TEXAS INSTRUMENTS
TPS65266-1
JAJSCK0 – OCTOBER 2016

TPS65266-1 2.7～6V入力電圧、3A/2A/2A出力電流、トリプル同期整流降圧型コンバータ

1 特長

- 動作入力電圧範囲: 2.7～6V
- フィードバック基準電圧: 0.6V±1%
- 最大連続出力電流: 3A/2A/2A
- 専用のイネーブルおよびソフトスタート
- イネーブル・ピンの放電によるスタートアップ・タイミングの正確な制御
- 軽負荷時のパルス・スキップ・モード(PSM)
- サイクル単位の電流制限とヒカップ・モード過電流保護
- 可変クロック周波数: 250kHz～2.4MHz
- 外部クロック同期
- パワーグッド・インジケータ
- 過熱保護

2 アプリケーション

- プリンタおよびスキャナ
- デジタルTV
- セットトップ・ボックス
- ホーム・ゲートウェイおよびアクセス・ポイント・ネットワーク
- 監視機器

3 概要

TPS65266-1には3チャネルの高効率な同期整流バック・コンバータが搭載されており、入力電圧が6Vよりも低いアダプタやバッテリで動作するアプリケーション用に設計されています。

このバックDC/DCコンバータにはパワーMOSFETが内蔵されており、電力効率を最適化するとともに、外付け部品の数を減らすことができます。ピーク電流モード制御により補償が簡素化され、過渡応答が高速です。クロック周波数が高いため、小型で値の低いインダクタやコンデンサを使用できます。外部補償では、最適化されたループ補償と高速な過渡応答がサポートされます。このバック・コンバータは、軽負荷の状況ではPSMモードで動作し、システムへ供給される入力電力を低減します。サイクル単位の過電流制限とヒカップ・モードにより、短絡または過負荷のフォルト状況において、MOSFETの消費電力が制限されます。

TPS65266-1にはパワーグッド・スーパーバイザ回路があり、コンバータのすべての出力を監視します。PGOODピンは、各チャネルの出力電圧がレギュレートされた状態になり、シーケンシングが完了した後でアサートされます。

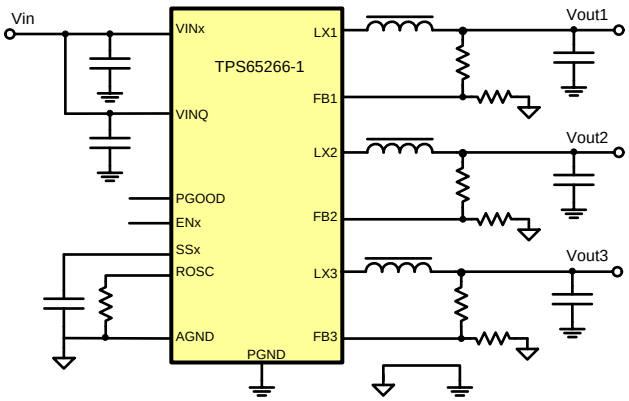
継続的な重い過負荷、または短絡により降圧コンバータの消費電力が増大すると、デバイスの損傷を防ぐため、内部の熱保護回路によりデバイスがシャットオフされます。デバイスの温度が十分に低下すると、自動的にサーマル・シャットダウンからの回復が行われます。

製品情報(1)

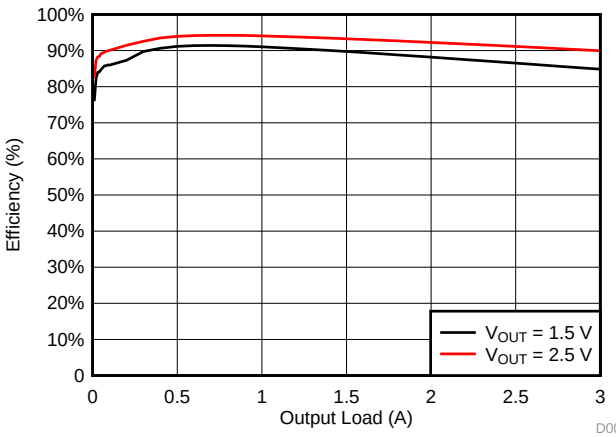
型番	パッケージ	本体サイズ(公称)
TPS65266-1	VQFN (32)	5.00mm×5.00mm

(1) 利用可能なすべてのパッケージについては、この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。

簡略化されたアプリケーション回路図



効率と出力負荷との関係



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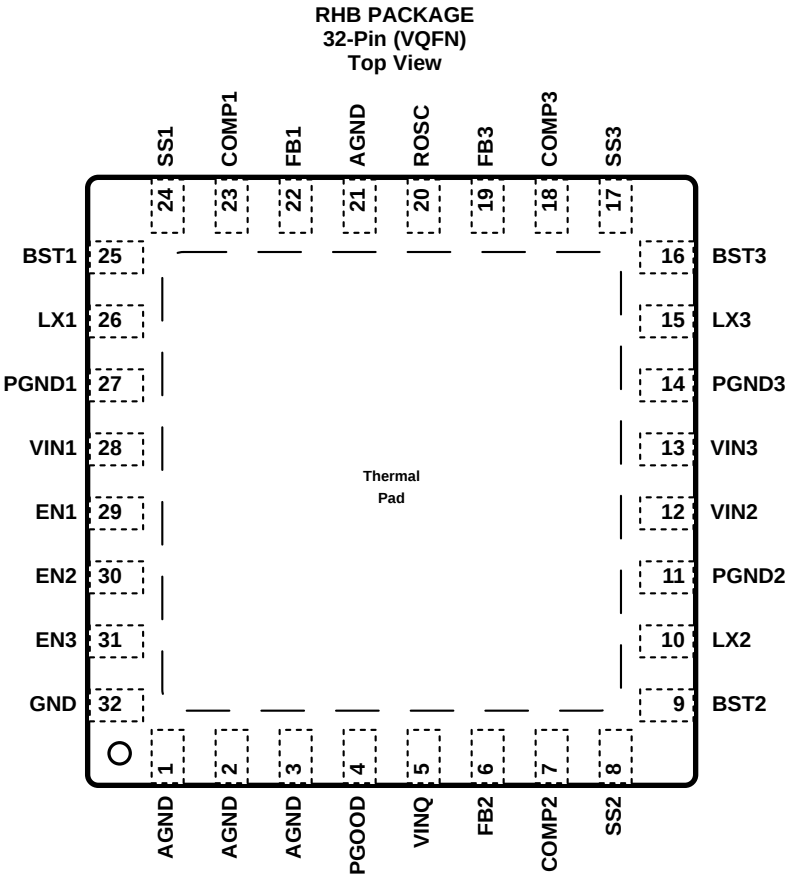
4 改訂履歴

日付	改訂内容	注
2016年10月	*	初版

5 Pin Configuration and Functions



A. There is no electric signal down bonded to thermal pad inside IC. Exposed thermal pad must be soldered to PCB for optimal thermal performance.

Pin Functions

PIN		DESCRIPTION
NO.	NAME	
1	AGND	Analog ground pin
2	AGND	Analog ground pin
3	AGND	Analog ground pin
4	PGOOD	An open-drain output; asserts low if output voltage of bucks beyond regulation range due to thermal shutdown, over-current, under-voltage, or ENx low.
5	VINQ	Input voltage of converter controller and reference power supply bias. TI recommends to connect a 1-µF capacitor from the pin to analog ground and put the capacitor as near as possible to this pin.
6	FB2	Feedback Kelvin sensing pin for buck2 output voltage. Connect this pin to buck2 feedback resistor divider.
7	COMP2	Error amplifier output and loop compensation pin for buck2. Connect a series resistor and capacitor to compensate the control loop of buck converter 2 with peak current PWM mode.
8	SS2	Soft-start and tracking input for buck2 converter. An internal 5.5-µA pullup current source is connected to this pin. The soft-start time of buck2 can be programmed by connecting a capacitor between this pin and ground.
9	BST2	Boot strapped supply to the high-side floating gate driver in buck2 converter. Connect a capacitor (47 nF recommended) from BST2 pin to LX2 pin.
10	LX2	Switching node connection to the inductor and bootstrap capacitor for buck2 converter. The voltage swing at this pin is from a diode voltage below the ground up to VIN2 voltage.
11	PGND2	Power ground connection of buck2. Connect PGND2 pin as close as practical to the (–) terminal of VIN2 input ceramic capacitor.

Pin Functions (continued)

PIN		DESCRIPTION
NO.	NAME	
12	VIN2	Input power supply for buck2. Connect VIN2 pin as close as practical to the (+) terminal of an input ceramic capacitor (10 µF suggested).
13	VIN3	Input power supply for buck3. Connect VIN3 pin as close as practical to the (+) terminal of an input ceramic capacitor (10 µF suggested).
14	PGND3	Power ground connection of buck3. Connect PGND3 pin as close as practical to the (–) terminal of VIN3 input ceramic capacitor.
15	LX3	Switching node connection to the inductor and bootstrap capacitor for buck3 converter. The voltage swing at this pin is from a diode voltage below the ground up to VIN3 voltage.
16	BST3	Boot strapped supply to the high-side floating gate driver in buck3 converter. Connect a capacitor (47 nF recommended) from BST3 pin to LX3 pin.
17	SS3	Soft-start and tracking input for buck3 converter. An internal 5.5-µA pullup current source is connected to this pin. The soft-start time of buck3 can be programmed by connecting a capacitor between this pin and ground.
18	COMP3	Error amplifier output and loop compensation pin for buck3. Connect a series resistor and capacitor to compensate the control loop of buck converter 3 with peak current PWM mode.
19	FB3	Feedback Kelvin sensing pin for buck3 output voltage. Connect this pin to buck3 feedback resistor divider.
20	ROSC	Oscillator frequency programmable pin. Connect an external resistor to set the switching frequency.
21	AGND	Analog ground common to buck controllers and other analog circuits. It must be routed separately from high-current power grounds to the (–) terminal of bypass capacitor of input voltage VINQ.
22	FB1	Feedback Kelvin sensing pin for buck1 output voltage. Connect this pin to buck1 feedback resistor divider.
23	COMP1	Error amplifier output and loop compensation pin for buck1. Connect a series resistor and capacitor to compensate the control loop of buck converter 1 with peak current PWM mode.
24	SS1	Soft-start and tracking input for buck1 converter. An internal 5.5-µA pullup current source is connected to this pin. The soft-start time of buck1 can be programmed by connecting a capacitor between this pin and ground.
25	BST1	Boot strapped supply to the high-side floating gate driver in buck1 converter. Connect a capacitor (47 nF recommended) from BST1 pin to LX1 pin.
26	LX1	Switching node connection to the inductor and bootstrap capacitor for buck1 converter. The voltage swing at this pin is from a diode voltage below the ground up to VIN1 voltage.
27	PGND1	Power ground connection of buck1. Connect PGND1 pin as close as practical to the (–) terminal of VIN1 input ceramic capacitor.
28	VIN1	Input power supply for buck1. Connect VIN1 pin as close as practical to the (+) terminal of an input ceramic capacitor (suggest 10 µF).
29	EN1	Enable for buck1 converter. Float to enable. Can use this pin to adjust the input undervoltage lockup of buck1 with resistors divider.
30	EN2	Enable for buck2 converter. Float to enable. Can use this pin to adjust the input undervoltage lockup of buck2 with resistors divider.
31	EN3	Enable for buck3 converter. Float to enable. Can use this pin to adjust the input undervoltage lockup of buck3 with resistors divider.
32	GND	Ground pin
—	Thermal PAD	No electric connection to any signal. Soldered to the ground in PCB for better thermal performance.

6 Specifications

6.1 Absolute Maximum Ratings

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

		MIN	MAX	UNIT
Voltage at	VIN1, VIN2, VIN3, VINQ	–0.3	7	V
	LX1, LX2, LX3 (maximum withstand voltage transient <20 ns)	–1.0	7	
	BST1, BST2, BST3 referenced to LX1, LX2, LX3 pins respectively	–0.3	7	
	EN1, EN2, EN3, PGOOD	–0.3	7	
	FB1, FB2, FB3, COMP1 , COMP2, COMP3, SS1, SS2, SS3, ROSC	–0.3	3.6	
	AGND, PGND1, PGND2, PGND3	–0.3	0.3	
T_J	Operating junction temperature	–30	125	°C
T_{stg}	Storage temperature	–55	150	°C

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

6.2 ESD Ratings

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

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

6.3 Recommended Operating Conditions

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

		MIN	MAX	UNIT
Voltage at	VIN1, VIN2, VIN3, VINQ	2.7	6	V
	LX1, LX2, LX3 (maximum withstand voltage transient <20 ns)	–0.8	6	
	BST1, BST2, BST3 referenced to LX1, LX2, LX3 pins respectively	–0.1	6	
	EN1, EN2, EN3, PGOOD	–0.1	6	
	FB1, FB2, FB3, COMP1 , COMP2, COMP3, SS1, SS2, SS3, ROSC	–0.1	3	
T_J	Operating junction temperature	–30	125	°C

6.4 Thermal Information

THERMAL METRIC(1)		TPS65266-1 RHB 32-PIN VQFN	UNIT
$R_{\theta JA}$	Junction-to-ambient thermal resistance	34.2	°C/W
$R_{\theta JC(top)}$	Junction-to-case (top) thermal resistance	27.5	°C/W
$R_{\theta JB}$	Junction-to-board thermal resistance	8.3	°C/W
ψ_{JT}	Junction-to-top characterization parameter	0.4	°C/W
ψ_{JB}	Junction-to-board characterization parameter	8.3	°C/W
$R_{\theta JC(bot)}$	Junction-to-case (bottom) thermal resistance	2.8	°C/W

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

6.5 Electrical Characteristics

T_J = –30°C to 125°C, V_{IN}= 5 V, typical values are at T_J = 25°C, (unless otherwise noted)

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
INPUT SUPPLY VOLTAGE						
V_{IN}	Input voltage range		2.7		6	V
UVLO	VIN undervoltage lockout	V_{IN} rising	2.35	2.45	2.6	V
		V_{IN} falling	2.15	2.25	2.37	V
$I_{DD(SDN)}$	Shutdown supply current	EN1 = EN2 = EN3 = 0 V		9		µA
$I_{DD(Q\_NSW)}$	Input quiescent current without buck1/2/3 switching	EN1 = EN2 = EN3 = 5 V, FB1 = FB2 = FB3 = 0.8 V		790		µA
$I_{DD(Q\_NSW1)}$		EN1 = 5 V, EN2 = EN3 = 0 V, FB1 = 0.8 V		340		µA
$I_{DD(Q\_NSW2)}$		EN2 = 5 V, EN1 = EN3 = 0 V, FB2 = 0.8 V		340		µA
$I_{DD(Q\_NSW3)}$		EN3 = 5 V, EN1 = EN2 = 0 V, FB3 = 0.8 V		340		µA
BUCK1, BUCK2, BUCK3						
V_{FB}	Feedback voltage	V_{COMP} = 1.2 V	0.594	0.6	0.606	V
$V_{EN(XH)}$	EN1/2/3 high-level input voltage			1.2	1.26	V
$V_{EN(XL)}$	EN1/2/3 low-level input voltage		1.1	1.15		V
$I_{EN(X1)}$	EN1/2/3 pullup current	ENx = 1 V	1.7	2.1	2.5	µA
$I_{EN(X2)}$	EN1/2/3 pullup current	ENx = 1.3 V		5.3		µA
$I_{EN(hys)}$	Hysteresis current			3.2		µA
I_{SSX}	Soft-start charging current		4.5	5.5	6.5	µA
$G_{(m\_PS1/2/3)}$	COMP1/2/3 voltage to inductor current G_m (1)	I_{LX} = 0.5 A		10		A/V
$I_{(LIMIT1)}$	Buck1 peak inductor current limit		3.55	4.6	5.6	A
$I_{(LIMITSINK1)}$	Buck1 low-side sink current limit			1.4		
$I_{(LIMIT2/3)}$	Buck2/3 peak inductor current limit		2.35	3.1	3.7	A
$I_{(LIMITSINK2/3)}$	Buck2/3 low-side sink current limit			1.2		A
$R_{ds(on)\_HS1}$	Buck1 high-side switch resistance (2)	VINQ = 5 V		45		mΩ
$R_{ds(on)\_LS1}$	Buck1 low-side switch resistance (2)	VINQ = 5 V		50		mΩ
$R_{ds(on)\_HS2}$	Buck2 high-side switch resistance (2)	VINQ = 5 V		60		mΩ
$R_{ds(on)\_LS2}$	Buck2 low-side switch resistance (2)	VINQ = 5 V		60		mΩ
$R_{ds(on)\_HS3}$	Buck3 high-side switch resistance (2)	VINQ = 5 V		60		mΩ
$R_{ds(on)\_LS3}$	Buck3 low-side switch resistance (2)	VINQ = 5 V		60		mΩ
POWER GOOD						
$V_{(th\_PG)}$	Feedback voltage threshold	FBx undervoltage falling		92.5		%VREF
		FBx undervoltage rising		95		%VREF
I_{PG}	PGOOD pin leakage				1	µA
$V_{(LOW\_PG)}$	PGOOD pin low voltage	$I_{(SINK)}$ = 1 mA			0.4	V

