TPS40170-Q1 車載用 4.5V～60V、広い入力電圧範囲、同期整流 PWM 降圧コントローラ

1 特長

- 新しい類似製品が利用可能:
  - LM5145-Q1 広い入力電圧範囲とデューティ・サイクル範囲の 75V 同期整流降圧コントローラ
  - LM5146-Q1 広いデューティ・サイクル範囲の 100V 同期整流降圧 DC/DC コントローラ
- 車載アプリケーション向け認定済み
- 下記結果で AEC-Q100 認定済み
  - デバイス温度グレード 1:-40℃～125℃の動作時周囲温度範囲
  - デバイス HBM ESD 分類レベル 1C
  - デバイス CDM ESD 分類レベル C4B
- 広い入力電圧範囲:4.5V～60V
- 精度 1% の 600mV 基準電圧
- プログラム可能な UVLO とヒステリシス
- 電圧フィード・フォワードによる電圧モード制御
- 100kHz～600kHz のプログラマブル周波数
- プライマリとセカンダリのオプションによる双方向周波数同期
- ローサイド FET センシング過電流保護と、内蔵の熱補償によるハイサイド FET センシング短絡保護
- プログラム可能な閉ループ・ソフト・スタート
- プリバイアス出力をサポート
- ヒステリシス付き 165℃でのサーマル・シャットダウン
- 電圧トラッキング
- パワー・グッド
- 1μA の低電流シャットダウン機能による ENABLE
- 8V および 3.3V の LDO 出力
- ブートストラップ・ダイオードを内蔵
- 20 ピン、4.5mm × 3.5mm VQFN (RGY) パッケージ

2 アプリケーション

- 先進運転支援システム (ADAS)
- 車載用インフォテインメントおよびクラスタ

3 概要

TPS40170-Q1 デバイスは、4.5V～60V の入力電圧で動作するフル機能の同期整流 PWM 降圧コントローラで、電力密度が高く信頼性の高い DC/DC コンバータ・アプリケーションに最適化されています。このコントローラは、入力電圧フィードフォワード補償による電圧モード制御を実装し、入力電圧の変化に対する瞬時の応答を可能にします。スイッチング周波数は 100kHz～600kHz の範囲で設定可能です。

TPS40170-Q1 デバイスは、プログラム可能な UVLO、ローサイド FET の検出によるプログラム可能な過電流保護 (OCP)、ハイサイド FET の検出による選択可能な短絡保護 (SCP)、サーマル・シャットダウンなど、一連のシステム保護および監視機能を備えています。ENABLE ピンにより、低電流 (代表値 1μA) モードでシステム・シャットダウンが可能です。このコントローラは、プリバイアス出力をサポートし、オープン・ドレインの PGOOD 信号を供給するほか、閉ループのソフトスタート、出力電圧トラッキング、適合型デッドタイム制御を備えています。

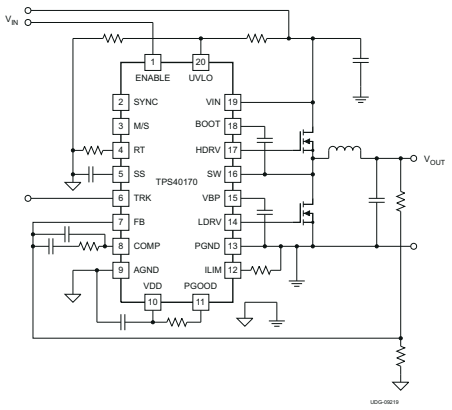
TPS40170-Q1 デバイスは、精度 1% 内の正確な出力電圧レギュレーションを実現します。さらに、このコントローラは 1 つのコントローラがプライマリ・コントローラとして動作し、他のダウンストリーム・コントローラが 1 次側に同相、または 180° 位相不一致に同期しているセカンダリ・コントローラとして動作している新しい双方向同期方式を実装しています。セカンダリ・コントローラは、フリーランニング・スイッチング周波数の ±30% 以内で外部クロックと同期できます。

新製品 (LM5145-Q1 および LM5146-Q1) には、BOM コストの削減、効率の向上、ソリューション・サイズの小型化など、多くの特長があります。

製品情報

部品番号	パッケージ (1)	パッケージ・サイズ (2)
TPS40170-Q1	RGY (VQFN、20)	4.50mm × 3.50mm

(1) 供給されているすべてのパッケージについては、セクション 10 を参照してください。
(2) パッケージ・サイズ (長さ×幅) は公称値であり、該当する場合はピンも含まれます。



概略回路図

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4 Pin Configuration and Functions

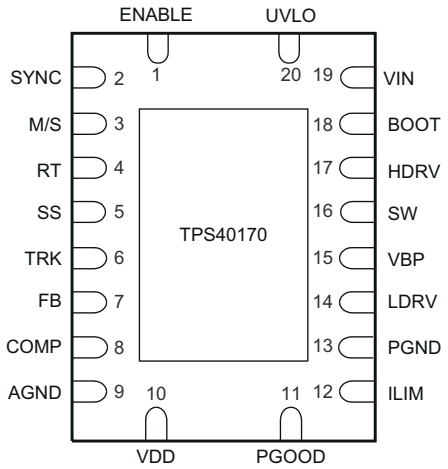


図 4-1. RGY PACKAGE QFN-20 (Top View)

表 4-1. Pin Functions

PIN		TYPE (1)	DESCRIPTION
NAME	NO.		
AGND	9	—	Analog signal ground. This pin must be electrically connected to power ground PGND externally.
BOOT	18	O	Boot-capacitor node for high-side FET gate driver. The boot capacitor is connected from this pin to SW.
COMP	8	O	Output of the internal error amplifier. The feedback loop compensation network is connected from this pin to the FB pin.
ENABLE	1	I	This pin must be high for the device to be enabled. If this pin is pulled low, the device is put in a low-power-consumption shutdown mode.
FB	7	I	Negative input to the error amplifier. The output voltage is fed back to this pin through a resistor-divider network.
HDRV	17	O	Gate-driver output for the high-side FET.
ILIM	12	I	A resistor from this pin to PGND sets the overcurrent limit. This pin provides source current used for the overcurrent-protection threshold setting.
LDRV	14	O	Gate driver output for the low-side FET. Also, a resistor from this pin to PGND sets the multiplier factor to determine the short-circuit current limit. If no resistor is present, the multiplier defaults to 7 times the ILIM pin voltage.
M/S	3	I	Primary- or secondary-mode selector pin for frequency synchronization. This pin must be tied to VIN for primary mode. In the secondary mode, this pin must be tied to AGND or left floating. If the pin is tied to AGND, the device synchronizes with a 180° phase shift. If the pin is left floating, the device synchronizes with a 0° phase shift.
PGND	13	—	Power ground. This pin must externally connect to the AGND at a single point.
PGOOD	11	O	Power-good indicator. This pin is an open-drain output pin, and TI recommends a 10-kΩ pullup resistor to be connected between this pin and VDD.
RT	4	I	A resistor from this pin to AGND sets the oscillator frequency. Even if operating in secondary mode, it is required to have a resistor at this pin to set the free-running switching frequency.
SS	5	I	Soft-start. A capacitor must be connected from this pin to AGND. The capacitor value sets the soft-start time.
SW	16	I	This pin must connect to the switching node of the synchronous buck converter. The high-side and low-side FET current sensing are also done from this node.
SYNC	2	I/O	Synchronization. This is a bidirectional pin used for frequency synchronization. In the primary mode, it is the SYNC output pin. In the secondary mode, it is a SYNC input pin. If unused, this pin can be left open.
TRK	6	I	Tracking. External signal at this pin is used for output voltage tracking. This pin goes directly to the internal error amplifier as a positive reference. The lesser of the voltages between V_{TRK} and the internal 600-mV reference sets the output voltage. If not used, this pin must be pulled up to VDD.

表 4-1. Pin Functions (続き)

PIN		TYPE (1)	DESCRIPTION
NAME	NO.		
UVLO	20	I	Undervoltage lockout. A resistor divider on this pin from VIN to AGND can be used to set the UVLO threshold.
VBP	15	O	8-V regulated output for gate driver. A ceramic capacitor with a value between 1 µF and 10 µF must be connected from this pin to PGND
VDD	10	O	3.3-V regulated output. A ceramic bypass capacitor with a value between 0.1 µF and 1 µF must be connected between this pin and the AGND pin and placed closely to this pin.
VIN	19	I	Input voltage for the controller, which is also the input voltage for the dc-dc converter. A 1-µF bypass capacitor from this pin to AGND must be added and placed closed to VIN.

(1) I = input, O = output

5 Specifications

5.1 Absolute Maximum Ratings

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

		MIN	MAX	UNIT
Input voltage	VIN	–0.3	62	V
	M/S	–0.3	VIN	
	UVLO	–0.3	16	
	SW	–5	VIN	
	BOOT		V_{SW} + 8.8	
Output voltage	HDRV	V_{SW}	BOOT	V
	BOOT-SW, HDRV-SW (differential from BOOT or HDRV to SW)	–0.3	8.8	
	VBP, LDRV, COMP, RT, ENABLE, PGOOD, SYNC	–0.3	8.8	
	VDD, FB, TRK, SS, ILIM	–0.3	3.6	
Grounding	AGND-PGND, PGND-AGND	200	200	mV
	PowerPAD to AGND (must be electrically connected external to device)		0	
Ambient temperature	T_A	–40	125	°C
Storage temperature	T_{stg}	–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.

5.2 ESD Ratings

				VALUE	UNIT
$V_{(ESD)}$	Electrostatic discharge	Human body model (HBM), per AEC Q100-002(1)		±1500	V
		Charged device model (CDM), per AEC Q100-011	Corner pins (SYNC, VDD, PGOOD, ILIM, VIN, UVLO, ENABLE)	±750	
			Other pins	±500	

(1) AEC Q100-002 indicates HBM stressing is done in accordance with the ANSI/ESDA/JEDEC JS-001 specification.

5.3 Recommended Operating Conditions

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

		MIN	NOM	MAX	UNIT
V_{IN}	Input voltage	4.5		60	V

5.4 Thermal Information

THERMAL METRIC(1)		TPS40170-Q1 RGY 20 PINS	UNIT
$R_{\theta JA}$	Junction-to-ambient thermal resistance	35.4	°C/W
$R_{\theta JC(top)}$	Junction-to-case(top) thermal resistance	38.1	
$R_{\theta JB}$	Junction-to-board thermal resistance	10.8	
ψ_{JT}	Junction-to-top characterization parameter	0.5	
ψ_{JB}	Junction-to-board characterization parameter	10.9	
$R_{\theta JC(bot)}$	Junction-to-case(bottom) thermal resistance	4.3	

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

5.5 Electrical Characteristics

These specifications apply for –40°C ≤ T_A ≤ +125°C, V_{VIN} = 12 V, unless otherwise noted.

