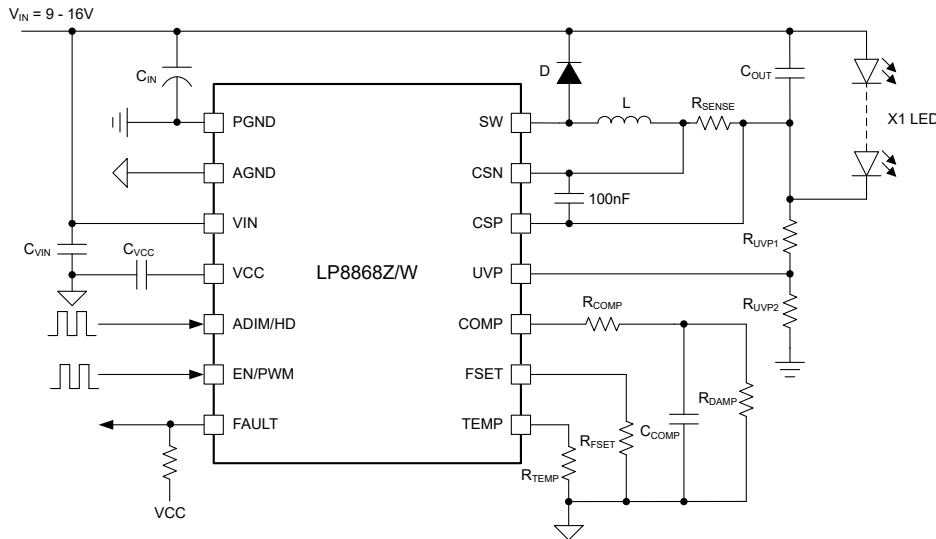


Figure 8-15. Typical Application for Buck Topology with LP8868Z

8.2.3.1 Design Requirements

For this design example, use the parameters in the following table.

Table 8-3. Design Parameters

PARAMETER	VALUE
Input voltage range	9 V -16 V
LED string	1 LED
Output voltage	3 V
Switching frequency	400 kHz
Maximum LED current	2 A
Inductor current ripple	40% of maximum inductor current
Dimming type	PWM dimming/ADIM dimming

8.2.3.2 Detailed Design Procedure

8.2.3.2.1 Inductor Selection

For this design, the input voltage is 9V to 16V. The output is single white LED and the inductor current ripple by requirement is less than 40% of maximum LED current. To choose a proper peak-to-peak inductor current ripple, the low-side FET current limit should not be violated when the converter works in no-load condition. This requires half of the peak-to-peak inductor current ripple to be lower than that limit. Another consideration is to ensure reasonable inductor core loss and copper loss caused by the peak-to-peak current ripple. Once this peak-to-peak inductor current ripple is chosen, use [Equation 24](#) to calculate the recommended value of the inductor L.

$$L = \frac{V_{OUT} \times (V_{IN(max)} - V_{OUT})}{V_{IN(max)} \times K_{IND} \times I_{L(max)} \times f_{SW}} \quad (24)$$

where

- K_{IND} is a coefficient that represents the amount of inductor ripple current relative to the maximum LED current.
- $I_{L(max)}$ is the maximum LED current.
- f_{SW} is the switching frequency.
- $V_{IN(max)}$ is the maximum input voltage.
- V_{OUT} is the sum of the voltage across LED load and the voltage across sense resistor.

With the chosen inductor value, the user can calculate the actual inductor current ripple using [Equation 25](#).

$$I_{L(ripple)} = \frac{V_{OUT} \times (V_{IN(max)} - V_{OUT})}{V_{IN(max)} \times L \times f_{SW}} \quad (25)$$

The design ratings of inductor RMS current and saturation current must be greater than those seen in the system requirement. This is to ensure no inductor overheat or saturation occurring. During power up, transient conditions or fault conditions, the inductor current may exceed its normal operating current and reach the current limit. Therefore, it is preferred to select a saturation current rating equal to or greater than the converter current limit. The peak-inductor-current and RMS current equations are shown in [Equation 26](#) and [Equation 27](#).

$$I_{L(peak)} = I_{L(max)} + \frac{I_{L(ripple)}}{2} \quad (26)$$

$$I_{L(rms)} = \sqrt{I_{L(max)}^2 + \frac{I_{L(ripple)}^2}{2}} \quad (27)$$

In this design, $V_{IN(max)} = 16V$, $V_{OUT} = 3V$, $I_{LED} = 2A$, considering the efficiency as 0.9, $I_{L(max)} = 0.741A$, $f_{SW} = 400kHz$, choose $K_{IND} = 0.4$, the calculated inductance is 20.57 μH . A 22- μH inductor is chosen. With this inductor, the ripple, peak, and rms currents of the inductor are 0.277 A, 0.879 A, 0.766 A, respectively.

