1.25-MHz Boost Converter White LED Driver with Internal Power Switch

APPLICATIONS

- Portable Phones and Game Devices
- PDAs and Palm-Top Computers
- Local Boost Regulator
- CCD Bias Supplies
- Digital Cameras
- TFT-LCD Displays
- DSL Modems
- PCMCIA Cards
- White LED Backlight
- OLED Driver

DESCRIPTION

The SiP12510/11 are a 1.25 MHz current-mode boost converter with a feedback reference voltage of 0.1 V which offers small size and high power conversion efficiency. Its input voltage range is from 2.5 V to 6 V, and output voltage can go up to 27.5 V for SiP12511 and 17 V for SiP12510. The internal frequency compensation minimizes number of external components. The integrated 33 V power switch can carry up to 0.55 A. The integrated power switch also features the over current limiting to protect itself. The internal soft-start circuit controls the rate of rise of the output voltage during start-up to prevent overshoot. The logic-level shutdown pin can be used to reduce quiescent current to < 1 µA and, effectively, extend battery life. Thermal shutdown at 165 °C is also included. The low FB voltage of 0.1 V improves the overall circuit efficiency. These features and more, make the SiP12510/11 an ideal power solution to white LED, OLED, LCD, and CCD applications operating from a single or dual cell lithium-ion battery.

SiP12510/11 are available in 6-pin TSOT23-6 package and are specified to operate over the industrial temperature range of - 40 °C to 85 °C.

FEATURES

- Output Voltage Range up to 27.5 V for SiP12511 and 17 V for SiP12510
- Current Mode Control with Internal Frequency Compensation
- 2.5 V to 6 V Input Voltage Range
- 1.25 MHz Switching Frequency
- Low Shutdown Current (< 1 µA)
- Under Voltage Lockout Protection
- Output Over Voltage Protection
- Thermal Shutdown Protection (165 °C)
- 0.55 A Switch Current Limiting
- High Efficiency up to 90 %
- Built-in Soft Start Control
- Minimum External Components
- TSOT23-6 Package

Pb-free
RoHS
COMPLIANT

TYPICAL APPLICATION CIRCUIT

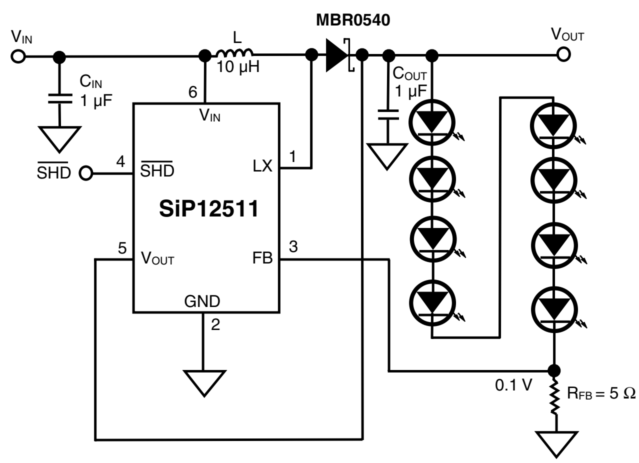


Figure 1. SiP12511 Typical Application Circuit

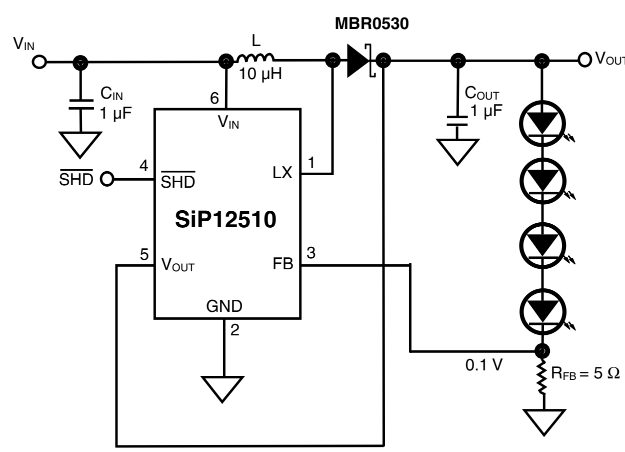


Figure 2. SiP12510 Typical Application Circuit

ABSOLUTE MAXIMUM RATINGS

Parameter	Limit	Unit
Input Voltage, V_{IN} to GND	- 0.3 to 12	V
V_{OUT}, LX Voltage	- 0.3 to 33	
$\overline{SHD}$ Voltage	- 0.3 to 12	
FB Voltage	- 0.3 to 12	
ESD (Human Body Model)[a]	2	kV
Maximum Junction Temperature	150	°C
Storage Temperature	- 55 to + 150	
Power Dissipation (T_A = 70 °C)[b]	367	mW
Junction to Ambient Thermal Impedance ($R_{\theta J}$)[c]	150	°C/W
Maximum Operating Junction Temperature	125	°C

Notes:
a. The human body model is a 100 pF capacitor discharged through a 1.5 kΩ resistor into each pin.
b. Derate 6.67 mW/°C above 70 °C.
c. Device mounted with all leads soldered or welded to PC board.

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.

RECOMMENDED OPERATING RANGE

Parameter	Limit		Unit
Input Voltage, V_{IN} to GND	2.5 to 6		V
$\overline{SHD}$	0 to V_{IN}		
V_{OUT}	SiP12511	SiP12510	
	V_{IN} to 27.5	V_{IN} to 17	V
LX	0 to 28		
Operating Temperature Range	- 40 to 85		°C

SPECIFICATIONS

Parameter	Symbol	Test Conditions Unless Specified V_{IN} = 5 V, $V_{\overline{SHD}}$ = 2 V, T_A = 25 °C	Temp	Min[a]	Typ[b]	Max[a]	Unit
Input Voltage	V_{IN}		Full	2.5		6	V
Switch Current Limit	I_{LIMIT}		Full	0.385	0.55	0.735	A
Switch On Resistance	$R_{DS (on)}$	I_{SW} = 100 mA	Full		0.75	1.25	Ω
$\overline{SHD}$ Input High Level	$V_{\overline{SHDH}}$		Full	1.5			V
$\overline{SHD}$ Input Low Level	$V_{\overline{SHDL}}$		Full			0.4	
Feedback Voltage	V_{FB}		Full	0.092	0.100	0.108	
Feedback Bias Current	I_{FB}				60		nA
Feedback Voltage Line Regulation	ΔV_{FB}/ (V_{FB} x ΔV_{IN})	5 ≥ V_{IN} ≥ 2.5 V and V_{OUT} = 15 V at 20 mA			0.2		%/V
Quiescent Current	I_Q	V_{FB} = 0 V (Switching)	Full		1.3	2	mA
		V_{FB} = 1.5 V (Not Switching)	Full		0.3	0.5	
		$V_{\overline{SHD}}$ = 0 V	Full			1	µA
Switching Frequency	F_{SW}		Full	1.05	1.25	1.45	MHz
Maximum Duty Cycle	D_{MAX}		Full	85	90		%
Switch Leakage	I_{LEAK}	Not Switching, V_{LX} = 5 V	Room			1	µA
			Full			10	
Thermal Shutdown	T_{SHD}				165		°C

Document Number: 74498
S-70770–Rev. A, 30-Apr-07

SPECIFICATIONS

Thermal Shutdown Hysteresis	T_{HYST}				20		°C
Under Voltage Lockout	V_{UVLO}		Full	2.00	2.24	2.48	V
UVLO Hysteresis	$V_{UVLOHYST}$				0.1		
Over Voltage Protection	V_{OVLO}	SiP12510	Full	17	19	21	
		SiP12511	Full	27.5	30	32	
OVLO Hysteresis	$V_{OVLOHYST}$				0.2		

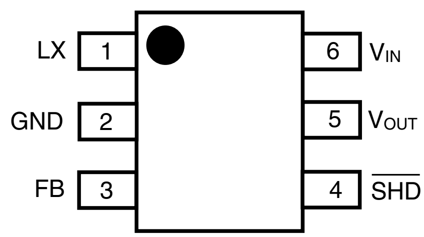
Notes:

a. Limits are guaranteed by testing.

b. Typical values are derived from the mean value of a large quantity of samples tested during characterization and represent the most likely expected value of the parameter.