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
INPUT SUPPLY						
V_{VIN}	Input voltage range		4.5		60	V
I_{SD}	Shutdown current	V_{ENABLE} < 100 mV		1	2.5	µA
I_{QQ}	Operating current, drivers not switching	V_{ENABLE} ≥ 2 V, f_{SW} = 300 kHz			4.5	mA
ENABLE						
V_{DIS}	ENABLE pin voltage to disable the device				100	mV
V_{EN}	ENABLE pin voltage to enable the device		600			mV
I_{ENABLE}	ENABLE pin source current				300	nA
8-V AND 3.3-V REGULATORS						
V_{VBP}	8-V regulator output voltage	V_{ENABLE} ≥ 2 V, 8.2 V < V_{VIN} ≤ 60 V, 0 mA < I_{IN} < 20 mA	7.8	8.0	8.3	V
V_{DO}	8-V regulator dropout voltage, $V_{VIN-VVBP}$	4.5 < V_{VIN} ≤ 8.2 V, V_{EN} ≥ 2 V, I_{IN} = 10 mA		110	200	mV
V_{VDD}	3.3-V regulator output voltage	V_{ENABLE} ≥ 2 V, 4.5 V < V_{VIN} ≤ 60 V, 0 mA < I_{IN} < 5 mA	3.22	3.3	3.42	V
FIXED AND PROGRAMMABLE UVLO						
V_{UVLO}	Programmable UVLO ON voltage (at UVLO pin)	V_{ENABLE} ≥ 2 V	878	900	919	mV
I_{UVLO}	Hysteresis current out of UVLO pin	V_{ENABLE} ≥ 2 V , UVLO pin > V_{UVLO}	4.06	5	6.2	µA
VBP_{ON}	VBP turnon voltage	V_{ENABLE} ≥ 2 V, UVLO pin > V_{UVLO}	3.85		4.4	V
VBP_{OFF}	VBP turnoff voltage		3.6		4.05	
VBP_{HYS}	VBP UVLO Hysteresis voltage		180		400	mV
REFERENCE						
V_{REF}	Reference voltage (+ input of the error amplifier)	T_J = 25°C, 4.5 V < V_{VIN} ≤ 60 V	594	600	606	mV
		–40°C ≤ T_J ≤ 125°C, 4.5 V < V_{VIN} ≤ 60 V	591	600	609	
OSCILLATOR						

5.5 Electrical Characteristics (続き)

These specifications apply for –40°C ≤ T_A ≤ +125°C, V_{VIN} = 12 V, unless otherwise noted.

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
f_{SW}	Switching frequency	Range (typical)	100		600	kHz
		R_{RT} = 100 kΩ, 4.5 V < V_{VIN} ≤ 60 V	90	100	110	
		R_{RT} = 31.6 kΩ, 4.5 V < V_{VIN} ≤ 60 V	270	300	330	
		R_{RT} = 14.3 kΩ, 4.5 V < V_{VIN} ≤ 60 V	540	600	660	
V_{VALLEY}	Valley voltage		0.7	1	1.2	V
K_{PWM} (1)	PWM gain (V_{VIN} / V_{RAMP})	4.5 V < V_{VIN} ≤ 60 V	14	15	16	V/V

5.5 Electrical Characteristics (続き)

These specifications apply for –40°C ≤ T_A ≤ +125°C, V_{VIN} = 12 V, unless otherwise noted.

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
PWM AND DUTY CYCLE						
$t_{ON(min)}$ (1)	Minimum controlled pulse	V_{VIN} = 4.5 V, f_{SW} = 300 kHz		100	150	ns
		V_{VIN} = 12 V, f_{SW} = 300 kHz		75	100	
		V_{VIN} = 60 V, f_{SW} = 300 kHz		50	80	
$t_{OFF(max)}$ (1)	Minimum OFF time	V_{VIN} = 12V, f_{SW} = 300 kHz		170	250	ns
D_{MAX} (1)	Maximum duty cycle	f_{SW} = 100 kHz, 4.5 V < V_{VIN} ≤ 60 V	95%			
		f_{SW} = 300 kHz, 4.5 V < V_{VIN} ≤ 60 V	91%			
		f_{SW} = 600 kHz, 4.5 V < V_{VIN} ≤ 60 V	82%			
ERROR AMPLIFIER						
GBWP(1)	Gain bandwidth product		7	10	13	MHz
A_{OL} (1)	Open-loop gain		80	90	95	dB
I_{IB}	Input bias current				100	nA
I_{EAOP}	Output source current	V_{VFB} = 0 V	2			mA
I_{EAOM}	Output sink current	V_{VFB} = 1 V	2			mA
PROGRAMMABLE SOFT START						
$I_{SS(source,start)}$	Soft-start source current	V_{SS} < 0.5 V, V_{SS} = 0.25 V	42	52	62	µA
$I_{SS(source,normal)}$	Soft-start source current	V_{SS} > 0.5 V, V_{SS} = 1.5 V	9.3	11.6	13.9	µA
$I_{SS(sink)}$	Soft-start sink current	V_{SS} = 1.5 V	0.77	1.05	1.33	µA
$V_{SS(fltH)}$	SS pin HIGH voltage during fault (OC or thermal) reset timing		2.38	2.5	2.61	V
$V_{SS(fltL)}$	SS pin LOW voltage during fault (OC or thermal) reset timing		235	300	375	mV
$V_{SS(steady\_state)}$	SS pin voltage during steady-state		3.25	3.3	3.5	V
$V_{SS(offst)}$	Initial offset voltage from SS pin to error amplifier input		525	650	775	mV
TRACKING						
$V_{TRK(ctrl)}$ (1)	Range of TRK which overrides V_{REF}	4.5 V < V_{IN} ≤ 60 V	0		600	mV
SYNCHRONIZATION (PRIMARY/SECONDARY)						
V_{MSTR}	M/S pin voltage in primary mode		3.9		VIN	V
$V_{SLV(0)}$	M/S pin voltage in secondary 0° mode		1.25		1.75	V
$V_{SLV(180)}$	M/S pin voltage in secondary 180° mode		0		0.75	V
$I_{SYNC(in)}$	SYNC pin pulldown current	M/S configured as secondary- 0° or secondary-180°	8	11	14	µA
$V_{SYNC(in\_high)}$	SYNC pin input high-voltage level		2			V
$V_{SYNC(in\_low)}$	SYNC pin input low-voltage level				0.8	V
$t_{SYNC(high\_min)}$	Minimum SYNC high pulse duration		40	50		ns
$t_{SYNC(low\_min)}$	Minimum SYNC low pulse duration		40	50		ns

5.5 Electrical Characteristics (続き)

These specifications apply for –40°C ≤ T_A ≤ +125°C, V_{VIN} = 12 V, unless otherwise noted.

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
GATE DRIVERS						
R_{HDHI}	High-side driver pullup resistance	C_{LOAD} = 2.2 nF, I_{DRV} = 300 mA, T_A = –40°C to 125°C	1.37	2.64	4	Ω
R_{HDLO}	High-side driver pulldown resistance		1.08	2.4	4	Ω
R_{LDHI}	Low-side driver pullup resistance		1.37	2.4	4	Ω
R_{LDLO}	Low-side driver pulldown resistance		0.44	1.1	1.7	Ω
$t_{NON-OVERLAP1}$	Time delay between HDRV fall and LDRV rise	C_{LOAD} = 2.2 nF, V_{HDRV} = 2 V, V_{LDRV} = 2 V		50		ns
$t_{NON-OVERLAP2}$	Time delay between HDRV rise and LDRV fall			60		ns
OVERCURRENT PROTECTION (LOW-SIDE MOSFET SENSING)						
I_{ILIM}	ILIM pin source current	4.5 V < V_{IN} < 60 V, T_A = 25°C	9	9.75	10.45	µA
		4.5 V < V_{IN} < 60 V, T_A = –40°C to 125°C	7		12	µA
$I_{ILIM,(ss)}$	ILIM pin source current during soft-start	4.5 V < V_{IN} < 60 V, T_A = 25°C		15		µA
		4.5 V < V_{IN} < 60 V, T_A = –40°C to 125°C	7		12	µA
$I_{ILIM, Tc}$ (1)	Temperature coefficient of ILIM current	4.5 V < V_{IN} < 60 V		1400		ppm
V_{ILIM} (1)	ILIM pin voltage operating range	4.5 V < V_{IN} < 60 V	50		300	mV
OCP_{TH}	Overcurrent protection threshold (voltage across low-side FET for detecting overcurrent)	R_{ILIM} = 10 kΩ, I_{ILIM} = 10 µA (V_{ILIM} = 100 mV)	–110	–100	–84	mV
SHORT CIRCUIT PROTECTION HIGH-SIDE MOSFET SENSING)						
$V_{LDRV(max)}$	LDRV pin maximum voltage during calibration	R_{LDRV} = open		300	360	mV
A_{OC3}	Multiplier factor to set the SCP based on OCP level setting at the ILIM pin	R_{LDRV} = 10 kΩ	2.75	3.2	3.6	V/V
A_{OC7}		R_{LDRV} = open	6.4	7.25	7.91	V/V
A_{OC15}		R_{LDRV} = 20 kΩ	13.9	16.4	18	V/V
THERMAL SHUTDOWN						
$T_{SD,set}$ (1)	Thermal shutdown set threshold	4.5 V < V_{VIN} < 60 V	155	165	175	°C
$T_{SD,reset}$ (1)	Thermal shutdown reset threshold		125	135	145	°C
T_{hyst} (1)	Thermal shutdown hysteresis			30		°C
POWER GOOD						
V_{OV}	FB pin voltage upper limit for power good	4.5 V < V_{VIN} < 60 V	627	647	670	mV
V_{UV}	FB pin voltage lower limit for power good		527	552	570	mV
$V_{PG,HYST}$	Power good hysteresis voltage at FB pin		8.5	20	32	mV
$V_{PG(out)}$	PGOOD pin voltage when FB pin voltage > V_{OV} or < V_{UV}, I_{PGD} = 2 mA				100	mV
$V_{PG(np)}$	PGOOD pin voltage when device power is removed	V_{VIN} is open, 10-kΩ to V_{EXT} = 5 V		1	1.5	V
BOOT DIODE						
V_{DFWD}	Bootstrap diode forward voltage	I = 20 mA	0.5	0.7	0.9	V

5.5 Electrical Characteristics (続き)

These specifications apply for –40°C ≤ T_A ≤ +125°C, V_{VIN} = 12 V, unless otherwise noted.

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
$R_{BOOT-SW}$	Discharge resistor from BOOT to SW			1		MΩ

(1) Not production tested.