8.2.3.2.2 Input Capacitor Selection

An input capacitor is required to reduce the surge current drawn from the input supply and the switching noise coming from the device. Electrolytic capacitors are recommended for energy storage. Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. For most applications, it is recommended to place a 10- μF capacitor along with a 0.1- μF capacitor from VIN to PGND/AGND to provide high-frequency filtering. The input capacitor voltage rating must be greater than the maximum input voltage. Use [Equation 28](#) to calculate the input ripple voltage, where ESR_{CIN} is the ESR of input capacitor, and K_{DR} is the derating coefficient of ceramic capacitance at the applied DC voltage.

$$V_{IN(ripple)} = I_{L(max)} \times \left(\frac{V_{OUT}}{K_{DR} \times C_{IN} \times f_{SW} \times V_{IN(max)}} + ESR_{CIN} \right) \quad (28)$$

In this design, a 33-μF, 100V electrolytic capacitor, a 2.2-μF, 100V X7R ceramic capacitor and a 0.1-μF, 100V X7R ceramic capacitor are chosen, yielding around 70-mV input ripple voltage.

8.2.3.2.3 Output Capacitor Selection

The output capacitor reduces the high-frequency current ripple through the LED string. Excessive current ripple increases the RMS current in the LED string, therefore increasing the LED temperature.

1. Calculate the total dynamic resistance of the LED string (R_{LED}) using the LED manufacturer's datasheet.
2. Calculate the required impedance of the output capacitor (Z_{COUT}) given the acceptable peak-to-peak ripple current through the LED string, $I_{LED(ripple)}$. $I_{L(ripple)}$ is the peak-to-peak inductor ripple current as calculated with the selected inductor.
3. Calculate the minimum effective output capacitance required.
4. Increase the output capacitance appropriately due to the derating effect of applied DC voltage.

See [Equation 29](#), [Equation 30](#), and [Equation 31](#).

$$R_{LED} = \frac{\Delta V_F}{\Delta I_F} \times \# \text{ of LEDs} \quad (29)$$

$$Z_{COUT} = \frac{R_{LED} \times I_{LED(ripple)}}{I_{L(max)} - I_{LED(ripple)}} \quad (30)$$

$$C_{COUT} = \frac{1}{2\pi \times f_{SW} \times Z_{COUT}} \quad (31)$$

Once the output capacitor is chosen, [Equation 32](#) can be used to estimate the peak-to-peak ripple current through the LED string.

$$I_{LED(ripple)} = \frac{Z_{COUT} \times I_{L(ripple)}}{Z_{COUT} + R_{LED}} \quad (32)$$

Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. In this design, a 10-μF, 35-V X7R ceramic capacitor is chosen.

8.2.3.2.4 Sense Resistor Selection

The maximum LED current is 2 A at 100% PWM duty and the corresponding V_{REF} is 200 mV. By using [Equation 33](#), the sense resistance is calculated as 100 mΩ. Note that the power consumption of the sense resistor is 400 mW, requiring enough margin of the resistor's power rating in selection.

$$R_{SENSE} = \frac{V_{REF}}{I_{LED_FS}} \quad (33)$$

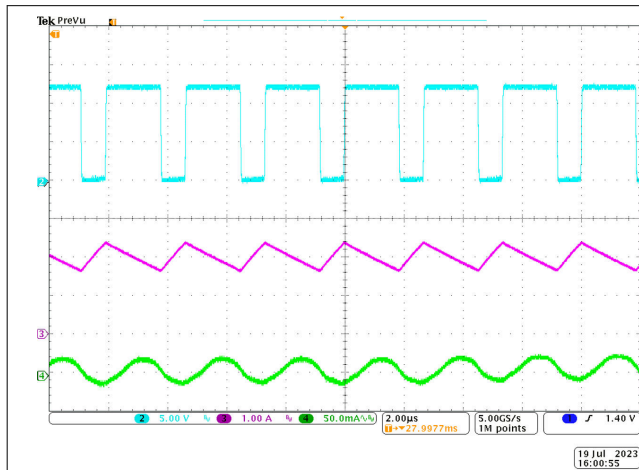
In buck topology, C_{sense} is optional to use,

8.2.3.2.5 Other External Components Selection

In this design, a 0.1-μF, 50-V X7R ceramic capacitor is chosen for high-frequency filtering of sense feedback.