PIN CONFIGURATION

TSOT23-6 Package



TOP VIEW

Figure 3.

PIN DESCRIPTION

Pin Number	Name	Function
1	LX	Drain Pin of the Internal Switch. Connect inductor/diode to LX. Minimize trace area at this pin to keep electromagnetic interference down to a minimum.
2	V_{IN}	Analog and power input of the controller IC. A bypass capacitor is required on this pin.
3	$\overline{SHD}$	Logic Controlled Shutdown Input. $\overline{SHD}$ = high: Normal operation. $\overline{SHD}$ = low: Shutdown.
4	FB	Voltage Feedback Pin, the inverting input of the voltage error amplifier. This is internally compared against a voltage of 0.1 V appearing on the voltage error amplifier's non-inverting input. External resistors are connected to this pin to set the regulated output voltage.
5	V_{OUT}	Output Voltage Pin, Output voltage sense for over voltage protection and the slop compensation.
6	GND	Signal and Power Ground, this pin acts as both the analog ground and the power ground for this part.

ORDERING INFORMATION

Part Number	Marking	Temperature Range	Package
SiP12510DT-T1-E3	M3WXX	- 40 °C to 85 °C	TSOT23-6
SiP12511DT-T1-E3	M4WXX		

XX = Lot Code
W = Work Week code

FUNCTIONAL BLOCK DIAGRAM

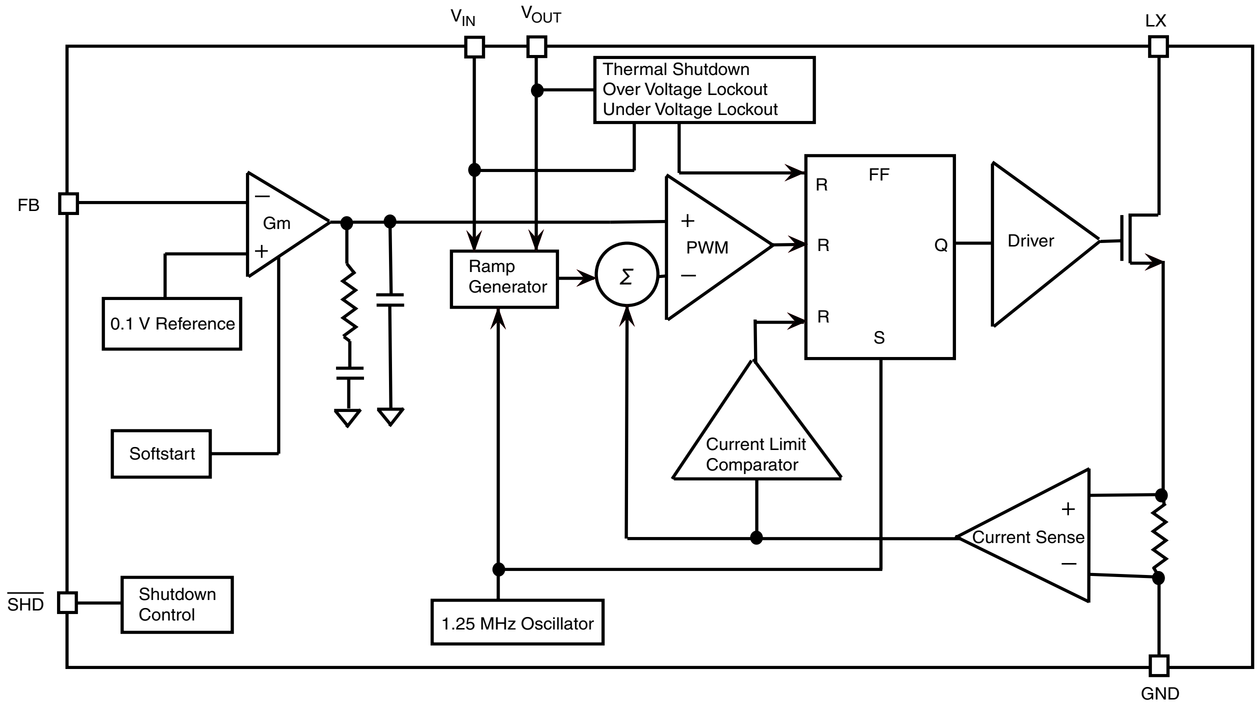


Figure 4. Internal Block Diagram

DETAILED OPERATION DESCRIPTION

The SiP12510/11 is a current mode, internally compensated, step-up switching converter that operates at a fixed frequency of 1.25 MHz. The current mode topology allows for fast transient response over a wide input range and provides a real-time, cycle-by-cycle current limiting function. The operation of the converter can be described through the interaction of two separate internal loops: the current sense loop and voltage sense loop.

Within the current sense loop, the switch MOSFET current is monitored by sensing the voltage across an internal current sense resistor, which is fed to the inputs of both the current limit amplifier and the pulse width modulation (PWM) comparator.

At the beginning of each switch cycle, the oscillator sets the S-R latch thereby turning on the MOSFET. As current through the switch increases, so does the voltage drop across the sense resistor. This voltage is summed with the ramp coming from the ramp generator and applied to the input of the PWM comparator. When this ramping voltage exceeds the output of the error amplifier, the latch changes state and turns off the MOSFET. The slope of the ramp generator is proportional to voltages on the V_{IN} and V_{OUT} pins, therefore, any sudden changes in input or output voltage can be corrected and accommodated for on a cycle-by-cycle basis. If the MOSFET current surpasses the current limit threshold, the current limit comparator will unconditionally turn off the internal power switch. At the beginning of the next oscillator cycle, the switch is allowed to turn on again.

The voltage feedback loop works by monitoring the LED drive current through a resistor divider on FB and comparing that voltage with an internal 0.1 V reference voltage (V_{ref}). If the LED current falls below the set current, the voltage on the feedback pin will drop slightly below V_{ref} causing the output of error amplifier to increase. This will keep the PWM comparator's output high for a greater portion of an oscillator cycle, thus ensuring that the MOSFET will stay on longer. This means that the duty cycle of the LED driver converter will increase. The output voltage of LED driver converter will increase. This, in turn, will allow more current to be delivered to the load. Following similar logic, should the LED current become higher than the set current, FB voltage will increase above Vref, the converter will decrease its duty cycle, which will lessen the energy delivered to the load at each cycle, and thereby, reduce LED current and maintain desired brightness.

In essence, by modifying the on time of the switch, the PWM comparator continually sets the correct maximum current through the MOSFET to regulate the LED current to a desired value.

SIP12510/11 FUNCTIONAL DESCRIPTION

SiP12510/11 is the combination of the high voltage low-side MOSFET and the current mode PWM controller. The current mode PWM controller consists of error amplifier, 0.1 V reference voltage, 1.25 MHz oscillator, ramp generator, current sense amplifier, PWM, and current limit comparator, thermal protection, over voltage protection and MOSFET driver.

