TPS4019x 4.5V～18V 入力、電圧モード、パワー・グッド付きの同期整流降圧コントローラ

1 特長

- 動作入力電圧範囲: 4.5V～18V
- 最大20Aの出力電流
- プリバイアス出力をサポート
- 0.5%、591mVの基準電圧
- スイッチング周波数
  - TPS40192: 600kHz
  - TPS40193: 300kHz
- 3つの熱補償付きの短絡保護レベルを選択可能
- フォルトからのヒカップ再起動
- 5Vレギュレータ内蔵
- ハイサイドおよびローサイドのMOSFETオン抵抗($R_{DS(on)}$) 電流センシング
- 10ピンの3mm×3mm SONパッケージ
- 4msの内部ソフトスタート時間
- 145℃でのサーマル・シャットダウン保護機能

2 アプリケーション

- ケーブル・モデムCPE
- デジタル・セットトップ・ボックス
- グラフィック/オーディオ・カード
- エントリ・レベルおよびミッドレンジのサーバー

3 概要

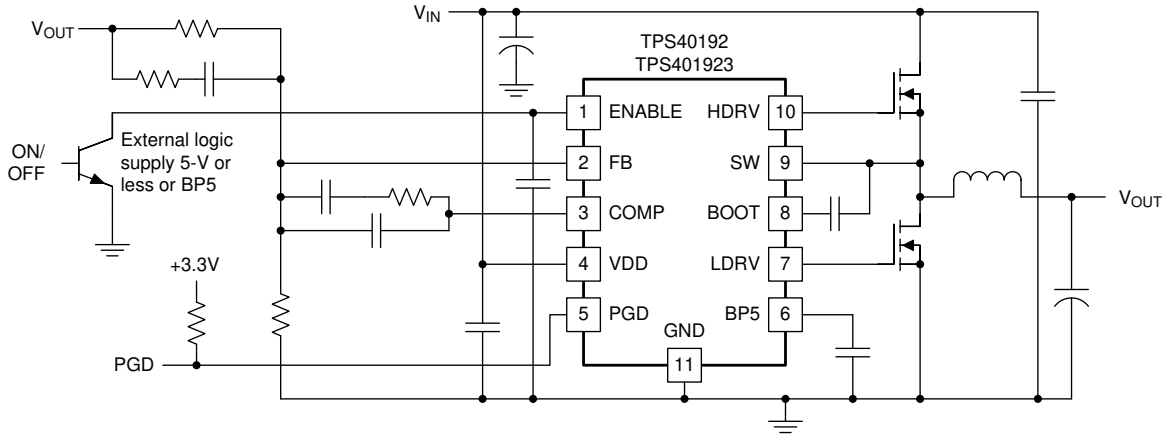
TPS40192およびTPS40193は、コストに対して最適化された同期整流降圧コントローラで、4.5V～18Vの入力電圧で動作します。これらのコントローラには電圧モード制御アーキテクチャが実装されており、スイッチング周波数は600kHz (TPS40192)または300kHz (TPS40193)に固定されています。スイッチング周波数が高いため、小型のインダクタおよび出力コンデンサを使用でき、コンパクトな電源ソリューションを構築できます。適応型のクロス導通防止方式により、パワーFETを流れる貫通電流が防止されます。

製品情報(1)

型番	パッケージ	本体サイズ(公称)
TPS40192	VSON (10)	3.00mm×3.00mm

(1) 提供されているすべてのパッケージについては、データシートの末尾にある注文情報を参照してください。

アプリケーション概略図



目次

4 改訂履歴

資料番号末尾の英字は改訂を表しています。その改訂履歴は英語版に準じています。

Revision G (January 2019) から **Revision H** に変更 **Page**

- 編集上の変更のみ、技術上の変更なし ... 1

Revision F (November 2016) から **Revision G** に変更 **Page**

- Deleted last sentence in *Enable Functionality* ... 11
- 削除 表 6からTPS40190を ... 27

Revision E (May 2013) から **Revision F** に変更 **Page**

- 「*ピン構成および機能*」セクション、「*ESD定格*」表、「*機能説明*」セクション、「*デバイスの機能モード*」セクション、「*アプリケーションと実装*」セクション、「*電源に関する推奨事項*」セクション、「*レイアウト*」セクション、「*デバイスおよびドキュメントのサポート*」セクション、「*メカニカル、パッケージ、および注文情報*」セクション 追加 ... 1

Revision D (July 2012) から **Revision E** に変更 **Page**

- Added clarity to Figure 15 ... 16
- Added note regarding high-resistance resistor ... 20

Revision C (August 2010) から **Revision D** に変更 **Page**

- Added text to the last paragraph in the Enable Functionality section. ... 11

Revision B (September 2007) から **Revision C** に変更 **Page**

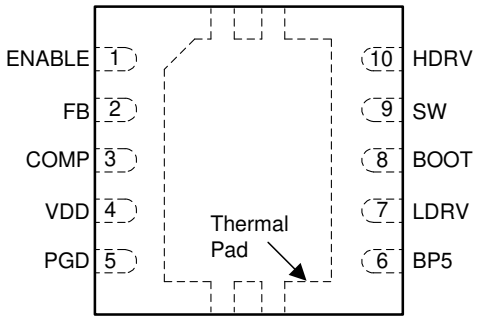
- Changed corrected label for pin 8 ... 4
- Changed corrected waveform ... 12

5 概要（続き）

短絡検出のため、ローサイドMOSFETがオンのときに、その両端での電圧降下が検出され、ユーザーの選択したスレッショルドである100mV、200mV、280mVのいずれかと比較されます。このスレッショルドは、COMPからGNDに接続されている、単一の外付け抵抗により設定されます。この抵抗はスタートアップ時に検出され、選択されたスレッショルドがラッチされます。パルス単位の制限(電流暴走抑止のため)は、ハイサイドMOSFETがオンのときに両端の電圧を検出し、電圧降下が固定スレッショルドの550mVを超過したときサイクルを中断することで行われます。コントローラにより出力の短絡が検出された場合、両方のMOSFETがオフになり、タイムアウト期間が終了するまで再起動は試みられません。この動作により、持続的なフォルトが発生しても消費電力が制限されます。

6 Pin Configuration and Functions

DRC Package
10-Pin VSON
Top View



Pin Functions

PIN		I/O	DESCRIPTION
NAME	NO.		
BOOT	8	I	Gate drive voltage for the high-side N-channel MOSFET. A 100-nF typical capacitor must be connected between this pin and SW.
BP5	6	O	Output bypass for the internal regulator. Connect a capacitor with a value of at least 1-μF from this pin to GND. Larger capacitors (up to 4.7 μF) can improve noise performance when using a low-side MOSFET with a gate charge of 25 nC or greater. Low power, low noise loads may be connected here if desired. The sum of the external load and the gate drive requirements must not exceed 50 mA. This regulator is turned off when ENABLE is pulled low.
COMP	3	O	Output of the error amplifier.
ENABLE	1	I	Logic level input which starts or stops the controller from an external user command. A high-level turns the controller on. A weak internal pullup holds this pin high so that the pin may be left floating if this function is not used.
FB	2	I	Inverting input to the error amplifier. In normal operation the voltage on this pin is equal to the internal reference voltage (591 mV typical)
HDRV	10	O	Bootstrapped output for driving the gate of the high-side N-channel FET.
LDRV	7	O	Output to the rectifier MOSFET gate
PGD	5	O	Open drain power good output
SW	9	I	Sense line for the adaptive anti-cross conduction circuitry. Serves as common connection for the flying high-side MOSFET driver
VDD	4	I	Power input to the controller
Thermal pad		G	Common reference for the device. Connect to the system GND.

7 Specifications

7.1 Absolute Maximum Ratings

over operating free-air temperature range unless otherwise noted(1)

		MIN	MAX	UNIT
Input voltage	VDD, ENABLE	–0.3	20	V
	SW	–5	25	
	BOOT, HDRV	–0.3	30	
	BOOT-SW, HDRV-SW (differential from BOOT or HDRV to SW)	–0.3	6	
	COMP, FB, BP5, LDRV, PGD	–0.3	6	
Operating junction temperature, T_J		–40	150	°C
Storage temperature, T_{stg}		–55	150	

(1) 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 under "recommended operating conditions" is not implied. Exposure to absolute-maximum-rated conditions for extended periods may affect device reliability.

7.2 ESD Ratings

			VALUE	UNIT
$V_{(ESD)}$	Electrostatic discharge	Human-body model (HBM), per ANSI/ESDA/JEDEC JS-001(1)	±2500	V
		Charged-device model (CDM), per JEDEC specification JESD22-C101(2)	±1500	

(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.

7.3 Recommended Operating Conditions

		MIN	MAX	UNIT
V_{VDD}	Input voltage	4.5	18	V
T_J	Operating Junction temperature	-40	125	°C

7.4 Thermal Information

THERMAL METRIC(1)		TPS40192 DRC (VSON) 10 PINS	UNIT
$R_{\theta JA}$	Junction-to-ambient thermal resistance	46.4	°C/W
$R_{\theta JC(top)}$	Junction-to-case (top) thermal resistance	51.4	°C/W
$R_{\theta JB}$	Junction-to-board thermal resistance	21.8	°C/W
ψ_{JT}	Junction-to-top characterization parameter	0.9	°C/W
ψ_{JB}	Junction-to-board characterization parameter	22.0	°C/W
$R_{\theta JC(bot)}$	Junction-to-case (bottom) thermal resistance	6.6	°C/W

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

7.5 Electrical Characteristics

T_J = –40°C to 85°C, V_{VDD}= 12 V_{dc}, all parameters at zero power dissipation (unless otherwise noted)

PARAMETER		TEST CONDITIONS		MIN	TYP	MAX	UNIT
REFERENCE							
V_{FB}	Feedback voltage range	0°C ≤ T_J ≤ 85°C		588	591	594	mV
		-40°C ≤ T_J ≤ 85°C		585	591	594	
INPUT SUPPLY							
V_{VDD}	Input voltage range			4.5		18	V
I_{VDD}	Operating current	V_{ENABLE} = 3 V			2.5	4	mA
		V_{ENABLE} = 0.6 V			45	70	µA
ON-BOARD REGULATOR							
V_{5VBP}	Output voltage	V_{VDD} > 6 V, I_{5VBP} ≤ 10 mA		5.1	5.3	5.5	V
V_{DO}	Regulator dropout voltage	V_{VDD} - V_{BP5} , V_{VDD} = 5 V, I_{BP5} ≤ 25 mA			350	550	mV
I_{SC}	Regulator current limit threshold			50			mA
I_{BP5}	Average current					50	
OSCILLATOR							
f_{SW}	Switching frequency		TPS40193	240	300	360	kHz
			TPS40192	500	600	700	
V_{RMP}	Ramp amplitude(1)				1		V
PWM							
D_{MAX}	Maximum duty cycle(1)			85%			
$t_{ON(min)}$	Minimum controlled pulse(1)					110	ns
t_{DEAD}	Output driver dead time	HDRV off to LDRV on			50		
		LDRV off to HDRV on			25		
SOFT START							
t_{SS}	Soft-start time			3	4	6	ms
t_{SSDLY}	Soft-start delay time				2		
t_{REG}	Time to regulation				6		
ERROR AMPLIFIER							
GBWP	Gain bandwidth product(1)			7	10		MHz
A_{OL}	DC gain(1)			60			dB
I_{IB}	Input bias current (current out of FB pin)					100	nA
I_{EAOP}	Output source current	V_{FB} = 0 V		1			mA
I_{EAOM}	Output sink current	V_{FB} = 2 V		1			
SHORT CIRCUIT PROTECTION							
$t_{PSS(min)}$	Minimum pulse during short circuit(1)				250		ns
t_{BLNK}	Blanking time(1)			60	90	120	
t_{OFF}	Off-time between restart attempts			30	50		ms
V_{ILIM}	Short circuit comparator threshold voltage	$R_{COMP(GND)}$ = OPEN, T_J = 25°C		160	200	240	mV
		$R_{COMP(GND)}$ = 4 kΩ, T_J = 25°C		80	100	120	
		$R_{COMP(GND)}$ = 12 kΩ, T_J = 25°C		228	280	342	
V_{ILIMH}	Short circuit threshold voltage on high-side MOSFET	T_J = 25°C		400	550	650	

(1) Specified by design. Not production tested.

Electrical Characteristics (continued)

T_J = –40°C to 85°C, V_{VDD}= 12 V_{dc}, all parameters at zero power dissipation (unless otherwise noted)

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
OUTPUT DRIVERS						
R_{HDHI}	High-side driver pullup resistance	V_{BOOT} - V_{SW} = 4.5 V, I_{HDRV} = -100 mA		3	6	Ω
R_{HDLO}	High-side driver pull-down resistance	V_{BOOT} - V_{SW} = 4.5 V, I_{HDRV} = 100 mA		1.5	3	
R_{LDHI}	Low-side driver pullup resistance	I_{LDRV} = -100 mA		2.5	5	
R_{LDLO}	Low-side driver pull-down resistance	I_{LDRV} = 100 mA		0.8	1.5	
t_{HRISE}	High-side driver rise time(1)	C_{LOAD} = 1 nF		15	35	ns
t_{HFALL}	High-side driver fall time(1)			10	25	
t_{LRISE}	Low-side driver rise time(1)			15	35	
t_{LFALL}	Low-side driver fall time(1)			10	25	
UVLO						
V_{UVLO}	Turn-on voltage		3.9	4.2	4.4	V
$UVLO_{HYST}$	Hysteresis		700	800	900	mV
SHUTDOWN						
V_{IH}	High-level input voltage, ENABLE			1.9	3	V
V_{IL}	Low-level input votlage, ENABLE		0.6			
POWER GOOD						
V_{OV}	Feedback voltage limit for power good			650		mV
V_{UV}	Feedback voltage limit for power good			525		
V_{PG_HYST}	Powergood hysteresis voltage at FB pin			30		
R_{PGD}	Pulldown resistance of PGD pin	V_{FB} = 0 V		7	50	Ω
I_{PDGLK}	Leakage current	V_{FB} = 0 V		7	12	µA
BOOT DIODE						
V_{DFWD}	Bootstrap diode forward voltage	I_{BOOT} = 5 mA	0.5	0.8	1.2	V
THERMAL SHUTDOWN						
T_{JSD}	Junction shutdown temperature(1)			145		°C
T_{JSDH}	Hysteresis(1)			20		

7.6 Dissipation Ratings

PACKAGE	AIRFLOW (LFM)	$R_{\theta JA}$ High-K Board(1) (°C/W)	Power Rating (W) T_A = 25°C	Power Rating (W) T_A = 85°C
DRC	0 (Natural Convection)	47.9	2.08	0.835
	200	40.5	2.46	0.987
	400	38.2	2.61	1.04

(1) Ratings based on JEDEC High Thermal Conductivity (High K) Board. For more information on the test method, see TI Technical Brief SZZA017.