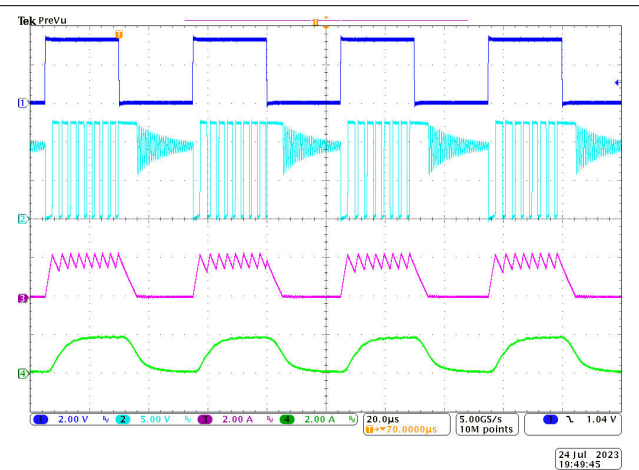
For loop stability, it is recommended to select a 1-nF, 10-V X7R ceramic capacitor for C_{COMP} and a 1-kΩ resistor for R_{COMP} . A 1-MΩ resistor is chosen for R_{DAMP} to suppress the overshoot current at rising edge of PWM on.

8.2.3.3 Application Curves



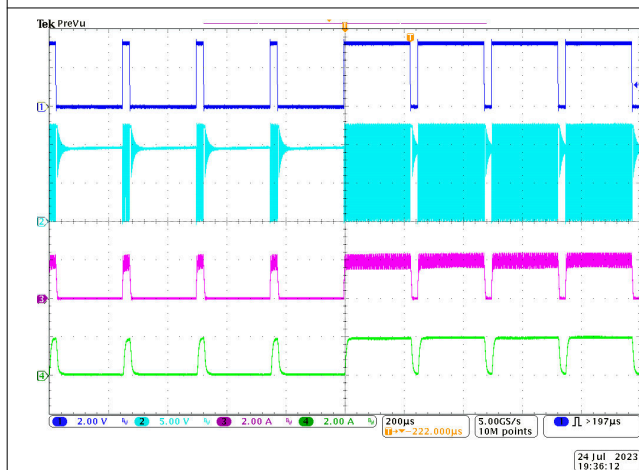
Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED Current
 (AC)

Figure 8-16. LED Current Ripple at ADIM = 100%, 1 kHz and $F_{sw} = 400$ kHz



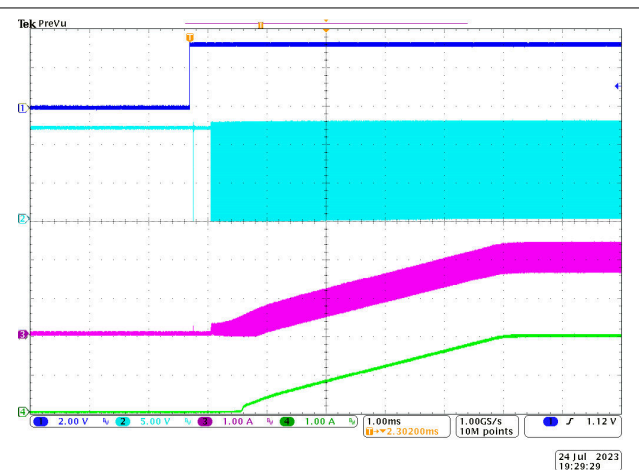
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

Figure 8-17. LED Current Ripple at PWM = 50%, 20 kHz and $F_{sw} = 400$ kHz



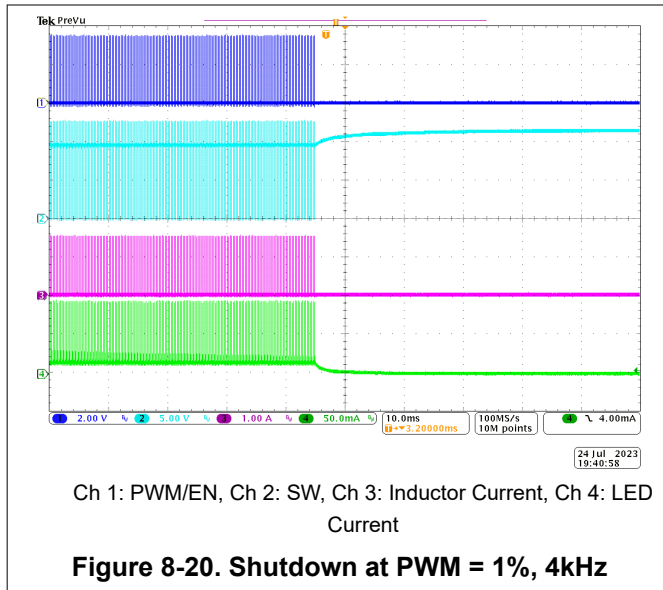
Ch 1: PWM/EN, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