OSCILLATOR

The typical oscillator frequency is internally set to 1.25 MHz. The output of the oscillator is not only used to drive the input of the built-in MOSFET driver. It is also set the operating frequency of the ramp generator for the slop compensation.

0.1 V REFERENCE VOLTAGE

The 0.1 V reference voltage is connected on the noninverting input of the error amplifier for output voltage regulation. The typical application of the SiP12510/11 boost converter is to driver the white LED. The output voltage of the boost converter is set by the forward drop voltage of the LED in series in the feedback voltage divider. The LED load is in series with the resistor of the voltage divider. The lower the reference is, the higher the efficiency of the converter circuit. Based on the efficiency measurement, the efficiency of the converter can achieve 90 % (see efficiency curve in the typical waveform section). So the efficiency is benefited from 0.1 V low reference voltage.

CURRENT LIMIT COMPARATOR AND CURRENT SENSE AMPLIFIER

The current limit comparator and the current sense amplifier are design to protect the built-in MOSFET and the converter from over current operation condition. The current sense amplifier not only monitors the switch current for cycle-by-cycle current mode operation. It also senses the switch current for cycle-by-cycle over current protection for the built-in switch. The typical maximum over current protection threshold is set to 0.55 A. Once the maximum current exceeds 0.55 A, the current limit comparator will shut down the gate drive signal for the built-in MOSFET to protect the built-in MOSFET.

OVER VOLTAGE PROTECTION

SiP12510/11 have a built-In output voltage protection to shut down the controller. The minimum over voltage protection threshold is 27.5 V for SiP12511 and 17 V for SiP12510. The protection will completely shut down the controller and restart if the over voltage fault occur. The maximum voltage rating for the built-in power MOSFET is 33 V for both SiP12510/11. Any potential over voltage on the output of the converter will not be able to damage built-in MOSFET.

In the event of an output open circuit (e.g. when the LEDs are either disconnected form the output or an LED fails), the feedback voltage will become zero causing the SiP12510/11 to go to maximum duty cycle. This would generally result in a high output voltage and, possibly, cause the voltage on the LX pin to exceed it's absolute maximum rating and damage the part. However, the SiP12510/11 have a built-in over-voltage protection circuitry that will clamp the output to 19 V typical for SiP12510 and 30 V typical for SiP12511. SiP12510/11 guarantee safe operation under open-circuit conditions.

THERMAL PROTECTION

The thermal protection circuit senses the die temperature. The temperature threshold is set to 165 °C typical with 20 °C hysteresis. The built-In MOSFET will be disabled when the temperature exceeds 165 °C remain disabled until the die temperature drop below 145 °C to re-enable.

START UP AND SOFT-START

When voltage is applied to the V_{IN} pin, the under voltage lockout (UVLO) circuit prevents the controller's output switch and oscillator circuit from turning on until the voltage on the V_{IN} pin exceeds 2.24 V. Provided the V_{IN} pin is above this threshold, when $\overline{SHD}$ pin is raised high, soft-start is initiated. Soft-start is achieved by slowly ramping up the internal reference. For a certain period of soft-start time (about 0.7 mS), the value of over-current protection threshold is being changed twice: it is about 40 % of its steady state value during the first phase of this period of soft-start time, and about 66 % of its steady state value during the second phase of soft-start period of time. The heavy load is applied on the output of SiP12511 WLED driver converter to show the current limiting phase in soft-start period in Figure 5. Once the soft-start time has elapsed, SiP12510/11 enters into a normal state of operation. The converter then operates continuously unless the voltage on V_{IN} drops below 2.24 V or $\overline{SHD}$ is set low. UVLO hysteresis prevents the converter from dropping in and out of start-up, unintentionally locking up the system.

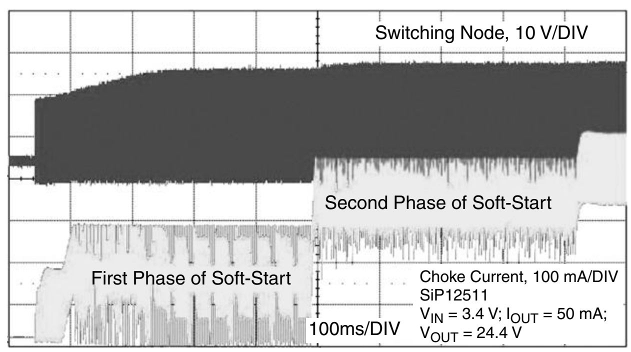


Figure 5.

KEY APPLICATION CALCULATION

INPUT CAPACITOR SELECTION

The input bypass capacitor acts as an energy reservoir that satisfies the transient inductor current needs each time the switch turns on. In effect, the input capacitor is responsible for reducing the input voltage ripple and the amount of EMI that is inevitably passed to other circuitry on that line. For this purpose, a 1 µF capacitance minimum is recommended as input capacitor. A ceramic capacitor is recommended. If preferred, tantalum capacitors may be used instead of ceramics.

OUTPUT CAPACITOR SELECTION

To curb output voltage ripple, a multi-layer ceramic capacitor should be used as the output filter capacitor. Ceramic capacitors are favored for their low ESR (equivalent series resistance) and high resonance frequency, which makes them ideal for high frequency switching converters. A high ESL (equivalent series inductance) can give rise to ringing in the low megahertz region and a high ESR could reduce phase margin and potentially cause instability of the design. In addition, the ripple current flowing through the capacitor's ESR causes power dissipation and heats up the capacitor internally. If the ripple current ratings of the capacitor are exceeded, the excessive temperature could shorten the expected life of the capacitor.

If a high value capacitor is required for improved transient response, to keep component costs down and to save PC board real estate, tantalum capacitor may be used in parallel with ceramics. If the maximum tolerated ripple current (I_{P-P}) and ripple voltage (ΔV_{OUT}) design specifications are known, the maximum tolerated ESR on the output capacitor and its value can be calculated using the following formulas:

$$C_{OUT(MIN)} = \frac{I_{OUT(MAX)} \times D_{MAX}}{f_{SW} \times \Delta V_{OUT}}$$

and

$$ESR_{(MAX)} = \frac{\Delta V_{OUT}}{\frac{1}{1 - D_{MAX}} \times I_{OUT(MAX)} + \frac{I_{P-P}}{2}}$$

Where
$I_{OUT(MAX)}$ is the maximum output current.
$D_{(MAX)}$ is the maximum duty cycle.
f_{SW} is switching frequency.
D_{VOUT} is the output ripple voltage.
$C_{OUT(MIN)}$ is the minimum output capacitance.
$ESR_{(MAX)}$ is the maximum equivalent resistance of the output capacitor.
I_{P-P} is the peak-to-peak value of the choke current.
The formulas above are used to figure out the minimum out capacitance value. To reduce the out ripple voltage and improve the stability, it is recommended to used the larger capacitance for output capacitor.

DUTY CYCLE CALCULATION

In continuous mode of operation, the maximum duty cycle of a boost switching regulator determines the maximum amount of boost (V_{OUT}/V_{IN}) attainable and can be calculated using the expression

$$D_{(MAX)} = \frac{V_{OUT} + V_{DIODE} - V_{IN(MIN)}}{V_{OUT} + V_{DIODE}}$$

Where V_{DIODE} is the forward bias voltage of the schottky diode and $V_{IN(MIN)}$ is the minimum operating input voltage of the converter.

$$D_{MAX} = 1 - \text{Efficiency} \times \frac{V_{IN}}{V_{OUT}}$$

The above equation yields only an approximation of the duty cycle since it ignores power loss terms resulting from wire losses in the inductor, switching losses of the internal FET, and capacitor ripple current losses due to their inherent non-zero ESR. A more accurate estimate of the duty cycle can be determined byAnd by using the provided efficiency curves to approximate efficiency for a given input and output voltage.