5.6 Typical Characteristics

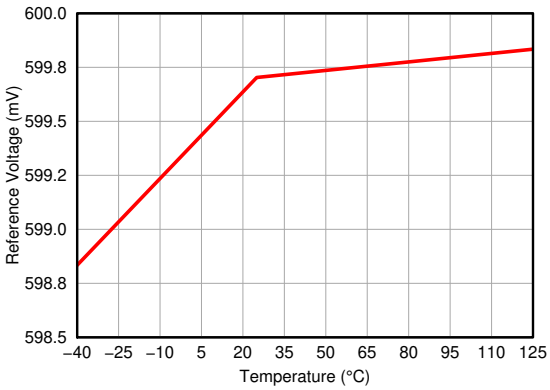


図 5-1. Reference Voltage vs Junction Temperature

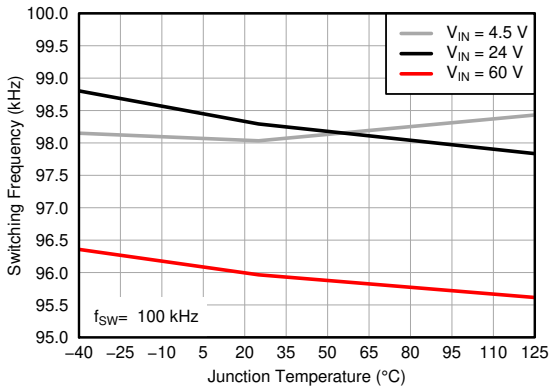


図 5-2. Switching Frequency vs Junction Temperature

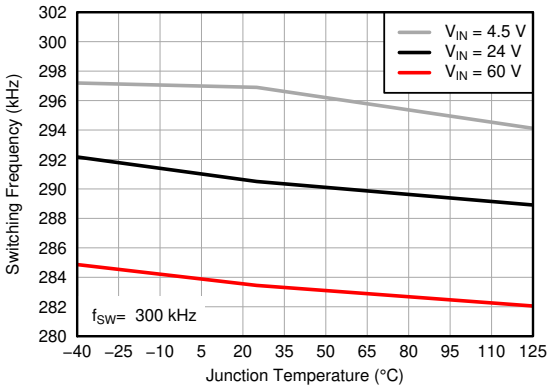


図 5-3. Switching Frequency vs Junction Temperature

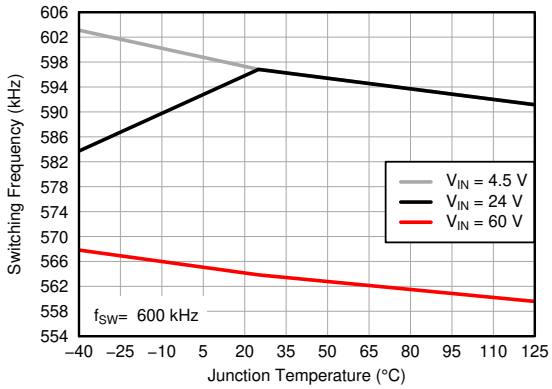


図 5-4. Switching Frequency vs Junction Temperature

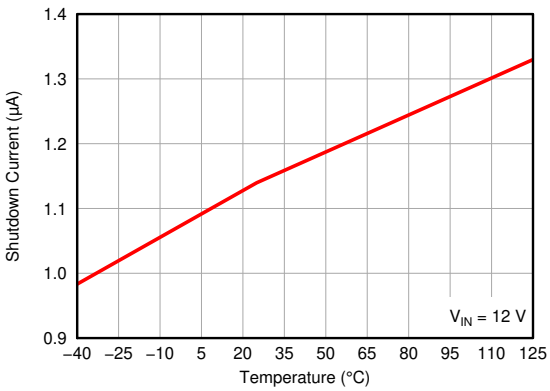


図 5-5. Shutdown Current vs Junction Temperature

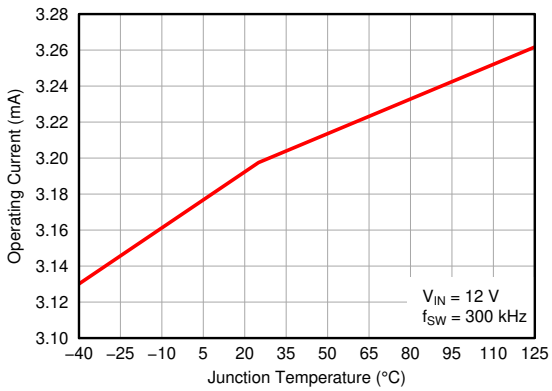


図 5-6. Operating Current vs Junction Temperature

5.6 Typical Characteristics (continued)

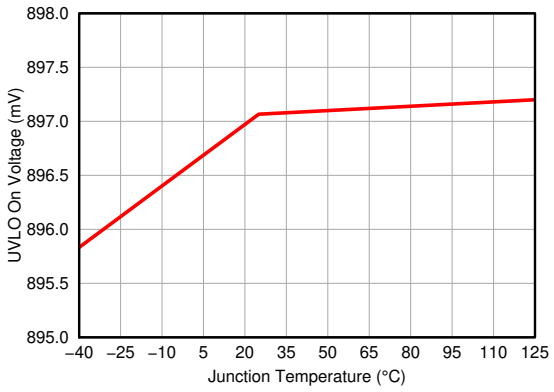


図 5-7. UVLO On Voltage vs Junction Temperature

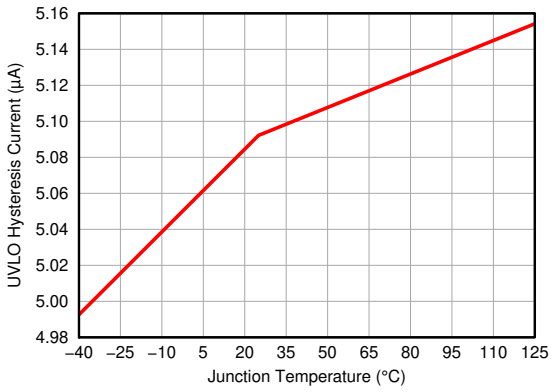


図 5-8. UVLO Hysteresis Current vs Junction Temperature

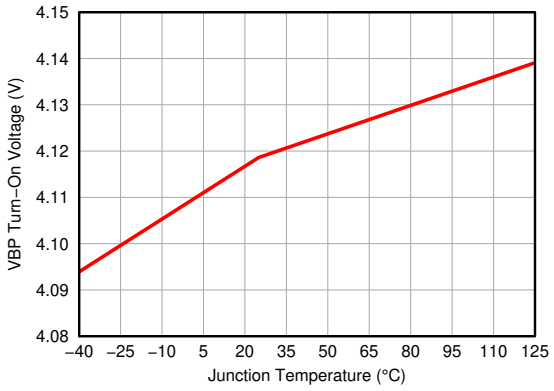


図 5-9. VBP Turnon Voltage vs Junction Temperature

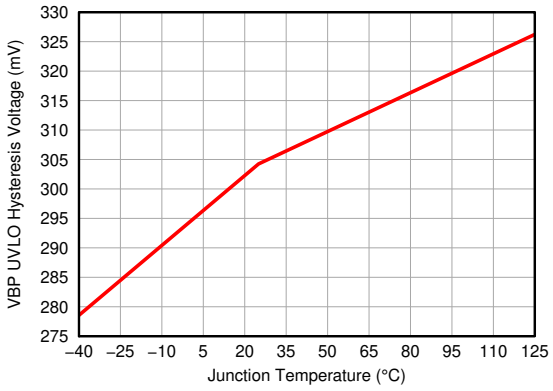


図 5-10. VBP UVLO Hysteresis Voltage

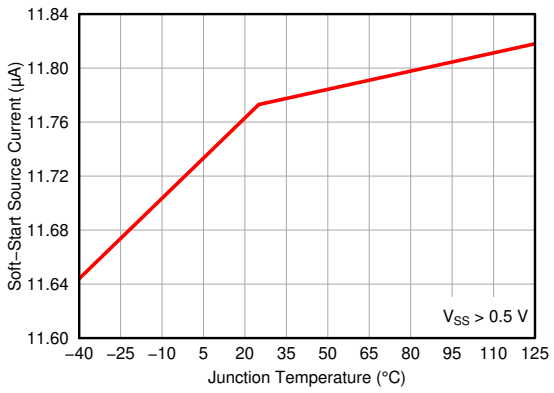


図 5-11. Soft-Start Source Current vs Junction Temperature

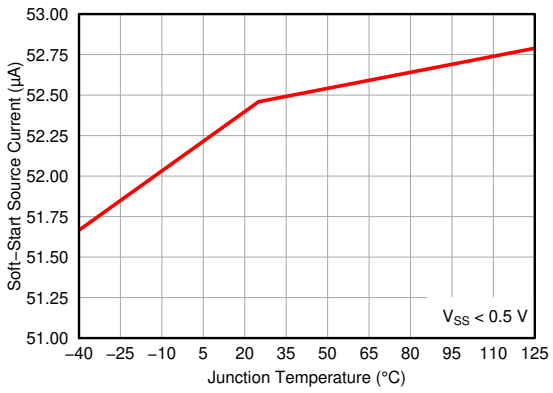


図 5-12. Soft-Start Source Current vs Junction Temperature

5.6 Typical Characteristics (continued)

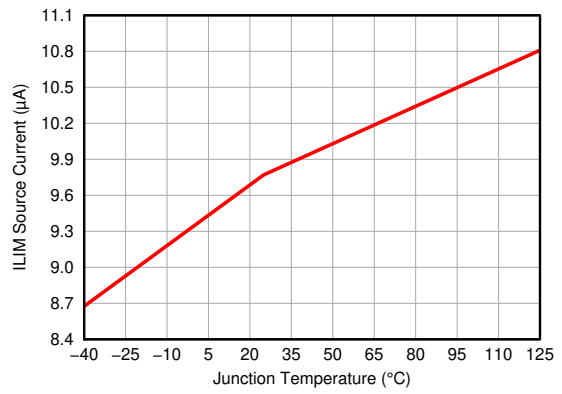


図 5-13. ILIM Source Current vs Junction Temperature

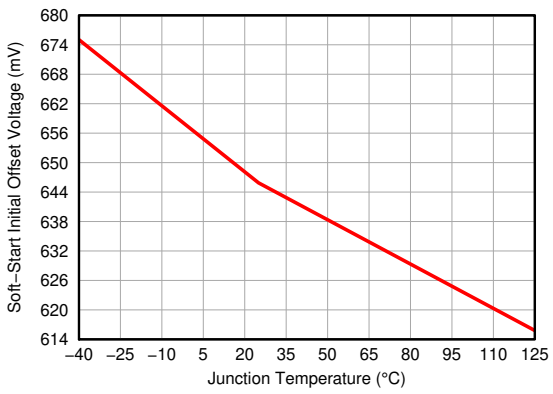


図 5-14. Soft-Start Initial Offset Voltage vs Junction Temperature

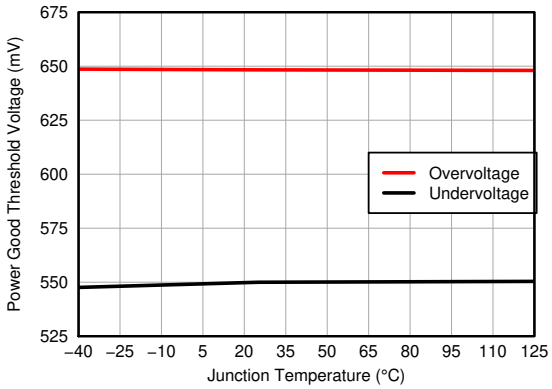


図 5-15. V_{OV}/V_{UV} Power Good Threshold Voltage

6 Detailed Description

6.1 Overview

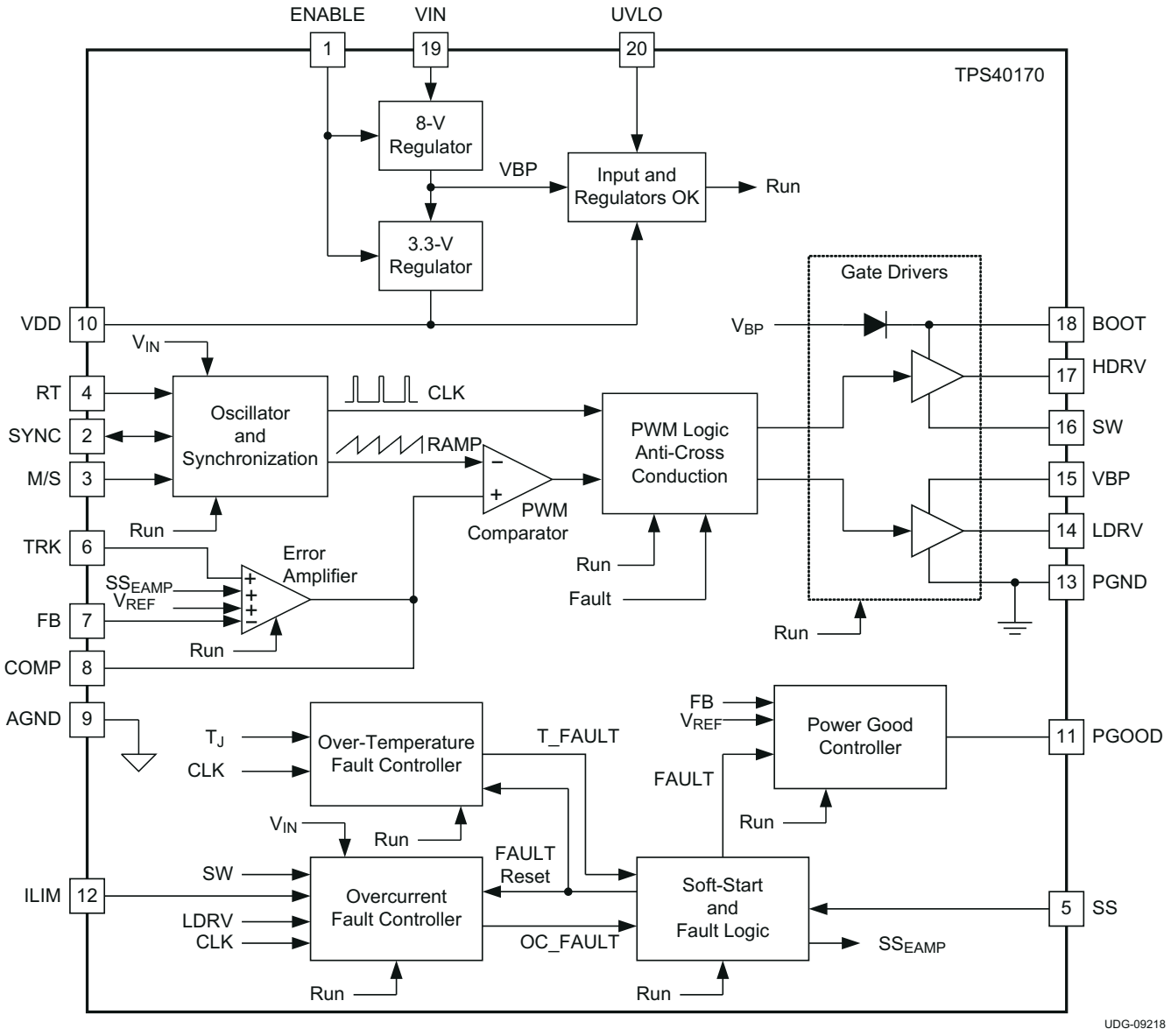
The TPS40170-Q1 device is a synchronous PWM buck controller that accepts a wide range of input voltages from 4.5 V to 60 V and features voltage-mode control with input-voltage feed-forward compensation. The switching frequency is programmable from 100 kHz to 600 kHz.

The TPS40170-Q1 device has a complete set of system protections such as programmable UVLO, programmable overcurrent protection (OCP), selectable short-circuit protection (SCP), and thermal shutdown. The ENABLE pin allows for system shutdown in a low-current (1-µA typical) mode. The controller supports pre-biased outputs, provides an open-drain PGOOD signal, and has closed-loop programmable soft-start, output-voltage tracking, and adaptive dead-time control.

The TPS40170-Q1 device provides accurate output voltage regulation within 1% accuracy.

Additionally, the controller implements a novel scheme of bidirectional synchronization with one controller acting as the primary and other downstream controllers acting as secondaries, synchronized to the primary in-phase or 180° out-of-phase. Secondary controllers can be synchronized to an external clock within ±30% of the internal switching frequency.

6.2 Functional Block Diagram



6.3 Feature Description

6.3.1 LDO Linear Regulators and Enable

The TPS40170-Q1 device has two internal low-dropout (LDO) linear regulators. One has a nominal output voltage of V_{VBP} and is present at the VBP pin. This is the voltage that is mainly used for the gate-driver output. The other linear regulator has an output voltage of V_{VDD} and is present at the VDD pin. This voltage can be used in external low-current logic circuitry. The maximum allowable current drawn from the VDD pin must not exceed 5 mA.

The TPS40170-Q1 device has a dedicated device-enable pin (ENABLE). This simplifies user-level interface design because no multiplexed functions exist. If the ENABLE pin of the TPS40170-Q1 device is higher than V_{EN}, then the LDO regulators are enabled. To ensure that the LDO regulators are disabled, the ENABLE pin must be pulled below V_{DIS}. By pulling the ENABLE pin below V_{DIS}, the device is completely disabled and the current consumption is very low (nominally, 1 µA). Both LDO regulators are actively discharged when the

ENABLE pin is pulled below V_{DIS}. A functionally equivalent circuit to the enable circuitry on the TPS40170-Q1 device is shown in 図 6-1.

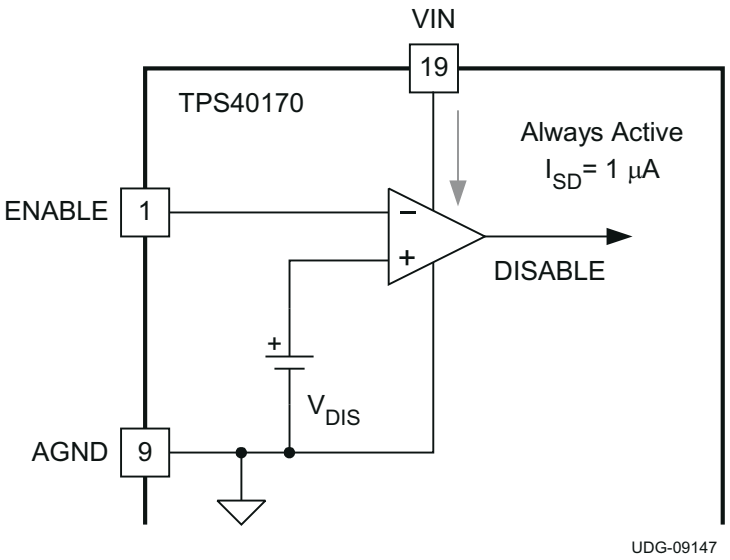


図 6-1. TPS40170-Q1 ENABLE Functional Block

The ENABLE pin must not be allowed to float. If the ENABLE function is not needed for the design, then it is suggested that the ENABLE pin be pulled up to VIN by a high-value resistor, ensuring that the current into the ENABLE pin does not exceed 10 µA. If it is not possible to meet this clamp current requirement, then it is suggested that a resistor divider from VIN to GND be used to connect to ENABLE pin. The resistor divider must be such that the ENABLE pin is higher than V_{EN} and lower than 8 V.

6.3.2 Input Undervoltage Lockout (UVLO)