Figure 8-18. LED Current Transient for a PWM Transition from 10% to 99%, 4 kHz



Ch 1: ADIM, Ch 2: SW, Ch 3: Inductor Current, Ch 4: LED
 Current

Figure 8-19. Startup at ADIM = 100%, 500Hz



8.3 Power Supply Recommendations

The device is designed to operate from an input voltage supply ranging between 4.5 V and 65 V. This input supply must be well regulated. The device requires an input capacitor to reduce the surge current drawn from the input supply and the switching noise from the device. Ceramic capacitors with X5R or X7R dielectrics are highly recommended because of their low ESR and small temperature coefficients. For most applications, a 10- μ F capacitor is enough.

8.4 Layout

The LP8868-Q1 family requires a proper layout for optimal performance. The following section gives some guidelines to ensure a proper layout.

8.4.1 Layout Guidelines

Examples of a proper layout for boost topology, buck-boost topology and buck topology of LP8868 family is shown below.

- Creating a large GND plane for good electrical and thermal performance is important.
- The IN and GND traces should be as wide as possible to reduce trace impedance. Wide traces have the additional advantage of providing excellent heat dissipation.
- Thermal vias can be used to connect the top-side GND plane to additional printed-circuit board (PCB) layers for heat dissipation and grounding.
- The input capacitors must be located as close as possible to the IN pin and the GND pin.
- The VCC capacitor should be placed as close as possible to VCC pin to ensure stable LDO output voltage.
- The SW trace must be kept as short as possible to reduce parasitic inductance and thereby reduce transient voltage spikes. Short SW trace also reduces radiated noise and EMI.
- Do not allow switching current to flow under the device.
- The routing of CSN and CSP traces are recommended to be in parallel and kept as short as possible and placed away from the high-voltage switching trace and the ground shield.
- The compensation capacitor must be placed as close as possible to COMP pin so as to prevent oscillation and system instability.