7.7 Typical Characteristics

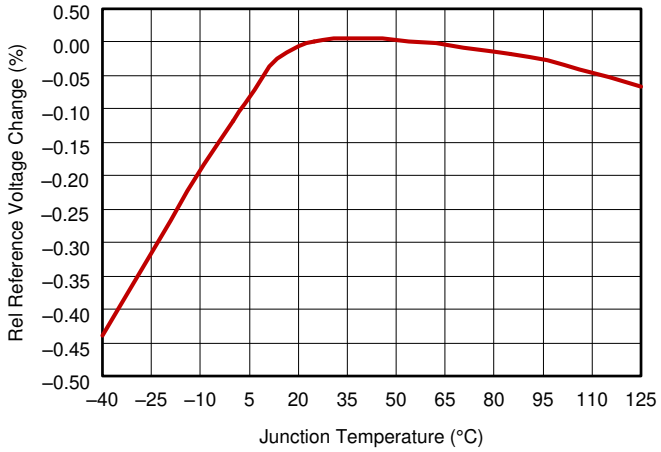


Figure 1. Relative Reference Feedback Voltage vs Junction Temperature

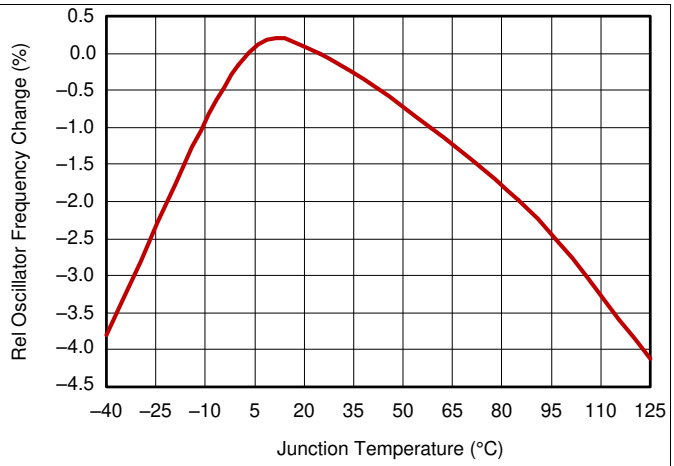


Figure 2. Relative Oscillator Frequency Change vs Junction Temperature

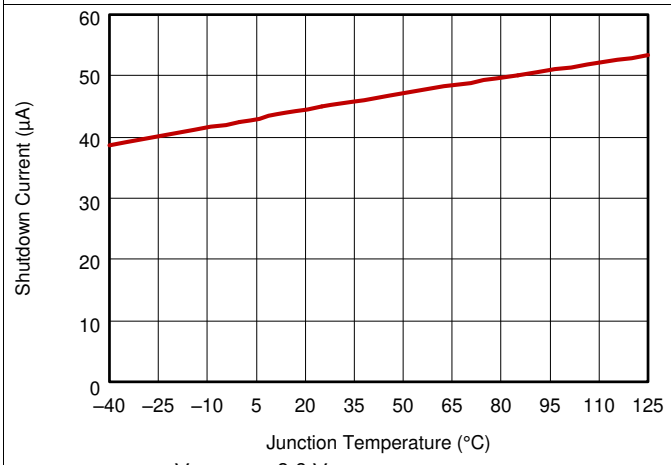


Figure 3. Shutdown Input Current vs Junction Temperature

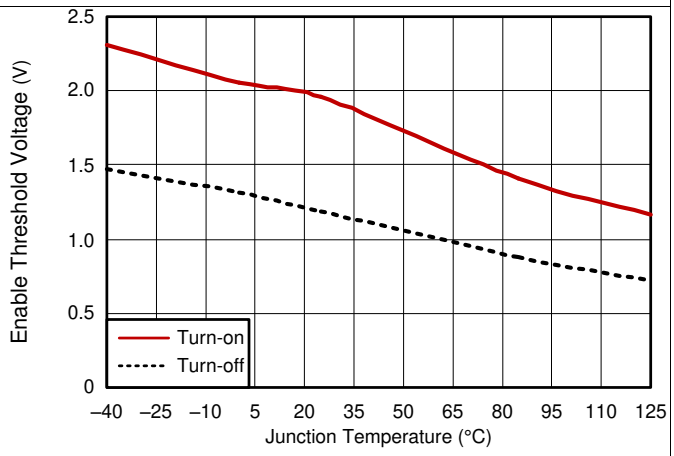


Figure 4. Enable Threshold Voltage vs Junction Temperature

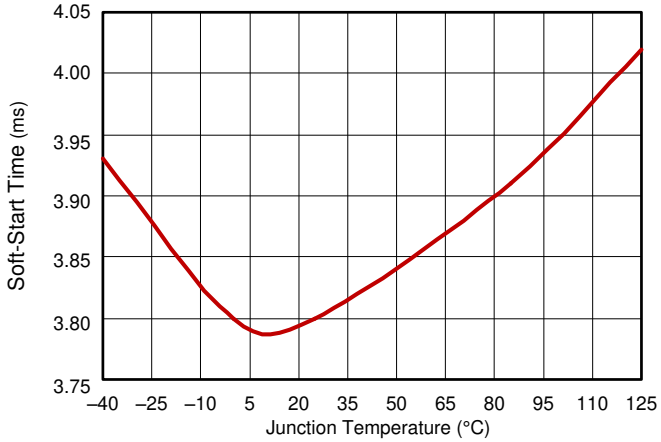


Figure 5. Soft-start Time vs Junction Temperature

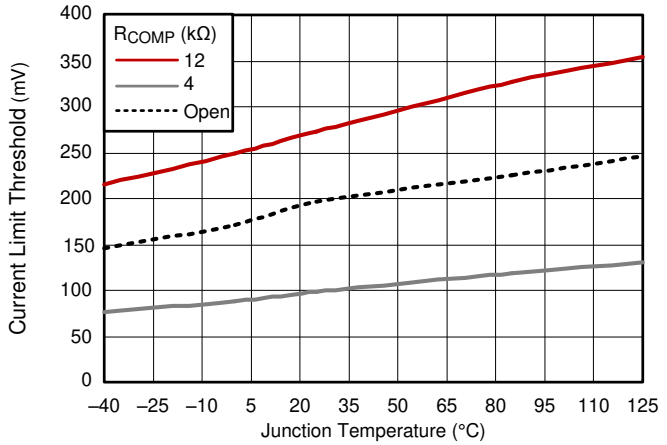


Figure 6. Low-Side MOSFET Current Limit Threshold vs Junction Temperature

Typical Characteristics (continued)

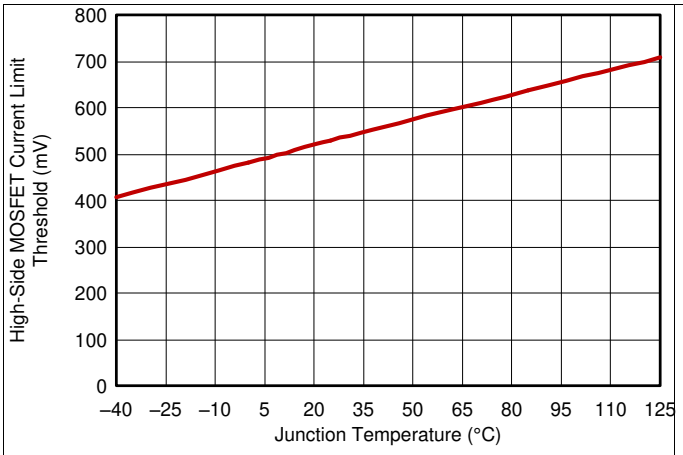


Figure 7. High-Side MOSFET Current Limit Threshold vs Junction Temperature

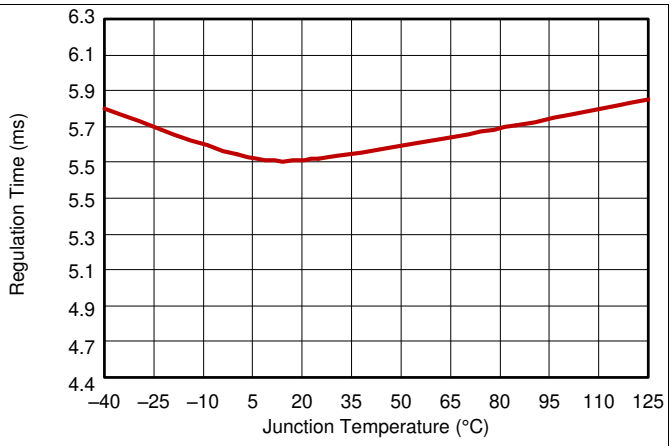


Figure 8. Total Time to Regulation vs Junction Temperature

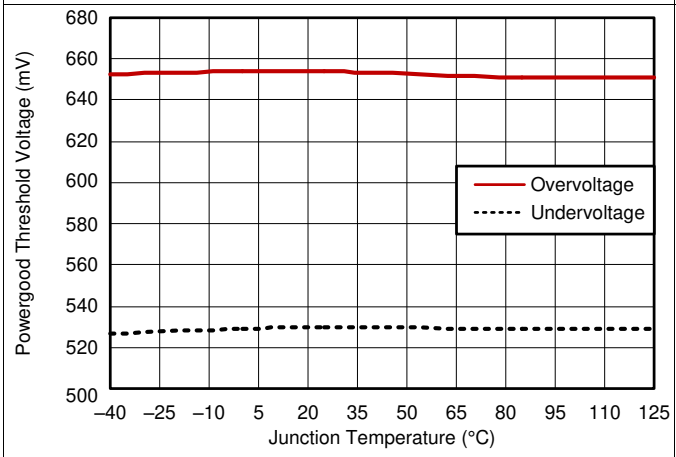


Figure 9. Power-Good Threshold voltage vs Junction Temperature

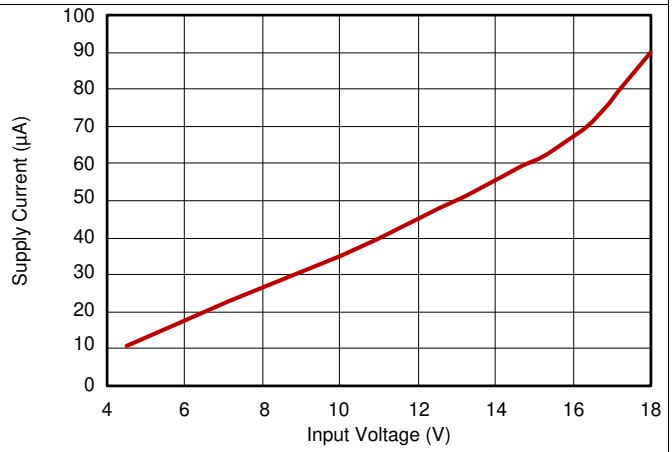


Figure 10. Shutdown Current vs Input Voltage

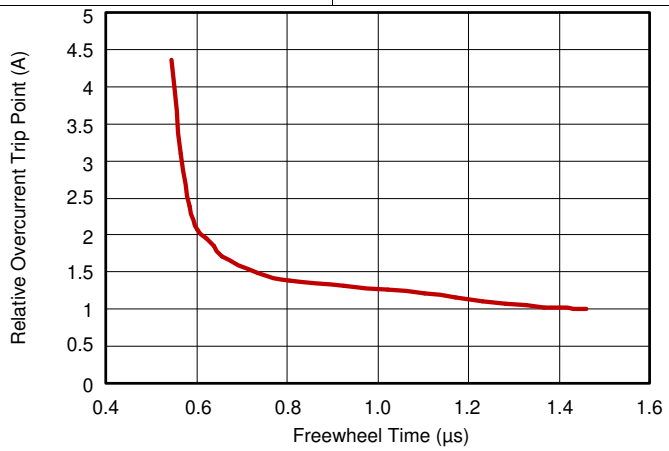


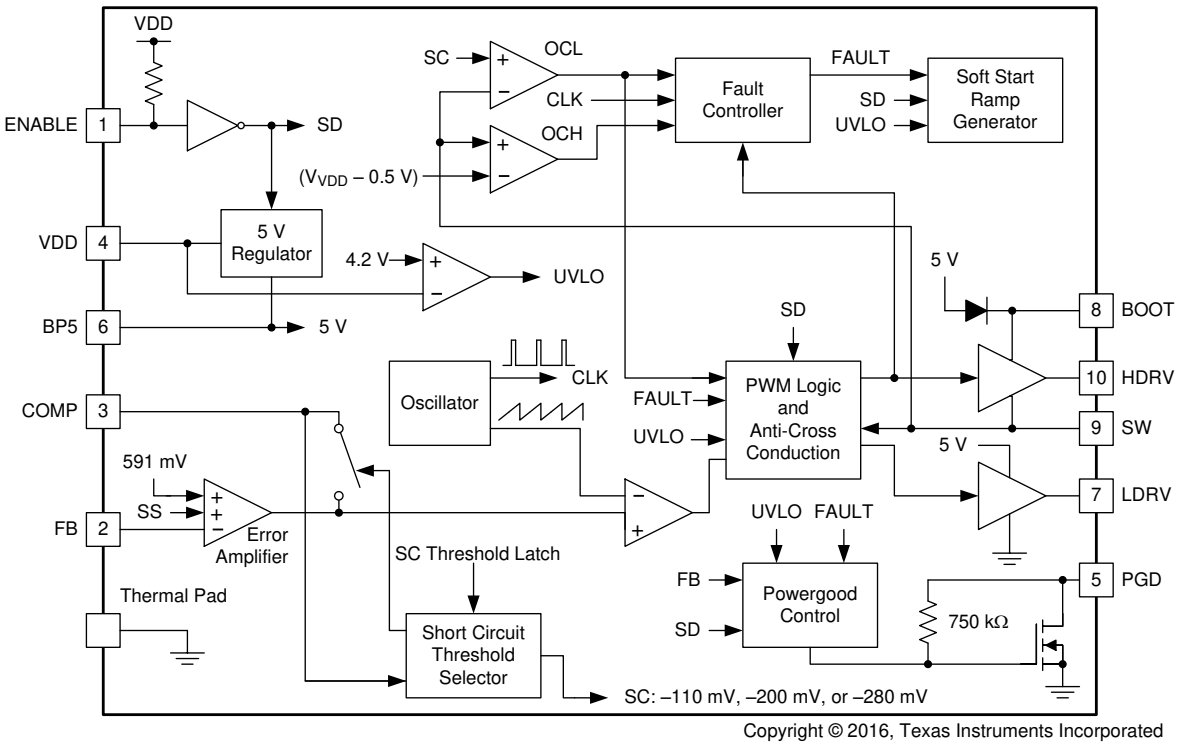
Figure 11. Relative Overcurrent Trip Point vs Freewheel Time