DIODE SELECTION

A schottky diode is recommended for use as the external rectifier. Schottky diodes are typically preferred in DC-DC conversion applications because of their low forward voltage drop and fast recovery time, which allows for high frequency switching. In choosing a diode, ensure that the diode's reverse breakdown voltage exceeds the intended V_{OUT} of design and that its current rating is greater then the peak inductor current. SiP12510/11 have the typical 0.55 A maximum over current limiting function for the integrated power MOSFET. The maximum current of the inductor is limited to 0.55 A maximum typically. The relationship for the maximum diode current, maximum switch current and output current can be expressed as the following:

$$I_{Q(MAX)} = I_{DIODE(MAX)} = I_{L(MAX)}$$

and

$$I_{DIODE(MAX)} = \frac{I_{OUT}}{1 - D_{MAX}} + \frac{(1 - D_{MAX}) \times D_{MAX} \times V_{OUT}}{2 \times L \times f_{SW}}$$

The average current rating of the output diode shall be equal to the output current. This is because the energy in the inductor only delivers to the output through the output diode when the switch is off.

$$I_{DIODE(AVG)} = I_{OUT}$$

For typical application of SiP12511, MBR0540 is recommended. For typical application of SiP12510, MBR0530 is recommended.

INDUCTOR SELECTION

An inductor is one of the energy storage components in a converter. Choosing an inductor means specifying its size, structure, material, inductance, saturation level, DC-resistance (DCR), and core loss. Choosing the right inductor is not a simple task and involves trade-offs in performance. The following are some key parameters that should be focused on. In PWM mode, inductance has a direct impact on the ripple current. The inductor value can be calculated as

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

Where V_{SW} is the voltage drop across the switch in its onstate, f_{sw} is the switching frequency, and D is the duty cycle. Higher inductance means lower ripple current, lower rms current, lower voltage ripple on both input and output, and higher efficiency, unless the resistive loss of the inductor dominates the overall conduction loss. However, higher inductance also means a larger inductor size and a slower transient response. For fixed line, load, and frequency conditions, higher inductance results in a lower peak current for each pulse and a higher load capability. The saturation current is another important parameter inchoosing inductors. Note that the saturation levels specified in data sheets are maximum currents. For a DC-DC converter operating in PWM mode, it is the maximum peak inductor current that is relevant, and which can be calculated using these equations:

$$I_{L(MAX)} = \frac{I_{OUT(MAX)}}{1 - D_{MAX}} + \frac{I_{P-P}}{2}$$

or

$$I_{L(MAX)} = \frac{I_{OUT(MAX)}}{1 - D_{MAX}} + \frac{D_{MAX} \times V_{IN(MIN)}}{2 \times f_{SW} \times L}$$

Where

$I_{L(MAX)}$ is the maximum current in the choke.

$I_{OUT(MAX)}$ is the maximum output current.

This peak current varies with inductance tolerance and other errors, and the rated saturation level varies over temperature. So a sufficient design margin is required when choosing current ratings. A high-frequency core material, such as ferrite, should be chosen, the core loss could lead to serious efficiency penalties. The DCR should be kept as low as possible to reduce conduction losses.

KEY APPLICATION CONSIDERATION

LAYOUT CONSIDERATIONS

In high frequency switching regulators such as the SiP12510/11, great attention must be given to the layout process in order to ensure stable operation and minimize noise. Since most power traces in step up converters carry pulsating current, energy stored in trace inductance during the pulse can cause high-frequency ringing with input and output capacitors. This effect can generally be curbed by minimizing the length and increasing the width of power traces.

To minimize stray capacitance and even more importantly, parasitic trace inductance, all components must be kept as close to the switcher as possible. Of special importance, is the path between the switching node LX, D1, C1,C2, and ground of the regulator; the length of this path must be kept as small as possible since any parasitic inductance in series with the diode and output capacitance will increase noise and produce ringing in the circuit. Pulsating currents in the ground trace can cause voltage drops due to trace resistance and cause ground bounce. For this reason, it is strongly recommended to use a separate ground plane.

As an example, Figure 6 and 7 demonstrate a recommended schematic and related layout of components. It is urged that this layout be followed closely as possible to obtain best performance.

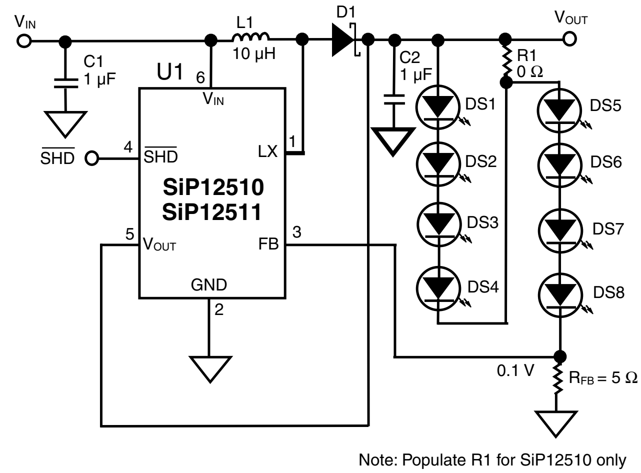


Figure 6.

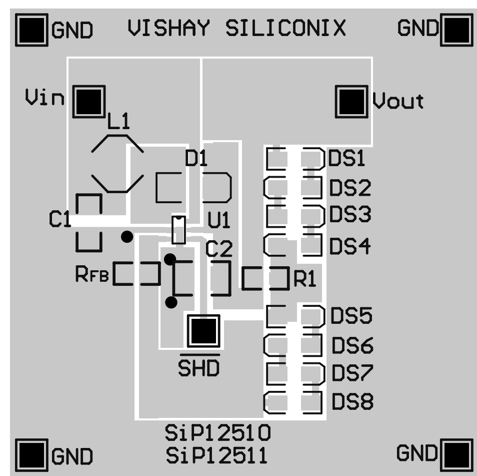


Figure 7.

LED CURRENT CONTROL

The SiP12510/11 is a white LED driver. The low feedback voltage of 0.1 V is designed to reduce losses outside of the white LEDs and thus improve overall circuit efficiency. The LED current is set by the small sense resistor on FB and can be calculated using the following expression:

$$I_{LED} = \frac{V_{REF}}{R_{FB}} = \frac{0.1\,V}{R_{FB}}$$

In order to have accurate LED current, use of 1 % precision resistor is recommended.

As shown in Figures 1 and 2, the SiP12510 can be used to drive four LEDs in series or to drive parallel strings of LEDs. And SiP12511 can be used to drive up to seven LED in series.

WHITE LED BRIGHTNESS CONTROL

Figures 8 and 9 delineate two possible brightness controlschemes. In Figure 8, a PWM signal is injected into the shutdown pin. The average LED current is proportional to the duty cycle of the PWM signal and thus, the brightness will vary from low to high as the duty cycle of the PWM signal is increased. The frequency of the PWM signal has to be low enough to allow the part to undergo soft-start and fully power up at each cycle. A frequency of 100 Hz to 500 Hz is, therefore, recommended. The magnitude of the PWM signal should be higher than the maximum enable voltage of $\overline{\text{SHD}}$ pin, in order to let the dimming control perform correctly.