8.4.2 Layout Example

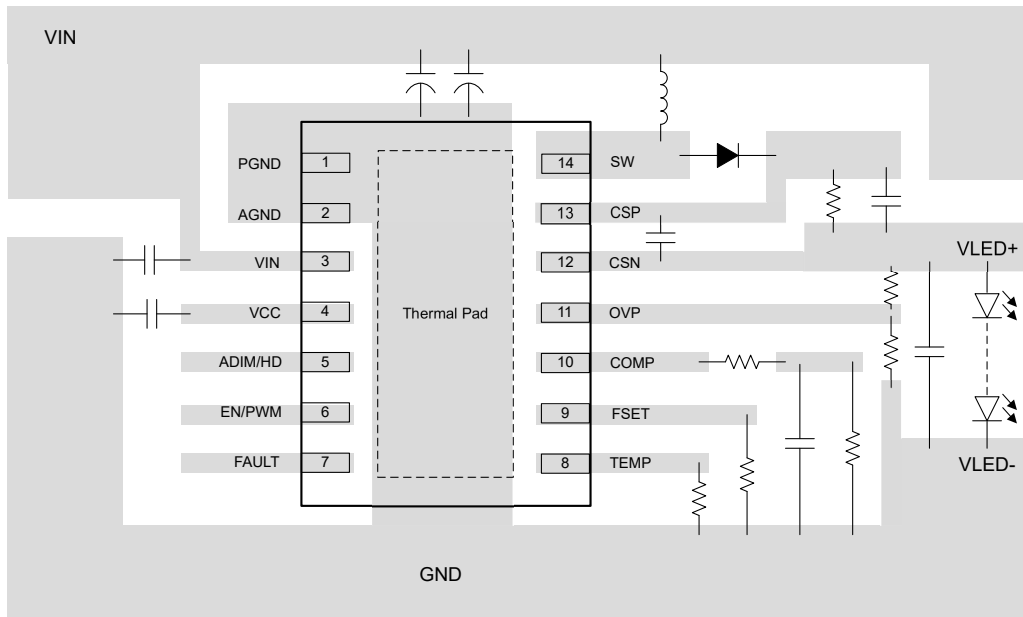


Figure 8-22. Boost Topology Top View Layout Example

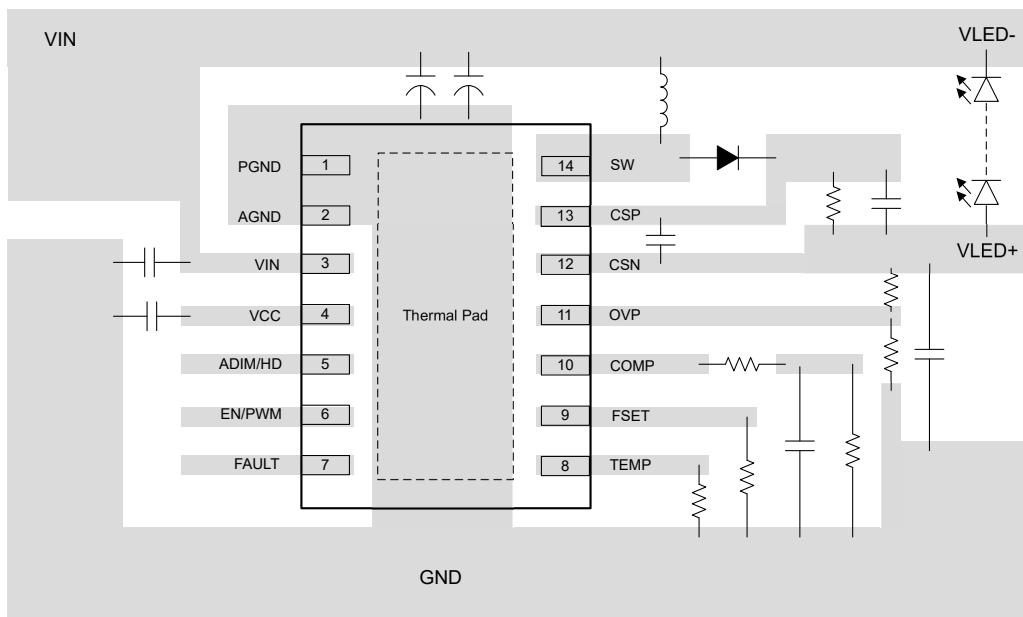


Figure 8-23. Buck-Boost Topology Top View Layout Example

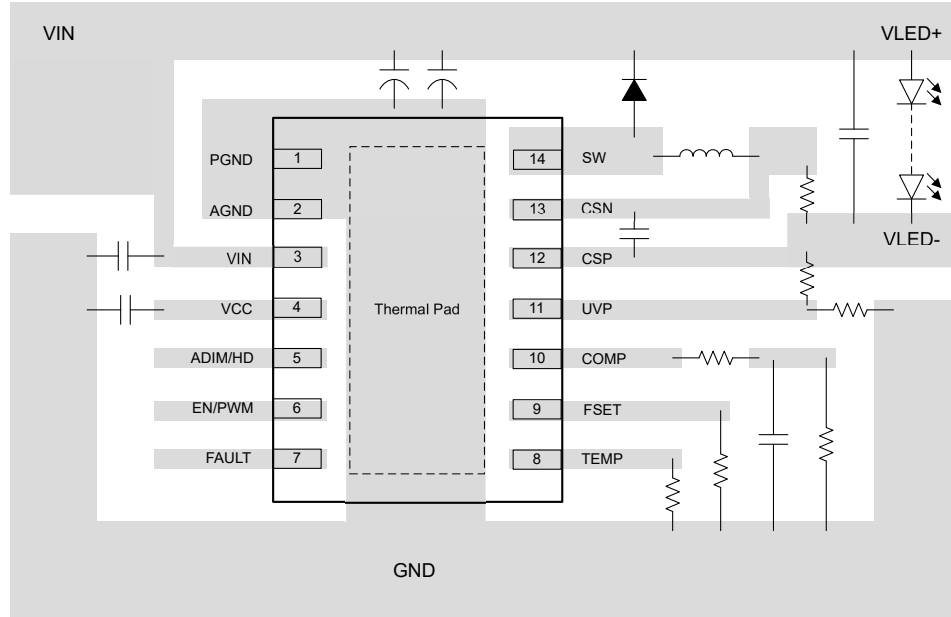


Figure 8-24. Buck Topology Top View Layout Example

9 Device and Documentation Support

TI offers an extensive line of development tools. Tools and software to evaluate the performance of the device, generate code, and develop solutions are listed below.

9.1 Receiving Notification of Documentation Updates

To receive notification of documentation updates, navigate to the device product folder on ti.com. Click on *Notifications* to register and receive a weekly digest of any product information that has changed. For change details, review the revision history included in any revised document.

9.2 Support Resources

TI E2E™ [support forums](#) are an engineer's go-to source for fast, verified answers and design help — straight from the experts. Search existing answers or ask your own question to get the quick design help you need.

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9.3 Trademarks

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9.4 Electrostatic Discharge Caution



This integrated circuit can be damaged by ESD. Texas Instruments recommends that all integrated circuits be handled with appropriate precautions. Failure to observe proper handling and installation procedures can cause damage.