The TPS40170-Q1 device has both fixed and programmable input undervoltage lockout (UVLO). In order for the device to turn ON, all of the following conditions must be met:

- The ENABLE pin voltage must be greater than V_{EN}.
- The VBP voltage (at VBP pin) must be greater than $V_{VBP(on)}$.
- The UVLO pin must be greater than V_{UVLO}.

In order for the device to turn OFF, any one of the following conditions must be met:

- The ENABLE pin voltage must be less than V_{DIS}.
- The VBP voltage (at the VBP pin) must be less than $V_{VBP(off)}$.
- The UVLO pin must be less than V_{UVLO}.

Programming the input UVLO can be accomplished using the UVLO pin. A resistor divider from the input voltage (VIN pin) to GND sets the UVLO level. Once the input voltage reaches a value that meets the V_{UVLO} level at the UVLO pin, then a small hysteresis current, I_{UVLO} at the UVLO pin is switched in. The programmable UVLO function is shown in 図 6-2.

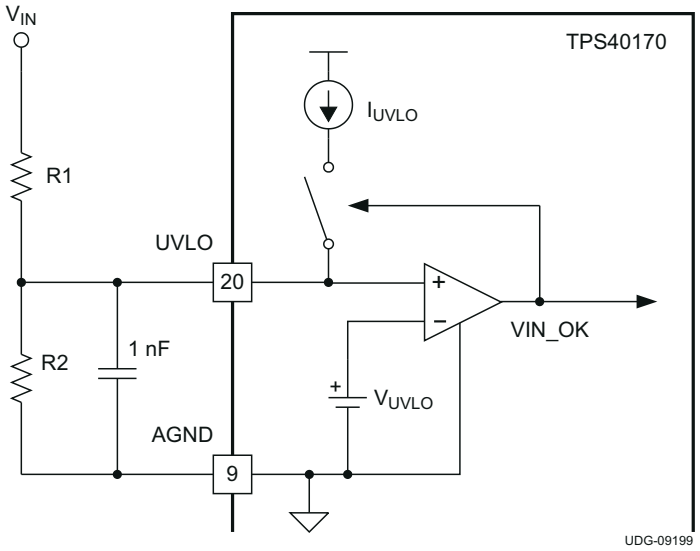


図 6-2. UVLO Functional Block Schematic

6.3.3 Equations for Programming the Input UVLO

Components R1 and R2 represent external resistors for programming UVLO and hysteresis; their values can be calculated in 式 1 and 式 2, respectively.

$$R_1 = \frac{V_{ON} - V_{OFF}}{I_{UVLO}} \tag{1}$$

$$R_2 = R_1 \times \frac{V_{UVLO}}{(V_{ON} - V_{UVLO})} \tag{2}$$

where

- V_{ON} is the desired turnon voltage of the converter.
- V_{OFF} is the desired turnoff voltage for the converter.
- I_{UVLO} is the hysteresis current generated by the device, 5 µA (typical).
- V_{UVLO} is the UVLO pin threshold voltage, 0.9 V (typical).

注

If the UVLO pin is connected to a voltage greater than 0.9 V, the programmable UVLO is disabled and the device defaults to an internal UVLO ($V_{VBP(on)}$ and $V_{VBP(off)}$). For example, the UVLO pin can be connected to VDD or the VBP pin to disable the programmable UVLO function.

A 1-nF ceramic bypass capacitor must be connected between the UVLO pin and GND.

6.3.4 Overcurrent Protection and Short-Circuit Protection (OCP and SCP)

The TPS40170-Q1 device has the capability to set a two-level overcurrent protection. The first level of overcurrent protection (OCP) is the normal overload setting based on low-side MOSFET voltage sensing. The second level of protection is the heavy overload setting, such as short-circuit based, on the high-side MOSFET voltage sensing. This protection takes effect immediately. The second level is termed short-circuit protection (SCP).

The OCP level is set by the ILIM pin voltage. A current (I_{ILIM}) is sourced into the ILIM pin from which a resistor R_{ILIM} is connected to GND. Resistor R_{ILIM} sets the first level of overcurrent limit. The OCP is based on the low-

side FET voltage at the switch-node (SW pin) when LDRV is ON after a blanking time, which is the product of inductor current and low-side FET turnon resistance $R_{DS(on)}$. The voltage is inverted and compared to ILIM pin voltage. If it is greater than the ILIM pin voltage, then a 3-bit counter inside the device increments the fault-count by 1 at the start of the next switching cycle. Alternatively, if it is less than the ILIM pin voltage, then the counter inside the device decrements the fault-count by 1. When the fault-count reaches 7, an overcurrent fault (OC_FAULT) is declared and both the HDRV and LDRV are turned OFF. Resistor R_{ILIM} can be calculated by 式 3.

$$R_{ILIM} = \frac{I_{OC} \times R_{DS(on)}}{I_{ILIM}} = \frac{I_{OC} \times R_{DS(on)}}{9.0\,\mu A} \tag{3}$$

The SCP level is set by a multiple of the ILIM pin voltage. The multiplier has three discrete values, 3, 7, or 15 times, which can be selected by choosing a 10-kΩ, open-circuit, or 20-kΩ resistor, respectively, from the LDRV pin to GND. This multiplier AOC information is translated during the t_{CAL} time, which starts after the enable and UVLO conditions are met.

The SCP is based on sensing the high-side FET voltage drop from V_{VIN} to V_{SW} when HDRV is ON after a blanking time, which is product of inductor current and high-side FET turnon resistance $R_{DS(on)}$. The voltage is compared to the product of the multiplier and the ILIM pin voltage. If the voltage exceeds the product, then the fault-count is immediately set to 7 and the OC_FAULT is declared. HDRV is terminated immediately without waiting for the duty cycle to end. When an OC_FAULT is declared, both the HDRV and LDRV are turned OFF. The appropriate multiplier (A), can be selected using 式 4.

$$A = \frac{I_{SC} \times R_{DS(on)HS}}{I_{OC} \times R_{DS(on)LS}} \tag{4}$$

図 6-3 is a functional block diagram of the two-level overcurrent protection.