(1) Lab validation result
(2) Typical value without bonding wires; from design simulation

Electrical Characteristics (continued)

T_J = –30°C to 125°C, V_{IN}= 5 V, typical values are at T_J = 25°C, (unless otherwise noted)

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
OSCILLATOR						
F_{SW}	Switching frequency	$R_{(OSC)}$ = 51.1 kΩ	920	1000	1080	kHz
$F_{SW(range)}$	Switching frequency		250		2400	kHz
$F_{(SYNC)}$	Clock sync frequency range		250		2400	kHz
$F_{(SYNC\_HI)}$	Clock sync high threshold				2	V
$V_{(SYNC\_LO)}$	Clock sync low threshold		0.4			V
THERMAL PROTECTION						
$T_{(TRIP\_OTP)}$ (1)	Thermal protection trip point	Temperature rising		160		°C
$T_{(HYST\_OTP)}$ (1)		Hysteresis		20		°C

6.6 Timing Requirements

			MIN	NOM	MAX	UNIT
BUCK1, BUCK2, BUCK3						
t_{ON_MIN}	Minimum on-time			80	115	ns
G_{m_EA}	Error amplifier transconductance	–2 µA < $I_{(COMPX)}$ < 2 µA		290		µS
HICCUP TIMING						
t_{Hiccup_wait}	Overcurrent wait time(1)			512		cycles
t_{Hiccup_re}	Hiccup time before restart(1)			16382		cycles
POWER GOOD						
$t_{DEGLITCH(PG)\_F}$	PGOOD falling edge deglitch time			128		cycles
$t_{RDEGLITCH(PG)\_R}$	PGOOD rising edge deglitch time			16350		cycles
OSCILLATOR						
t_{SYNC_w}	Clock sync minimum pulse duration		80			ns

(1) Lab validation result

6.7 Typical Characteristics

T_A = 25°C, V_{IN} = 5 V, V_{OUT1} = 1.0 V, V_{OUT2} = 1.5 V, V_{OUT3} = 1.8 V F_{SW} = 1 MHz (unless otherwise noted)

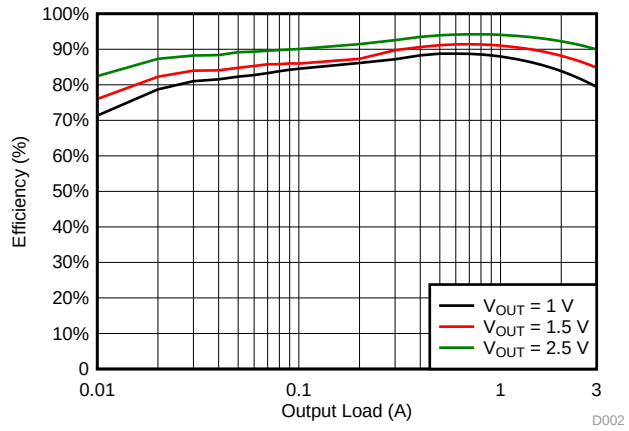


Figure 1. Buck1 Efficiency

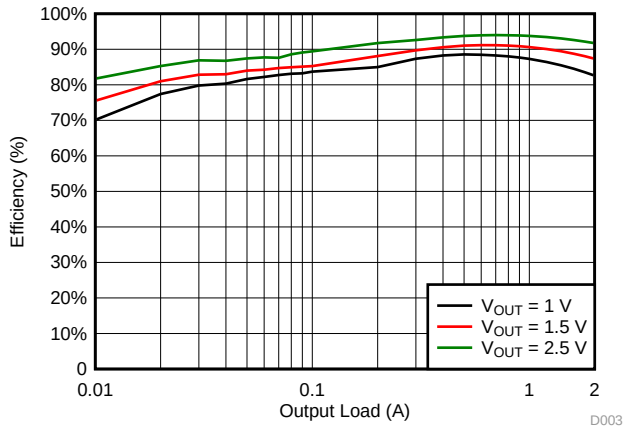


Figure 2. Buck2 Efficiency

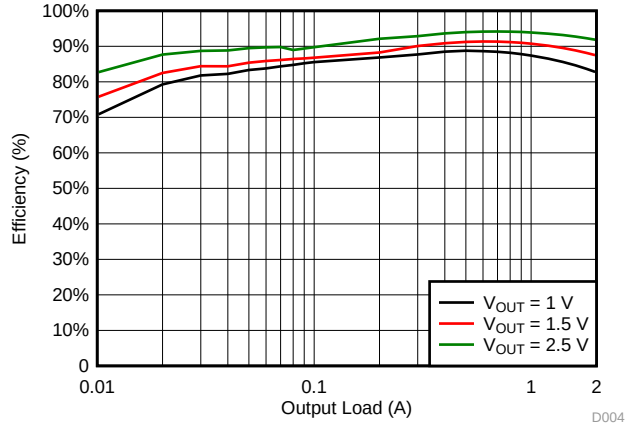


Figure 3. Buck3 Efficiency

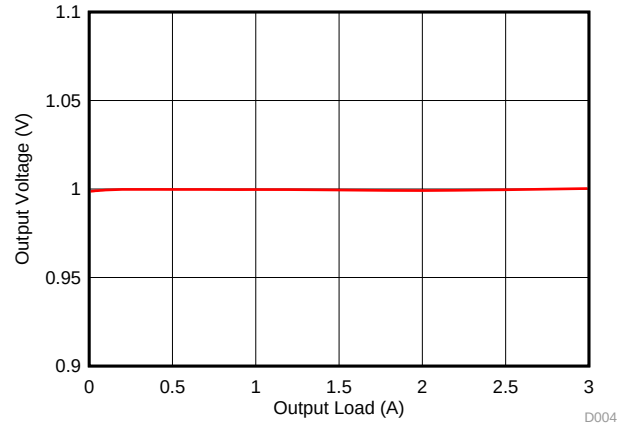


Figure 4. Buck1, Load Regulation

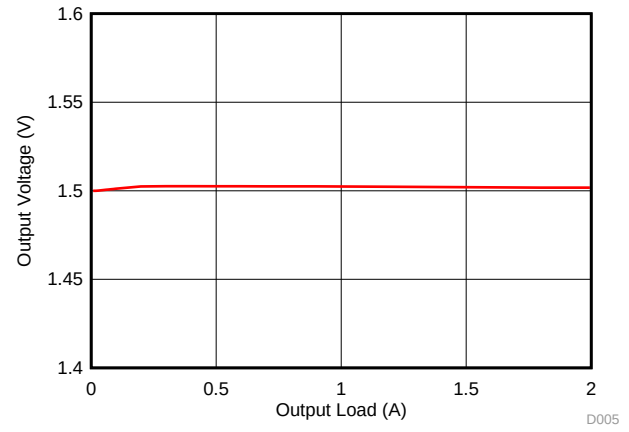


Figure 5. Buck2, Load Regulation

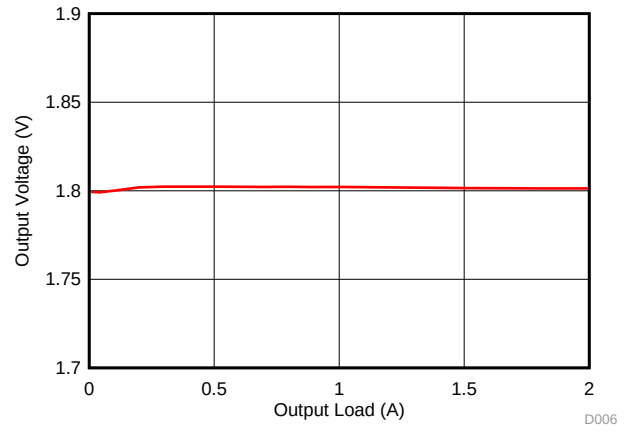


Figure 6. Buck3, Load Regulation

Typical Characteristics (continued)

T_A = 25°C, V_{IN} = 5 V, V_{OUT1} = 1.0 V, V_{OUT2} = 1.5 V, V_{OUT3} = 1.8 V F_{SW} = 1 MHz (unless otherwise noted)

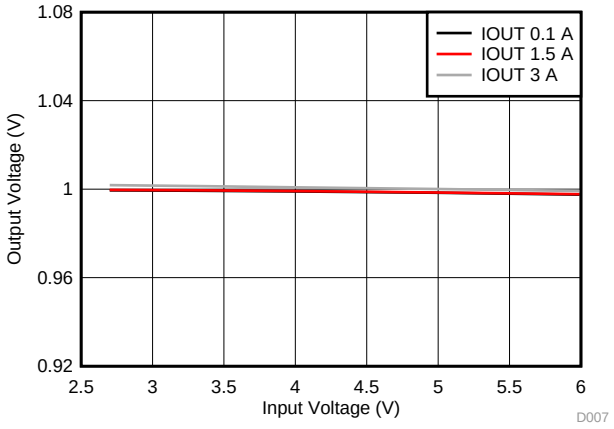


Figure 7. Buck1, Line Regulation

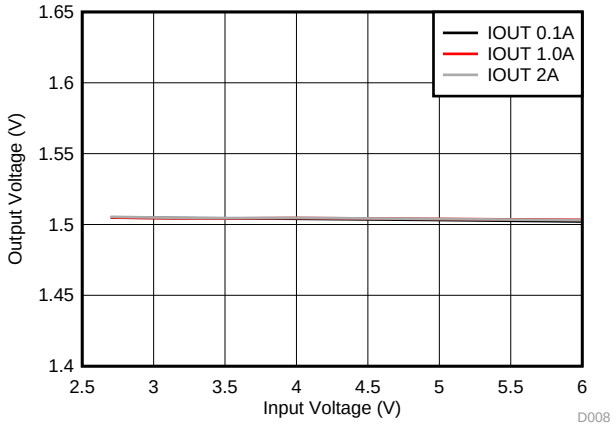


Figure 8. Buck2, Line Regulation

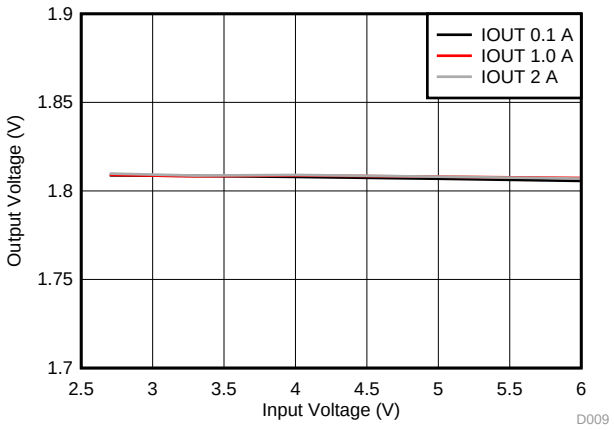


Figure 9. Buck3, Line Regulation

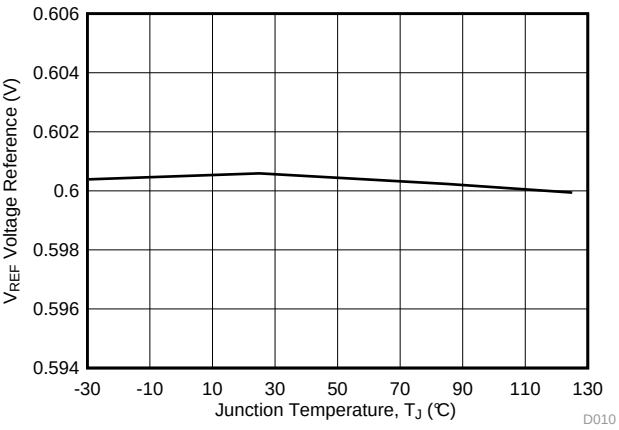


Figure 10. Voltage Reference vs Temperature

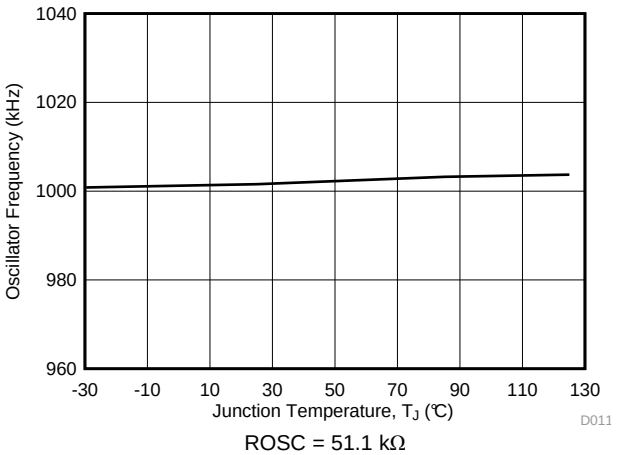


Figure 11. Oscillator Frequency vs Temperature

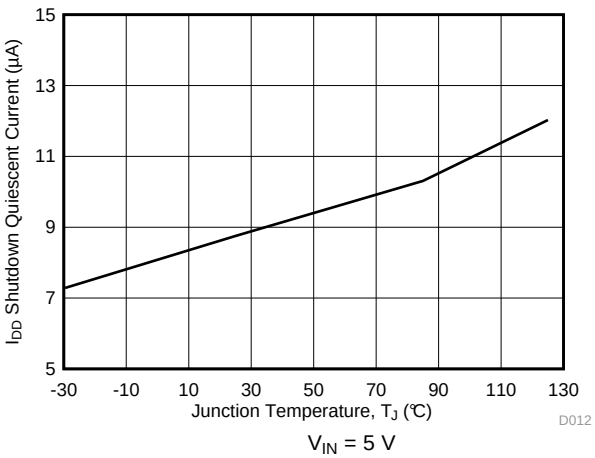


Figure 12. Shutdown Quiescent Current vs Temperature

Typical Characteristics (continued)

T_A = 25°C, V_{IN} = 5 V, V_{OUT1} = 1.0 V, V_{OUT2} = 1.5 V, V_{OUT3} = 1.8 V F_{SW} = 1 MHz (unless otherwise noted)