In Figure 9, a more analog approach to brightness control. As the control voltage V_{CTRL} is increased from 0 V, the voltage drop across R2 and R3 increases driving voltage on node A low thereby reducing current through the White LEDs and dimming brightness. Reducing V_{CTRL} to about 0 V, will turn the LEDs fully on with 20 mA of current. The equation for the LED current can be expressed as

$$I_{LED} = \frac{0.1\,V}{5\,\Omega} + \frac{R2}{R3} \times \frac{(0.1\,V - V_{CTRL})}{5\,\Omega}$$

Figure 10 demonstrates a more practical approach for dimming control,which is really the synthesis of the two ideas demonstrated above. In this approach, a filtered PWM signal acts as a DC voltage to control the brightness of the LEDs. It is recommended that PWM signal with frequency higher than 22 kHz be used. Figures 11 illustrate another ideal to power up WLED and additional load, which required constant output voltage.

POWER DISSIPATION CONSIDERATIONS

An important consideration when designing power converters is the maximum allowable power dissipation of a part. The maximum power dissipation in any application is dependant on the maximum junction temperature, $T_{J(MAX)}$ = 125 °C, the junction-to-ambient thermal resistance for the TSOT-23 package, $\theta_{J\text{-}A}$ = 150 °C/W, and the ambient temperature, T_A, which may be formulaically expressed as:

$$P_{(MAX)} = \frac{T_{J(MAX)} - T_A}{\theta_{J-A}} = \frac{125\,°C - T_A}{150\,°C/W}$$

It then follows that, assuming an ambient temperature of 70 °C, the maximum power dissipation will be limited to about 0.37 W. In the event that the power dissipation exceeds the value specified above and the die temperature reaches 165 °C, the internal thermal protection circuitry will ensure safe operation by turning off the internal MOSFET, thereby maintaining junction temperature at a safe level. In this state, only the system monitor circuitry will be active. Once the temperature of the SiP12510/11 drops below 145 °C, the SiP12510/11 re-enters soft-start mode and resumes normal operation.

WHITE LED BRIGHTNESS CONTROL SCHEMATIC

PWM ON/OFF CONTROL

When low frequency PWM signal is available.

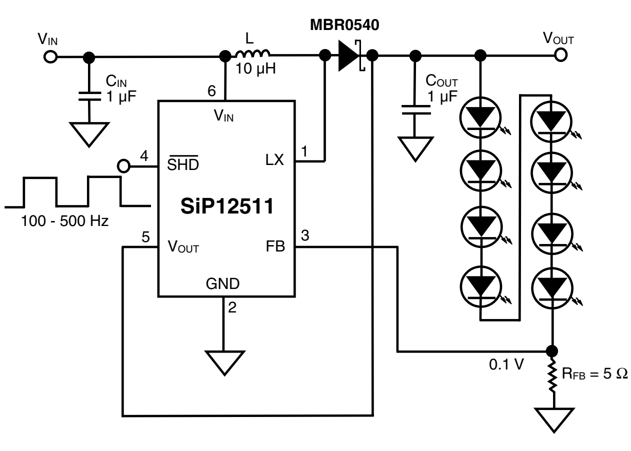


Figure 8.

PWM FEEDBACK VOLTAGE CONTROL

When high frequency PWM signal is available.

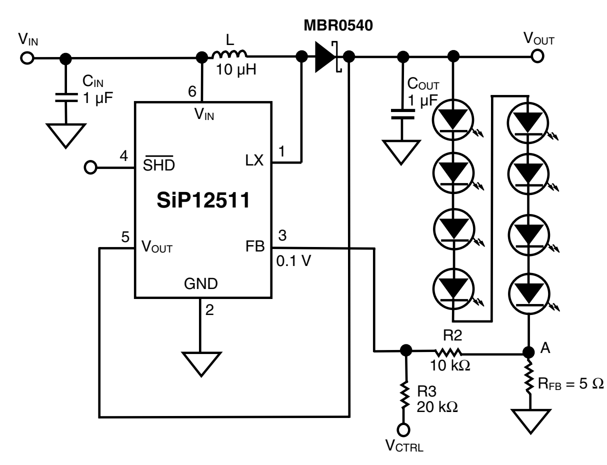


Figure 9.

LINEAR FEEDBACK VOLTAGE CONTROL
When PWM signal is not available.

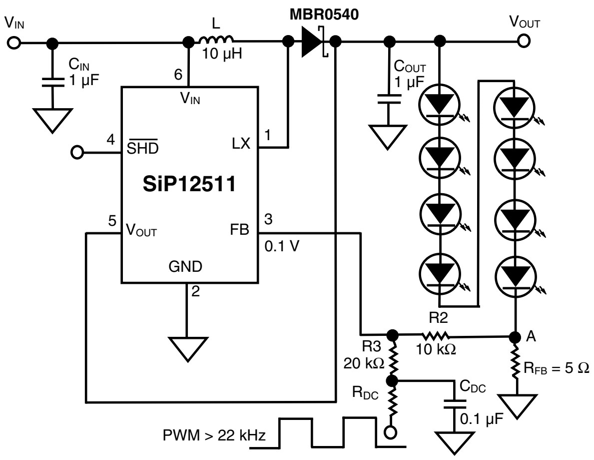


Figure 10.

PWM CONTROL WITH CONSTANT OUTPUT VOLTAGE
When the output is used to power up another load.

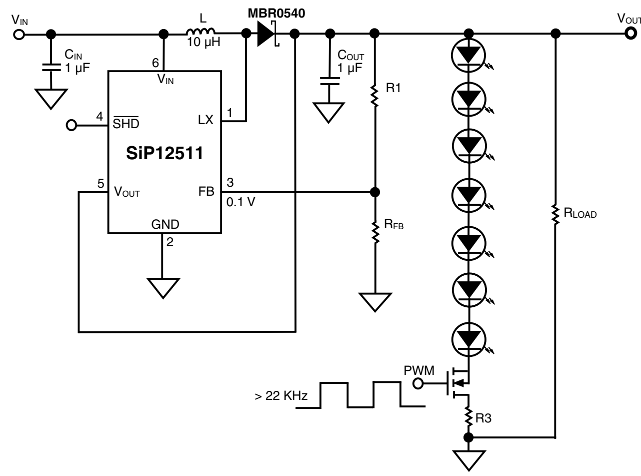
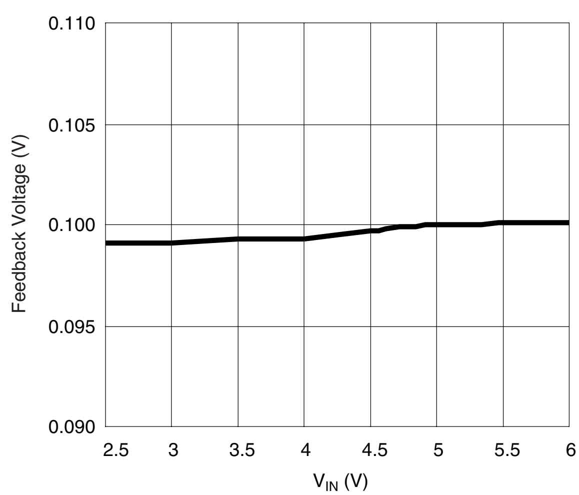
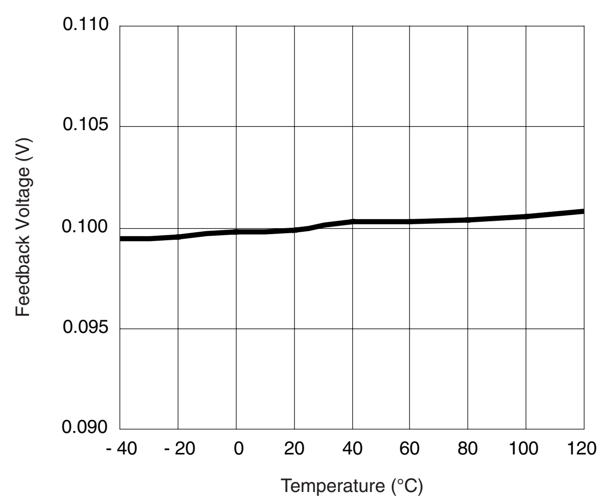


Figure 11.