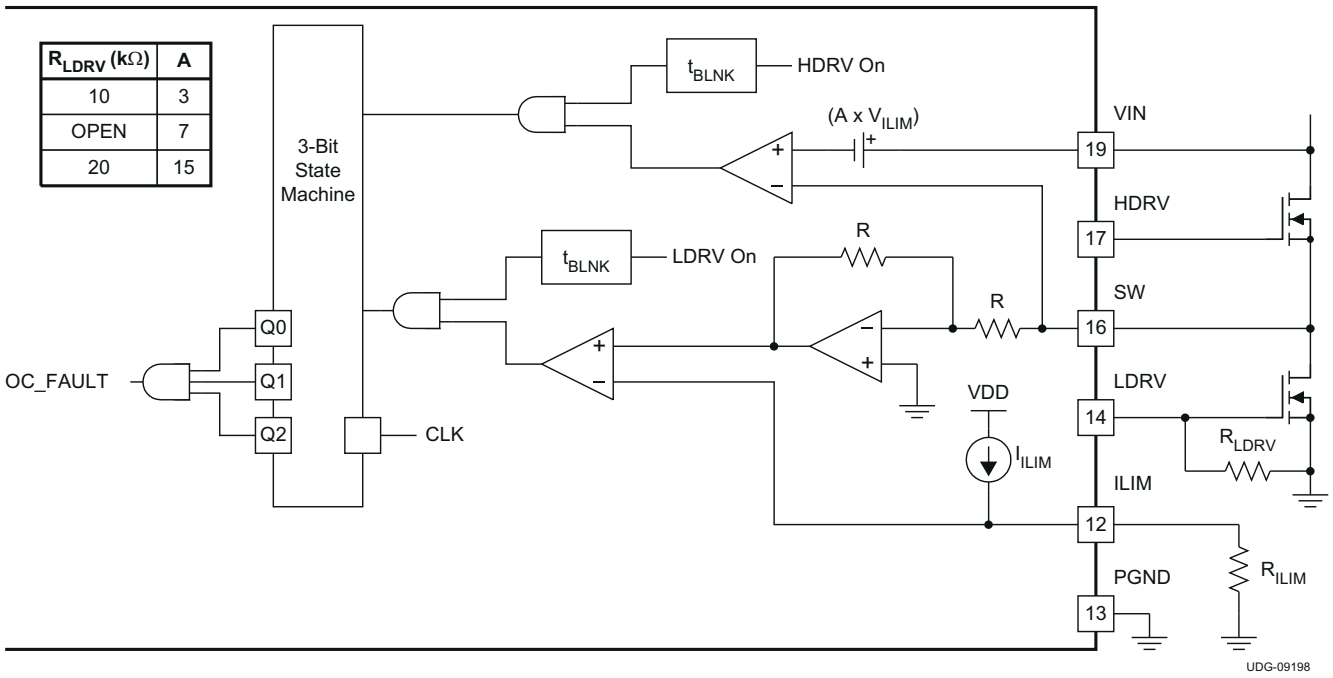


図 6-3. OCP and SCP Protection Functional Block Diagram

注

Both OCP and SCP are based on low-side and high-side MOSFET voltage sensing at the SW node. Excessive ringing on the SW node can have a negative impact on the accuracy of OCP and SCP. Adding an R-C snubber from the SW node to GND helps minimize the potential impact.

6.3.5 Oscillator and Voltage Feed-Forward

TPS40170-Q1 device implements an oscillator with input-voltage feed-forward compensation that enables instant response to input voltage changes. 図 6-4 shows the oscillator timing diagram for the TPS40170-Q1 device. The resistor from the RT pin to GND sets the free-running oscillator frequency. Voltage V_{RT} on the RT pin is made proportional to the input voltage (see 式 5).

$$V_{RT} = \frac{V_{IN}}{K_{PWM}} \tag{5}$$

where

- K = 15

The resistor at the RT pin sets the current in the RT pin. The proportional current charges an internal 100-pF oscillator capacitor. The ramp voltage on this capacitor is compared with the RT pin voltage, V_{RT}. Once the ramp voltage reaches V_{RT}, the oscillator capacitor is discharged. The ramp that is generated by the oscillator (which is proportional to the input voltage) acts as voltage feed-forward ramp to be used in the PWM comparator.

The time between the start of the discharging oscillator capacitor and the start of the next charging cycle is fixed at 170 ns (typical). During the fixed discharge time, the PWM output is maintained as OFF. This is the minimum OFF-time of the PWM output.

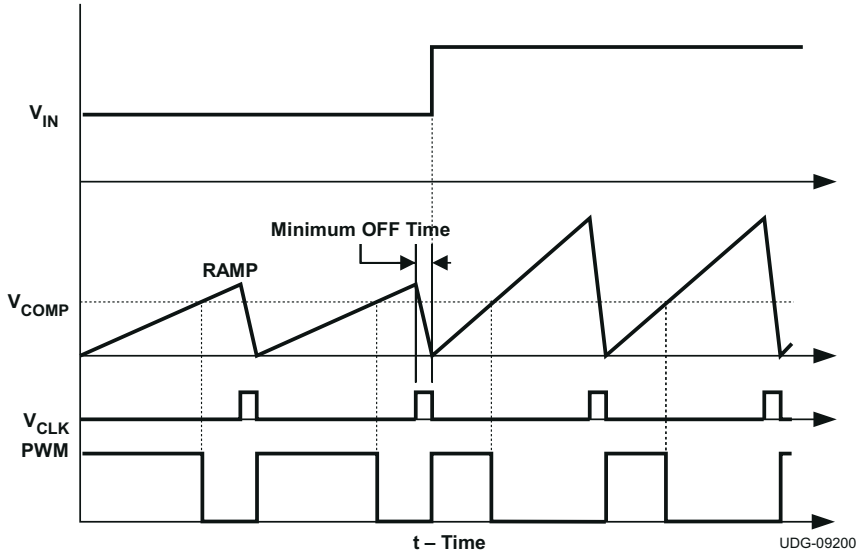


図 6-4. Feed-Forward Oscillator Timing Diagram

6.3.5.1 Calculating the Timing Resistance (R_{RT})

$$R_{RT} = \left(\frac{10^4}{f_{SW}}\right) - 2\,(k\Omega) \tag{6}$$

where

- f_{SW} is the switching frequency in kHz.
- R_{RT} is the resistor connected from RT pin to GND in kΩ.

6.3.6 Feed-Forward Oscillator Timing Diagram

注

The switching frequency can be adjusted between 100 kHz and 600 kHz. The maximum switching frequency before skipping pulses is determined by the input voltage, output voltage, FET resistances, DCR of the inductor, and the minimum on-time of the TPS40170-Q1 device. Use 式 7 to determine the maximum switching frequency. For further details, see application note SLYT293.

$$f_{SW(max)} = \frac{V_{OUT(min)} + \left(I_{OUT(min)} \times \left(R_{DS2} + R_{LOAD}\right)\right)}{t_{ON(min)} \times \left(V_{IN(max)} - I_{OUT(min)} \times \left(R_{DS1} - R_{DS2}\right)\right)} \quad (7)$$

where

- $f_{SW(max)}$ is the maximum switching frequency.
- $V_{OUT(min)}$ is the minimum output voltage.
- $V_{IN(max)}$ is the maximum input voltage.
- $I_{OUT(min)}$ is the minimum output current.
- R_{DS1} is the high-side FET resistance.
- R_{DS2} is the low-side FET resistance.
- R_{LOAD} is the inductor series resistance.

6.3.7 Soft-Start and Fault-Logic

A capacitor from the SS pin to GND defines the SS time, t_{SS}. The TPS40170-Q1 device enters into soft-start immediately after completion of the overcurrent calibration. The SS pin goes through the internal level-shifter circuit of the device before reaching one of the positive inputs of the error amplifier. The SS pin must reach approximately 0.65 V before the input to the error amplifier begins to rise above 0 V. To charge the SS pin from 0 V to 0.65 V faster, an extra charging current (40.4 µA, typical.) is switched-in to the SS pin at the beginning of the soft-start in addition to the normal charging current (11.6 µA, typical.). As the SS capacitor reaches 0.5 V, the extra charging current is turned off and only the normal charging current remains. 図 6-5 shows the soft-start function block.

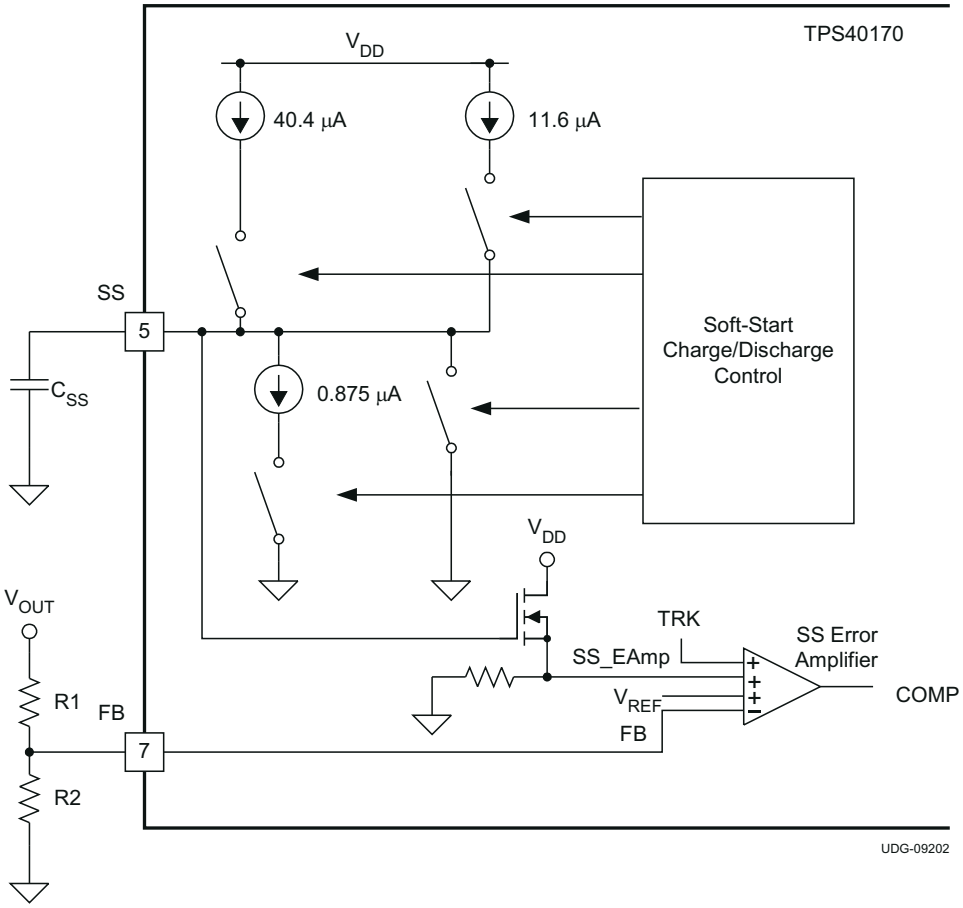


図 6-5. Soft-Start Schematic Block

As the SS pin voltage approaches 0.65 V, the positive input to the error amplifier begins to rise (see 図 6-6). The output of the error amplifier (the COMP pin) starts rising. The rate of rise of the COMP voltage is mainly limited by the feedback-loop compensation network. Once V_{COMP} reaches the V_{valley} of the PWM ramp, the switching begins. The output is regulated to the error amplifier input through the FB pin in the feedback loop. Once the FB pin reaches the 600-mV reference voltage, the feedback node is regulated to the reference voltage, V_{REF}. The SS pin continues to rise and is clamped to VDD.