8 Detailed Description

8.1 Overview

The TPS40192 and TPS40193 devices are cost-optimized controllers providing all the necessary features to construct a high performance DC/DC converter while keeping costs to a minimum. Support for pre-biased outputs eliminates concerns about damaging sensitive loads during start-up. Strong gate drivers for the high-side and rectifier N-channel MOSFETs decrease switching losses for increased efficiency. Adaptive gate drive timing prevents shoot through and minimizes body diode conduction in the rectifier MOSFET, also increasing efficiency. Selectable short circuit protection thresholds and hiccup recovery from a short circuit increase design flexibility and minimize power dissipation in the event of a prolonged output fault. The dedicated ENABLE pin allows the converter to be placed in a very low quiescent current shutdown mode. Internally fixed switching frequency and soft-start time reduce external component count, simplifying design and layout, as well as reducing footprint and cost. The 3-mm × 3- mm package size also contributes to a reduced overall converter footprint.

8.2 Functional Block Diagram



8.3 Feature Description

8.3.1 Voltage Reference

The band gap cell is designed with a trimmed 591-mV output. The 0.5% tolerance on the reference voltage allows the user to design a very accurate power-supply.

8.3.2 Oscillator

The TPS40192 has a fixed internal switching frequency of 600 kHz. Tthe TPS40193 operates at a switching frequency of 300 kHz.

Feature Description (continued)

8.3.3 UVLO

When the input voltage is below the UVLO threshold, the device holds all gate drive outputs in the low (OFF) state. When the input rises above the UVLO threshold, and the ENABLE pin is above the turn ON threshold, the oscillator begins to operate and the start-up sequence is allowed to begin. The UVLO level is internally fixed at 4.2 V.

8.3.4 Enable Functionality

A dedicated ENABLE pin simplifies a user-level interface design where no multiplexed functions exist. Another benefit is a true low power shutdown mode of operation. When the ENABLE pin is pulled to GND, all unnecessary functions, including the BP5 regulator, are turned off, reducing the device supply (I_{DD}) current to 45-μA. A functionally equivalent circuit of the enable circuitry shown in Figure 12.

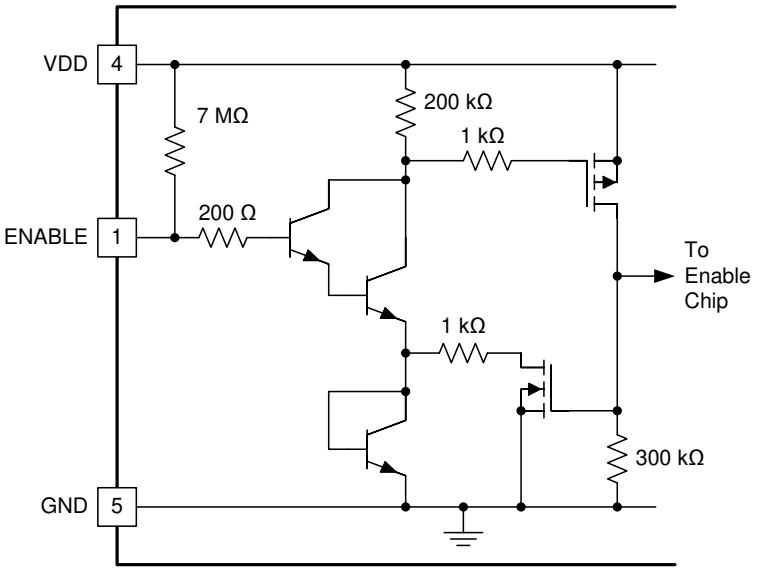


Figure 12. Enable Pin Internal Circuitry

If the ENABLE pin is left floating, the chip starts automatically. The pin must be pulled to less than 600 mV to ensure that the TPS40192 and TPS40193 devices is in shutdown mode. Note that the ENABLE pin is relatively high impedance. Some applications generate enough nearby noise to cause the ENABLE pin to swing below the 600 mV threshold and give an erroneous shutdown commands to the rest of the device. There are two solutions to solve this problem.

1. Place a capacitor from ENABLE to GND. A side effect of this is to delay the start of the converter while the capacitor charges past the enable threshold
2. Place a resistor from VDD to ENABLE. This causes more current to flow in the shutdown mode, but does not delay converter start-up. If a resistor is used, the total current into the ENABLE pin should be limited to no more than 500 μA.

The ENABLE pin is self-clamping. The clamp voltage can be as low as 1 V with a 1-kΩ ground impedance. Due to this self-clamping feature, the pullup impedance on the ENABLE pin should be selected to limit the sink current to less than 500 μA. Driving the ENABLE pin with a low-impedance source voltage can result in damage to the device. Because of the self-clamping feature, it requires care when connecting multiple ENABLE pins together.

Feature Description (continued)

8.3.5 Start-Up Sequence and Timing

After input power is applied, the 5-V onboard regulator comes up. Once this regulator comes up, the device goes through a period where it samples the impedance at the COMP pin and determines the short circuit protection threshold voltage, by placing 400 mV on the COMP pin for approximately 1 ms. During this time, the current is measured and compared against internal thresholds to select the short circuit protection threshold. After this, the COMP pin is brought low for 1 ms. This ensures that the feedback loop is preconditioned at start-up and no sudden output rise occurs at the output of the converter when the converter is allowed to start switching. After these initial two milliseconds, the internal soft-start circuitry is engaged and the converter is allowed to start. See Figure 13.

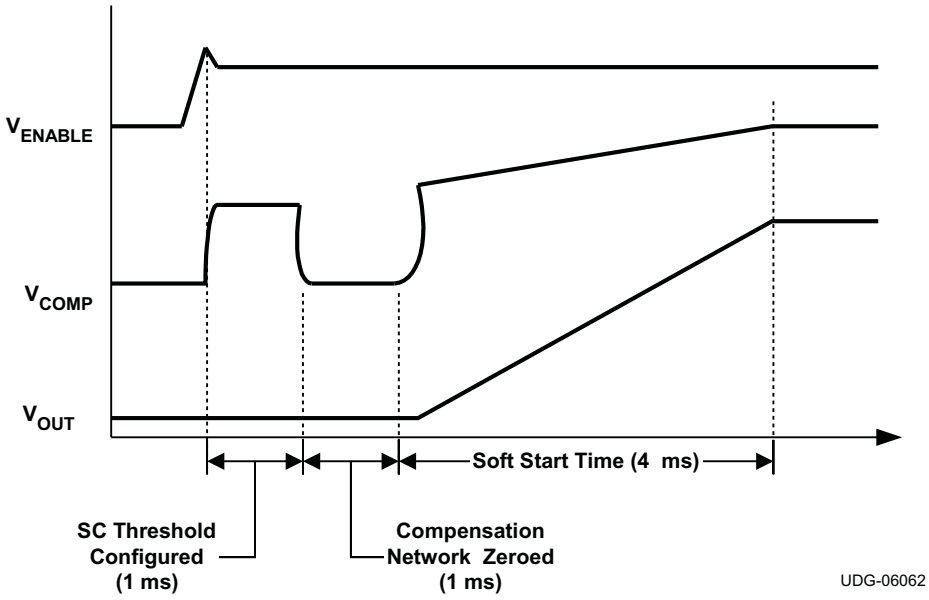


Figure 13. Start-Up Sequence

8.3.6 Selecting the Short Circuit Current

A short circuit in the devices is detected by sensing the voltage drop across the low-side MOSFET when it is on, and across the high-side MOSFET when it is on. If the voltage drop across either MOSFET exceeds the short circuit threshold in any given switching cycle, a counter increments one count. If the voltage across the high-side MOSFET was higher that the short circuit threshold, that MOSFET is turned off early. If the voltage drop across either MOSFET does not exceed the short circuit threshold during a cycle, the counter is decremented for that cycle. If the counter fills up (a count of 7) a fault condition is declared and the drivers turn off both MOSFETs. After a timeout of approximately 50 ms, the controller attempts to restart. If a short circuit is still present at the output, the current quickly ramps up to the short circuit threshold and another fault condition is declared and the process of waiting for the 50 ms an attempting to restart repeats. The low-side threshold increases as the low-side on time decreases due to blanking time and comparator response time. See Figure 11 for changes in the threshold as the low-side MOSFET conduction time decreases.

These devices provide three selectable short circuit protection thresholds for the low-side MOSFET: 100 mV, 200 mV and 280 mV. The particular threshold is selected by connecting a resistor from COMP to GND. Table 1 shows the short circuit thresholds for corresponding resistors from COMP to GND. When designing the compensation for the feedback loop, remember that a low impedance compensation network combined with a long network time constant can cause the short circuit threshold setting to not be as expected. The time constant and impedance of the network connected from COMP to FB should be as in Equation 1 to ensure no interaction with the short circuit threshold setting.

Feature Description (continued)

$$\frac{0.4\ \text{V}}{\text{R1}} \times e^{\left(\frac{-t}{\text{R1} \times \text{C1}}\right)} < 10\ \mu\text{A} \quad (1)$$

where

- t is 1 ms, the sampling time of the short circuit threshold setting circuit
- R1 and C1 are the values of the components in Figure 14

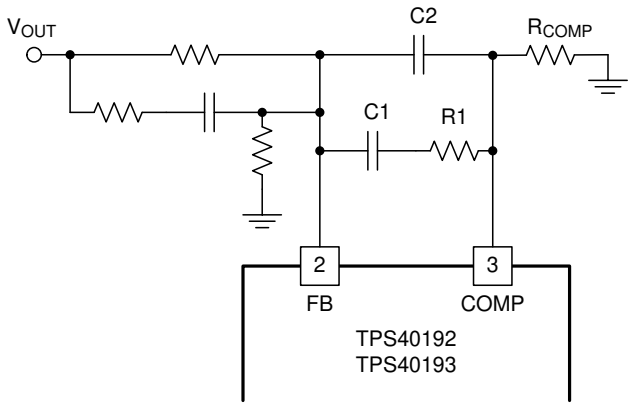


Figure 14. Short Circuit Threshold Feedback Network

Table 1. Short Circuit Threshold Voltage Selection

COMPARATOR RESISTANCE R_{COMP} (kΩ)	CURRENT LIMIT THRESHOLD VOLTAGE (mV) V_{ILIM}(V)
12 ±10%	280
Open	200
4 ±10%	100

The range of short circuit current thresholds that can be expected is shown in Equation 2 and Equation 3.

$$I_{SCP\,(max)} = \frac{V_{ILIM\,(max)}}{R_{DS\,(on)min}} \quad (2)$$

$$I_{SCP\,(min)} = \frac{V_{ILIM\,(min)}}{R_{DS\,(on)max}}$$

where

- I_{SCP} is the short circuit current
- V_{ILIM} is the short circuit threshold for the low-side MOSFET
- $R_{DS(on)}$ is the channel ON-resistance of the low-side MOSFET (3)

Due to blanking time considerations, overcurrent threshold accuracy may fall off for duty cycle greater than 75% with the TPS40192, or 88% with the TPS40193. Specifically, the overcurrent comparator has only a very short time to sample the SW pin voltage under these conditions. As a result, the comparator may not have time to respond to voltages very near the threshold.

The short circuit protection threshold for the high-side MOSFET is fixed at 550 mV typical, 400 mV minimum. This threshold is in place to provide a maximum current output using pulse by pulse current limit in the case of a fault. The pulse terminates when the voltage drop across the high-side MOSFET exceeds the short circuit threshold. The maximum amount of current that can be specified to be sourced from a converter is found by Equation 4.

$$I_{OUT\,(max)} = \frac{V_{ILIM\,(min)}}{R_{DS\,(on)max}}$$

where

- $I_{OUT(max)}$ is the maximum current that the converter is specified to source
- $V_{ILIMH(min)}$ is the short circuit threshold for the high-side MOSFET (400 mV)
- $R_{DS(on)max}$ is the maximum resistance of the high-side MOSFET (4)

If the required current from the converter is greater than the calculated $I_{OUT(max)}$, a lower resistance high-side MOSFET must be chosen. Both the high-side and low-side thresholds use temperature compensation to approximate the change in resistance for a typical power MOSFET. This helps to counteract shifts in overcurrent thresholds as temperature increases. For this feature to be effective, the MOSFETs and the device must be well coupled thermally.