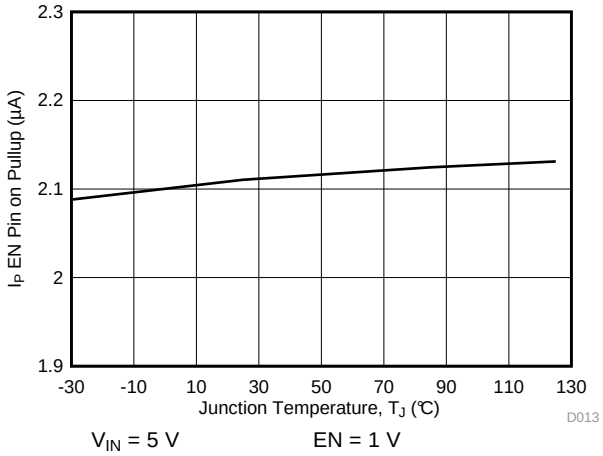


Figure 13. EN Pin Pullup Current vs Temperature, EN = 1 V

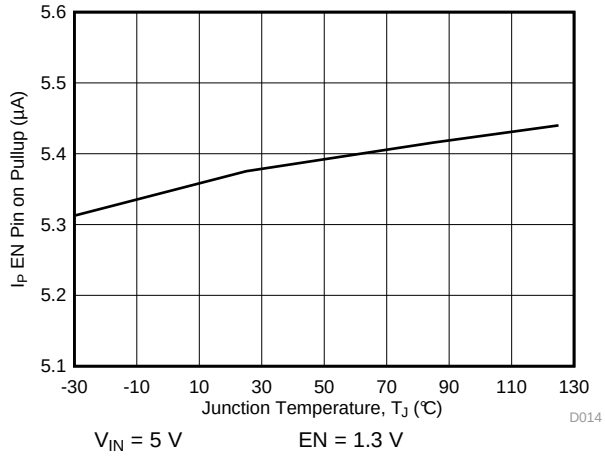


Figure 14. EN Pin Pullup Current vs Temperature, EN = 1.3 V

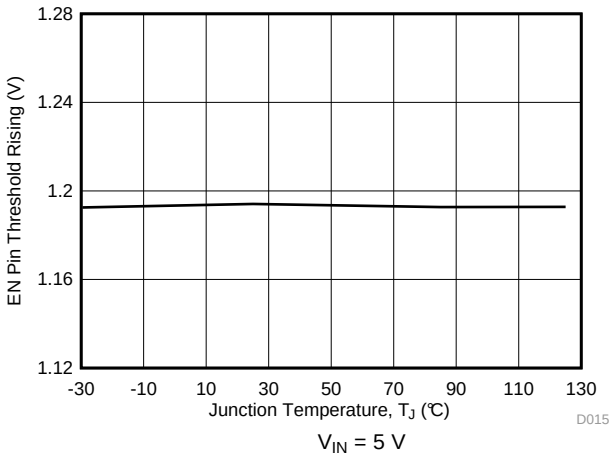


Figure 15. EN Pin Threshold Rising vs Temperature

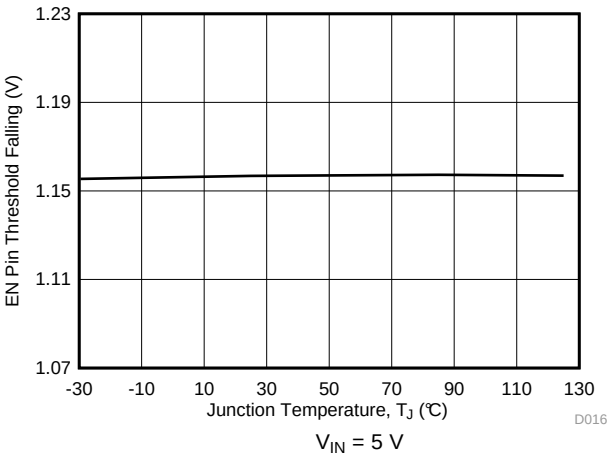


Figure 16. EN Pin Threshold Falling vs Temperature

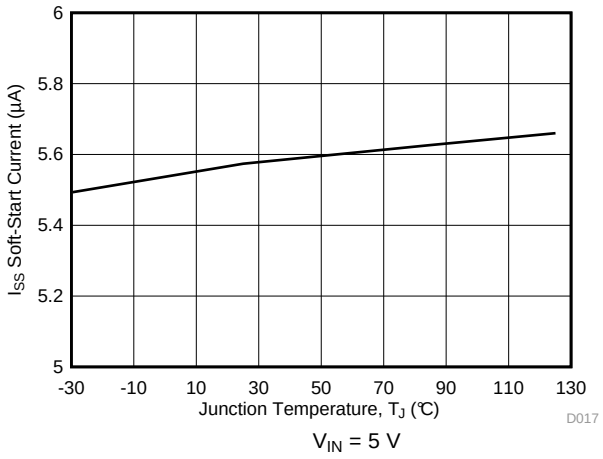


Figure 17. SS Pin Charge Current vs Temperature

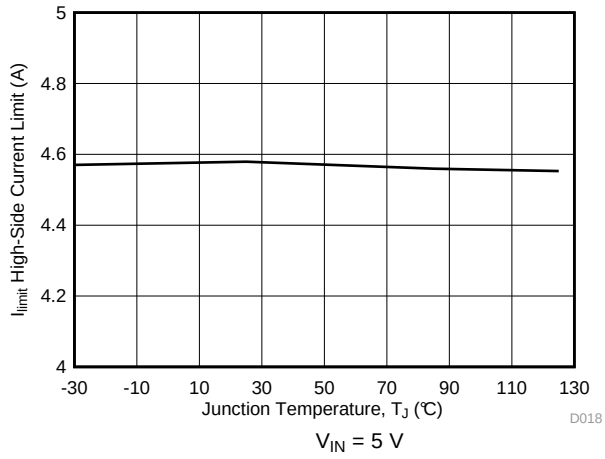


Figure 18. Buck1 High-Side Current Limit vs Temperature

Typical Characteristics (continued)

T_A = 25°C, V_{IN} = 5 V, V_{OUT1} = 1.0 V, V_{OUT2} = 1.5 V, V_{OUT3} = 1.8 V F_{SW} = 1 MHz (unless otherwise noted)

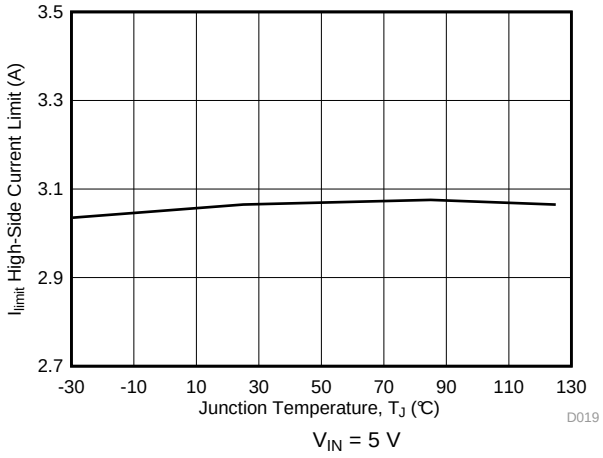


Figure 19. Buck2 High-Side Current Limit vs Temperature

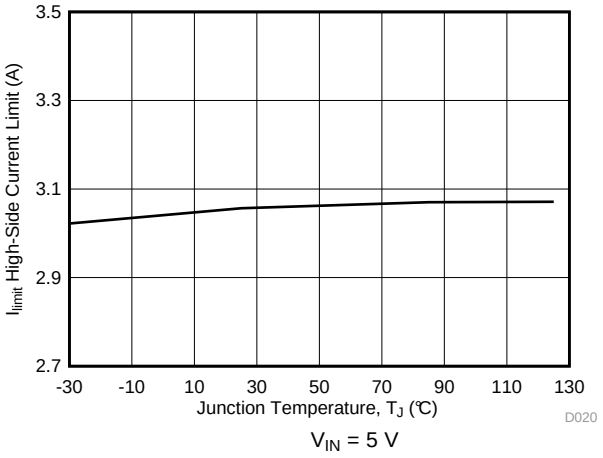


Figure 20. Buck3 High-Side Current Limit vs Temperature

7 Detailed Description

7.1 Overview

The TPS65266-1 is a triple 3-A/2-A/2-A output current, synchronous step-down (buck) converter for applications operating off the adaptor or battery with input voltage lower than 6 V. The feedback voltage reference for each buck is 0.6 V. Each buck is independent with dedicated enable, soft-start, and loop compensation. The TPS65266-1 implements a constant frequency, peak current mode control that simplifies external loop compensation. The switch clock of buck1 is 180° out-of-phase operation from the clock of buck2 and buck3 channels to reduce input current ripple, input capacitor size and power-supply-induced noise.

The TPS65266-1 has been designed for safe monotonic startup into prebiased loads. The default start-up is when VIN is typically 2.45 V. The ENx pin has an internal pullup current source that can be used to adjust the input voltage undervoltage lockout (UVLO) with two external resistors. In addition, the EN pin can be floating for automatically starting up the converters with the internal pullup current.

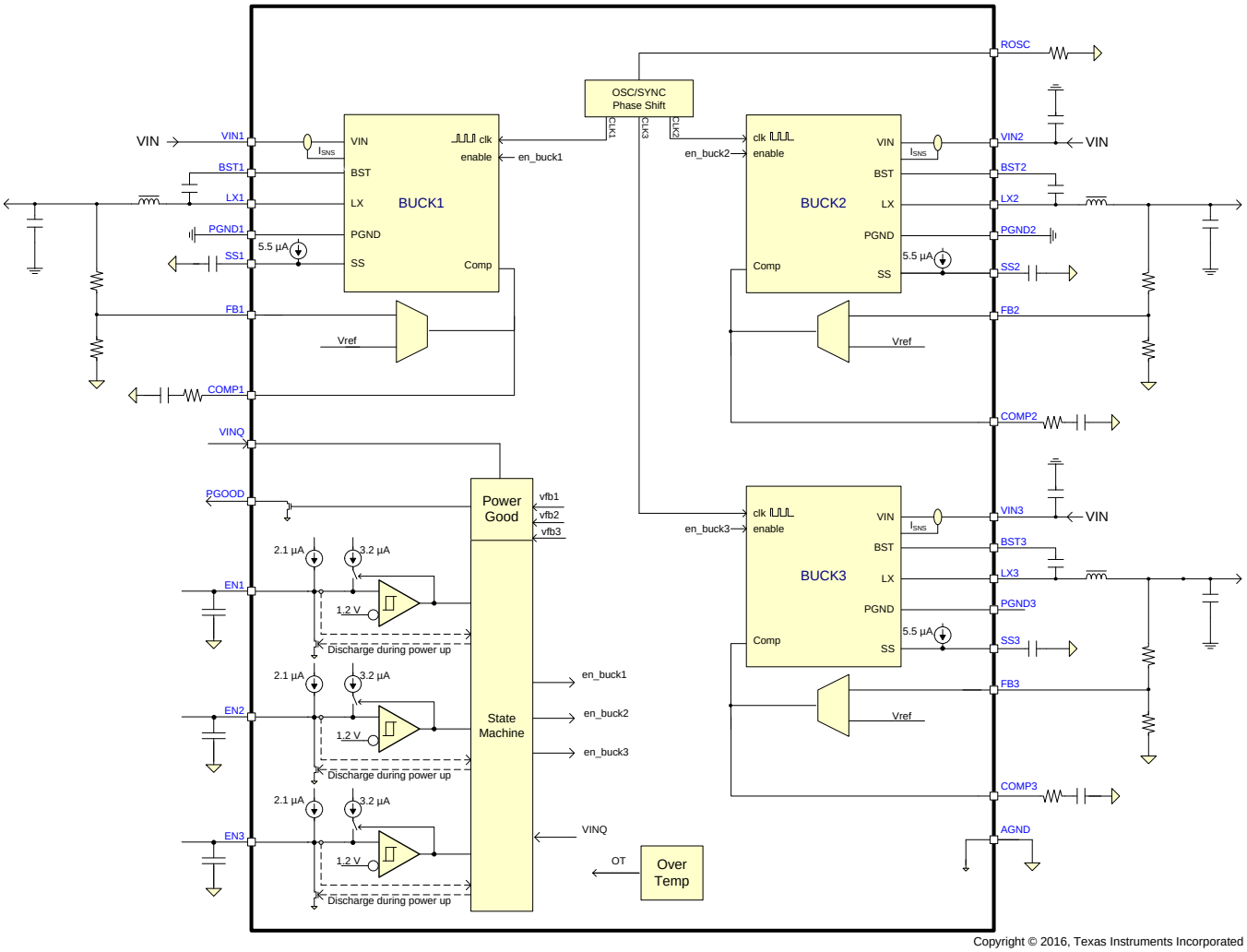
The TPS65266-1 features PGOOD pin to supervise output voltages of buck converter. The TPS65266-1 has power-good comparators with hysteresis, which monitor the output voltages through internal feedback voltages. When all bucks are in regulation range and power sequence is done, PGOOD is asserted high.

The SS (soft-start/tracking) pin is used to minimize inrush currents or provide power-supply sequencing during power up. A small-value capacitor or resistor divider should be coupled to the pin for soft start or critical power-supply sequencing requirements.

The TPS65266-1 is protected from overload and thermal fault conditions.

At light load, TPS65266-1 automatically operates in the pulse skipping mode (PSM) to save power.

7.2 Functional Block Diagram



7.3 Feature Description

7.3.1 Adjusting the Output Voltage

The output voltage of each buck is set with a resistor divider from the output of buck to the FB pin. TI recommends to use 1% tolerance or better resistors.

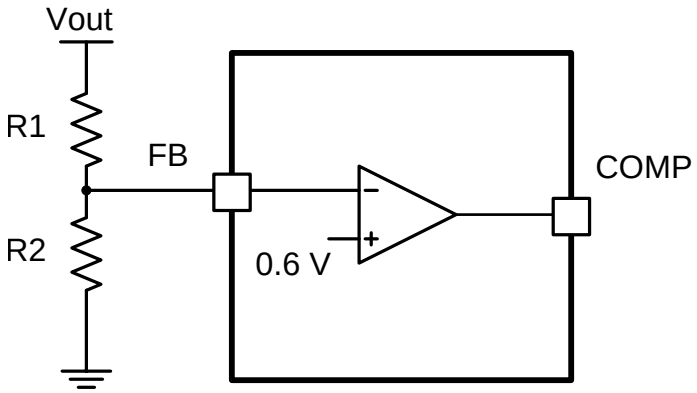


Figure 21. Voltage Divider Circuit

Feature Description (continued)

$$R_2 = R_1 \times \frac{0.6}{V_{out} - 0.6} \quad (1)$$

To improve efficiency at light loads, consider using larger-value resistors. If the values are too high, the regulator is more sensitive to noise. Table 1 shows the recommended resistor values.

Table 1. Output Resistor Divider Selection

Output Voltage (V)	R1 (kΩ)	R2 (kΩ)
1	10	15
1.2	10	10
1.5	15	10
1.8	20	10
2.5	31.6	10
3.3	45.3	10
3.3	22.6	4.99
5	73.2	10
5	36.5	4.99

7.3.2 Enable and Adjusting UVLO

The EN1/2/3 pin provides electrical on and off control of the device. After the EN1/2/3 pin voltage exceeds the threshold voltage, the device starts operation. If each ENx pin voltage is pulled below the threshold voltage, the regulator stops switching and enters low I_q state.

The EN pin has an internal pullup current source, allowing the user to float the EN pin for enabling the device. If an application requires controlling the EN pin, use open-drain or open-collector output logic to interface with the pin.

The device implements internal UVLO circuitry on the VINQ pin. The device is disabled when the VINQ pin voltage falls below the internal VIN UVLO threshold. The internal VIN UVLO threshold has a hysteresis of 200 mV. If an application requires either a higher UVLO threshold on the VINQ pin or a secondary UVLO on the VINx, in split-rail applications, then the ENx pin can be configured as shown in Figure 22, Figure 23, and Figure 24. When using the external UVLO function, TI recommends to set the hysteresis to be >200 mV.