The SS pin is discharged through an internal switch during the following conditions:

- Input (VIN) undervoltage lock out UVLO pin less than V_{UVLO}
- Overcurrent protection calibration time (t_{CAL})
- VBP less than threshold voltage ($V_{BP(off)}$)

Because it is discharged through an internal switch, the discharging time is relatively fast compared with the discharging time during the fault restart, which is discussed in the *セクション 6.3.7.1* section.

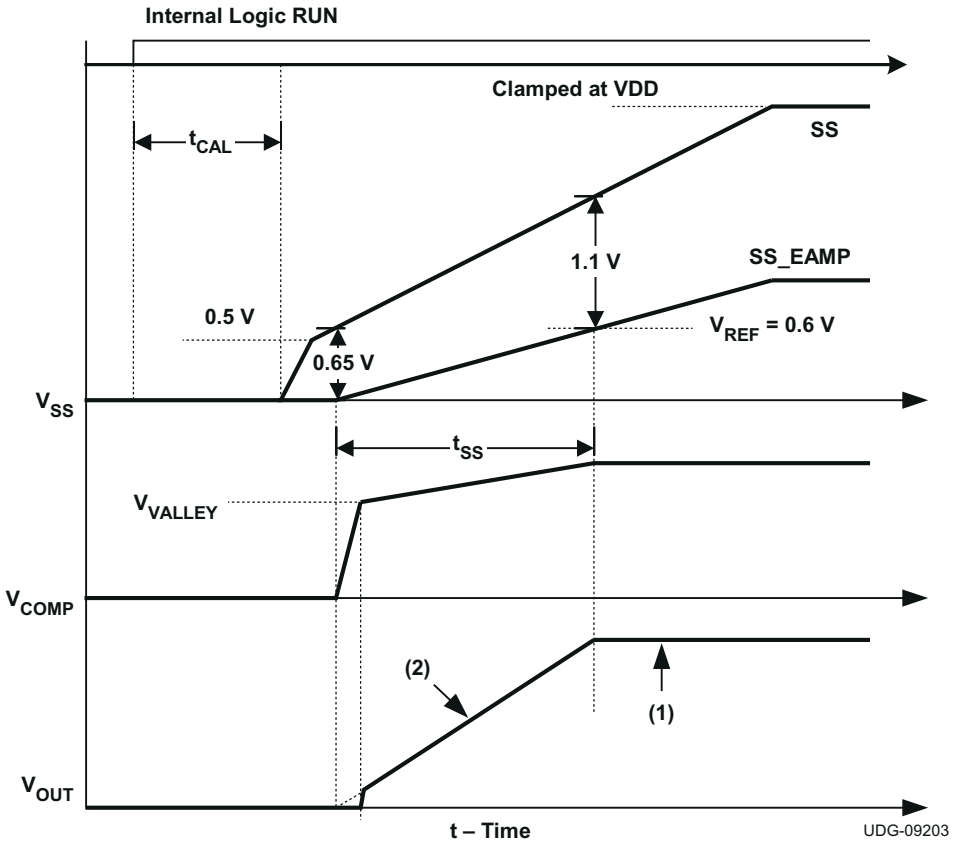


図 6-6. Soft-Start Waveforms

注

Referring to 図 6-6:

(1) VREF dominates the positive input of the error amplifier.

(2) SS_EAMP dominates the positive input of the error amplifier.

For $0 < V_{SS\_EAMP} < V_{REF}$

$$V_{OUT} = V_{SS(EAMP)} \times \frac{(R1+R2)}{R2} \quad (8)$$

For $V_{SS\_EAMP} > V_{REF}$

$$V_{OUT} = V_{REF} \times \frac{(R1+R2)}{R2} \quad (9)$$

6.3.7.1 Soft-Start During Overcurrent Fault

The soft-start block also has a role to control the fault-logic timing. If an overcurrent fault (OC_FAULT) is declared, the soft-start capacitor is discharged internally through the device by a small current $I_{SS(sink)}$ (1.05 µA, typical). Once the SS pin capacitor is discharged to below $V_{SS(flt,low)}$ (300 mV, typical), the soft-start capacitor begins charging again. If the fault is persistent, a fault is declared which is determined by the overcurrent-protection state machine. If the soft-start capacitor is below $V_{SS(flt,high)}$ (2.5 V, typical), then the soft-start

capacitor continues to charge until it reaches $V_{SS(flt,high)}$ before a discharge cycle is initiated. This ensures that the re-start time-interval is always constant. 図 6-7 shows the restart timing.

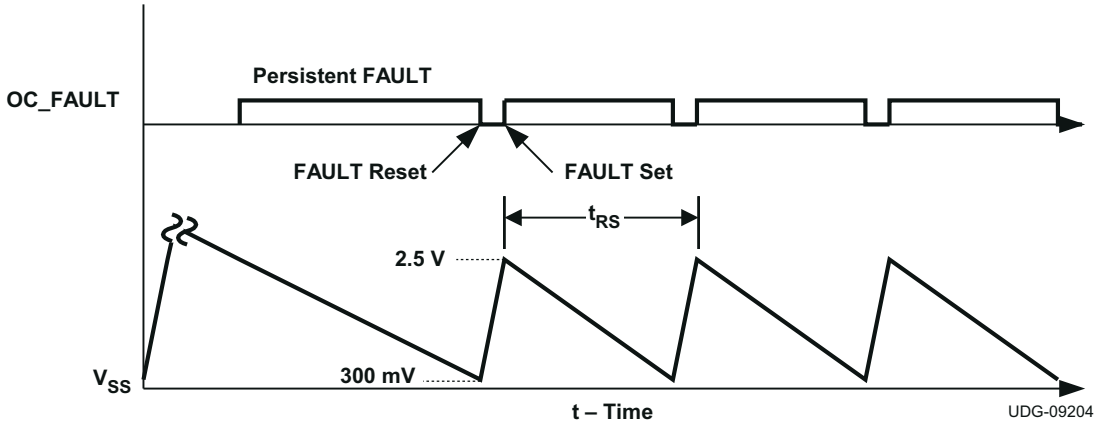


図 6-7. Overcurrent Fault Restart Timing

注

For the feedback to be regulated to the SS_EAMP voltage, the TRK pin must be pulled high directly or through a resistor to VDD.

6.3.7.2 Equations for Soft-Start and Restart Time

The soft-start time (t_{SS}) is defined as the time taken for the internal SS_EAMP node to go from 0 V to the 0.6-V V_{REF} voltage. SS_EAMP starts rising as the SS pin goes beyond 0.65 V. The offset voltage between SS and SS_EAMP starts increasing as the SS pin voltage starts rising. 図 6-6 shows that the SS time can be defined as the time taken for the SS pin voltage to change by 1.05 V (see 式 10).

$$C_{SS} = \frac{t_{SS}}{0.09} \tag{10}$$

The restart time (t_{RS}) is defined in 式 11 as the time taken for the soft-start capacitor (C_{SS}) to discharge from 2.5 V to 0.3 V and to then recharge up to 2.5 V.

$$t_{RS} \approx 2.28 \times C_{SS} \tag{11}$$

where

- C_{SS} is the soft-start capacitance in nF
- t_{SS} is the soft-start time in ms
- t_{RS} is the restart time in ms

注

During soft-start (V_{SS} < 2.5 V), the overcurrent protection limit is 1.5 times the normal overcurrent protection limit. This allows a higher output capacitance to charge fully without activating overcurrent protection.

6.3.8 Overtemperature Fault

図 6-8 shows the over temperature protection scheme. If the junction temperature of the device reaches the thermal shutdown limit of $t_{SD(set)}$ (165°C, typical) and SS charging is completed, an overtemperature FAULT is

declared. The soft-start capacitor begins to be discharged. During soft-start discharging period, the PWM switching is terminated; therefore, both HDRV and LDRV are driven low, turning off both MOSFETs.

The soft-start capacitor begins to charge and an overtemperature fault is reset whenever the soft-start capacitor is discharged below $V_{SS(flt,low)}$ (300 mV, typical). During each restart cycle, PWM switching is turned on. When SS is fully charged, PWM switching is terminated. These restarts repeat until the temperature of the device has fallen below the thermal reset level, $t_{SD(reset)}$ (135°C typical). PWM switching continues and the system returns to normal regulation.

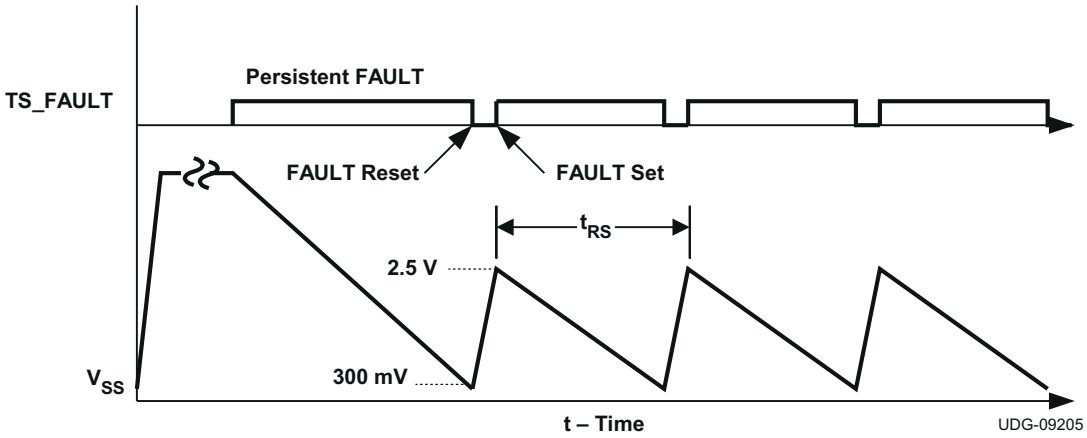


図 6-8. Overtemperature Fault Restart Timing

The soft-start timing during an overtemperature fault is the same as the soft-start timing during an overcurrent fault. See the *Equations for Soft-Start and Restart Time* section.

6.3.9 Tracking

The TRK pin is used for output voltage tracking. The output voltage is regulated so that the FB pin equals the lowest of the internal reference voltage (V_{REF}) or the level-shifted SS pin voltage (SS_{EAMP}) or the TRK pin voltage. Once the TRK pin goes above the reference voltage, then the output voltage is no longer governed by the TRK pin, but it is governed by the reference voltage.

If the voltage tracking function is used, then it must be noted that the SS pin capacitor must remain connected to SS pin and is also used for FAULT timing. For proper tracking using the TRK pin, the tracking voltage must be allowed to rise only after SS_{EAMP} has exceeded V_{REF}, so that there is no possibility of the TRK pin voltage being higher than the SS_{EAMP} voltage. From 図 6-6, for SS_{EAMP} = 0.6 V, the SS pin voltage is typically 1.7 V.

The maximum slew rate on the TRK pin must be determined by the output capacitance and feedback loop bandwidth. A higher slew rate can possibly trip overcurrent protection.

図 6-9 shows the tracking functional block. For SS_{EAMP} voltages greater than TRK pin voltage, the V_{OUT} is given by 式 12 and 式 13.