ESD damage can range from subtle performance degradation to complete device failure. Precision integrated circuits may be more susceptible to damage because very small parametric changes could cause the device not to meet its published specifications.

9.5 Glossary

[TI Glossary](#) This glossary lists and explains terms, acronyms, and definitions.

10 Revision History

Changes from Revision * (July 2023) to Revision A (November 2023)	Page
• Added LP8868U-Q1, LP8868V-Q1, LP8868W-Q1 to Production Data.....	1

11 Mechanical, Packaging, and Orderable Information

The following pages include mechanical, packaging, and orderable information. This information is the most-current data available for the designated s. This data is subject to change without notice and without revision of this document. For browser-based versions of this data sheet, see the left-hand navigation pane.

PACKAGING INFORMATION

Orderable part number	Status (1)	Material type (2)	Package Pins	Package qty Carrier	RoHS (3)	Lead finish/ Ball material (4)	MSL rating/ Peak reflow (5)	Op temp (°C)	Part marking (6)
LP8868UQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868U
LP8868UQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868U
LP8868VQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868V
LP8868VQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868V
LP8868WQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868W
LP8868WQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868W
LP8868XQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868X
LP8868XQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868X
LP8868YQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868Y
LP8868YQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868Y
LP8868ZQDMTRQ1	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868Z
LP8868ZQDMTRQ1.A	Active	Production	VSON (DMT) 14	3000 LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 125	L868Z

(1) **Status:** For more details on status, see our [product life cycle](#).

(2) **Material type:** When designated, preproduction parts are prototypes/experimental devices, and are not yet approved or released for full production. Testing and final process, including without limitation quality assurance, reliability performance testing, and/or process qualification, may not yet be complete, and this item is subject to further changes or possible discontinuation. If available for ordering, purchases will be subject to an additional waiver at checkout, and are intended for early internal evaluation purposes only. These items are sold without warranties of any kind.

(3) **RoHS values:** Yes, No, RoHS Exempt. See the [TI RoHS Statement](#) for additional information and value definition.

(4) **Lead finish/Ball material:** Parts may have multiple material finish options. Finish options are separated by a vertical ruled line. Lead finish/Ball material values may wrap to two lines if the finish value exceeds the maximum column width.

(5) **MSL rating/Peak reflow:** The moisture sensitivity level ratings and peak solder (reflow) temperatures. In the event that a part has multiple moisture sensitivity ratings, only the lowest level per JEDEC standards is shown. Refer to the shipping label for the actual reflow temperature that will be used to mount the part to the printed circuit board.

(6) **Part marking:** There may be an additional marking, which relates to the logo, the lot trace code information, or the environmental category of the part.

Multiple part markings will be inside parentheses. Only one part marking contained in parentheses and separated by a "-" will appear on a part. If a line is indented then it is a continuation of the previous line and the two combined represent the entire part marking for that device.

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TAPE AND REEL INFORMATION

QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE


*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Reel Diameter (mm)	Reel Width W1 (mm)	A0 (mm)	B0 (mm)	K0 (mm)	P1 (mm)	W (mm)	Pin1 Quadrant
LP8868UQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1
LP8868VQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1
LP8868WQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1
LP8868XQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1
LP8868YQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1
LP8868ZQDMTRQ1	VSON	DMT	14	3000	330.0	12.4	3.3	4.8	1.2	8.0	12.0	Q1

TAPE AND REEL BOX DIMENSIONS


*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
LP8868UQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0
LP8868VQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0
LP8868WQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0
LP8868XQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0
LP8868YQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0
LP8868ZQDMTRQ1	VSON	DMT	14	3000	367.0	367.0	35.0