8.3.7 5-V Regulator

An on board 5-V regulator that allows the parts to operate from a single voltage feed. No separate 5-V feed to the part is required. This regulator needs to have a minimum of 1-µF of capacitance on the BP5 pin for stability. A ceramic capacitor is suggested for this purpose.

This regulator can also be used to supply power to nearby circuitry, eliminating the need for a separate LDO in some cases. If this pin is used for external loads, be aware that this is the power supply for the internals of the TPS40192 and TPS40193 devices . While efforts have been made to reduce sensitivity, any noise induced on this line has an adverse effect on the overall performance of the internal circuitry and shows up as increased pulse jitter, or skewed reference voltage. Also, when the device is disabled by pulling the EN pin low, this regulator is turned off and cannot supply power.

The amount of power available from this pin varies with the size of the power MOSFETs that the drivers must operate. Larger MOSFETs require more gate drive current and reduce the amount of power available on this pin for other tasks. The total current that this pin can draw from both the gate drive and external loads cannot exceed 50 mA. The device uses up to 4 mA from the regulator and the total gate drive current can be found from Equation 5.

For regulator stability, a 1-µF capacitor is required to be connected from BP5 to GND. In some applications using higher gate charge MOSFETs, a larger capacitor is required for noise suppression. For a total gate charge of both the high and low-side MOSFETs greater than 20 nC, a 2.2-µF or larger capacitor is recommended.

$$I_G = f_{SW} \times \left(Q_{G\,(high\,)} + Q_{G\,(low\,)}\right)$$

where

- I_G is the required gate drive current
- f_{SW} is the switching frequency (600 kHz for TPS40192, and 300 kHz for TPS40193)
- $Q_{G(high)}$ is the gate charge requirement for the high-side MOSFET when V_{GS}= 5 V
- $Q_{G(low)}$ is the gate charge requirement for the low-side MOSFET when V_{GS}= 5 V (5)

8.3.8 Prebias Start-Up

The TPS40192 and TPS40193 devices contains a unique circuit to prevent current from being *pulled* from the output during start-up in the condition the output is pre-biased. When the soft-start commands a voltage higher than the pre-bias level (internal soft-start becomes greater than feedback voltage [V_{FB}]), the controller slowly activates synchronous rectification by starting the first LDRV pulses with a narrow on-time. It then increments that on-time on a cycle-by-cycle basis until it coincides with the time dictated by (1-D), where D is the duty cycle of the converter. This scheme prevents the initial sinking of the pre-bias output, and ensures that the out voltage (V_{OUT}) starts and ramps up smoothly into regulation and the control loop is given time to transition from pre-biased start-up to normal mode operation with minimal disturbance to the output voltage. The amount of time from the start of switching until the low-side MOSFET is turned on for the full (1-D) interval is defined by 32 clock cycles.

8.3.9 Drivers

The drivers for the external HDRV and LDRV MOSFETs are capable of driving a gate-to-source voltage of 5 V. The LDRV driver switches between VDD and GND, while HDRV driver is referenced to SW and switches between BOOT and SW. The drivers have non-overlapping timing that is governed by an adaptive delay circuit to minimize body diode conduction in the synchronous rectifier. The drivers are capable of driving MOSFETS that are appropriate for a 15-A (TPS40192) or 20-A (TPS40193) converter.

8.3.10 Power Good

The TPS40192 and TPS40193 devices provides an indication that output power is good for the converter. This is an open drain signal and pulls low when any condition exists that would indicate that the output of the supply might be out of regulation. These conditions include:

- V_{FB} is more than ±10% from nominal
- soft-start is active
- an undervoltage condition exists for the device
- a short circuit condition has been detected
- die temperature is over (145°C)

NOTE

When there is no power to the device, PGOOD is not able to pull close to GND if an auxiliary supply is used for the power good indication. In this case, a built in resistor connected from drain to gate on the PGOOD pull down device makes the PGOOD pin look approximately like a diode to GND.

8.3.11 Thermal Shutdown

If the junction temperature of the device reaches the thermal shutdown limit of 145°C, the PWM and the oscillator are turned off and the HDRV pin and the LDRV pin are driven low, turning off both FETs. When the junction cools to the required level (125°C nominal), the PWM inititates soft start as during a normal power up cycle.

8.4 Device Functional Modes

8.4.1 Continuous Conduction Mode

The TPS40192 and TPS40193 devices devices operate in continuous conduction mode, regardless of the output current. Following the first 32 cycles, during which the low-side MOSFET on-time is slowly increased to prevent current sinking due to a pre-biased output, the high-side MOSFET and low-side MOSFET on-times are fully complementary.

8.4.2 Low-Quiescent Shutdown

When the ENABLE pin of the TPS40192, and TPS40193 devices is held below 0.6 V, the device enters a low quiescent current shutdown mode, drawing only 45 µA typically from the VDD pin.

9 Application and Implementation

NOTE

Information in the following applications sections is not part of the TI component specification, and TI does not warrant its accuracy or completeness. TI's customers are responsible for determining suitability of components for their purposes. Customers should validate and test their design implementation to confirm system functionality.

9.1 Application Information

This example illustrates the design process and component selection for a 12 V to 1.8 V point-of-load synchronous buck regulator using the TPS40192. A definition of symbols used can be found in 表 7 of this data sheet.

9.2 Typical Application

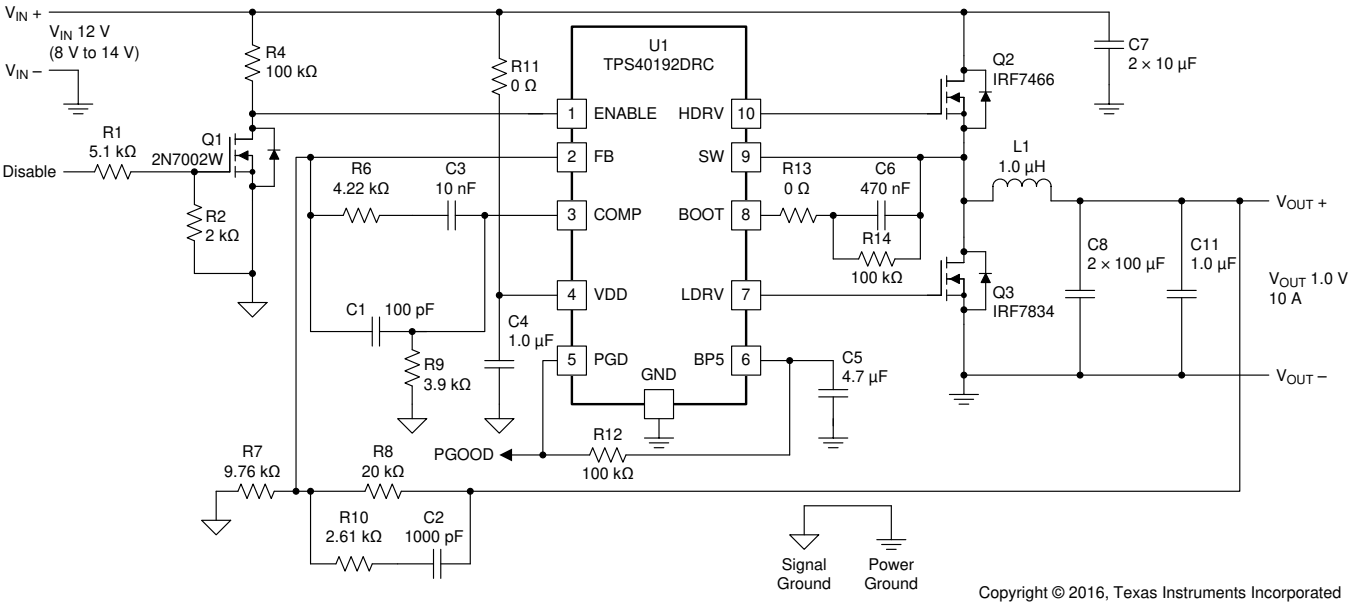


Figure 15. TPS40192 Design Example Schematic

Typical Application (continued)

9.2.1 Design Requirements

The requirements for this design are summarized in Table 2.

Table 2. Design Requirements

Parameter		Notes and Conditions	Min	Nom	Max	Units
Input Characteristics						
V_{IN}	Input voltage		8	12	14	V
I_{IN}	Input current	V_{IN} = 8 V, I_{OUT} = 10 A		2.7	2.85	A
	No load input current	V_{IN} = 8 V, I_{OUT} = 0 A		48	60	mA
VIN_UVLO	Input UVLO	0 A ≤ I_{OUT} ≤ 10 A	3.9	4.2	4.4	V
Output Characteristics						
V_{OUT}	Output voltage	V_{IN} = 12 V, I_{OUT} = 6 A		1.8		V
	Line regulation	8 V ≤ V_{IN} ≤ 14 V, I_{OUT} = 6 A			0.5%	
	Load regulation	V_{IN} = 12 V, 0 A ≤ I_{OUT} ≤ 10 A			0.5%	
$V_{OUT(ripple)}$	Output voltage ripple	V_{IN} = 12 V, I_{OUT} = 10 A			40	mVpp
I_{OUT}	Output current	8 V ≤ V_{IN} ≤ 14	0	6	10	A
I_{OCP}	Output overcurrent inception point	V_{IN} = 12 V, VOUT = VOUT – 5%		19		A
Transient Response						
ΔI	Load step	0.75 × $I_{OUT(max)}$ to 0.25 × $I_{OUT(max)}$		5		A
	Load slew rate			5		A/µsec
	Overshoot				50	mV
Systems Characteristics						
f_{SW}	Switching frequency		480	600	720	kHz
ηpk	Peak efficiency	V_{IN} = 8 V, 0 A ≤ I_{OUT} ≤ 10 A		89%		
η	Full-load efficiency	V_{IN} = 8 V, I_{OUT} ≤ 10 A		86%		
T_J	Operating temperature range	8 V ≤ V_{IN} ≤ 14 V, 0 A ≤ I_{OUT} ≤ 10 A	-40	25	60	°C

9.2.2 Detailed Design Procedure

9.2.2.1 Selecting the Switching Frequency

Choose a switching f_{SW} of 600 kHz to reduce the required inductor and capacitor sizes.

9.2.2.2 Inductor Selection

The inductor is typically sized for approximately 30% peak-to-peak ripple current (I_{RIPPLE}). Given this target ripple current, the required inductor size can be calculated by Equation 6.

$$L \approx \frac{V_{IN(max)} - V_{OUT}}{I_{RIPPLE}} \times \frac{V_{OUT}}{V_{IN(max)}} \times \frac{1}{f_{SW}} = \frac{14\ V - 1.8\ V}{0.3 \times 10\ A} \times \frac{1.8\ V}{14\ V} \times \frac{1}{600\ kHz} = 0.87\ \mu H \quad (6)$$

A standard value of 1 µH is selected. Solving for I_{RIPPLE} using an inductor value of 1 µH, results in 2.6-A peak-to-peak ripple.

The RMS current through the inductor is approximated by Equation 7.

$$I_{L(rms)} = \sqrt{\left(I_{L(avg)}\right)^2 + \frac{1}{12} \times (I_{RIPPLE})^2} = \sqrt{(I_{OUT})^2 + \frac{1}{12} \times (I_{RIPPLE})^2}$$

$$= \sqrt{(10\ A)^2 + \frac{1}{12} \times (2.6\ A)^2} \approx 10.03\ A \quad (7)$$

Using Equation 7, the maximum RMS current in the inductor is approximately 10.03 A

9.2.2.3 Output Capacitor Selection (C8)

The selection of the output capacitor is typically driven by the output transient response. The Equation 8 and Equation 9 overestimate the voltage deviation to account for delays in the loop bandwidth and can be used to determine the required output capacitance.

$$V_{OVER} < \frac{I_{TRAN}}{C_{OUT}} \times \Delta t = \frac{I_{TRAN}}{C_{OUT}} \times \frac{I_{TRAN} \times L}{V_{OUT}} = \frac{(I_{TRAN})^2 \times L}{V_{OUT} \times C_{OUT}}$$
$$V_{UNDER} < \frac{I_{TRAN}}{C_{OUT}} \times \Delta t = \frac{I_{TRAN}}{C_{OUT}} \times \frac{I_{TRAN} \times L}{(V_{IN} - V_{OUT})} = \frac{(I_{TRAN})^2 \times L}{(V_{IN} - V_{OUT}) \times C_{OUT}} \quad (8)$$

If

- $V_{IN(min)} > 2 \times V_{OUT}$, use overshoot to calculate minimum output capacitance.
- $V_{IN(min)} < 2 \times V_{OUT}$, use undershoot to calculate minimum output capacitance. (9)

$$C_{OUT\,(min)} = \frac{(I_{TRAN\,(max)})^2 \times L}{V_{OUT} \times V_{OVER}} = \frac{(4\text{ A})^2 \times 1.0\ \mu\text{H}}{1.8\text{ V} \times 50\text{ mV}} = 178\ \mu\text{F} \quad (10)$$

With a minimum capacitance, the maximum allowable ESR is determined by the maximum ripple voltage and is approximated by Equation 11.

$$ESR_{MAX} < \frac{V_{RIPPLE\,(tot)} - V_{RIPPLE\,(cap)}}{C_{OUT}} = \frac{V_{RIPPLE\,(tot)} - \left(\frac{I_{RIPPLE}}{C_{OUT} \times f_{SW}}\right)}{I_{RIPPLE}}$$
$$= \frac{36\text{ mV} - \left(\frac{2.6\text{ A}}{178\ \mu\text{F} \times 600\text{ kHz}}\right)}{2.6\text{ A}} = 4.4\ \text{m}\Omega \quad (11)$$

Two 1206 100-µF, 6.3-V X5R ceramic capacitors are selected to provide more than 178-µF of minimum capacitance and less than 4.4 mΩ of ESR (2.5 mΩ each).