The EN pin has a small pullup current I_p, which sets the default state of the pin to enable when no external components are connected. The pullup current is also used to control the voltage hysteresis for the UVLO function because it increases by I_h after the EN pin crosses the enable threshold. Calculate the UVLO thresholds using Equation 2 and Equation 3.

$$R_1 = \frac{V_{START}\left(\frac{V_{ENFALLING}}{V_{ENRISING}}\right) - V_{STOP}}{I_P\left(1 - \frac{V_{ENFALLING}}{V_{ENRISING}}\right) + I_h}$$

$$R_2 = \frac{R_1 \times V_{ENFALLING}}{V_{STOP} - V_{ENFALLING} + R_1(I_h + I_p)} \quad (2)$$

where

- I_h = 3.2 µA
- I_p = 2.1 µA
- $V_{ENRISING}$ = 1.2 V
- $V_{ENFALLING}$ = 1.15 V (3)

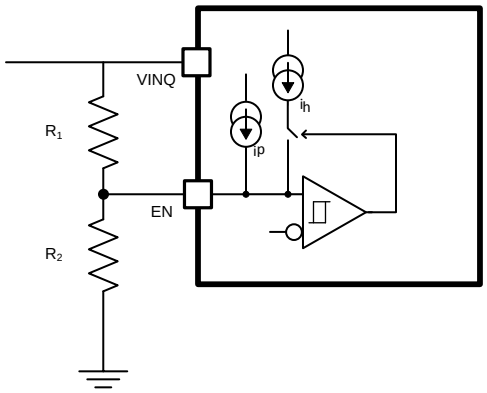


Figure 22. Adjustable VINQ UVLO

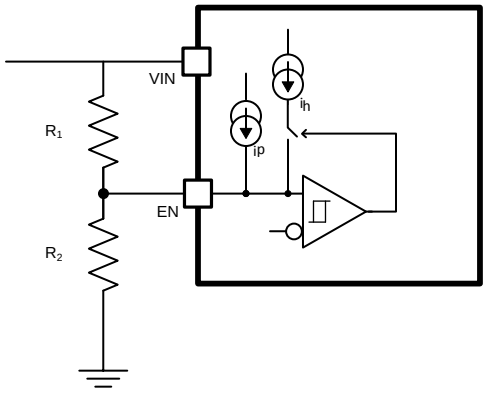


Figure 23. Adjustable VIN UVLO, VINQ > 2.7 V

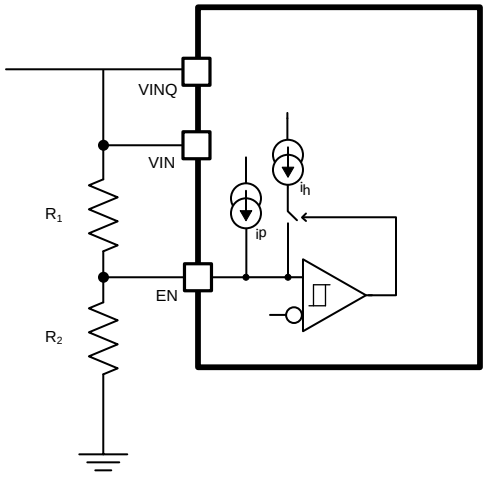


Figure 24. Adjustable VIN and VINQ UVLO

7.3.3 Soft-Start Time

The voltage on the respective SS pin controls the start-up of buck output. When the voltage on the SS pin is less than the internal 0.6-V reference, the TPS65266-1 regulates the internal feedback voltage to the voltage on the SS pin instead of 0.6 V. The SS pin can be used to program an external soft-start function or to allow output of buck to track another supply during start-up. The device has an internal pullup current source of 5.5 μA (typical) that charges an external soft-start capacitor to provide a linear ramping voltage at SS pin. The TPS65266-1 regulates the internal feedback voltage to the voltage on the SS pin, allowing VOUT to rise smoothly from 0 V to its regulated voltage without inrush current. Calculate the approximate soft-start time with Equation 4.

$$t_{ss}\,(ms) = \frac{Css\,(nF) \times Vref\,(V)}{Iss\,(\mu A)} \tag{4}$$

Many of the common power-supply sequencing methods can be implemented using the SSx and ENx pins. Figure 25 shows the method implementing ratiometric sequencing by connecting the SSx pins of three buck channels together. The regulator outputs ramp up and reach regulation at the same time. When calculating the soft-start time, the pullup current source must be tripled in Equation 4.

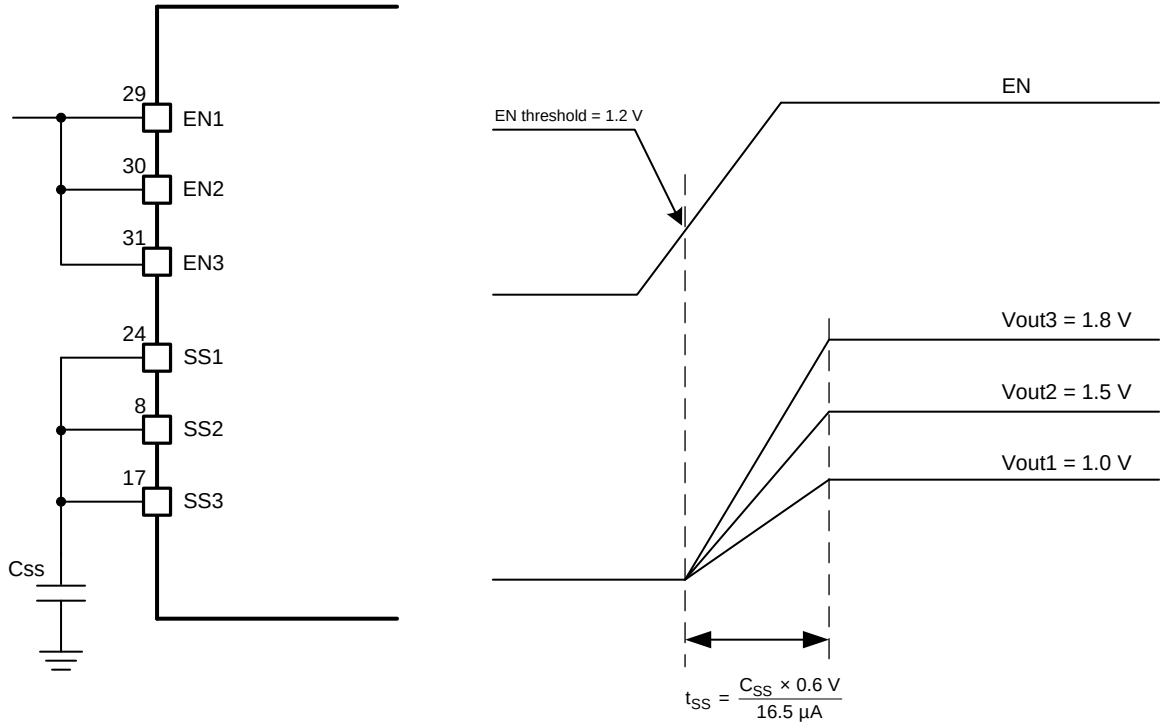


Figure 25. Ratiometric Power-Up Using SSx Pins

Simultaneous power-supply sequencing can be implemented by connecting capacitor to SSx pin, shown in Figure 26. Using Equation 4 and Equation 5, calculate the capacitors.

$$\frac{Css1}{Vout1} = \frac{Css2}{Vout2} = \frac{Css3}{Vout3} \tag{5}$$

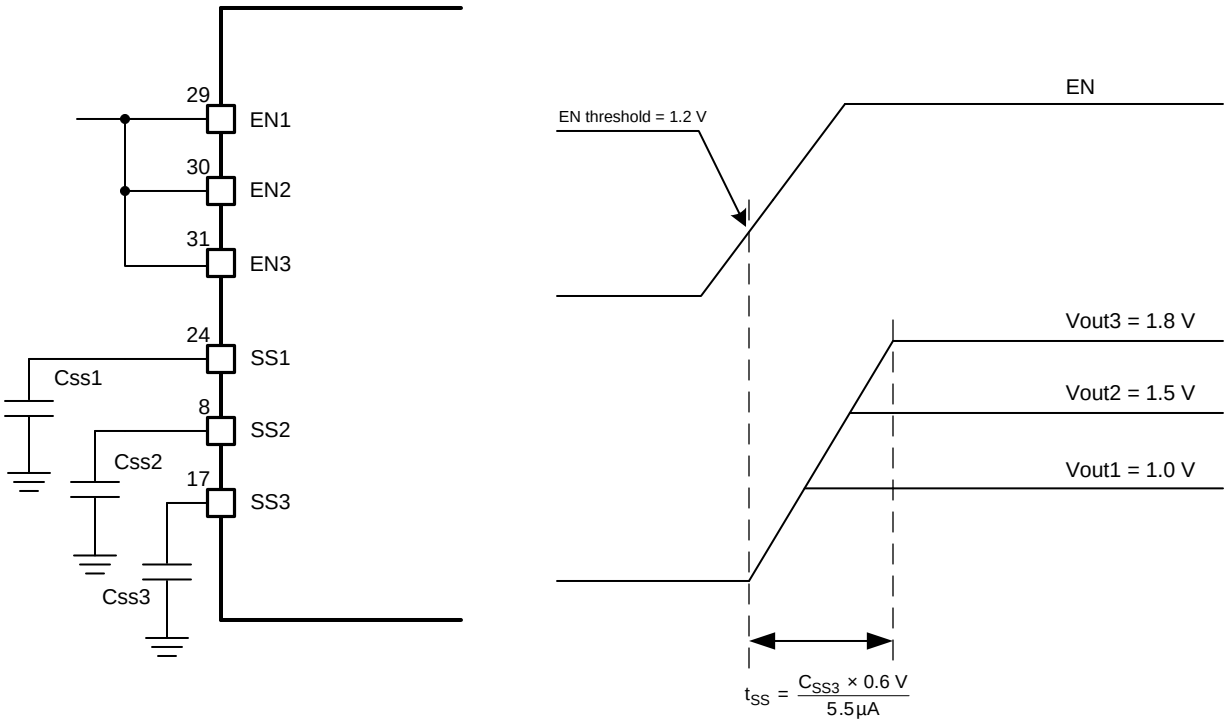


Figure 26. Simultaneous Startup Sequence Using SSx Pins

7.3.4 Power-Up Sequencing

The TPS65266-1 has a dedicated enable pin and soft-start pin for each converter. The converter enable pins are biased by a current source that allows for easy sequencing by the addition of an external capacitor. Enable pins have a discharge function, which ensures power-up sequencing is effective at quickly powering down and up status. Disabling the converter with an active pulldown transistor on the ENs pin allows for a predictable power-down timing operation. Figure 27 shows the timing diagram of a typical buck power-up sequence with connecting a capacitor at ENx pin.

When VINQ pin voltage rises to about 1 V, the internal EN turns on and a typical 1.4-µA current is charging ENx pin from input supply. If any of the EN pin voltages reaches 0.5 V when powered up, three EN pin discharge functions are triggered and keep 2 ms with discharge resistor around 1.2 kΩ to GND, then a 2.1-µA pullup current is sourcing ENx. After ENx pin voltage reaches to ENx enabling threshold, 3.2-µA hysteresis current sources to the pin to improve noise sensitivity.

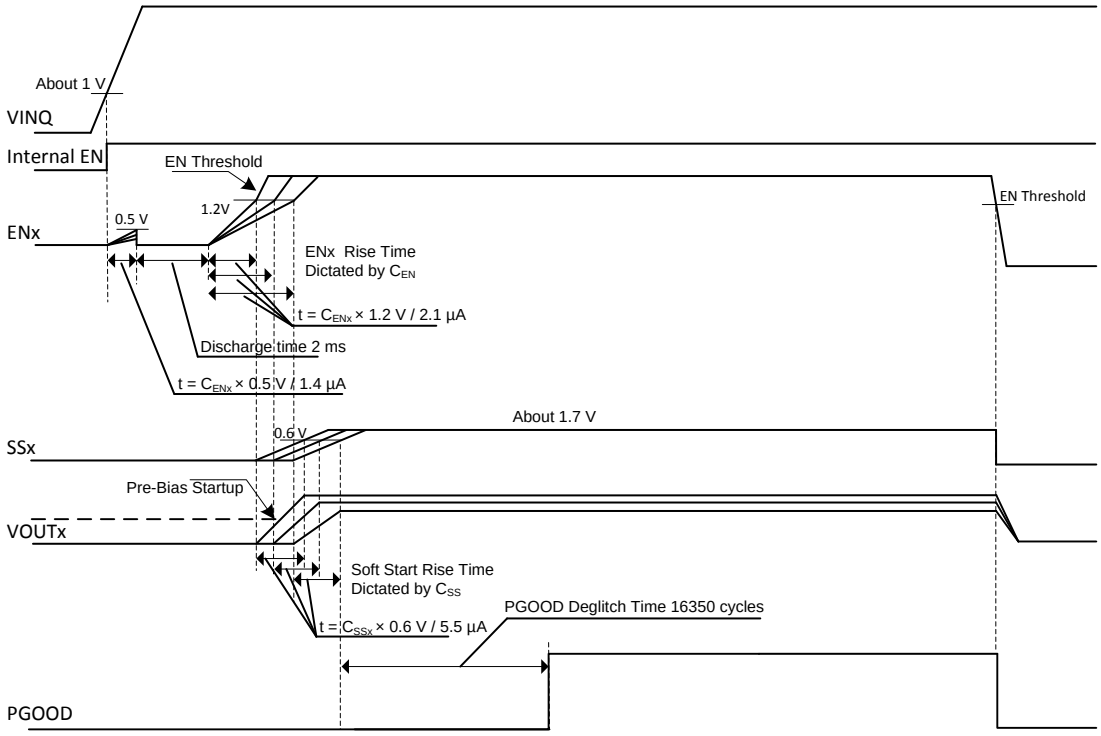


Figure 27. Startup Power Sequence

7.3.5 Bootstrap Voltage and BST-LX UVLO

Each high-side MOSFET driver is biased from the floating bootstrap capacitor, CB, shown in Figure 28, which is normally recharged during each cycle through an internal low-side MOSFET or the body diode of a low-side MOSFET when the high-side MOSFET turns off. The boot capacitor is charged when the BST pin voltage is less than VIN and BST-LX voltage is below regulation. The recommended value of this ceramic capacitor is 47 nF. TI recommends a ceramic capacitor with an X7R- or X5R-grade dielectric with a voltage rating of 10 V or higher because of the stable characteristics over temperature and voltage.

To improve dropout, the device is designed to operate at 100% duty cycle as long as the BST to LX pin voltage is greater than the BST-LX UVLO threshold, which is typically 2.1 V. When the voltage between BST and LX drops below the BST-LX UVLO threshold, the high-side MOSFET is turned off and the low-side MOSFET is turned on allowing the boot capacitor to be recharged.

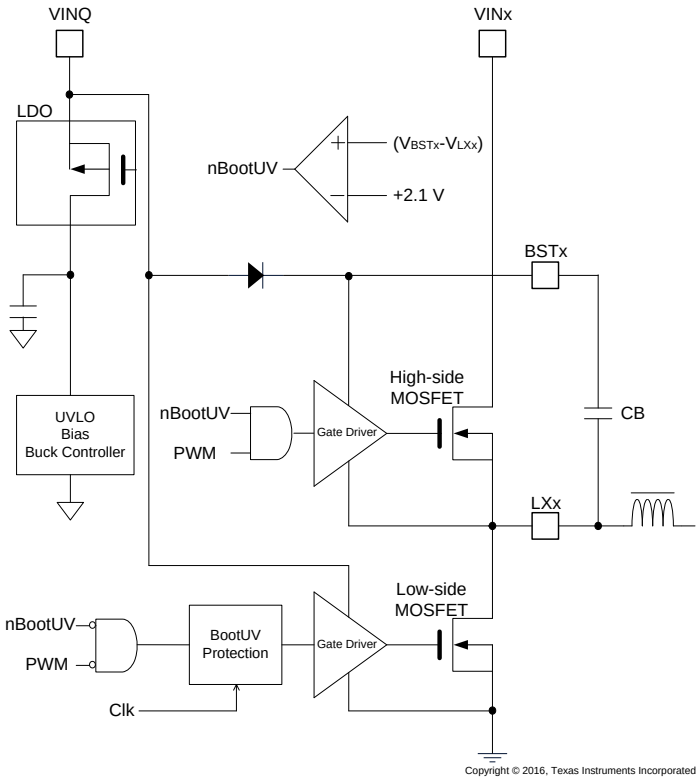


Figure 28. Bootstrap Voltage and Diagram

7.3.6 Out of Phase Operation

To reduce input ripple current, the switch clock of buck1 is 180° out-of-phase from the clock of buck2 and buck3. This enables the system by having less input current ripple to reduce input capacitors' size, cost, and EMI.