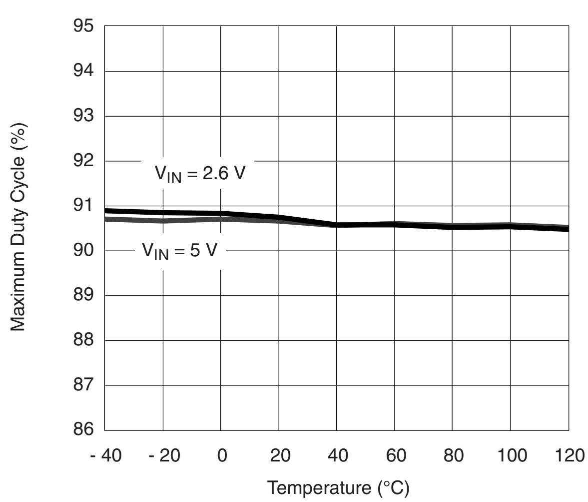
TYPICAL CHARACTERISTICS



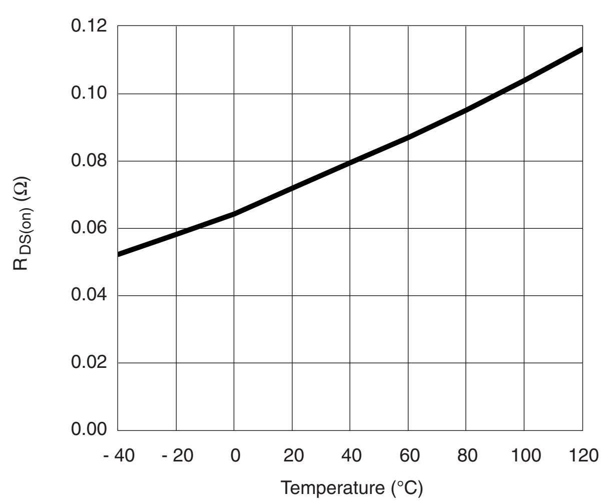
Feedback vs. Input Voltage



Feedback Voltage vs. Temperature

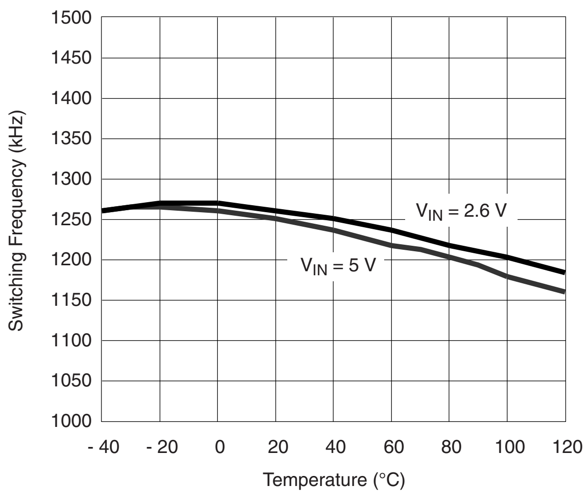


Maximum Duty Cycle vs. Temperature

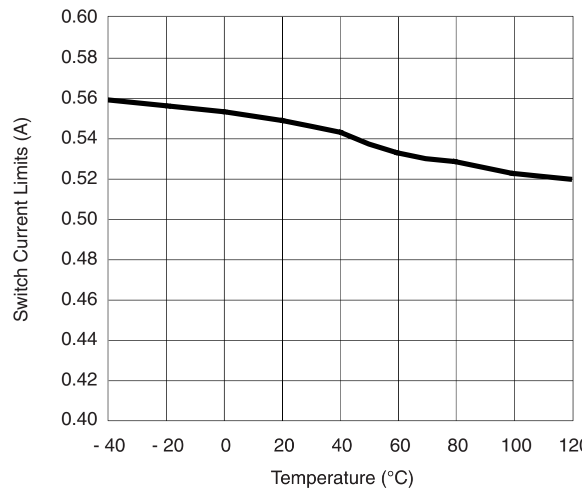


Switching on Resistance vs. Temperature

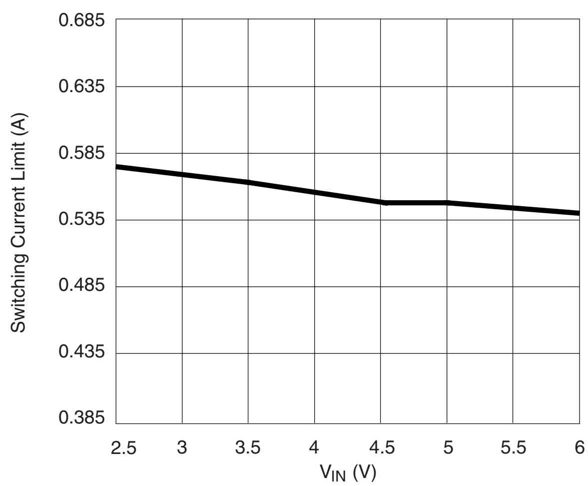
TYPICAL CHARACTERISTICS



Switching Frequency vs. Temperature

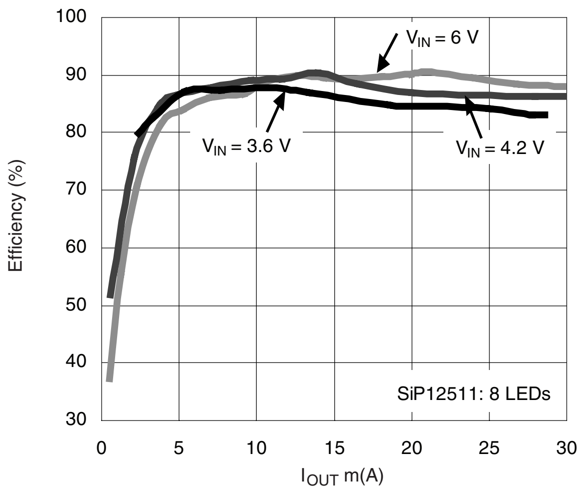


Switch Current Limit vs. Temperature

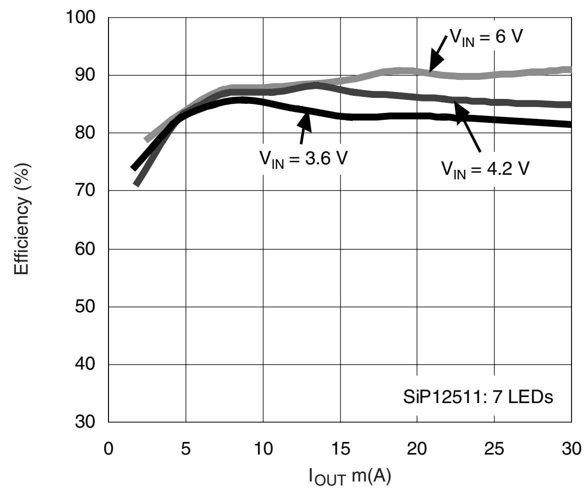


Switch Current Limit vs. Input Voltage

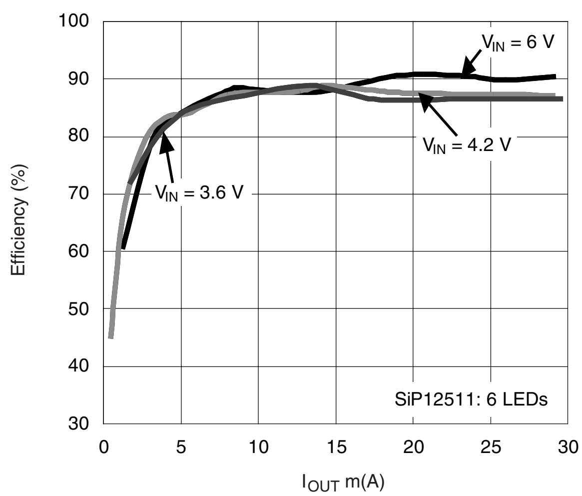
TYPICAL WAVEFORMS



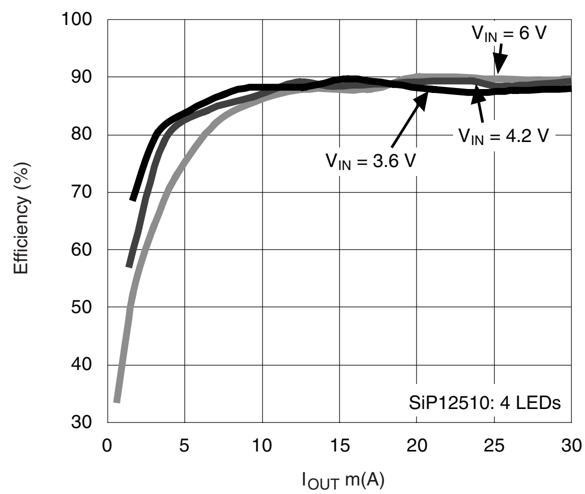