9.2.2.4 Peak Current Rating of the Inductor

With output capacitance, it is possible to calculate the charge current during start-up and determine the minimum saturation current rating for the inductor. The start-up charging current is approximated by Equation 12.

$$I_{CHARGE} = \frac{V_{OUT} \times C_{OUT}}{t_{SS}} = \frac{1.8\text{ V} \times 200\ \mu\text{F}}{3.0\text{ ms}} = 120\text{ mA} \quad (12)$$

$$I_{L(peak)} = I_{OUT\,(max)} + \left(\frac{1}{2} \times I_{RIPPLE}\right) + I_{CHARGE} = 10\text{ A} + \left(\frac{1}{2} \times 2.6\text{ A}\right) + 120\text{ mA} = 11.4\text{ A} \quad (13)$$

Table 3. Inductor Requirements

PARAMETER	SYMBOL	VALUE	UNITS
Inductance	L	1	µH
RMS current (thermal rating)	$I_{L(rms)}$	10.03	A
Peak current (saturation rating)	$I_{L(peak)}$	11.4	

A PG0083.102 1-µH is selected for its small size, low DCR (6.6 mΩ) and high current handling capability (12 A thermal, 17 A saturation)

9.2.2.5 Input Capacitor Selection (C7)

The input voltage ripple is divided between capacitance and ESR. For this design $V_{RIPPLE(cap)}$ = 400 mV and $V_{RIPPLE(ESR)}$ = 200 mV. Use Equation 14 to estimate the minimum capacitance and maximum ESR.

$$C_{IN\,(min)} = \frac{I_{OUT} \times V_{OUT}}{V_{RIPPLE\,(cap)} \times V_{IN\,(MIN)} \times f_{SW}} = \frac{10\text{ A} \times 1.8\text{ V}}{400\text{ mV} \times 8\text{ V} \times 600\text{ kHz}} = 9.375\ \mu\text{F} \quad (14)$$

$$ESR_{MAX} = \frac{V_{RIPPLE\ (esr)}}{I_{OUT} + \left(\frac{1}{2} \times I_{RIPPLE}\right)} = \frac{200\ mV}{10\ A + \left(\frac{1}{2} \times 2.6\ A\right)} = 17.7\ m\Omega \quad (15)$$

For this design $C_{IN(min)}$> 9.375 µF and ESR < 17.7 mΩ . Use Equation 16 to estimate the RMS current in the input capacitors.

$$I_{RMS\,(cin)} = I_{IN\,(rms)} - I_{IN\,(avg)} = \left(I_{OUT} + \frac{1}{12} \times I_{RIPPLE}\right) \times \sqrt{\frac{V_{OUT}}{V_{IN}}} - \frac{V_{OUT} \times I_{OUT}}{V_{IN}}$$

$$= \left(10\ A + \frac{1}{12} \times 2.6\ A\right) \times \sqrt{\frac{1.8\ V}{14\ V}} - \frac{1.8\ V \times 10\ A}{14\ V} = 2.37\ A \quad (16)$$

The total input capacitance must support 2.37 A of RMS ripple current.

Two 1210 10-µF, 25 V, X5R ceramic capacitors with approximately 2 mΩ ESR and a 2-A_{RMS} current rating are selected. Higher voltage capacitors minimize capacitance loss at the DC bias voltage to ensure the capacitors have sufficient capacitance at the working voltage.

9.2.2.6 MOSFET Switch Selection (Q1, Q2)

The switching losses for the high-side MOSFET are estimated by Equation 17.

$$P_{G1SW} = \frac{1}{2} \times V_{IN} \times I_{OUT} \times \Delta t_{SW} \times f_{SW} = \frac{1}{2} \times V_{IN} \times I_{OUT} \times \frac{Q_{GD1}}{\frac{V_{DRV} - V_{TH}}{R_{DRV}}} \times f_{SW} \quad (17)$$

Switching losses in this design are highest at high-line. Designing for 1 W of total loss in each MOSFET and 60% of the total high-side MOSFET losses in switching losses, estimate the maximum gate-drain charge for the design by using Equation 18.

$$Q_{GD1\,(max)} = \frac{P_{G1SW}}{V_{IN} \times I_{OUT}} \times \frac{V_{DRV} - V_{TH}}{R_{DRV}} \times \frac{1}{f_{SW}} = \frac{600\ mW}{14\ V \times 10\ A} \times \frac{5\ V - 2\ V}{2.5\ \Omega} \times \frac{1}{600\ kHz} = 8.6\ nC \quad (18)$$

The switching losses of the synchronous rectifier are lower than the switching losses of the main MOSFET because the voltage across the MOSFET at the point of switching is reduced to the forward voltage drop across the body diode of the SR MOSFET and are estimated by using Equation 19. The conduction losses in the main MOSFET are estimated by the RMS current through the MOSFET times its $R_{DS(on)}$.

$$P_{G1\,(CON)} = I_{L(rms)} \times R_{DS\,(on)Q1} \times \frac{V_{OUT}}{V_{IN}} \quad (19)$$

Estimating about 40% of total MOSFET losses to be high-side conduction losses, the maximum $R_{DS(on)}$ of the high-side MOSFET can be estimated by using Equation 20.

$$R_{DS\,(on)Q1,MAX} = \frac{P_{Q1CON}}{\left(I_{L(rms)}\right)^2 \times \frac{V_{OUT}}{V_{IN}}} = \frac{400\ mW}{(10.03\ A)^2 \times \frac{1.8\ V}{14\ V}} = 30.9\ m\Omega \quad (20)$$

Estimating 80% of total low-side MOSFET losses in conduction losses, repeat the calculation for the synchronous rectifier, whose losses are dominated by the conduction losses. Calculate the maximum $R_{DS(on)}$ of the synchronous rectifier by Equation 21.

$$R_{DS\,(on)Q2\_MAX} = \frac{P_{Q2CON}}{\left(I_{L(rms)}\right)^2 \times \left(1 - \frac{V_{OUT}}{V_{IN}}\right)} = \frac{800\ mW}{10.03^2 \times \left(1 - \frac{1.8\ V}{14\ V}\right)} = 9.1\ m\Omega \quad (21)$$

Table 4. Power MOSFET Requirements

PARAMETER	SYMBOL	VALUE	UNITS
High-side MOSFET on-resistance	$R_{DS(on)Q1}$	30.9	mΩ
High-side MOSFET gate-to-drain charge	Q_{GD1}	8.5	nC
Low-side MOSFET on-resistance	$R_{DS(on)Q2}$	8.8	mΩ

The IRF7466 has an $R_{DS(on)MAX}$ of 30.9 mΩ at 4.5-V gate drive and only 8.0-nC V_{GD} "Miller" charge with a 4.5-V gate drive, and is chosen as a high-side MOSFET. The IRF7834 has an $R_{DS(on)Q1,MAX}$ of 5.5 mΩ at 4.5-V gate drive and 44 nC of total gate charge. These two FETs have maximum total gate charges of 23 nC and 44 nC respectively, which draws 40.2-mA from the 5-V regulator, less than its 50-mA minimum rating.

9.2.2.7 Boot Strap Capacitor

To ensure proper charging of the high-side MOSFET gate, limit the ripple voltage on the boost capacitor to less than 50 mV.

$$C_{BOOST} = 20 \times Q_{G1} = 20 \times 23\ nC = 460\ nF \quad (22)$$

Use the next higher standard value of 470 nF for the value of the bootstrap capacitor.

NOTE

It is recommended to add a high-resistance resistor in parallel with the bootstrap capacitor. Adding a small amount of load to the bootstrap capacitor (100 kΩ for a 100-nF typical capacitor) creates a discharge time constant for the bootstrap voltage following a shutdown event. This prevents the possibility of an inadvertent turn-on of the high-side MOSFET following shutdown via the ENABLE pin, due to leakage paths within the driver stage which can slowly transfer the bootstrap voltage to the HDRV pin following the shutdown. (See Figure 15)

9.2.2.8 Input Bypass Capacitor (C6)

As suggested, select a 1.0-μF ceramic bypass capacitor for VDD.

9.2.2.9 BP5 Bypass Capacitor (C5)

The recommended minimum 1.0-μF ceramic capacitance stabilizes the 5-V regulator. To limit regulator noise to less than 10 mV, the bypass capacitor is sized by using Equation 23.

$$C_{BP5} = 100 \times MAX(Q_{G1}, Q_{G2}) = 100 \times MAX(23\ nC, 44\ nC) \cong 4.4\ \mu F \quad (23)$$

Because the Q2 gate charge is larger than Q1 and the total gate charge of Q2 is 44 nC, a BP5 capacitor of 4.4-μF is calculated, and the next larger standard value of 4.7 μF is selected to limit noise on the BP5 regulator.

9.2.2.10 Input Voltage Filter Resistor (R11)

Because the minimum input voltage ($V_{IN(min)}$) is greater than 6.0 V, place a 0-Ω resistor in the VDD resistor location. If $V_{IN(min)}$ was < 6.0 V, an optional series VDD resistor with a value between 1 Ω and 2 Ω filters switching noise from the device. Limit the voltage drop across this resistor to less than 50 mV.

$$R_{VDD} = \frac{V_{RVDD\ (max)}}{I_{DD}} = \frac{50\ mV}{3\ mA + (Q_{G1,tot} + Q_{G2,tot}) \times f_{SW}} = \frac{50\ mV}{3\ mA + (44\ nC + 23\ nC) \times 600\ kHz)}$$
$$= \frac{50\ mV}{43\ mA} \cong 1\Omega \quad (24)$$

Driving the two FETs with 23 nC and 44 nC respectively, the maximum I_{VDD} current calculation of 43 mA yields a resistor value of approximately 1 Ω.

9.2.2.11 Short Circuit Protection (R9)

The devices use the negative drop across the low-side MOSFET during the OFF time to measure the inductor current. Equation 25 approximates the voltage drop across the low-side MOSFET.

$$V_{CS(max)} = I_{L(peak)} \times R_{DS(on),Q2,MAX} = 11.4\ A \times 5.5\ m\Omega \cong 62.7\ mV \quad (25)$$

The internal temperature coefficient of the TPS40192 device helps compensate for the MOSFET on-resistance ($R_{DS(on)}$) temperature coefficient. For this design select the short circuit protection voltage threshold of 110 mV by selecting R9 = 3.9 kΩ.

9.2.2.12 Feedback Compensation (Modeling the Power Stage)

The DC gain of the modulator is given by Equation 26.

$$A_{MOD} = \frac{dV_{OUT}}{dV_{COMP}} = \frac{dD}{V_{COMP}} \times V_{IN} = \frac{dt}{dV_{RAMP}} \times \frac{1}{t_{SW}} \times V_{IN} \quad (26)$$

Because the peak-to-peak ramp voltage given in the *Electrical Characteristics* table is projected from the ramp slope over a full switching period, the modulator gain can be calculated as Equation 27. The maximum modulator gain for this design is found to be 14 (23 dB).

$$A_{MOD(max)} = \frac{V_{IN(max)}}{V_{RAMP(pp)}} = \frac{14\ V}{1\ V_{PP}} = 14 \quad (27)$$

The L-C filter applies a double pole at the resonance frequency described in Equation 28.

$$f_{RES} = \frac{1}{2\pi \times \sqrt{L \times C}} = \frac{1}{2\pi \times \sqrt{1\ \mu H \times 200\ \mu F}} \cong 11.3\ kHz \quad (28)$$

At any frequency lower than this (11.3 kHz), the power stage has a DC gain of 23 dB and at any higher frequency the power stage gain drops off at -40 dB per decade. The ESR zero is approximated in Equation 29.

$$f_{ESR} = \frac{1}{2\pi \times C_{OUT} \times R_{ESR}} = \frac{1}{2\pi \times (2 \times 100\ \mu F) \times \left(\frac{2.5\ m\Omega}{2}\right)} = 636\ kHz \quad (29)$$

Using two 100 µF, 2.5 mΩ ESR ceramic output capacitors, the calculated f_{ESR} of 636 kHz is greater than 1/5th the switching frequency, and therefore outside the scope of the error amplifier design. The gain of the power stage would change to –20 dB per decade above f_{ESR}. The straight line approximation the power stage gain is described in Figure 16.

The following compensation design procedure assumes $f_{ESR} > f_{RES}$. For designs using large high-ESR bulk capacitors on the output where $f_{ESR} < f_{RES}$. Type-II compensation can be used but is not described in this data sheet.