7.3.7 Output Overvoltage Protection (OVP)

The device incorporates an OVP circuit to minimize output voltage overshoot. When the output is overloaded, the error amplifier compares the actual output voltage to the internal reference voltage. If the FB pin voltage is lower than the internal reference voltage for a considerable time, the output of the error amplifier demands maximum output current. After the condition is removed, the regulator output rises and the error amplifier output transitions to the steady-state voltage. In some applications with small output capacitance, the load can respond faster than the error amplifier. This leads to the possibility of an output overshoot. Each buck compares the FB pin voltage to the OVP threshold. If the FB pin voltage is greater than the OVP threshold, the high-side MOSFET is turned off preventing current from flowing to the output and minimizing output overshoot. When the FB voltage drops lower than the OVP threshold, the high-side MOSFET turns on at the next clock cycle.

7.3.8 Slope Compensation

To prevent the subharmonic oscillations when the device operates at duty cycles greater than 50%, the TPS65266-1 adds built-in slope compensation, which is a compensating ramp to the switch current signal.

7.3.9 Overcurrent Protection

The device is protected from overcurrent conditions by cycle-by-cycle current limiting on both the high-side MOSFET and low-side MOSFET.

7.3.9.1 High-Side MOSFET Overcurrent Protection

The device implements current mode control, which uses the COMP pin voltage to control the turn off of the high-side MOSFET and the turn-on of the low-side MOSFET on a cycle-by-cycle basis. Each cycle the switch current and the current reference generated by the COMP pin voltage are compared, when the peak switch current intersects the current reference, the high-side switch is turned off.

7.3.9.2 Low-Side MOSFET Overcurrent Protection

While the low-side MOSFET is turned on, its conduction current is monitored by the internal circuitry. During normal operation the low-side MOSFET sources current to the load. At the end of every clock cycle, the low-side MOSFET sourcing current is compared to the internally set low-side sourcing current limit. If the low-side sourcing current is exceeded, the high-side MOSFET is not turned on and the low-side MOSFET stays on for the next cycle. The high-side MOSFET is turned on again when the low-side current is below the low-side sourcing current limit at the start of a cycle.

The low-side MOSFET may also sink current from the load. If the low-side sinking current limit is exceeded, the low-side MOSFET is turned off immediately for the rest of that clock cycle. In this scenario, both MOSFETs are off until the start of the next cycle.

Furthermore, if an output overload condition (as measured by the COMP pin voltage) has lasted for more than the hiccup wait time, which is programmed for 512 cycles (typical) shown in Figure 29, the device shuts down itself and restarts after the hiccup time of 16382 cycles (typical). The hiccup mode helps to reduce the device power dissipation under a severe overcurrent condition.

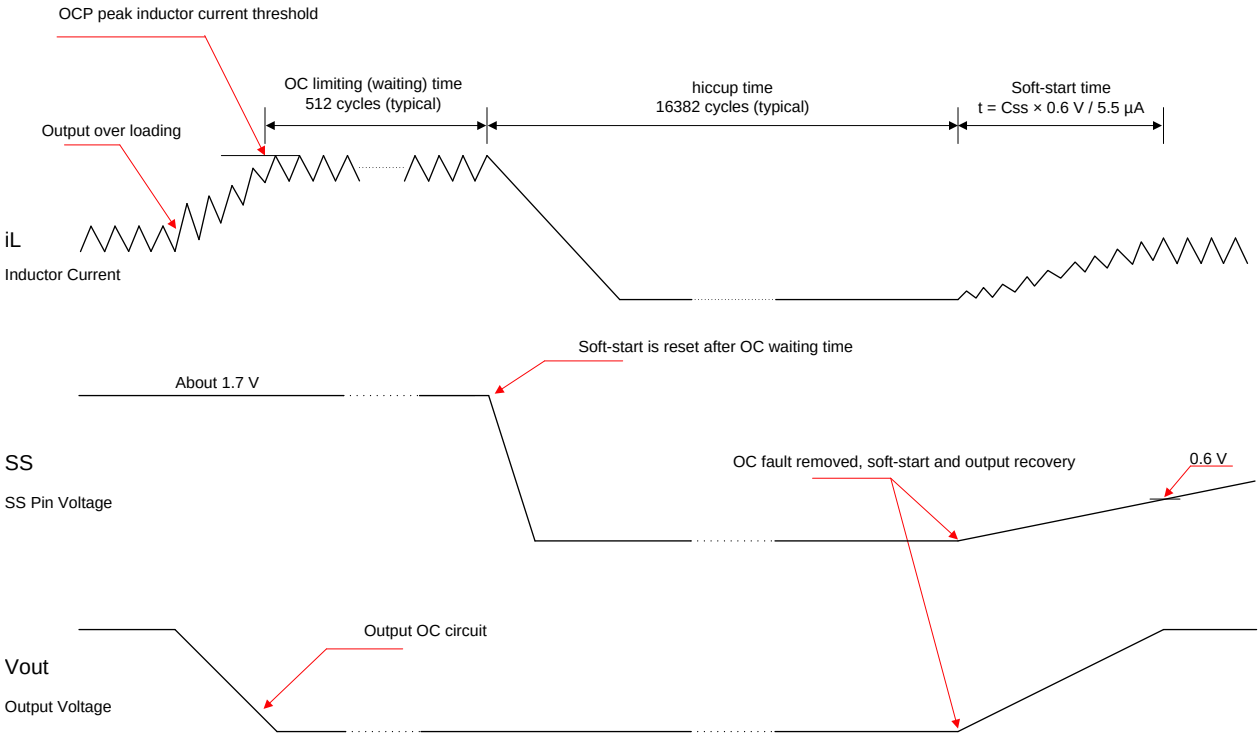


Figure 29. Overcurrent Protection

7.3.10 Power Good

The PGOOD pin is an open-drain output. After the feedback voltage of each buck is higher than 95% (rising) of the internal voltage reference, the PGOOD pin pulldown is deasserted and the pin floats. TI recommends to use a pullup resistor between the values of 10 kΩ and 100 kΩ to a voltage source that is 5.0 V or less.

The PGOOD pin is pulled low when any feedback voltage of a buck is lower than 92.5% (falling) of the nominal internal reference voltage. Also, the PGOOD is pulled low if the input voltage is undervoltage locked up, thermal shutdown is asserted, the EN pin is pulled low, or the converter is in a soft-start period.

7.3.11 Adjustable Switching Frequency

The ROSC pin can be used to set the switching frequency by connecting a resistor to GND. The switching frequency of the device is adjustable from 250 kHz to 2.4 MHz.

To determine the ROSC resistance for a given switching frequency, use Equation 6 or the curve in Figure 30. To reduce the solution size, the user should set the switching frequency as high as possible, but consider the tradeoffs of the supply efficiency and minimum controllable on-time.

$$f_{osc}\,(\text{kHz}) = 46657 \times R\,(\text{k}\Omega)^{-0.976} \quad (6)$$

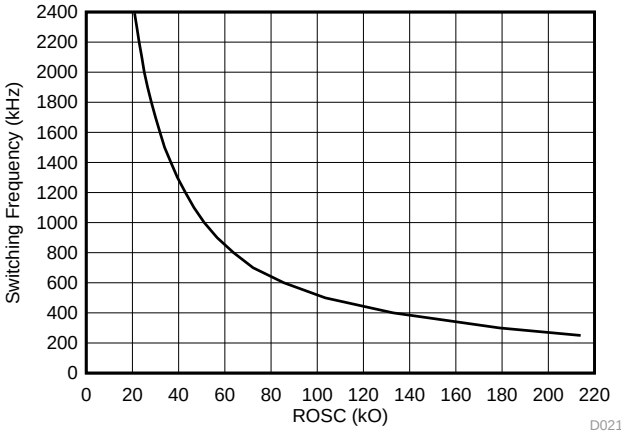


Figure 30. ROSC vs Switching Frequency

When an external clock applies to the ROSC pin, the internal phase locked loop (PLL) has been implemented to allow internal clock synchronizing to an external clock between 250 kHz and 2.4 MHz. To implement the clock synchronization feature, connect a square wave clock signal to the ROSC pin with a duty cycle between 20% to 80%. The clock signal amplitude must transition lower than 0.4 V and higher than 2.0 V. The start of the switching cycle is synchronized to the falling edge of the ROSC pin.

In applications where both resistor mode and synchronization mode are needed, the device can be configured as shown in Figure 31. Before an external clock is present, the device works in resistor mode and ROSC resistor sets the switching frequency. When an external clock is present, the synchronization mode overrides the resistor mode. The first time the ROSC pin is pulled above the ROSC high threshold (2.0 V), the device switches from the resistor mode to the synchronization mode and the ROSC pin becomes high impedance as the PLL starts to lock onto the frequency of the external clock. TI does not recommended to switch from synchronization mode back to resistor mode because the internal switching frequency drops to 100 kHz first before returning to the switching frequency set by ROSC resistor.

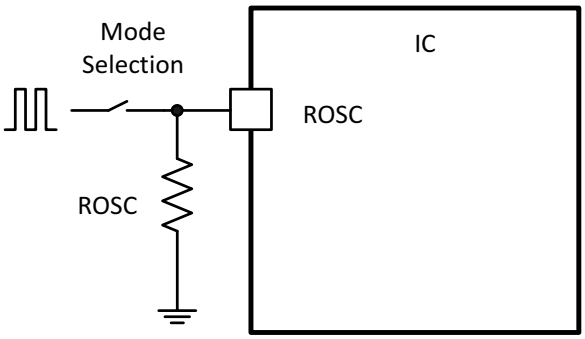


Figure 31. Works With Resistor Mode and Synchronization Mode

7.3.12 PSM

The TPS65266-1 can enter high-efficiency PSM operation at light load current.

When a controller is enabled for PSM operation, the peak inductor current is sensed and compared with 250-mA current typically. Because the integrated current comparator catches the peak inductor current only, the average load current entering PSM varies with the applications and external output filters. In PSM, the sensed peak inductor current is clamped at 250 mA (see Figure 32).

When a controller operates in PSM, the inductor current is not allowed to reverse. The reverse current comparator turns off the low-side MOSFET when the inductor current reaches 0, preventing it from reversing and going negative.

Due to the delay in the circuit and current comparator tdly (typical 50 nS at V_{IN} = 5 V), the real peak inductor current threshold to turn off high-side power MOSFET could shift higher depending on inductor inductance and input/output voltages. Calculate the threshold of peak inductor current to turn off high-side power MOSFET with Equation 7.

$$IL_{PEAK} = 250\ mA + \frac{V_{IN} - V_{OUT}}{L} \times tdly \quad (7)$$

When the charge accumulated on V_{OUT} capacitor is more than loading need, COMP pin voltage drops to low voltage driven by error amplifier. There is an internal comparator at COMP pin. If comp voltage is < 0.35 V, the power stage stops switching to save power.

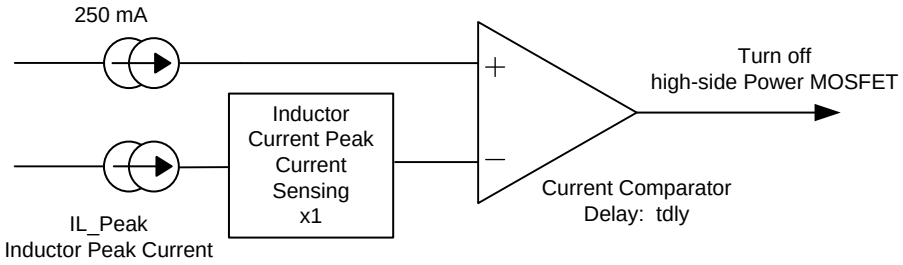


Figure 32. PSM Current Comparator

7.3.13 Thermal Shutdown

The internal thermal shutdown circuitry forces the device to stop switching if the junction temperature exceeds 160°C typically. The device reinitiates the power-up sequence when the junction temperature drops below 140°C typically.

7.4 Device Functional Modes

7.4.1 Operation With V_{IN} < 2.6 V (Minimum V_{IN})

The device operates with input voltages above 2.6 V. The maximum UVLO voltage is 2.6 V and will operate at input voltages above 2.6 V. The typical UVLO voltage is 2.45 V and the device may operate at input voltages above that point. The device also may operate at lower input voltages, the minimum UVLO voltage is 2.35 V (rising) and 2.15V (falling). At input voltages below the UVLO minimum voltage, the devices will not operate.

7.4.2 Operation With EN Control

The enable rising edge threshold voltage is 1.2 V typical and 1.26 V maximum. With EN held below that voltage the device is disabled and switching is inhibited. The IC quiescent current is reduced in this state. When input voltage is above the UVLO threshold and the EN voltage is increased above the rising edge threshold, the device becomes active. Switching is enabled, and the soft start sequence is initiated. The device will start at the soft start time determined by the external soft start capacitor as shown in Figure 35 to Figure 37.

8 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.

8.1 Application Information

The device is a triple-synchronous step-down DC/DC converter. It is typically used to convert a higher DC voltage to lower DC voltages with a continuously available output current of 3 A/2 A/2 A. The following design procedure can be used to select component values for the TPS65266-1. This section presents a simplified discussion of the design process.

8.2 Typical Application

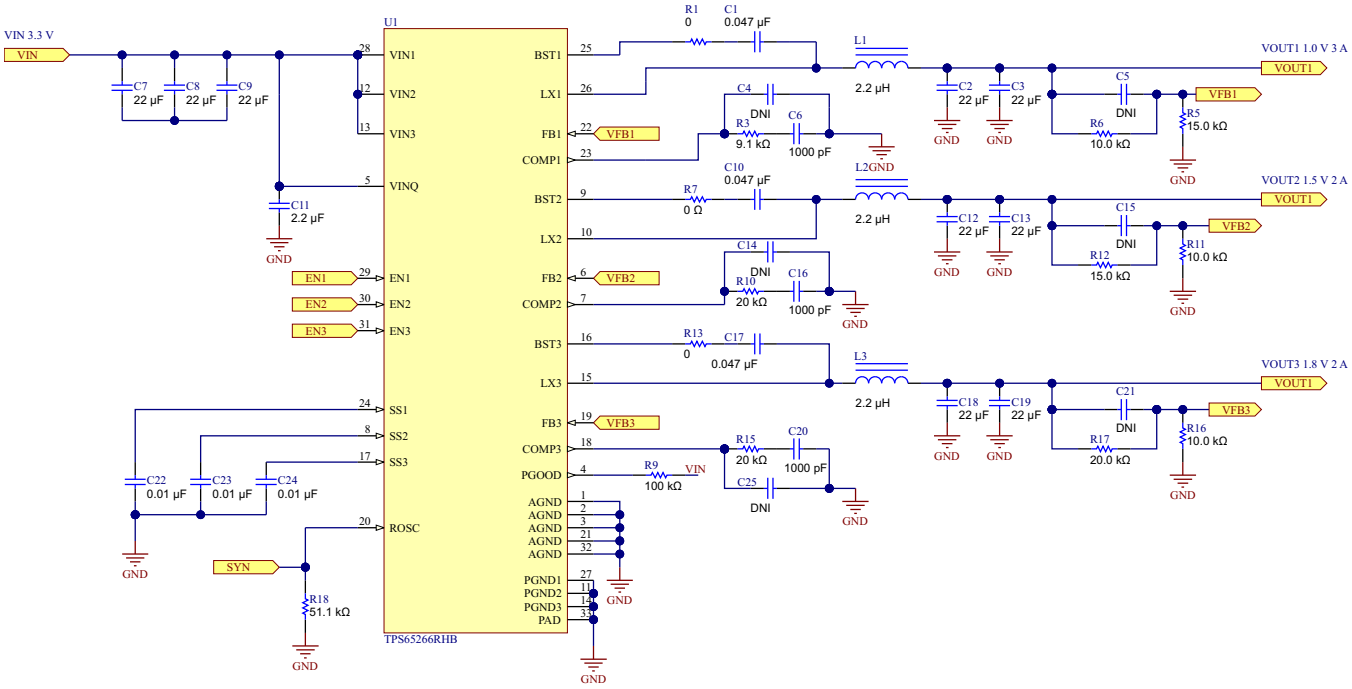


Figure 33. Typical Application Schematic

Typical Application (continued)

8.2.1 Design Requirements

This example details the design of a triple-synchronous step-down converter. The designer must know a few parameters to start the design process. These parameters are typically determined at the system level. For this example, start with the following known parameters in Table 2.