Efficiency vs. Output Current



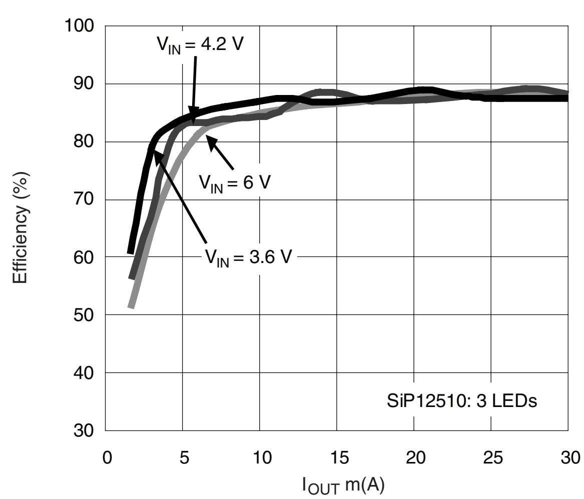
Efficiency vs. Output Current



Efficiency vs. Output Current

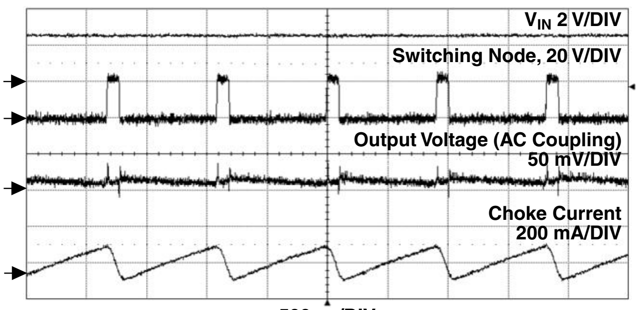


Efficiency vs. Output Current

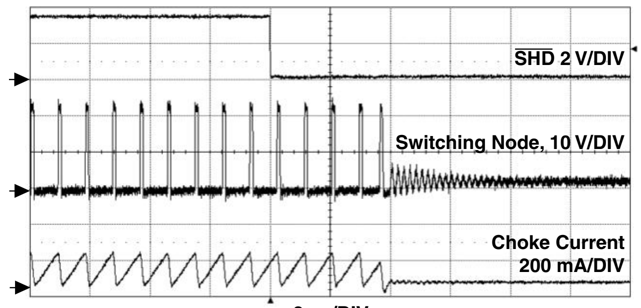


Efficiency vs. Output Current

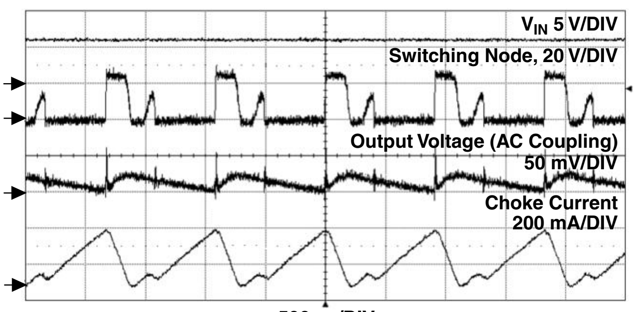
TYPICAL WAVEFORMS



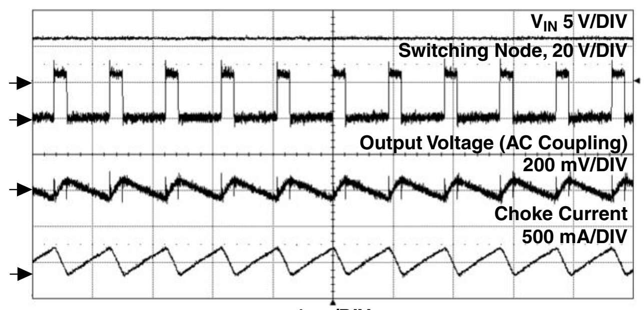
500 ns/DIV
Steady State Operation
V_{IN} = 2.5 V, 7 LEDs In Series. I_{OUT} = 20 mA



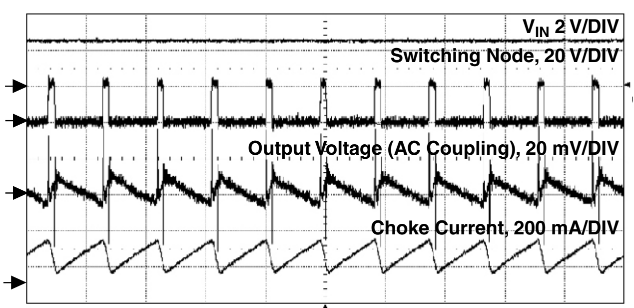
2 µs/DIV
Power-Down
V_{IN} = 2.5 V, 7 LEDs In Series. I_{OUT} = 20 mA



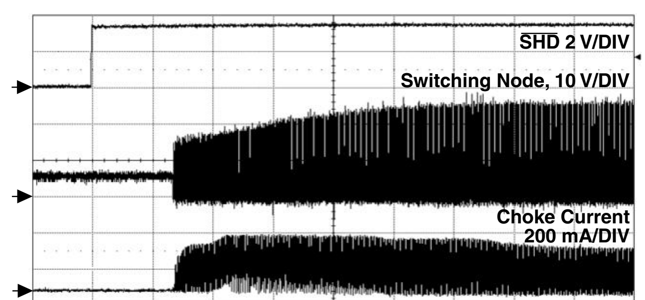
500 ns/DIV
Steady State Operation
V_{IN} = 6 V, 7 LEDs In Series. I_{OUT} = 20 mA