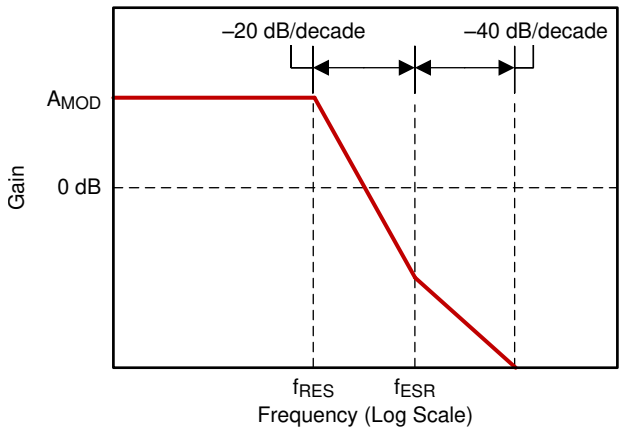


Figure 16. Approximation of Power Stage Gain

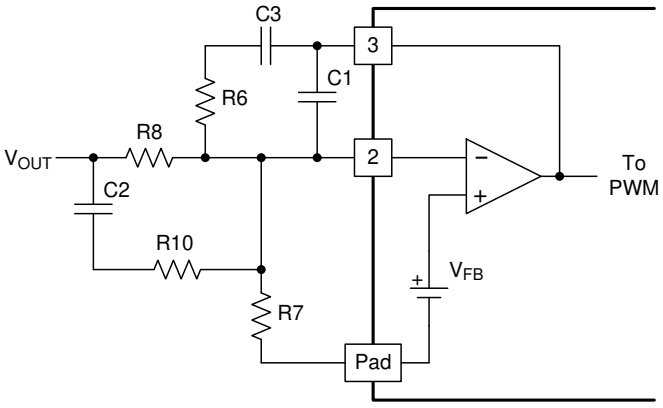


Figure 17. Type-III Compensator Used with TPS40040 or TPS40041

9.2.2.13 Feedback Divider (R7, R8)

Select a value for R8 between 10 kΩ and 100 kΩ. For this design, select 20 kΩ. R7 is then selected to produce the desired output voltage when V_{FB} = 0.591 V using Equation 30.

$$R7 = \frac{V_{FB} \times R8}{V_{OUT} - V_{FB}} = \frac{0.591\ V \times 20\ k\Omega}{1.8\ V - 0.591\ V} = 9.78\ k\Omega \quad (30)$$

Select the closest standard value of 9.76 kΩ. A slightly lower nominal value increases the nominal output voltage slightly to compensate for some trace impedance at load.

9.2.2.14 Error Amplifier Compensation (R6, R10, C1, C2, C3)

Place two zeros at 50% and 100% of the resonance frequency to boost the phase margin before resonance frequency generates –180° of phase shift. For f_{RES} = 11.7 kHz, f_{Z1} = 5.8 kHz and f_{Z2} = 11 kHz. Selecting the crossover frequency (f_{CO}) of the control loop between 3 times the LC filter resonance and 1/5th the switching frequency. For most applications 1/10th the switching frequency provides a good balance between ease of design and fast transient response.

- If $f_{ESR} < f_{CO}$; $f_{P1} = f_{ESR}$ and $f_{P2} = 4 \times f_{CO}$
- If $f_{ESR} > 2 \times f_{CO}$; $f_{P1} = f_{CO}$ and $f_{P2} = 8 \times f_{CO}$.

  For this design
- f_{SW} = 600 kHz
- f_{RES} = 11.7 kHz
- f_{ESR} = 636 kHz
- f_{CO} = 60 kHz and because
- $f_{ESR} > 2 \times f_{CO}$, $F_{P1} = f_{CO}$ = 60 kHz and $f_{P2} = 4 \times f_{CO}$ = 500 kHz.

Because $f_{CO} < f_{ESR}$ the power stage gain at the desired crossover can be approximated in Equation 31.

$$A_{PS\,(fco)} = A_{MOD\,(dc)} - 40\ \log\left(\frac{f_{CO}}{f_{RES}}\right) = 10^{APS\,(FCC)/20} = 10^{5.4\ dB/20} = 1.86 \quad (31)$$

Table 5. Error Amplifier Design Parameters

PARAMETER	SYMBOL	VALUE	UNITS
First zero frequency	f_{Z1}	5.8	kHz
Second zero frequency	f_{Z2}	11.0	
First pole frequency	f_{P1}	60	
Second pole frequency	f_{P2}	500	
Midband gain	$A_{MID(band)}$	1.86	V/V

Approximate C2 with the formula described in Equation 32.

$$C2 = \frac{1}{2\pi \times R8 \times f_{Z2}} = \frac{1}{2\pi \times 20\ k\Omega \times 11\ kHz} = 723\ pF \quad (32)$$

For a calculated value for C2 of 723 pF, the closest standard capacitor value is 1000 pF.

Approximate R10 using Equation 33.

$$R10 = \frac{1}{2\pi \times C2 \times f_{P1}} = \frac{1}{2\pi \times 1000\ pF \times 60\ kHz} = 2.65\ k\Omega \quad (33)$$

For a calculated value for R10 of 2.65 kΩ, the closest standard resistor value is 2.61 kΩ.

Calculate R6 using Equation 34.

$$R6 = \frac{A_{MID\,(band)} \times R10 \times R8}{R10 + R8} = \frac{1.86 \times 2.61\ k\Omega \times 20\ k\Omega}{2.61\ k\Omega + 20\ k\Omega} = 4.29\ k\Omega \quad (34)$$

For a calculated value for R6 of 4.29 kΩ, the closest standard resistor value is 4.22 kΩ.

Calculate C1 and C3 using Equation 35 and Equation 36.

$$C3 = \frac{1}{2\pi \times R6 \times f_{Z1}} = \frac{1}{2\pi \times 4.22\ k\Omega \times 5.8\ kHz} = 6.5\ nF \quad (35)$$

$$C1 = \frac{1}{2\pi \times R6 \times f_{P2}} = \frac{1}{2\pi \times 4.22\ k\Omega \times 500\ kHz} = 75\ pF \quad (36)$$

Using the a standard value close to 75 pF, select C1 = 100 pF. Likewise, using a standard value close to 6.5 nF, select C3 = 10 nF.

The error amplifier straight line approximation transfer function is described in Figure 18.

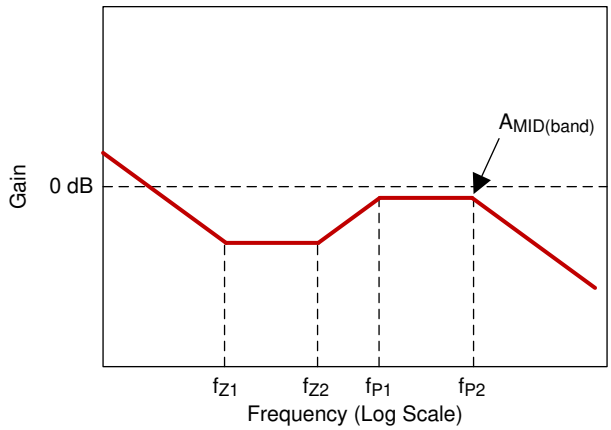


Figure 18. Error Amplifier Transfer Function Approximation

9.2.3 Application Curves

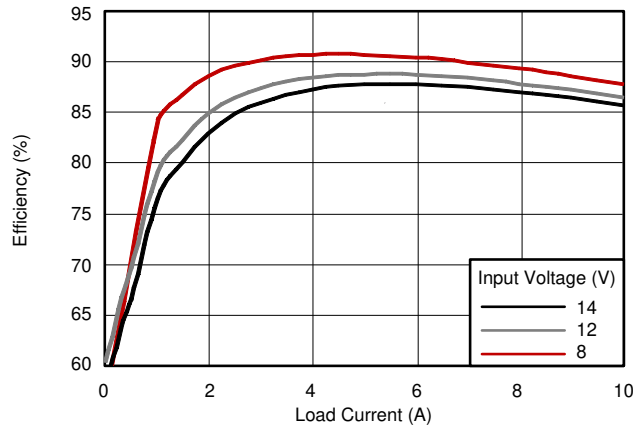


Figure 19. Efficiency vs. Output Current

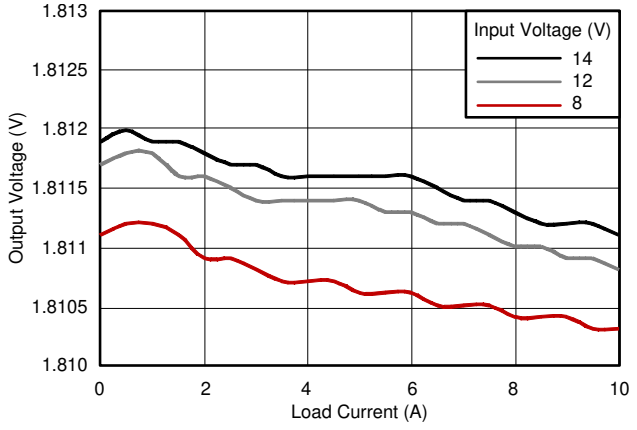


Figure 20. Output Voltage Regulation

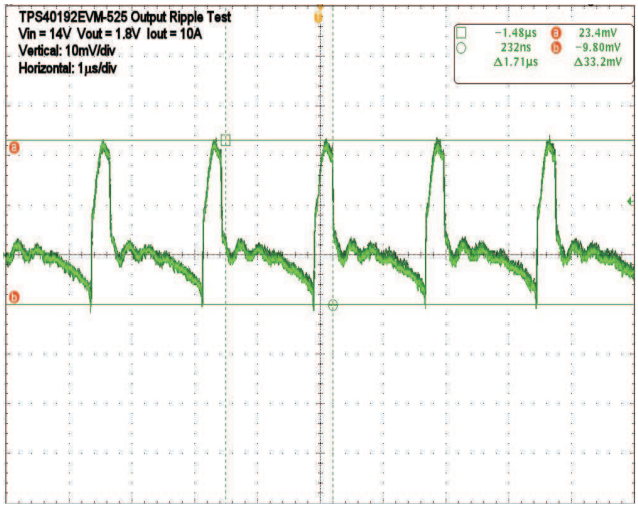


Figure 21. Output Voltage Ripple

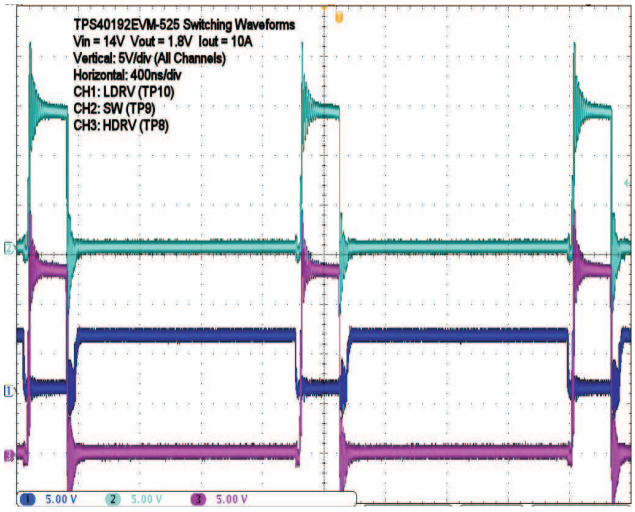


Figure 22. Switching Waveforms

10 Power Supply Recommendations

The TPS40192, and TPS40193 devices are designed to operate from a supply on the VDD pin, ranging from 4.5 V to 18 V. This supply must be well regulated and bypassed for proper operation of the devices. The VDD pin must be connected to the same supply as the power stage conversion input voltage for accurate high-side current sensing. The BP5 pin is the output of an internal low dropout regulator which is used to supply the gate drive voltages, and must also have good local bypassing for proper operation of the devices.

11 Layout

11.1 Layout Guidelines

- Optionally use R11 as a VDD filter resistor
- Locate the bypass capacitors (C7) near the power MOSFETs.
- Terminate signal components to a signal ground island separate from power ground
- Connect signal ground island to thermal pad with a single 10-mil wide trace.
- Connect power ground to the source of the synchronous rectifier.
- The thermal pad serves as the only ground for the controller.
- PowerPAD must be connected to signal ground and power ground at a single point only. Connect the PowerPad to the system ground.
- PowerPad™ should be directly connected to SYNC MOSFET (Q3) source with short, wide trace.
- Locate 3-5 vias in PowerPad™ land to remove heat from the device.
- Connect input capacitors (C7 and C9) and output capacitors (C8 andC10) grounds directly to SYNC MOSFET (Q3) source with wide copper trace or solid power ground island.
- Locate input capacitors (C7 and C9), MOSFETs (Q2 and Q3), inductor (L1) and output capacitor (C8 andC10) over power ground island.
- Use short, wide traces for LDRV and HDRV MOSFET connections.
- Route SW trace near HDRV trace.
- Route GND trace near LDRV trace.
- Use separate analog ground island under feedback components (C1, C2, C3, R5, R6, R7, R8 and R10).
- Connect ground islands at PowerPad™ with 10-mil wide trace opposite SYNC MOSFET (Q2) source connection.