Table 2. Design Parameters

Parameter	Value
Vout1	1.0 V
Iout1	3 A
Vout2	1.5 V
Iout2	2 A
Vout3	1.8 V
Iout3	2 A
Transient response 1-A load step	±5%
Input voltage	5.0 V normal, 2.7 to 6 V
Output voltage ripple	±1%
Switching frequency	1 MHz

8.2.2 Detailed Design Procedure

8.2.2.1 Output Inductor Selection

To calculate the value of the output inductor, use Equation 8. LIR is a coefficient that represents the amount of inductor ripple current relative to the maximum output current. The inductor ripple current is filtered by the output capacitor. Therefore, choosing high inductor ripple currents impact the selection of the output capacitor because the output capacitor must have a ripple current rating equal to or greater than the inductor ripple current. In general, the inductor ripple value is at the discretion of the designer; however, LIR is normally from 0.1 to 0.3 for the majority of applications.

$$L = \frac{V_{inmax} - V_{out}}{I_o \times LIR} \times \frac{V_{out}}{V_{inmax} \times f_{sw}} \quad (8)$$

For the output filter inductor, it is important that the RMS current and saturation current ratings not be exceeded. The RMS and peak inductor current can be found from Equation 10 and Equation 11.

$$I_{ripple} = \frac{V_{inmax} - V_{out}}{L} \times \frac{V_{out}}{V_{inmax} \times f_{sw}} \quad (9)$$

$$I_{Lrms} = \sqrt{I_O^2 + \frac{\left(\frac{V_{out} \times (V_{inmax} - V_{out})}{V_{inmax} \times L \times fsw}\right)^2}{12}} \quad (10)$$

$$I_{Lpeak} = I_{out} + \frac{I_{ripple}}{2} \quad (11)$$

The current flowing through the inductor is the inductor ripple current plus the output current. During power-up, faults, or transient load conditions, the inductor current can increase above the calculated peak inductor current level calculated previously. In transient conditions, the inductor current can increase up to the switch current limit of the device. For this reason, the most conservative approach is to specify an inductor with a saturation current rating equal to or greater than the switch current limit rather than the peak inductor current.

8.2.2.2 Output Capacitor Selection

The three primary considerations for selecting the value of the output capacitor are: the output capacitor determines the modulator pole, the output voltage ripple, and how the regulator responds to a large change in load current. The output capacitance must be selected based on the most stringent of these three criteria.

The first criterion is the desired response to a large change in the load current. The output capacitor needs to supply the load with current when the regulator cannot. This situation would occur if there are desired hold-up times for the regulator where the output capacitor must hold the output voltage above a certain level for a specified amount of time after the input power is removed. The regulator is also temporarily not able to supply sufficient output current if there is a large, fast increase in the current needs of the load such as a transition from no load to full load. The regulator usually needs two or more clock cycles for the control loop to see the change in load current and output voltage and adjust the duty cycle to react to the change. The output capacitor must be sized to supply the extra current to the load until the control loop responds to the load change. The output capacitance must be large enough to supply the difference in current for two clock cycles while only allowing a tolerable amount of droop in the output voltage. Equation 12 shows the minimum output capacitance necessary to accomplish this.

$$C_o = \frac{2 \times \Delta I_{out}}{f_{sw} \times \Delta V_{out}}$$

where

- ΔIout is the change in output current.
- f_{sw} is the regulator's switching frequency.
- ΔVout is the allowable change in the output voltage. (12)

Equation 13 calculates the minimum output capacitance needed to meet the output voltage ripple specification.

$$C_o > \frac{1}{8 \times f_{sw}} \times \frac{1}{\frac{V_{oripple}}{I_{oripple}}}$$

where

- f_{sw} is the switching frequency.
- V_{ripple} is the maximum allowable output voltage ripple.
- I_{ripple} is the inductor ripple current. (13)

Equation 14 calculates the maximum ESR an output capacitor can have to meet the output voltage ripple specification.

$$R_{ESR} < \frac{V_{oripple}}{I_{oripple}} \quad (14)$$

Additional capacitance deratings for aging, temperature, and DC bias should be factored in, which increase this minimum value. Capacitors generally have limits to the amount of ripple current they can handle without failing or producing excess heat. An output capacitor that can support the inductor ripple current must be specified. Some capacitor data sheets specify the root mean square (RMS) value of the maximum ripple current. Equation 15 can be used to calculate the RMS ripple current the output capacitor needs to support.

$$I_{corms} = \frac{V_{out} \times (V_{inmax} - V_{out})}{\sqrt{12} \times V_{inmax} \times L \times f_{sw}} \quad (15)$$

8.2.2.3 Input Capacitor Selection

The TPS65266-1 requires a high-quality ceramic, type X5R or X7R, input decoupling capacitor of at least 10 µF of effective capacitance on the VIN input voltage pins. In some applications, additional bulk capacitance may also be required for the VIN input. The effective capacitance includes any DC bias effects. The voltage rating of the input capacitor must be greater than the maximum input voltage. The capacitor must also have a ripple current rating greater than the maximum input current ripple of the TPS65266-1. Calculate the input ripple current using Equation 16.

$$I_{inrms} = I_{out} \times \sqrt{\frac{V_{out}}{V_{inmin}} \times \frac{(V_{inmin} - V_{out})}{V_{inmin}}} \quad (16)$$

The value of a ceramic capacitor varies significantly over temperature and the amount of DC bias applied to the capacitor. The capacitance variations due to temperature can be minimized by selecting a dielectric material that is stable over temperature. X5R and X7R ceramic dielectrics are usually selected for power regulator capacitors because they have a high capacitance-to-volume ratio and are fairly stable over temperature. The output capacitor must also be selected with the DC bias taken into account. The capacitance value of a capacitor decreases as the DC bias across a capacitor increases. The input capacitance value determines the input ripple voltage of the regulator. Calculate the input voltage ripple using Equation 17.

$$\Delta V_{in} = \frac{I_{outmax} \times 0.25}{C_{in} \times f_{sw}} \quad (17)$$

8.2.2.4 Loop Compensation

The TPS65266-1 incorporates a peak current mode control scheme. The error amplifier is a transconductance amplifier with a gain of 290 µS. A typical type II compensation circuit adequately delivers a phase margin between 30° and 90°. C_b adds a high-frequency pole to attenuate high-frequency noise when needed. To calculate the external compensation components, follow these steps.

1. Select switching frequency, f_{SW}, that is appropriate for application depending on L and C sizes, output ripple, EMI, and so forth. Switching frequency between 500 kHz to 1.5 MHz gives best trade-off between performance and cost. To optimize efficiency, lower switching frequency is desired.
2. Set up crossover frequency, f_c, which is typically between 1 / 5 and 1 / 20 of f_{SW}.
3. R_C can be determined by:

$$R_C = \frac{2\pi \times f_C \times V_O \times C_O}{G_{m-EA} \times Vref \times G_{m-PS}} \quad (18)$$

   where
   - G_{m_EA} is the error amplifier gain (290 µS).
   - G_{m_PS} is the power stage voltage to current conversion gain (10 A/V).

4. Calculate C_C by placing a compensation zero at or before the dominant pole $f_p = \frac{1}{C_O \times R_L \times 2\pi}$

$$C_C = \frac{R_L \times C_O}{R_C} \quad (19)$$

5. Optional C_b can be used to cancel the zero from the ESR associated with C_O.

$$C_b = \frac{R_{ESR} \times C_O}{R_C} \quad (20)$$

6. Type III compensation can be implemented with the addition of one capacitor, C_1. This allows for slightly higher loop bandwidths and higher phase margins. If used, C_1 is calculated from Equation 21.

$$C_1 = \frac{1}{2\pi \times R_1 \times f_c} \quad (21)$$

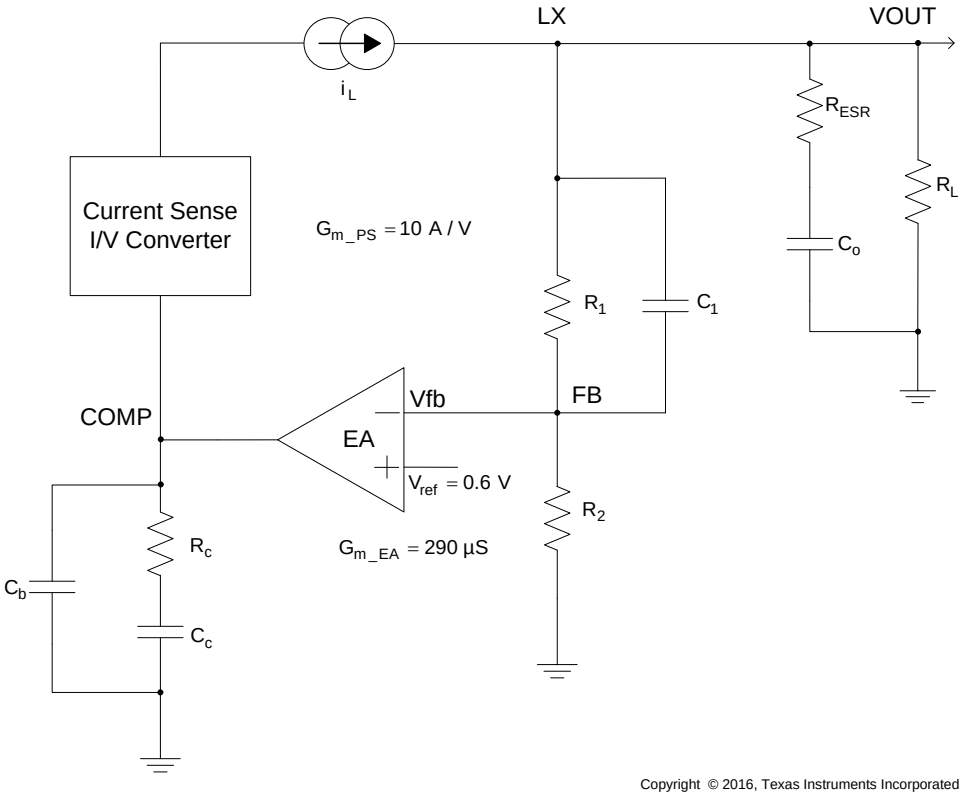


Figure 34. DC/DC Loop Compensation

8.2.3 Application Curves

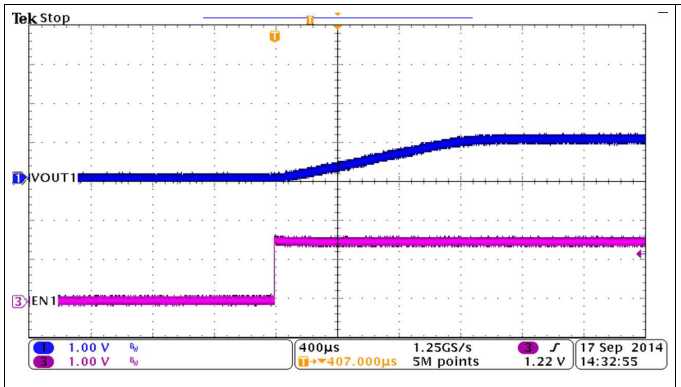


Figure 35. Buck1, Soft-Start, Iout = 3 A

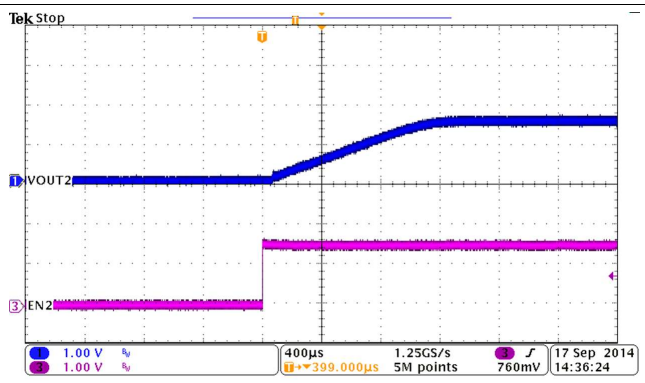


Figure 36. Buck2, Soft-Start, Iout = 2 A

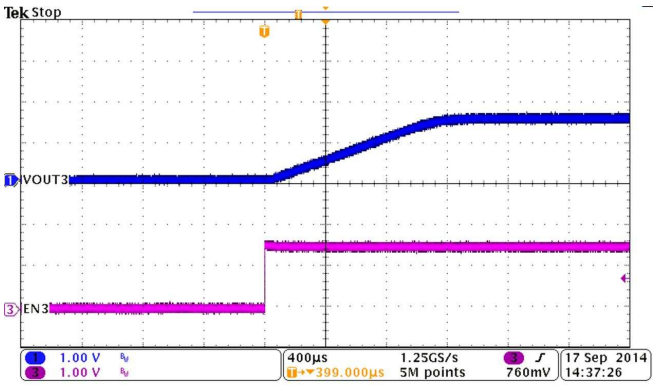


Figure 37. Buck3, Soft-Start, Iout = 2 A

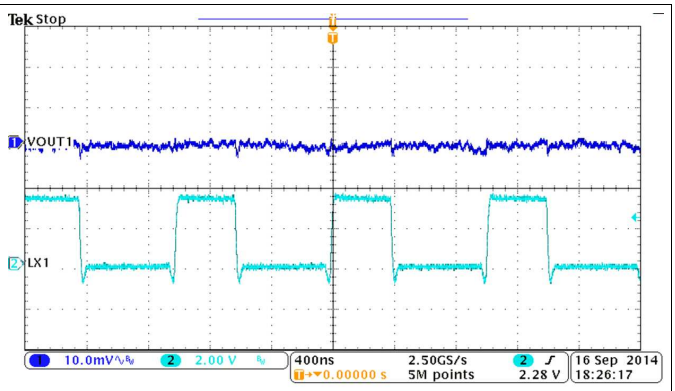


Figure 38. Buck1, Output Voltage Ripple, Iout = 3 A

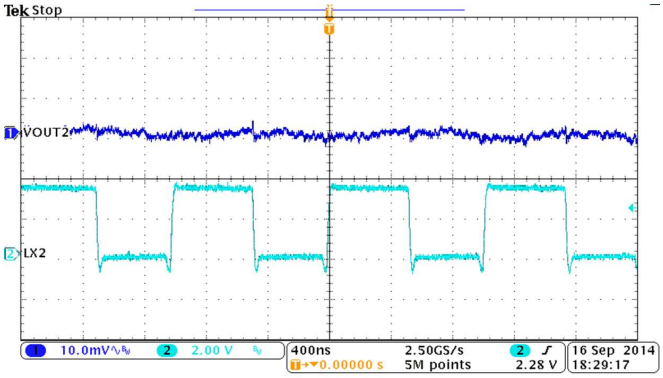


Figure 39. Buck2, Output Voltage Ripple, Iout = 2 A

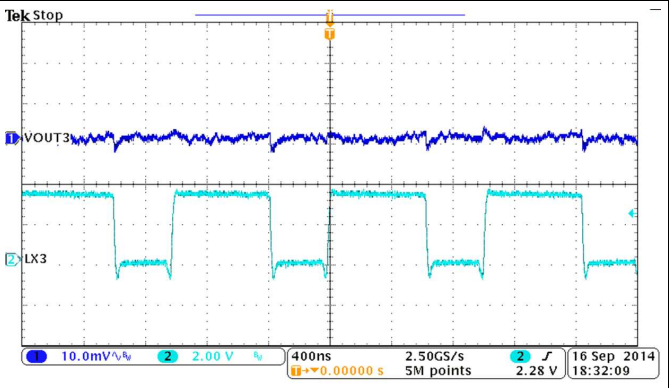
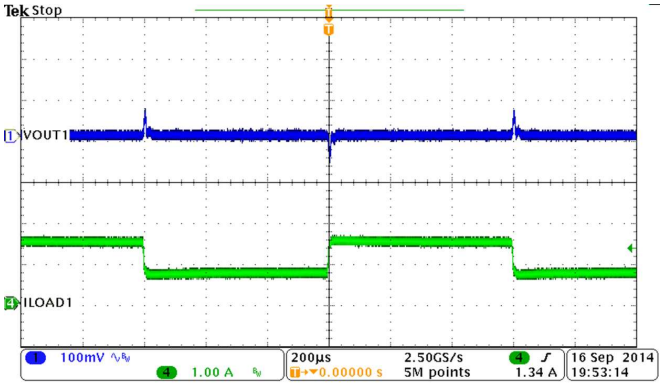
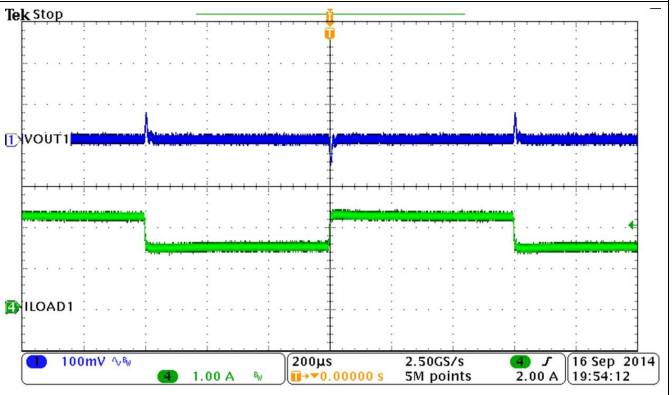