For $0\ V < V_{TRK} < V_{REF}$

$$V_{OUT} = V_{TRK} \times \frac{(R1+R2)}{R2} \tag{12}$$

For $V_{TRK} > V_{REF}$

$$V_{OUT} = V_{REF} \times \frac{(R1+R2)}{R2} \tag{13}$$

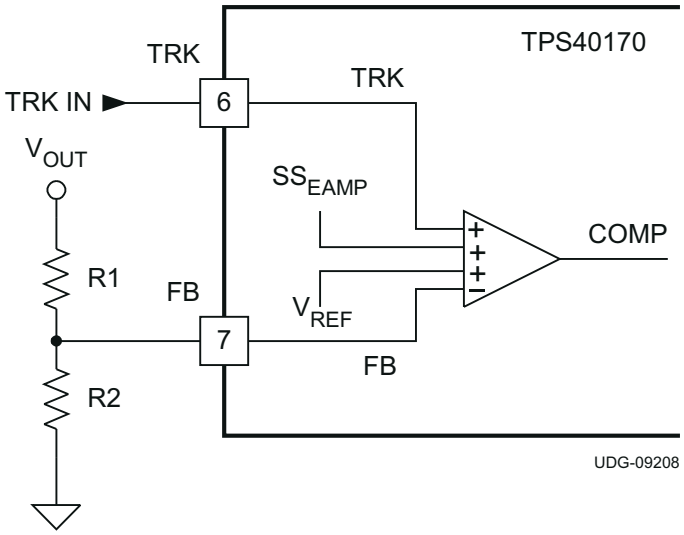


図 6-9. Tracking Functional Block

There are three potential applications for the tracking function.

- Simultaneous voltage tracking
- Ratiometric voltage tracking
- Sequential start-up mode

The tracking function configurations and waveforms are shown in 図 6-10, 図 6-11, 図 6-12, 図 6-13, 図 6-14, and 図 6-15 respectively.

In simultaneous voltage tracking, shown in 図 6-10, tracking signals VTRK1 and VTRK2 of two modules, POL1 and POL2, start up at the same time, and their output voltages VOUT1initial and VOUT2initial are approximately the same during initial startup. Because VTRK1 and VTRK2 are less than V_{REF} (0.6 V, typical), 式 12 is used. As a result, components selection must meet 式 14.

$$\left(\frac{(R_1+R_2)}{R_1}\right)\times V_{TRK1}=\left(\frac{(R_3+R_4)}{R_3}\right)\times V_{TRK2}\Rightarrow\frac{R_5}{R_6}=\left(\frac{\left(\frac{R_1}{(R_1+R_2)}\right)}{\left(\frac{R_3}{(R_3+R_4)}\right)}-1\right) \tag{14}$$

After the lower output voltage setting reaches the output-voltage V_{OUT1} set point, where V_{TRK1} increases above V_{REF}, the output voltage of the other one (V_{OUT2}) continues increasing until it reaches its own set point, where V_{TRK2} increases above V_{REF}. At that time, 式 13 is used. As a result, the resistor settings must meet 式 15 and 式 16.

$$V_{OUT1}=\left(\frac{(R_1+R_2)}{R_1}\right)\times V_{REF} \tag{15}$$

$$V_{OUT2}=\left(\frac{(R_3+R_4)}{R_3}\right)\times V_{REF} \tag{16}$$

式 14 can be simplified into 式 17 by substituting terms from 式 15 and 式 16.

$$\left(\frac{R_5}{R_6}\right) = \left(\left(\frac{V_{OUT2}}{V_{OUT1}}\right) - 1\right) \quad (17)$$

If 5-V V_{OUT2} and 2.5-V V_{OUT1} are required, according to 式 15, 式 16, and 式 17, the selected components can be as follows:

- $R_5 = R_6 = R_4 = R_2 = 10\ k\Omega$
- $R_1 = 3.16\ k\Omega$
- $R_3 = 1.37\ k\Omega$

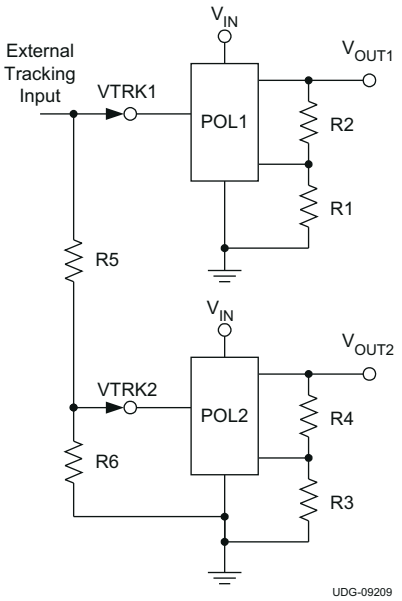


図 6-10. Simultaneous Voltage-Tracking Schematic

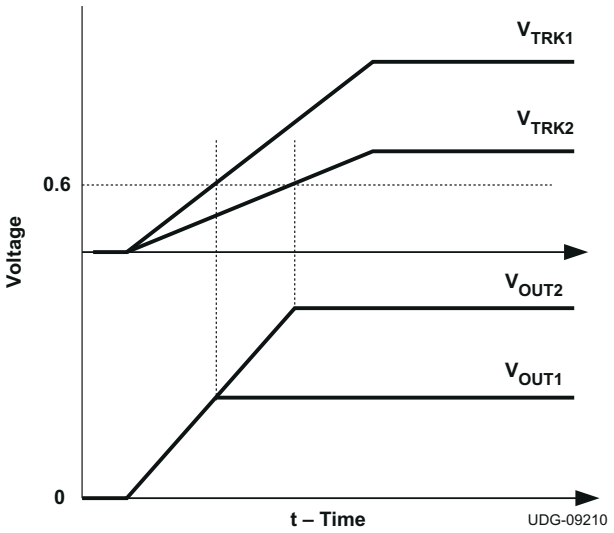


図 6-11. Simultaneous Voltage-Tracking Waveform

In ratiometric voltage tracking as shown in 図 6-12, the two tracking voltages, VTRK1 and VTRK2, for two modules, POL1 and POL2, are the same. Their output voltages, VOUT1 and VOUT2, are different with different voltage dividers, R2–R1 and R4–R3. VOUT1 and VOUT2 increase proportionally and reach their output voltage set points at approximately the same time.

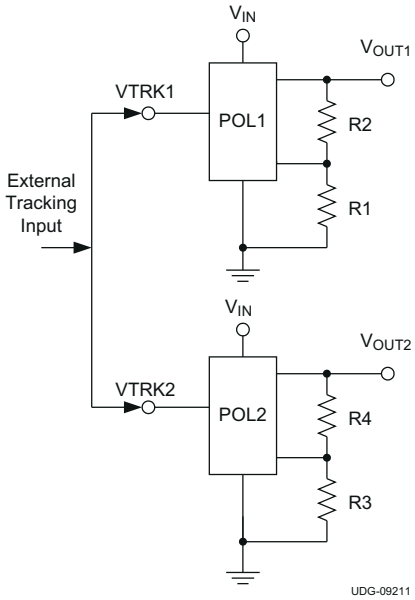


図 6-12. Ratiometric Voltage-Tracking Schematic

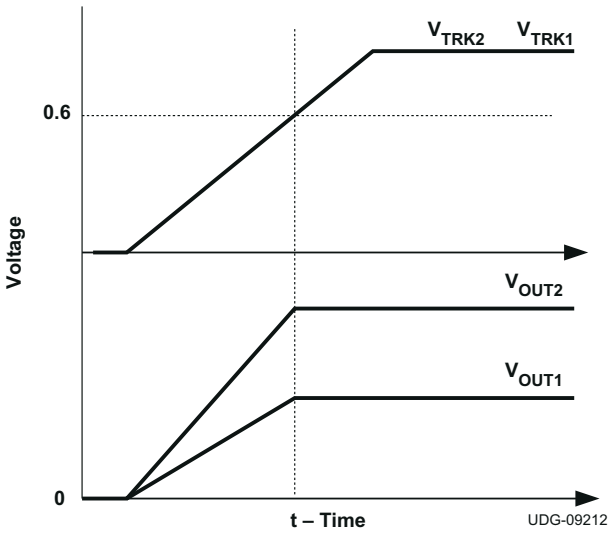


図 6-13. Ratiometric Voltage-Tracking Waveform

Sequential start-up is shown in 図 6-14. During start-up of the first module, POL1, its PGOOD1 is pulled to low. Because PGOOD1 is connected to soft-start SS2 of the second module, POL2, is not able to charge its soft-start capacitor. After output voltage VOUT1 of POL1 reaches its setting point, PGOOD1 is released. POL2 starts charging its soft-start capacitor. Finally, output voltage V_{OUT2} of POL2 reaches its setting point.

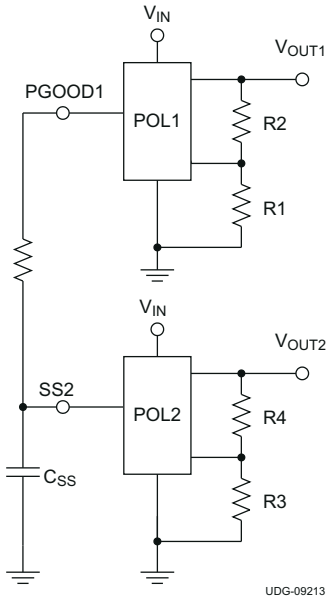


図 6-14. Sequential Start-Up Schematic

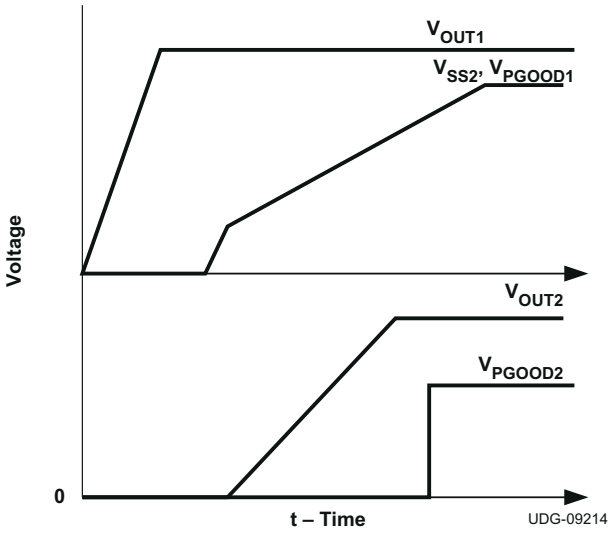


図 6-15. Sequential Start-Up Waveform

注

The TRK pin has high impedance, so it is a noise-sensitive terminal. If the tracking function is used, TI recommends a small R-C filter at the TRK pin to filter out high-frequency noise.

If the tracking function is not used, the TRK pin must be pulled up directly or through a resistor (with a value between 10 kΩ and 100 kΩ) to VDD.

6.3.10 Adaptive Drivers

The drivers for the external high-side and low-side MOSFETs are capable of driving a gate-to-source voltage, V_{VBP}. The LDRV driver for the low-side MOSFET switches between VBP and PGND, while the HDRV driver for the high-side MOSFET 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.

6.3.11 Start-Up Into Pre-Biased Output

The TPS40170-Q1 device contains a circuit to prevent current from being pulled out of the output during startup, in case 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_{VFB}]), the controller slowly activates synchronous rectification by starting the first LDRV pulses with a narrow on-time (see 図 6-16), where:

- V_{IN} = 5 V
- V_{OUT} = 3.3 V
- V_{PRE} = 1.4 V
- f_{SW} = 300 kHz
- L = 0.6 µH

LDRV pulses then increments the 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 output voltage (V_{OUT}) starts and ramps up smoothly into regulation and the control loop is given time to transition from pre-biased startup to normal mode operation with minimal disturbance to the output voltage. The time from the start of switching until the low-side MOSFET is turned on for the full (1 – D) interval is between approximately 20 and 40 clock cycles.