11.2 Layout Examples

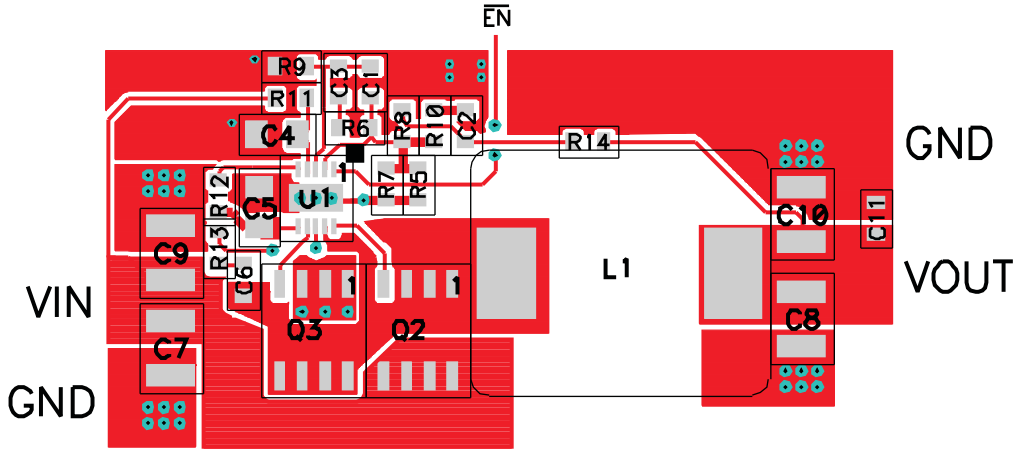


Figure 23. TPS40192 and TPS40193 Devices Sample Layout - Component Placement and Top Side Copper

Layout Examples (continued)

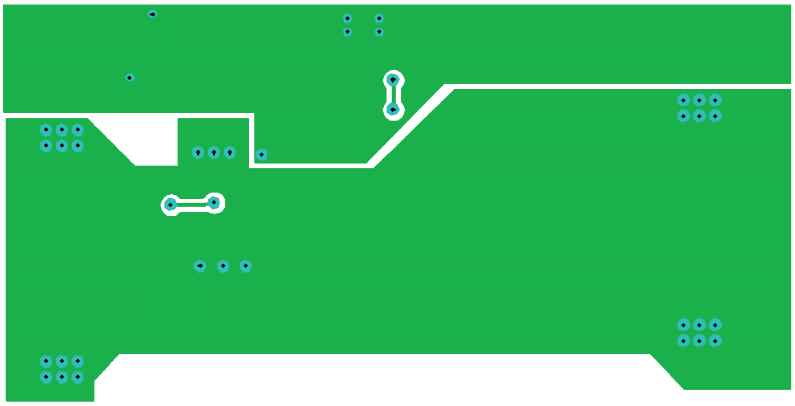


Figure 24. TPS40192 and TPS40193 Devices Sample Layout - Bottom Side Copper (X-Ray view from Top)

12 デバイスおよびドキュメントのサポート

12.1 デバイス・サポート

12.1.1 開発サポート

12.1.1.1 関連デバイス

以下のデバイスは、TPS40192およびTPS40193と特性が似ているものです。

表 6. 関連デバイス

デバイス	概要
TPS40100	ミッドレンジ入力同期整流コントローラ、高度なシーケンシングおよび出力マージン搭載
TPS40075	入力範囲の広い同期整流コントローラ、電圧フィード・フォワード搭載

12.1.2 デバイスの項目表記

表 7. 記号の定義

記号	説明
$V_{IN(max)}$	最大動作入力電圧
$V_{IN(min)}$	最小動作入力電圧
$V_{IN(ripple)}$	V_{IN}上のピーク・ツー・ピークACリップル電圧
V_{OUT}	対象出力電圧
$V_{OUT(ripple)}$	V_{OUT}上のピーク・ツー・ピークACリップル電圧
$I_{OUT(max)}$	最大動作負荷電流
I_{RIPPLE}	インダクタを流れるピーク・ツー・ピーク・リップル電流
$I_{L(peak)}$	インダクタを流れるピーク電流
$I_{L(rms)}$	インダクタを流れる2乗平均平方根電流
$I_{RMS(Cin)}$	入力コンデンサを流れる2乗平均平方根電流
f_{SW}	スイッチング周波数
f_{CO}	望ましい制御ループ・クロスオーバー周波数
A_{MOD}	PWMモジュレータの低周波数ゲイン($V_{OUT}/V_{CONTROL}$)
$V_{CONTROL}$	PWM制御電圧(エラー・アンプの出力電圧V_{COMP})
f_{RES}	L-Cフィルタの共振周波数
f_{ESR}	出力コンデンサのESRゼロ周波数
f_{P1}	エラー・アンプ補償の最初の極周波数
f_{P2}	エラー・アンプ補償の2番目の極周波数
f_{Z1}	エラー・アンプ補償の最初のゼロ周波数
f_{Z2}	エラー・アンプ補償の2番目のゼロ周波数
Q_{G1}	メインMOSFETの総ゲート電荷
Q_{G2}	同期整流MOSFETの総ゲート電荷
$R_{DS(on)Q1}$	メインMOSFETの「オン」ドレイン-ソース間抵抗
$R_{DS(on)Q2}$	同期整流MOSFETの「オン」ドレイン-ソース間抵抗
$P_{Q1C(on)}$	メイン・スイッチングMOSFETの伝導損失
P_{Q1SW}	メイン・スイッチングMOSFETのスイッチング損失
$P_{Q2C(on)}$	同期整流MOSFETの伝導損失
Q_{GD}	同期整流MOSFETのゲート-ドレイン間電荷
Q_{GS}	同期整流MOSFETのゲート-ソース間電荷

12.2 ドキュメントのサポート

ここに記載されている設計ソフトウェア、設計ツール、追加資料へのリンクなどの参照資料は、www.power.ti.comで入手できます。

- *『低電圧DC/DCコンバータの内部構造』*、2002セミナー・シリーズ、SEM1500トピック5
- *『スイッチモード電源における降圧電力ステージについて』*、1999年3月
- *『高速MOSFETゲート・ドライブ回路用の設計およびアプリケーション・ガイド』*、2001セミナー・シリーズ、SEM 1400
- *『安定した制御ループの設計』*、2001セミナー・シリーズ、SEM 1400
- *『PowerPAD™熱特性強化型パッケージ』*アプリケーション・レポート
- *『PowerPAD™の簡単な使用法』*アプリケーション・レポート
- *『QFN/SONのPCB実装』*アプリケーション・レポート

12.3 関連リンク

次の表に、クイック・アクセス・リンクを示します。カテゴリには、技術資料、サポートおよびコミュニティ・リソース、ツールとソフトウェア、およびサンプル注文またはご購入へのクイック・アクセスが含まれます。

表 8. 関連リンク

製品	プロダクト・フォルダ	サンプルとご購入	技術資料	ツールとソフトウェア	サポートとコミュニティ
TPS40192	ここをクリック	ここをクリック	ここをクリック	ここをクリック	ここをクリック
TPS40193	ここをクリック	ここをクリック	ここをクリック	ここをクリック	ここをクリック

12.4 ドキュメントの更新通知を受け取る方法

ドキュメントの更新についての通知を受け取るには、ti.comのデバイス製品フォルダを開いてください。右上の「アラートを受け取る」をクリックして登録すると、変更されたすべての製品情報に関するダイジェストを毎週受け取れます。変更の詳細については、修正されたドキュメントに含まれている改訂履歴をご覧ください。

12.5 コミュニティ・リソース

The following links connect to TI community resources. Linked contents are provided "AS IS" by the respective contributors. They do not constitute TI specifications and do not necessarily reflect TI's views; see TI's Terms of Use.

TI E2E™ Online Community ***TI's Engineer-to-Engineer (E2E) Community.*** Created to foster collaboration among engineers. At e2e.ti.com, you can ask questions, share knowledge, explore ideas and help solve problems with fellow engineers.

Design Support ***TI's Design Support*** Quickly find helpful E2E forums along with design support tools and contact information for technical support.

12.6 商標

E2E is a trademark of Texas Instruments.
All other trademarks are the property of their respective owners.

12.7 静電気放電に関する注意事項

これらのデバイスは、限定的なESD（静電破壊）保護機能を内 蔵しています。保存時または取り扱い時は、MOSゲートに対す る静電破壊を防止するために、リード線同士をショートさせて おくか、デバイスを導電フォームに入れる必要があります。

12.8 Glossary

SLYZ022 — *TI Glossary*.

This glossary lists and explains terms, acronyms, and definitions.

13 メカニカル、パッケージ、および注文情報

以降のページには、メカニカル、パッケージ、および注文に関する情報が記載されています。この情報は、そのデバイスについて利用可能な最新のデータです。このデータは予告なく変更されることがあり、ドキュメントが改訂される場合もあります。本データシートのブラウザ版を使用されている場合は、画面左側の説明をご覧ください。

PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead finish/ Ball material (6)	MSL Peak Temp (3)	Op Temp (°C)	Device Marking (4/5)	Samples
TPS40192DRCR	ACTIVE	VSON	DRC	10	3000	RoHS & Green	NIPDAU | NIPDAUAG	Level-2-260C-1 YEAR	-40 to 85	0192	Samples
TPS40192DRCT	ACTIVE	VSON	DRC	10	250	RoHS & Green	NIPDAU | NIPDAUAG	Level-2-260C-1 YEAR	-40 to 85	0192	Samples
TPS40193DRCR	ACTIVE	VSON	DRC	10	3000	RoHS & Green	NIPDAU | NIPDAUAG	Level-2-260C-1 YEAR	-40 to 85	0193	Samples
TPS40193DRCRG4	ACTIVE	VSON	DRC	10	3000	RoHS & Green	NIPDAU	Level-2-260C-1 YEAR	-40 to 85	0193	Samples
TPS40193DRCT	ACTIVE	VSON	DRC	10	250	RoHS & Green	NIPDAU | NIPDAUAG	Level-2-260C-1 YEAR	-40 to 85	0193	Samples

(1) The marketing status values are defined as follows:
ACTIVE: Product device recommended for new designs.
LIFEBUY: TI has announced that the device will be discontinued, and a lifetime-buy period is in effect.
NRND: Not recommended for new designs. Device is in production to support existing customers, but TI does not recommend using this part in a new design.
PREVIEW: Device has been announced but is not in production. Samples may or may not be available.
OBSOLETE: TI has discontinued the production of the device.

(2) **RoHS:** TI defines "RoHS" to mean semiconductor products that are compliant with the current EU RoHS requirements for all 10 RoHS substances, including the requirement that RoHS substance do not exceed 0.1% by weight in homogeneous materials. Where designed to be soldered at high temperatures, "RoHS" products are suitable for use in specified lead-free processes. TI may reference these types of products as "Pb-Free".
RoHS Exempt: TI defines "RoHS Exempt" to mean products that contain lead but are compliant with EU RoHS pursuant to a specific EU RoHS exemption.
Green: TI defines "Green" to mean the content of Chlorine (Cl) and Bromine (Br) based flame retardants meet JS709B low halogen requirements of <=1000ppm threshold. Antimony trioxide based flame retardants must also meet the <=1000ppm threshold requirement.

(3) MSL, Peak Temp. - The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

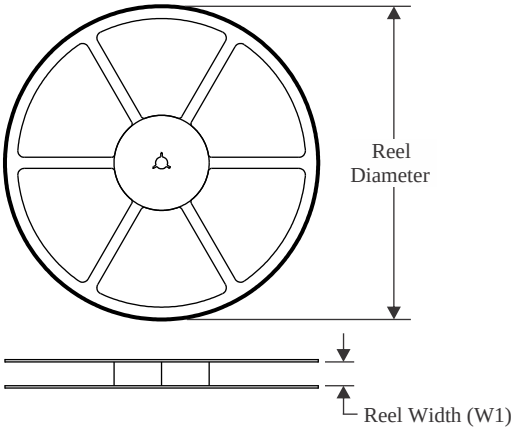
(4) There may be additional marking, which relates to the logo, the lot trace code information, or the environmental category on the device.

(5) Multiple Device Markings will be inside parentheses. Only one Device Marking contained in parentheses and separated by a "~" will appear on a device. If a line is indented then it is a continuation of the previous line and the two combined represent the entire Device Marking for that device.

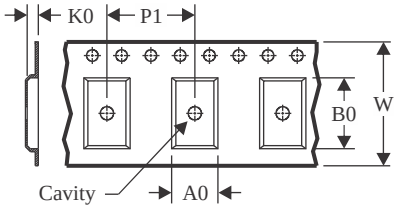
(6) Lead finish/Ball material - Orderable Devices may have multiple material finish options. Finish options are separated by a vertical ruled line. Lead finish/Ball material values may wrap to two lines if the finish value exceeds the maximum column width.

TAPE AND REEL INFORMATION

REEL DIMENSIONS

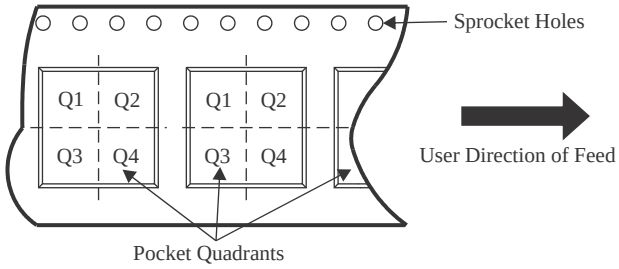


TAPE DIMENSIONS



A0	Dimension designed to accommodate the component width
B0	Dimension designed to accommodate the component length
K0	Dimension designed to accommodate the component thickness
W	Overall width of the carrier tape
P1	Pitch between successive cavity centers

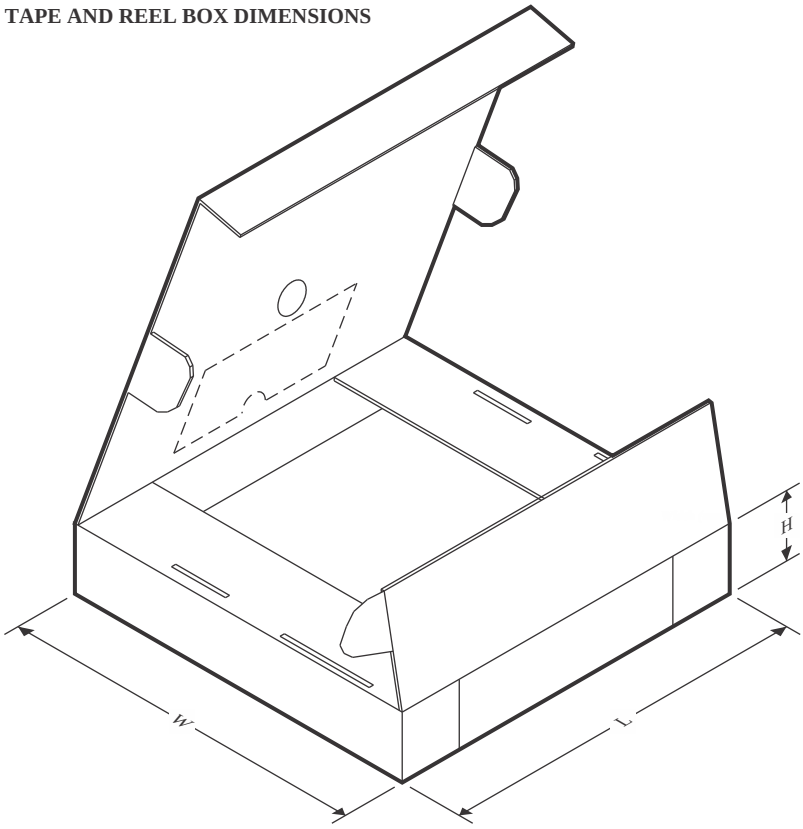
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE



*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
TPS40192DRCR	VSON	DRC	10	3000	330.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2
TPS40192DRCT	VSON	DRC	10	250	180.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2
TPS40193DRCR	VSON	DRC	10	3000	330.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2
TPS40193DRCT	VSON	DRC	10	250	180.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2

TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS40192DRCR	VSON	DRC	10	3000	346.0	346.0	33.0
TPS40192DRCT	VSON	DRC	10	250	210.0	185.0	35.0
TPS40193DRCR	VSON	DRC	10	3000	346.0	346.0	33.0
TPS40193DRCT	VSON	DRC	10	250	210.0	185.0	35.0

GENERIC PACKAGE VIEW

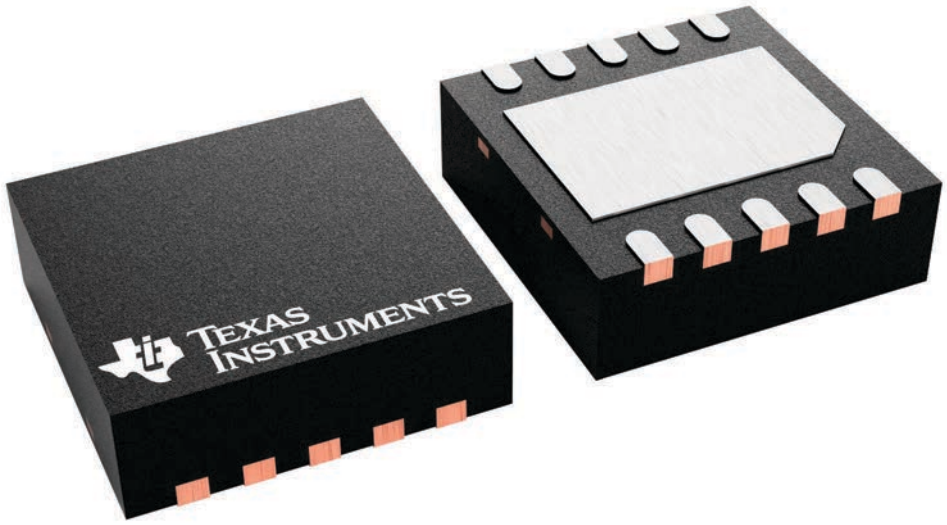
DRC 10

VSON - 1 mm max height

3 x 3, 0.5 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.



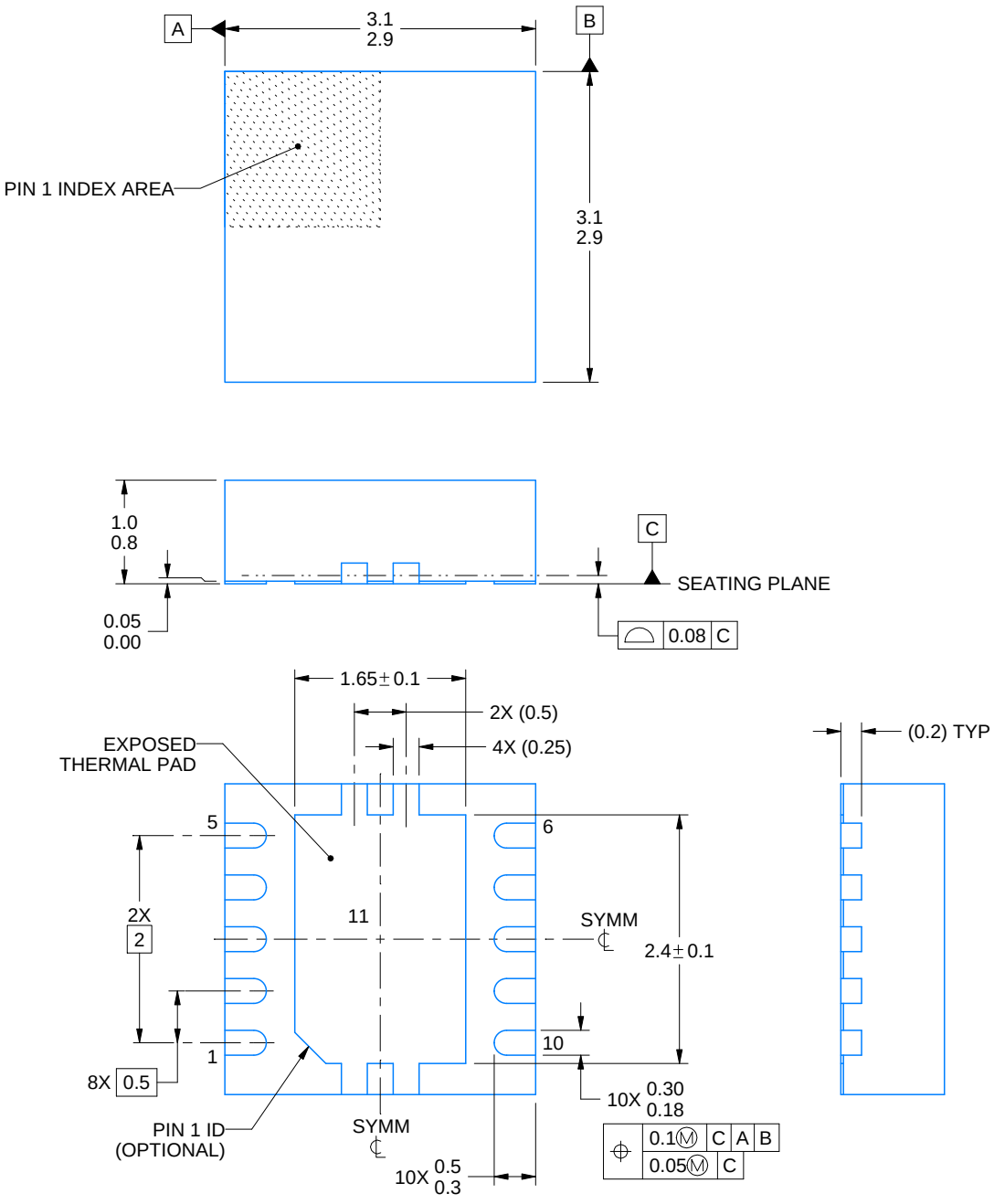
4226193/A

DRC0010J

PACKAGE OUTLINE

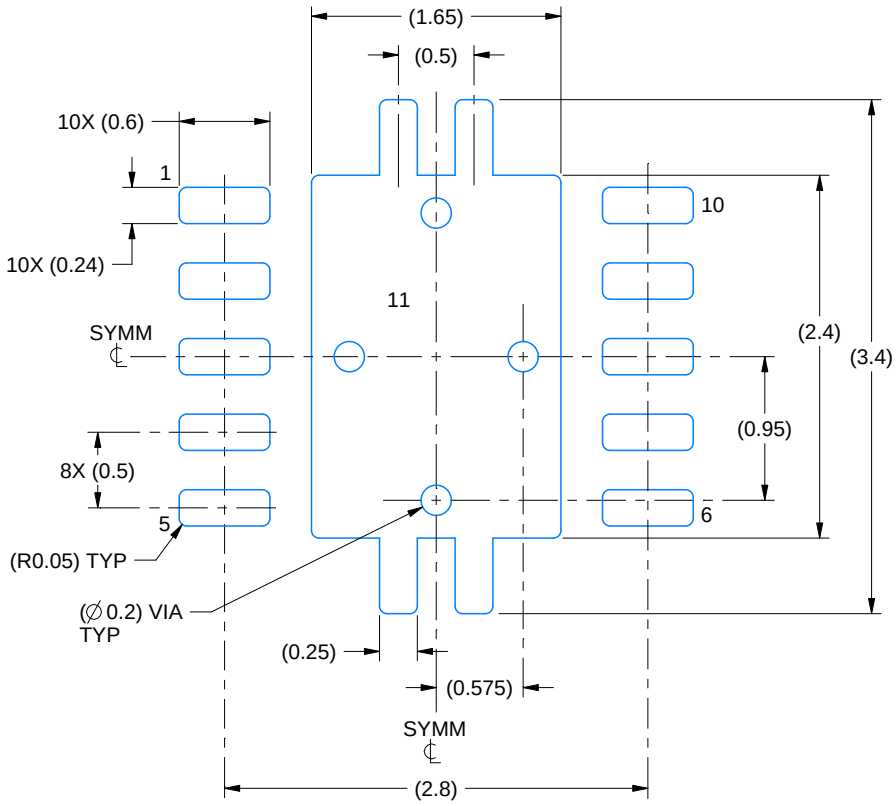
VSON - 1 mm max height

PLASTIC SMALL OUTLINE - NO LEAD

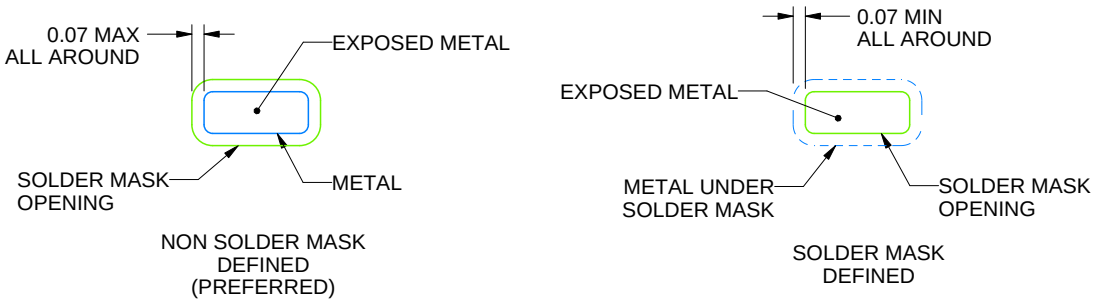


NOTES:

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 optimal thermal and mechanical performance.



LAND PATTERN EXAMPLE
EXPOSED METAL SHOWN
SCALE:20X



SOLDER MASK DETAILS

4218878/B 07/2018

NOTES: (continued)

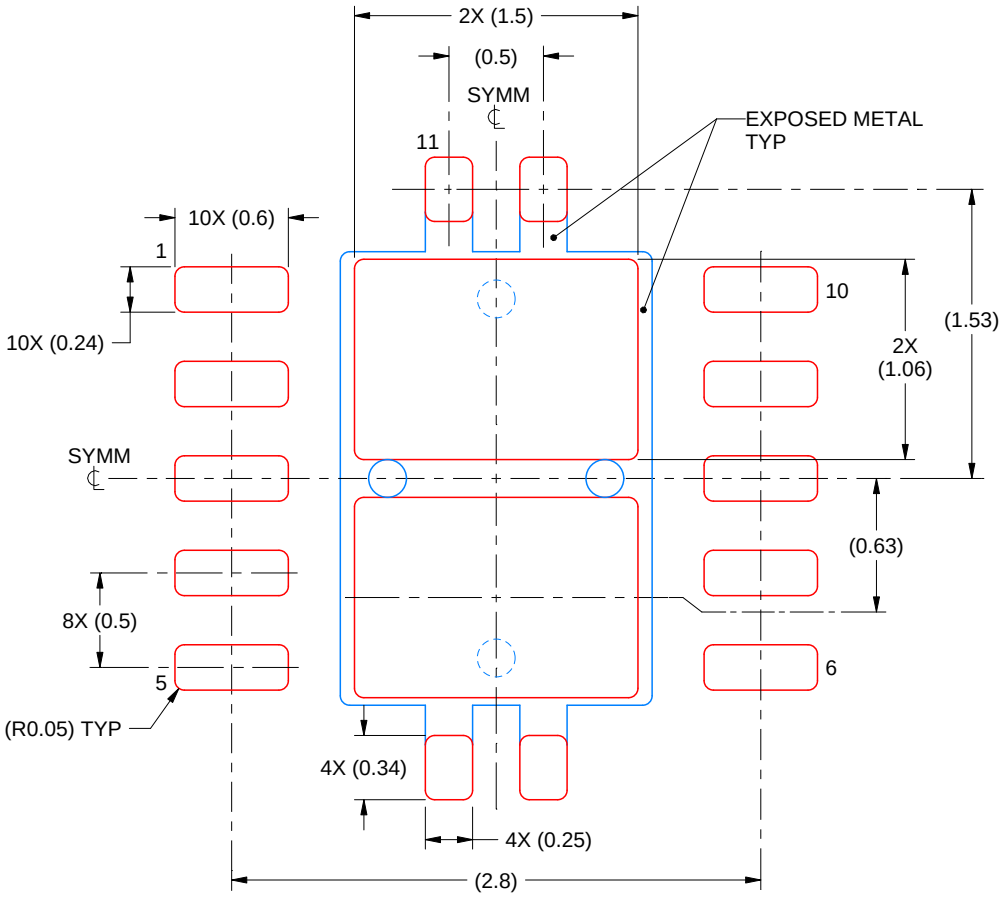
4. This package is designed to be soldered to a thermal pad on the board. For more information, see Texas Instruments literature number SLUA271 (www.ti.com/lit/slua271).
5. Vias are optional depending on application, refer to device data sheet. If any vias are implemented, refer to their locations shown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

EXAMPLE STENCIL DESIGN

DRC0010J

VSON - 1 mm max height

PLASTIC SMALL OUTLINE - NO LEAD



SOLDER PASTE EXAMPLE
BASED ON 0.125 mm THICK STENCIL

EXPOSED PAD 11:
80% PRINTED SOLDER COVERAGE BY AREA
SCALE:25X

4218878/B 07/2018

NOTES: (continued)

6. Laser cutting apertures with trapezoidal walls and rounded corners may offer better paste release. IPC-7525 may have alternate design recommendations.

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