Figure 40. Buck3, Output Voltage Ripple, Iout = 2 A



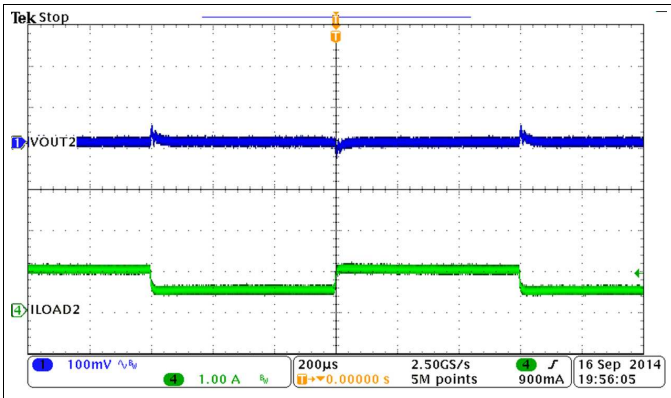
SR = 0.25 A/µs

Figure 41. Buck1, Load Transient, 0.75 to 1.5 A



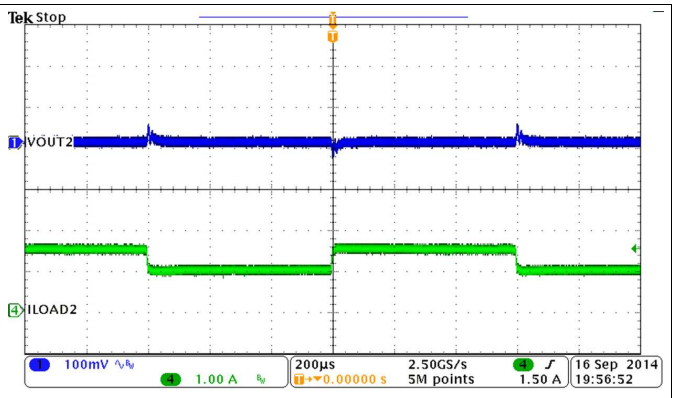
SR = 0.25 A/µs

Figure 42. Buck1, Load Transient, 1.5 to 2.25 A



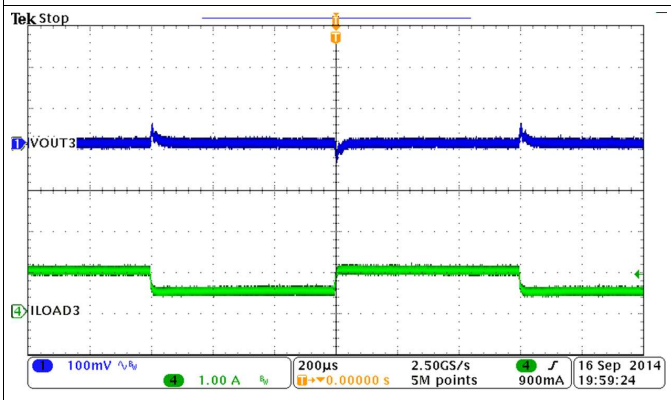
SR = 0.25 A/µs

Figure 43. Buck2, Load Transient, 0.5 to 1.0 A



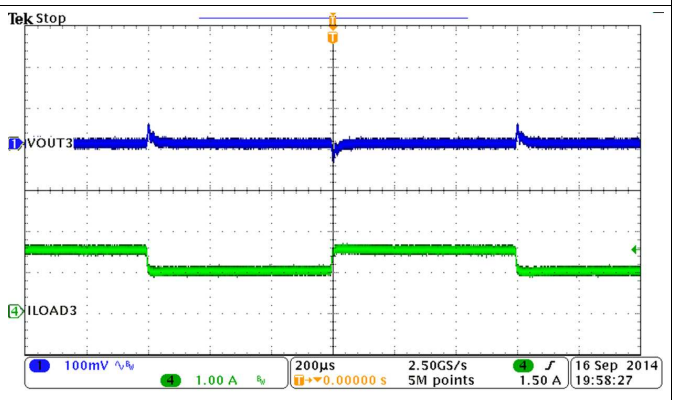
SR = 0.25 A/µs

Figure 44. Buck2, Load Transient, 1.0 to 1.5 A



SR = 0.25 A/µs

Figure 45. Buck3, Load Transient, 0.5 to 1.0 A



SR = 0.25 A/µs

Figure 46. Buck3, Load Transient, 1.0 to 1.5 A

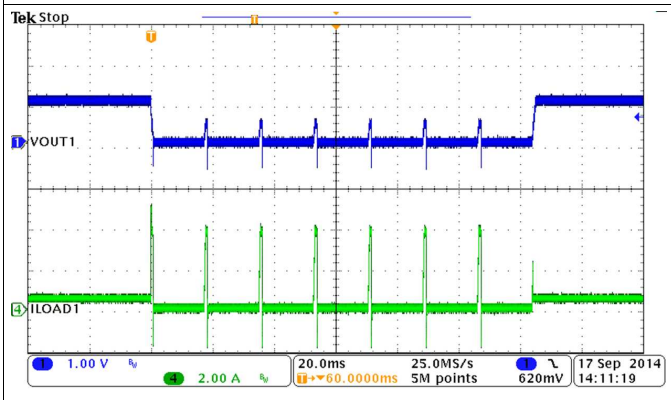


Figure 47. Buck1, Hiccup and Recovery

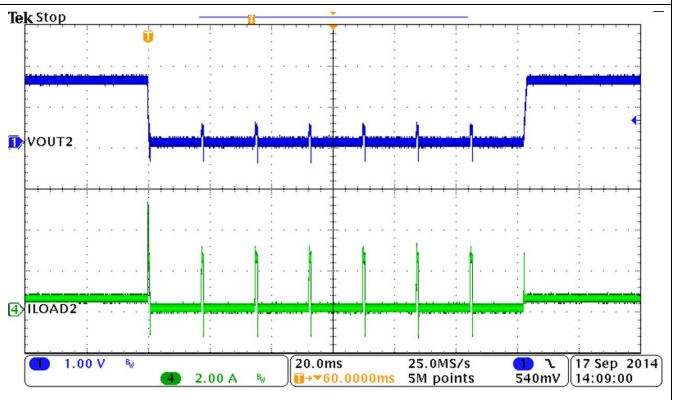


Figure 48. Buck2, Hiccup and Recovery

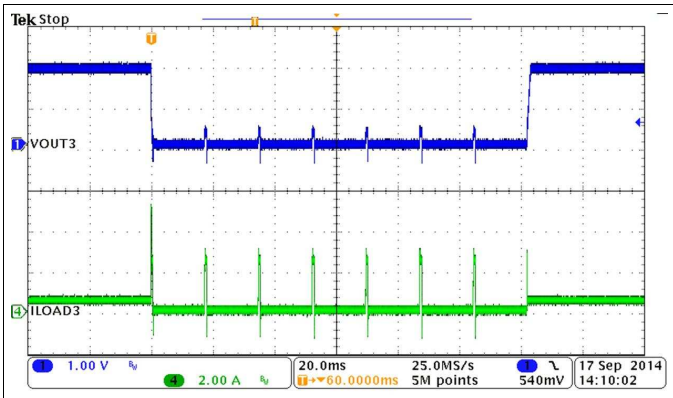


Figure 49. Buck3, Hiccup and Recovery

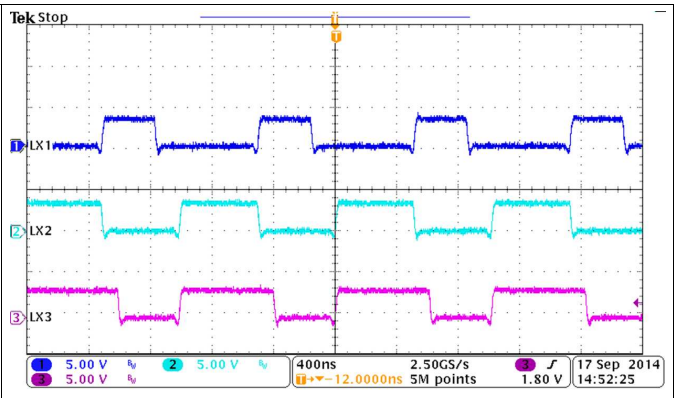


Figure 50. 180° Out of Phase

9 Power Supply Recommendations

The devices are designed to operate from an input voltage supply range between 2.7 and 6 V. This input power supply should be well regulated. If the input supply is located more than a few inches from the TPS65266-1 converter, additional bulk capacitance may be required in addition to the ceramic bypass capacitors. An electrolytic capacitor with a value of 47 μF is a typical choice.

10 Layout

10.1 Layout Guidelines

The TPS65266-1 supports a 2-layer PCB layout, shown in Figure 51.

Layout is a critical portion of good power supply design. See Figure 51 for a PCB layout example. The top contains the main power traces for VIN, VOUT, and LX. The top layer also has connections for the remaining pins of the TPS65266-1 and a large top-side area filled with ground. The top-layer ground area should be connected to the bottom layer ground using vias at the input bypass capacitor, the output filter capacitor, and directly under the TPS65266-1 device to provide a thermal path from the exposed thermal pad land to ground. The bottom layer acts as ground plane connecting analog ground and power ground.

For operation at full-rated load, the top-side ground area together with the bottom-side ground plane must provide an adequate heat dissipating area. Several signals paths conduct fast changing currents or voltages that can interact with stray inductance or parasitic capacitance to generate noise or degrade the power supplies' performance. To help eliminate these problems, the VIN pin should be bypassed to ground with a low-ESR ceramic bypass capacitor with X5R or X7R dielectric. Take care to minimize the loop area formed by the bypass capacitor connections, the VIN pins, and the ground connections. The VIN pin must also be bypassed to ground using a low-ESR ceramic capacitor with X5R or X7R dielectric.

Because the LX connection is the switching node, the output inductor should be located close to the LX pins, and the area of the PCB conductor minimized to prevent excessive capacitive coupling. The output filter capacitor ground should use the same power ground trace as the VIN input bypass capacitor. Try to minimize this conductor length while maintaining adequate width. The small signal components should be grounded to the analog ground path.

The FB and COMP pins are sensitive to noise so the resistors and capacitors should be located as close as possible to the IC and routed with minimal lengths of trace. Place the additional external components approximately as shown in Figure 51.

10.2 Layout Example

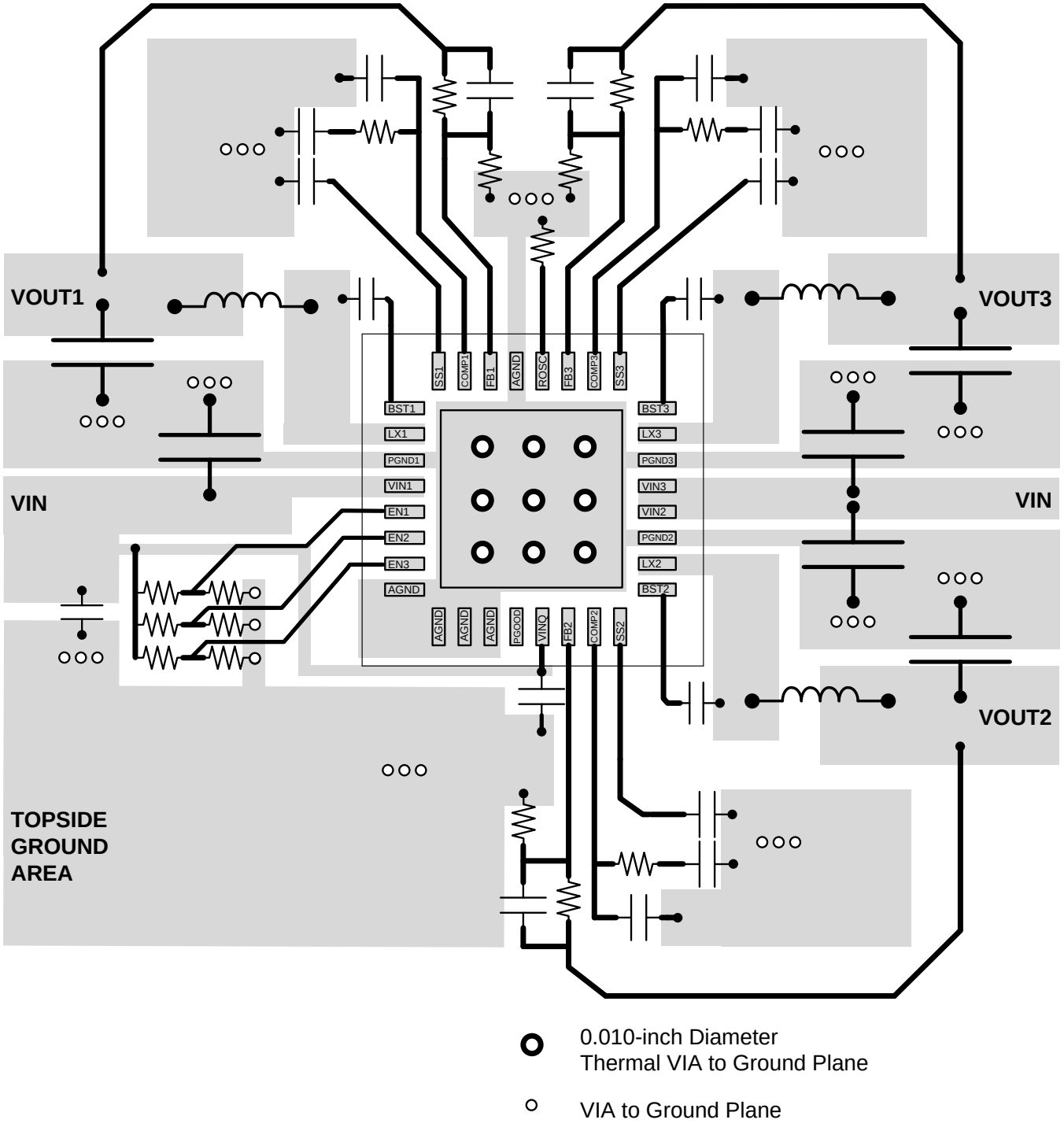


Figure 51. PCB Layout

11 デバイスおよびドキュメントのサポート

11.1 デバイス・サポート

11.1.1 関連製品

型番	説明	名称
TPS65261 TPS65261-1	4.5～18V、トリプル・バック、入力電圧パワー障害インジケータ付き	トリプル・バック3A/2A/2A出力電流、オープン・ドレインRESET信号による入力パワー障害監視、自動パワー・シーケンシング
TPS65262 TPS65262-1	4.5～18V、トリプル・バック、デュアル可変LDO付き	トリプル・バック3A/1A/1A出力電流、自動パワー・シーケンシング デュアルLDO: TPS65262、200mA/100mA TPS65262-1、350mA/150mA
TPS65263	4.5～18V、トリプル・バック、I^2Cインターフェイス付き	トリプル・バック、3A/2A/2A出力電流、I^2C制御の動的電圧スケーリング(DVS)

11.2 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。右上の隅にある「*通知を受け取る*」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。

11.3 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.