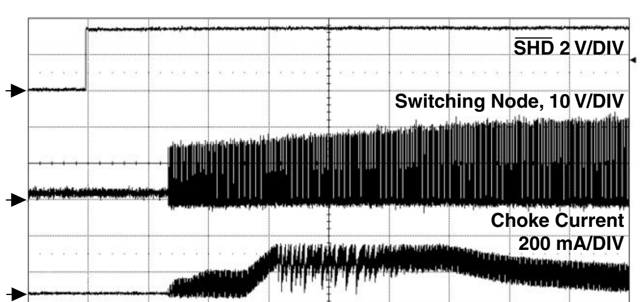
1 µs/DIV
Steady State Operation
V_{IN} = 6 V, 7 LEDs In Series, 5 Series In parallel, I_{OUT} = 100 mA



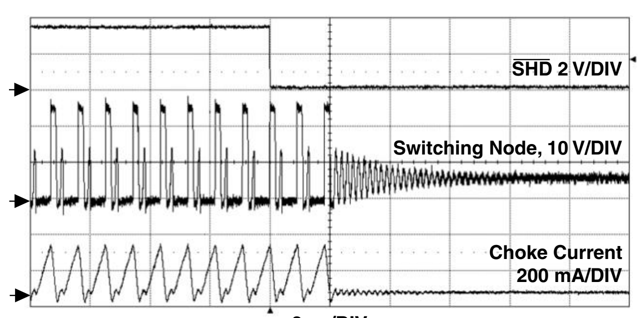
1 µs/DIV
Steady State Operation
V_{IN} = 2.5 V, 7 LEDs In Series, 5 Series In parallel, I_{OUT} = 100 mA



50 µs/DIV
Start-Up
V_{IN} = 6 V, 7 LEDs In Series. I_{OUT} = 20 mA

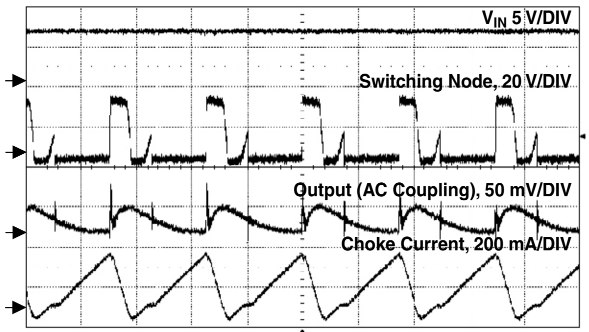


50 µs/DIV
Start-Up
V_{IN} = 2.5 V, 7 LEDs In Series. I_{OUT} = 20 mA

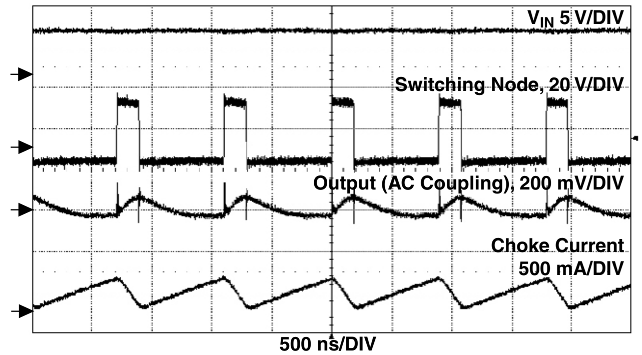


2 µs/DIV
Power-Down
V_{IN} = 6 V, 7 LEDs In Series. I_{OUT} = 20 mA

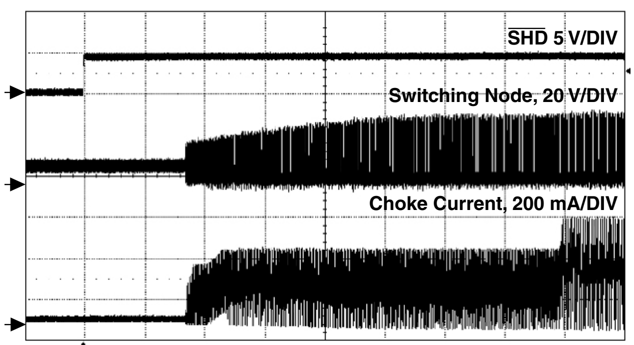
TYPICAL WAVEFORMS



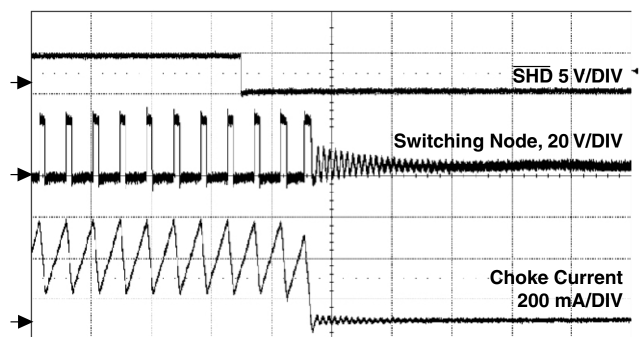
Steady State Operation
V_{IN} = 6 V, 8 LEDs In Series, I_{OUT} = 20 mA



Steady State Operation
V_{IN} = 6 V, 8 LEDs In Series, 3 Series in parallel, I_{OUT} = 60 mA



Start-Up
V_{IN} = 6 V, 8 LEDs In Series, 3 Series in parallel, I_{OUT} = 60 mA



Power-Down
V_{IN} = 6 V, 8 LEDs In Series, 3 Series In parallel, I_{OUT} = 60 mA

DESIGN CHECKLIST AND REVIEW

To design any power conversion circuit, it is necessary to verify the proper functionality on the bench to ensure the solidity of the design. The following are the some tests recommend proving the proper functionality of SiP12510/11 design.

1. Stability: Power up the SiP12510/11 white LED driver design. Monitor the switching node waveform in LX pin. Use the current probe to monitor the current waveform in the choke. Change the input voltage and white LEDs load in the designed circuit. Make sure the waveforms are stable in any designed line/load condition. Instable waveforms sometimes are caused by the improper design of the choke current. The current limiting circuit tries to shut down itself.

2. Current limit threshold: Make sure the peak current of the choke does not exceed the minimum current limit of the built-in switch in any designed line/load conditions. It is recommended to have some design margin. Exceed of the current limit of the switch will result instable switching waveforms.

3. Duty Cycle: Make sure the duty cycle is not locked in any designed line/load condition. The operating duty cycle in any designed line/load condition must be less than 85 %, which is the minimum value of the maximum duty cycle.

4. Output Voltage Check: Monitor the output voltage of the converter. Make sure the output voltage is less than the minimum over voltage protection threshold voltage, which is 17 V for SiP12510 and 27.5 V for SiP12511.

Disclaimer

All product specifications and data are subject to change without notice.

Vishay Intertechnology, Inc., its affiliates, agents, and employees, and all persons acting on its or their behalf (collectively, “Vishay”), disclaim any and all liability for any errors, inaccuracies or incompleteness contained herein or in any other disclosure relating to any product.

Vishay disclaims any and all liability arising out of the use or application of any product described herein or of any information provided herein to the maximum extent permitted by law. The product specifications do not expand or otherwise modify Vishay’s terms and conditions of purchase, including but not limited to the warranty expressed therein, which apply to these products.

No license, express or implied, by estoppel or otherwise, to any intellectual property rights is granted by this document or by any conduct of Vishay.

The products shown herein are not designed for use in medical, life-saving, or life-sustaining applications unless otherwise expressly indicated. Customers using or selling Vishay products not expressly indicated for use in such applications do so entirely at their own risk and agree to fully indemnify Vishay for any damages arising or resulting from such use or sale. Please contact authorized Vishay personnel to obtain written terms and conditions regarding products designed for such applications.

Product names and markings noted herein may be trademarks of their respective owners.