GENERIC PACKAGE VIEW

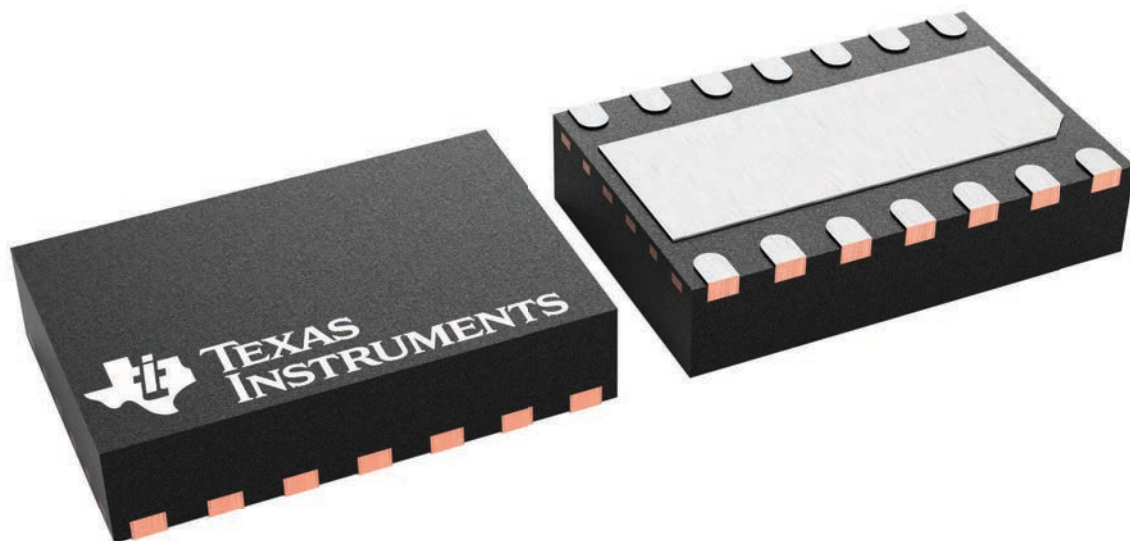
DMT 14

VSON - 0.9 mm max height

3 x 4.5, 0.65 mm pitch

PLASTIC SMALL OUTLINE - NO LEAD

This image is a representation of the package family, actual package may vary.
Refer to the product data sheet for package details.