11.4 商標

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11.6 Glossary

SLYZ022 — *TI Glossary*.

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

12 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 part number	Status (1)	Material type (2)	Package | Pins	Package qty | Carrier	RoHS (3)	Lead finish/ Ball material (4)	MSL rating/ Peak reflow (5)	Op temp (°C)	Part marking (6)
TPS65266-1RHBR	Active	Production	VQFN (RHB) | 32	3000 | LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 85	TPS 65266-1
TPS65266-1RHBR.A	Active	Production	VQFN (RHB) | 32	3000 | LARGE T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 85	TPS 65266-1
TPS65266-1RHBT	Active	Production	VQFN (RHB) | 32	250 | SMALL T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 85	TPS 65266-1
TPS65266-1RHBT.A	Active	Production	VQFN (RHB) | 32	250 | SMALL T&R	Yes	NIPDAU	Level-1-260C-UNLIM	-40 to 85	TPS 65266-1

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

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

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

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

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

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

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

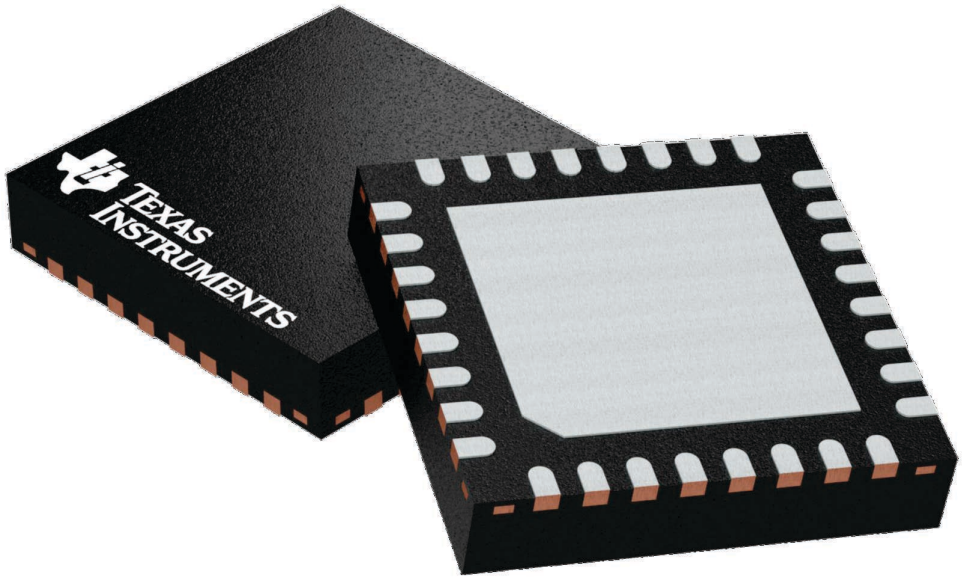
GENERIC PACKAGE VIEW

RHB 32

VQFN - 1 mm max height

5 x 5, 0.5 mm pitch

PLASTIC QUAD FLATPACK - NO LEAD



Images above are just a representation of the package family, actual package may vary.
Refer to the product data sheet for package details.

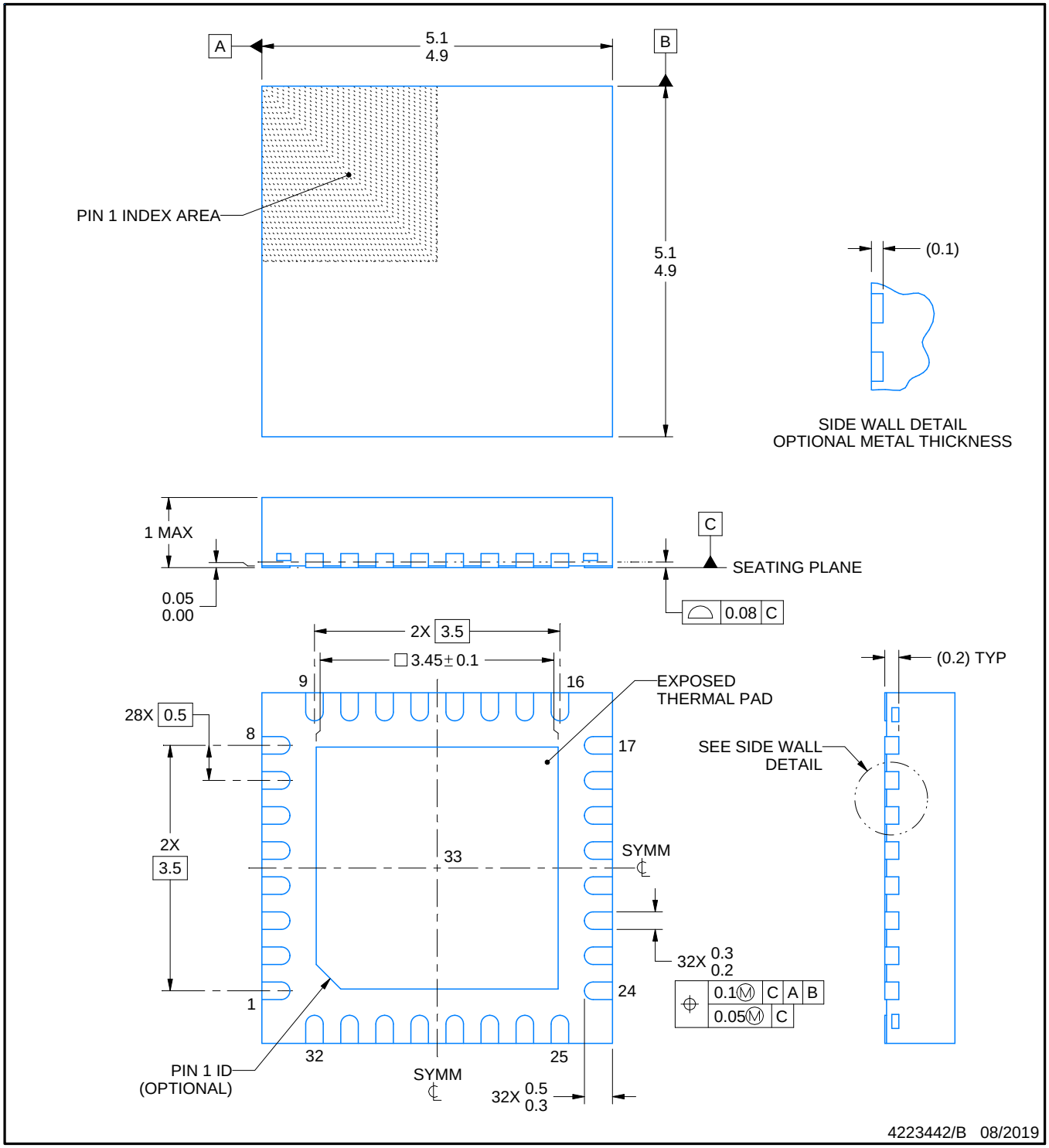
4224745/A

RHB0032E

PACKAGE OUTLINE

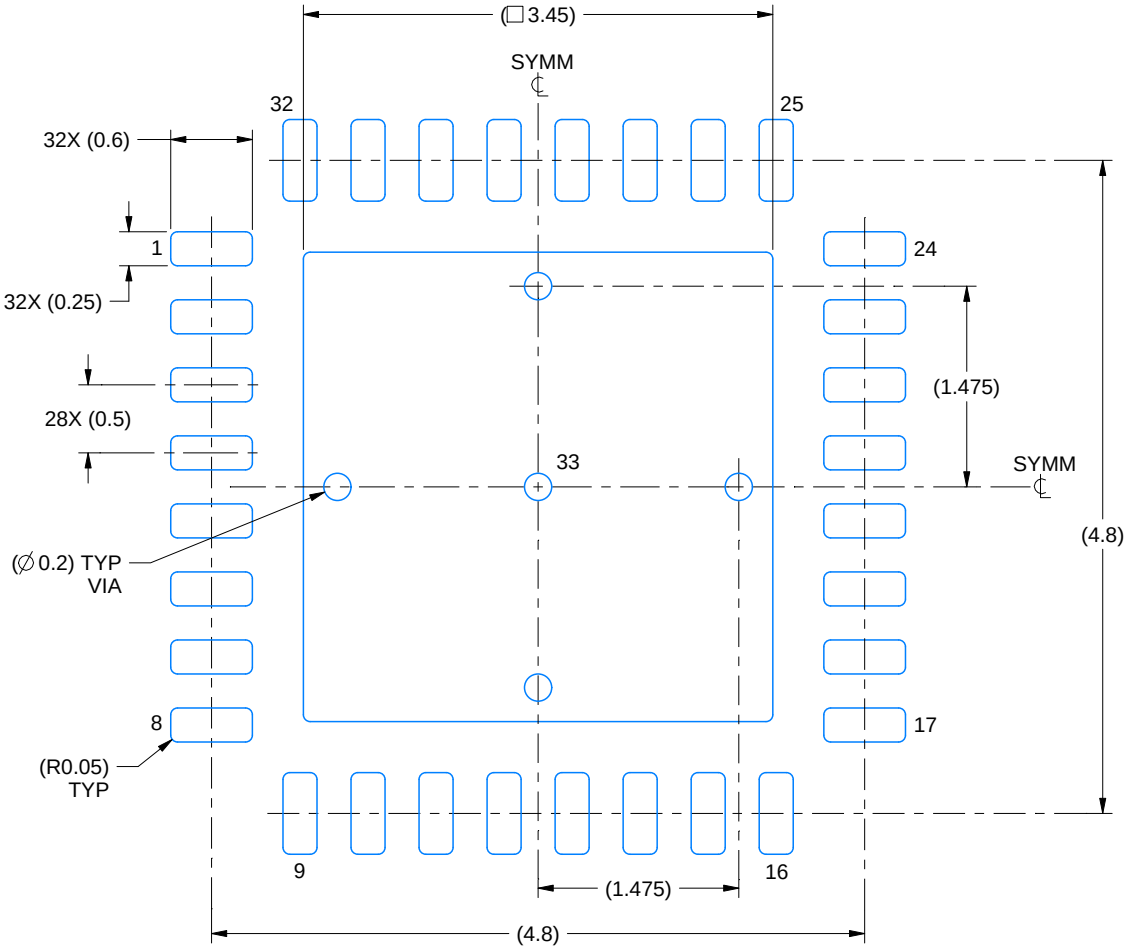
VQFN - 1 mm max height

PLASTIC QUAD FLATPACK - NO LEAD

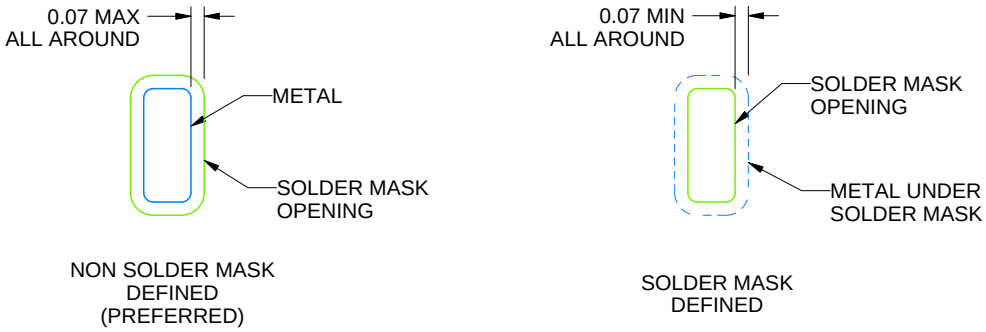


NOTES:

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 thermal and mechanical performance.



LAND PATTERN EXAMPLE
SCALE:18X



SOLDER MASK DETAILS

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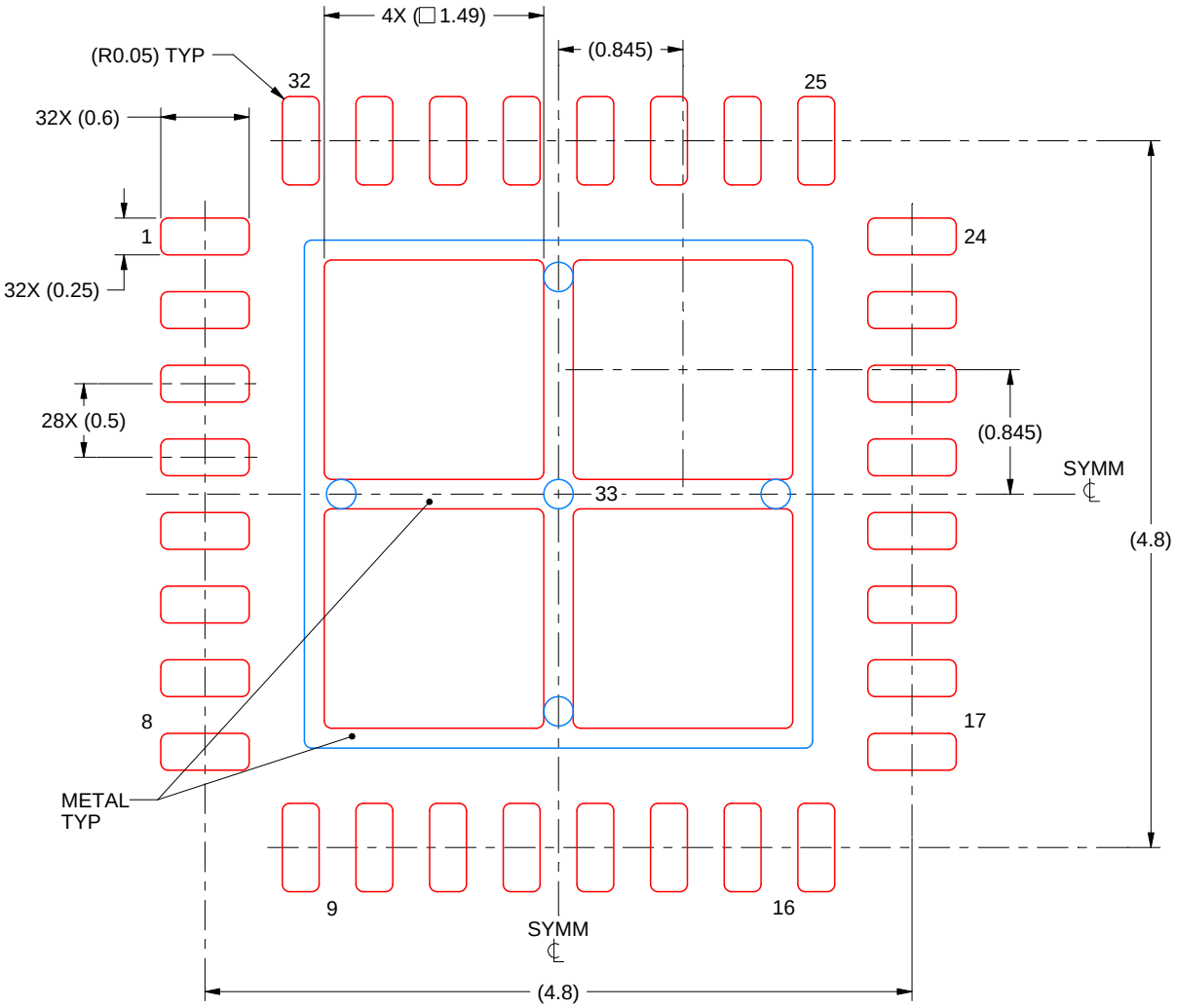
NOTES: (continued)

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.



SOLDER PASTE EXAMPLE
BASED ON 0.125 mm THICK STENCIL

EXPOSED PAD 33:
75% PRINTED SOLDER COVERAGE BY AREA UNDER PACKAGE
SCALE:20X

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NOTES: (continued)

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

重要なお知らせと免責事項

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