4225088/A

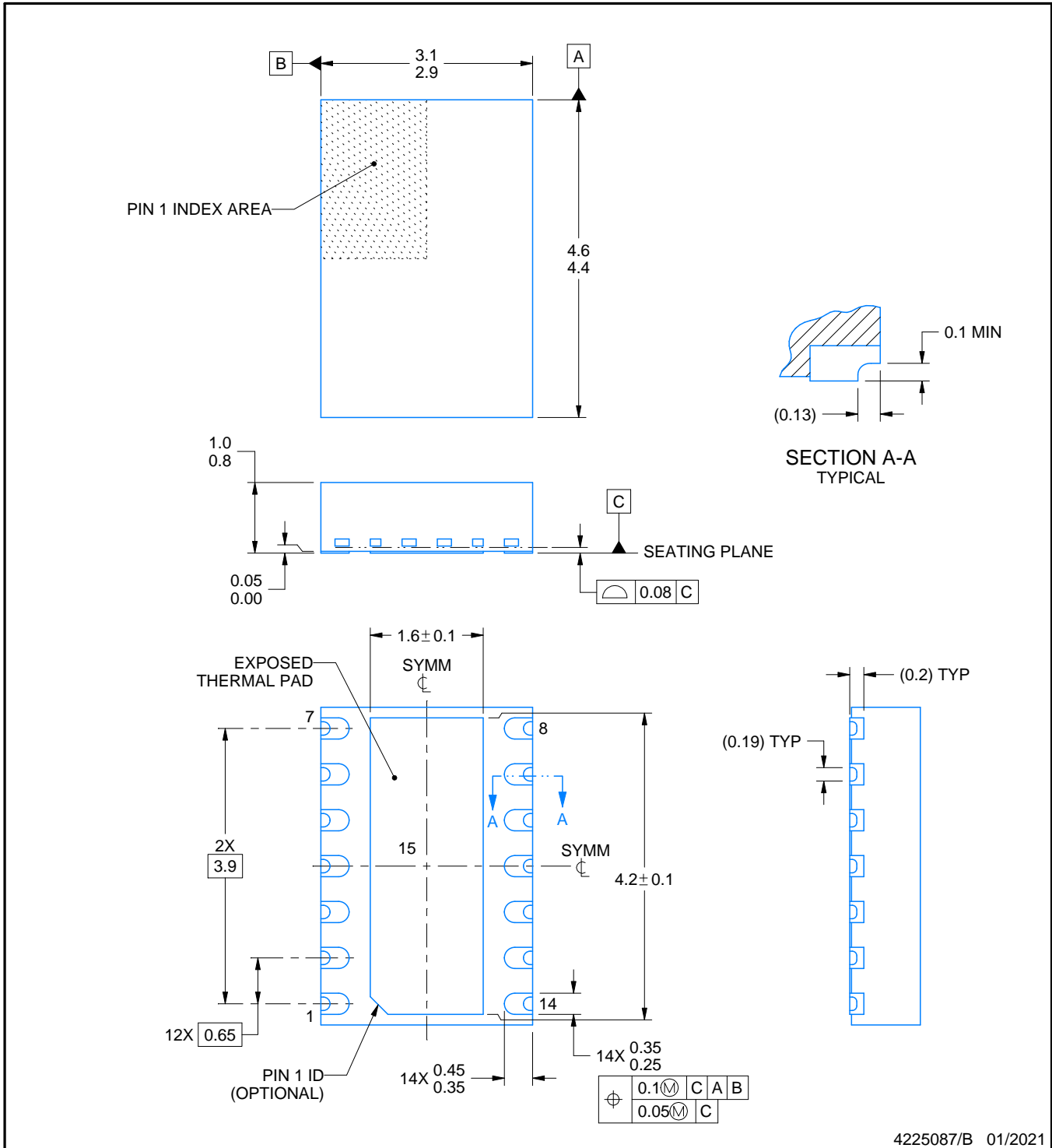
DMT0014B



PACKAGE OUTLINE

VSON - 1 mm max height

PLASTIC SMALL OUTLINE - NO LEAD



4225087/B 01/2021

NOTES:

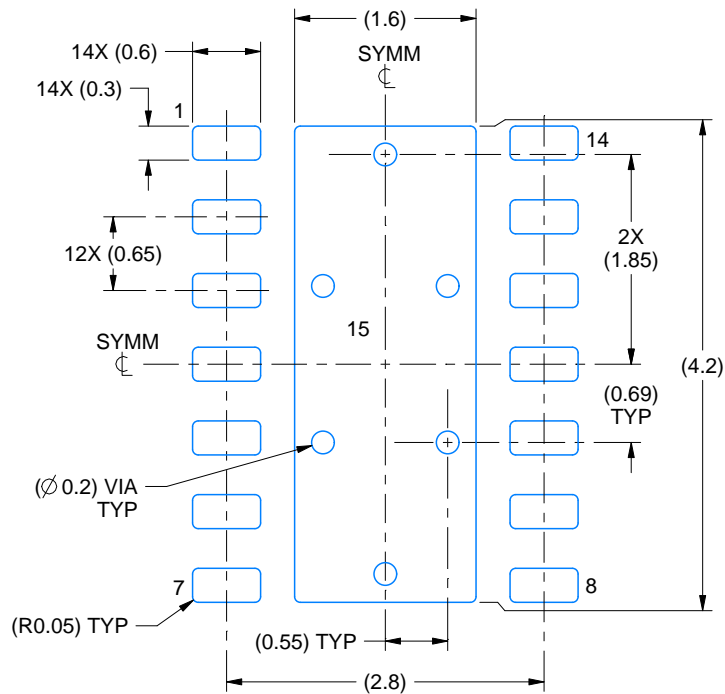
1. All linear dimensions are in millimeters. Any dimensions in parenthesis are for reference only. Dimensioning and tolerancing per ASME Y14.5M.
2. This drawing is subject to change without notice.
3. The package thermal pad must be soldered to the printed circuit board for thermal and mechanical performance.

EXAMPLE BOARD LAYOUT

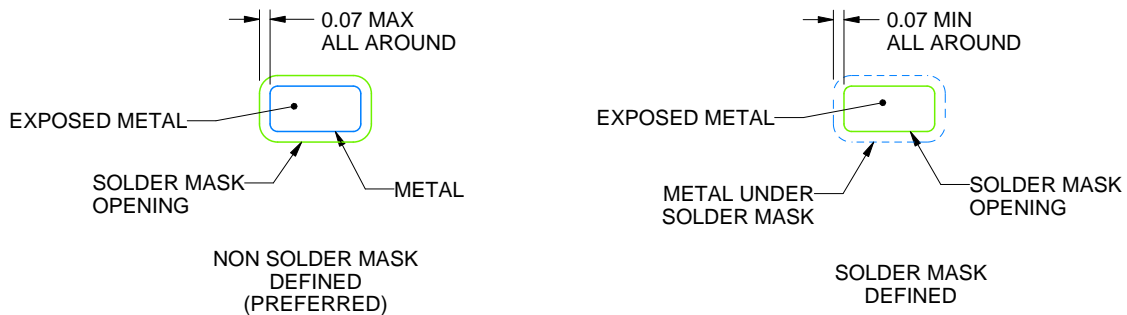
DMT0014B

VSON - 1 mm max height

PLASTIC SMALL OUTLINE - NO LEAD



LAND PATTERN EXAMPLE
EXPOSED METAL SHOWN
SCALE:15X



SOLDER MASK DETAILS

4225087/B 01/2021

NOTES: (continued)

- This package is designed to be soldered to a thermal pad on the board. For more information, see Texas Instruments literature number SLUA271 (www.ti.com/lit/slua271).
- Vias are optional depending on application, refer to device data sheet. If any vias are implemented, refer to their locations shown on this view. It is recommended that vias under paste be filled, plugged or tented.

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