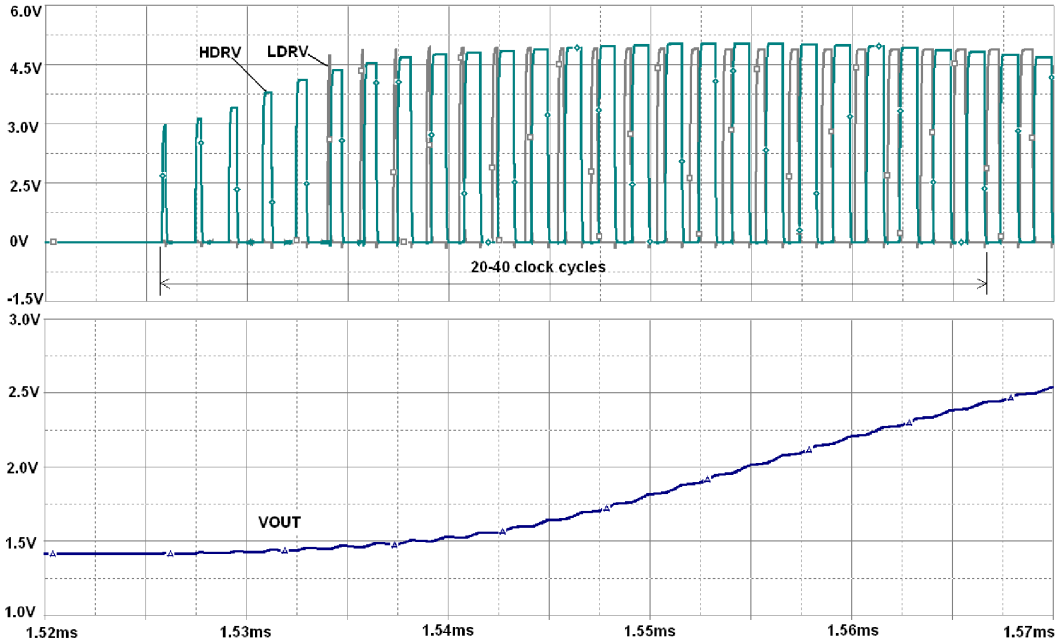


図 6-16. Start-Up Switching Waveform During Pre-Biased Condition

6.3.12

If the output is pre-biased to a voltage higher than the voltage commanded by the reference, then the PWM switching does not start.

注

When output is pre-biased at $V_{PREBIAS}$, that voltage also applies to the SW node during start-up. When the pre-bias circuitry commands the first few high-side pulses before the first low-side pulse is initiated, the gate voltage for the high-side MOSFET is as described in 式 18. Alternatively, if the pre-bias level is high, it is possible that SCP can be tripped due to high the turnon resistance of the high-side MOSFET with low gate voltage. Once tripped, the device resets and then attempts to restart. The device can not be able to start up until the output is discharged to a lower voltage level either by an active load or through feedback resistors.

In the case of a high pre-bias level, a low gate-threshold-voltage-rated device is recommended for the high-side MOSFET, and increasing the SCP level also helps alleviate the problem.

$$V_{GATE(hs)} = \left(V_{BP} - V_{DFWD} - V_{PRE-BIAS}\right) \quad (18)$$

where

- $V_{GATE(hs)}$ is the gate voltage for the high-side MOSFET.
- V_{BP} is the BP regulator output.
- V_{DFWD} is bootstrap diode forward voltage.

6.3.13 Power Good (PGOOD)

The TPS40170-Q1 device provides an indication that the output voltage of the converter is within the specified limits of regulation as measured at the FB pin. The PGOOD pin is an open-drain signal and pulls low when any

condition exists which can indicate that the output of the supply can be out of regulation. These conditions include:

- V_{VFB} is not within the PGOOD threshold limits.
- Soft-start is active, that is, the SS pin voltage is below $V_{SS,FLT,HIGH}$ limit.
- An undervoltage condition exists for the device.
- An overcurrent or short-circuit fault is detected.
- An overtemperature fault is detected.

図 6-17 shows a situation where no fault is detected during the start-up, (the normal PGOOD situation). It shows that PGOOD goes high t_{PGD} (20 µs, typical) after all the conditions (previously listed) are met.

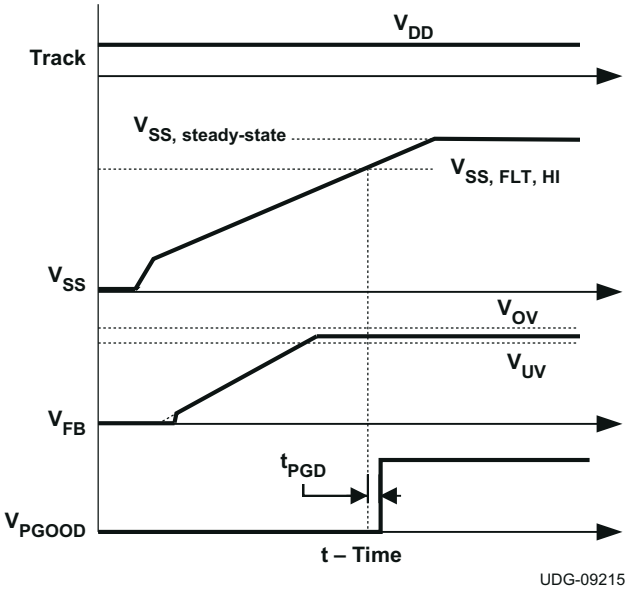


図 6-17. PGOOD Signal

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 pulldown device allows the PGOOD pin to operate as a diode to GND.

6.3.14 PGND and AGND

注

The TPS40170-Q1 device provides separate signal ground (AGND) and power ground (PGND) pins. PGND is primarily used for gate-driver ground return. AGND is an internal logic-signal ground return. These two ground signals are internally loosely connected by two anti-parallel diodes. PGND and AGND must be electrically connected externally.

6.3.15 Bootstrap Capacitor

A bootstrap capacitor with a value between 0.1 µF and 0.22 µF must be placed between the BOOT pin and the SW pin. It must be 10 times higher than MOSFET gate capacitance.

6.3.16 Bypass and Filtering

In an integrated circuit, supply bypassing is important for jitter-free operation. To decrease noise in the converter, ceramic bypass capacitors must be placed as close to the package as possible.

- VIN to GND: use a 0.1-µF ceramic capacitor

- BP to GND: use a 1-µF to 10-µF ceramic capacitor. It must be 10 times greater than the bootstrap capacitance
- VDD to GND: use a 0.1-µF to 1-µF ceramic capacitor

6.4 Device Functional Modes

6.4.1 Frequency Synchronization

The TPS40170-Q1 device has three modes.

- **Primary mode**: In this mode, the primary- or secondary-selector pin, (M/S) is connected to VIN. The SYNC pin emits a stream of pulses at the same frequency as the PWM switching frequency. The pulse stream at the SYNC pin is of 50% duty cycle and the same amplitude as V_{VBP}. Also, the falling edge of the voltage on SYNC pin is synchronized with the rising edge of HDRV.
- **Secondary–180° mode**: In this mode, the M/S pin is connected to GND. The SYNC pin of the TPS40170-Q1 device accepts a synchronization clock signal, and HDRV is synchronized with the rising edge of the incoming synchronization clock.
- **Secondary–0° mode**: In this mode, the M/S pin is left open. The SYNC pin of the TPS40170-Q1 device accepts a synchronization clock signal, and HDRV is synchronized with the falling edge of the incoming synchronization clock.

The two secondary modes can be synchronized to an external clock through the SYNC pin. They are shown in 図 6-18. The synchronization frequency must be within ±30% of its programmed free-running frequency.

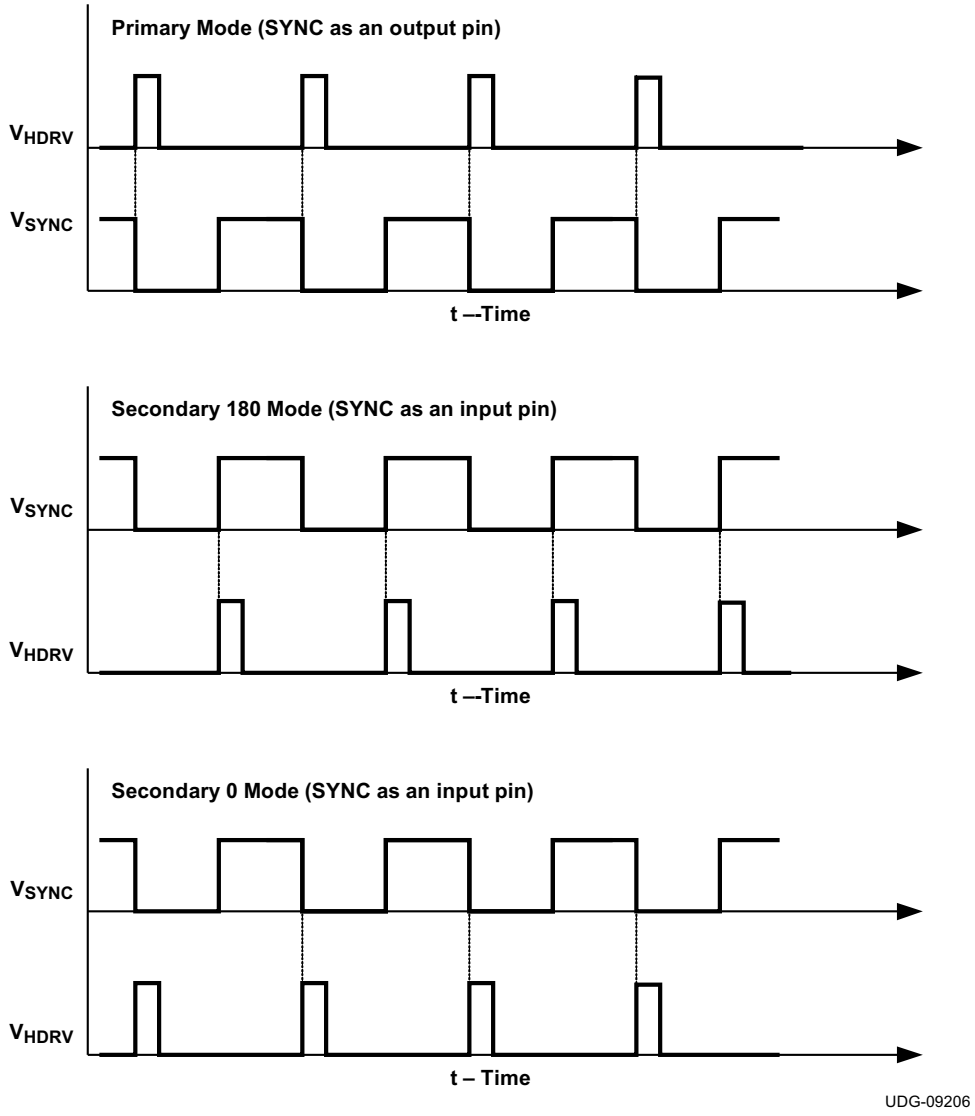


図 6-18. Frequency Synchronization Waveforms in Different Modes

TPS40170-Q1 device provides a smooth transition for the SYNC clock-signal loss in secondary mode. In secondary mode, a synchronization clock signal is provided externally through the SYNC pin to the device. The switching frequency is synchronized to the external SYNC clock signal. If for some reason the external clock signal is missing, the device switching frequency is automatically overridden by a transition frequency which is 0.7 times its programmed free-running frequency. This transition time is approximately 20 µs. After that, the device switching frequency is changed to its programmed free-running frequency. 図 6-19 shows this process.

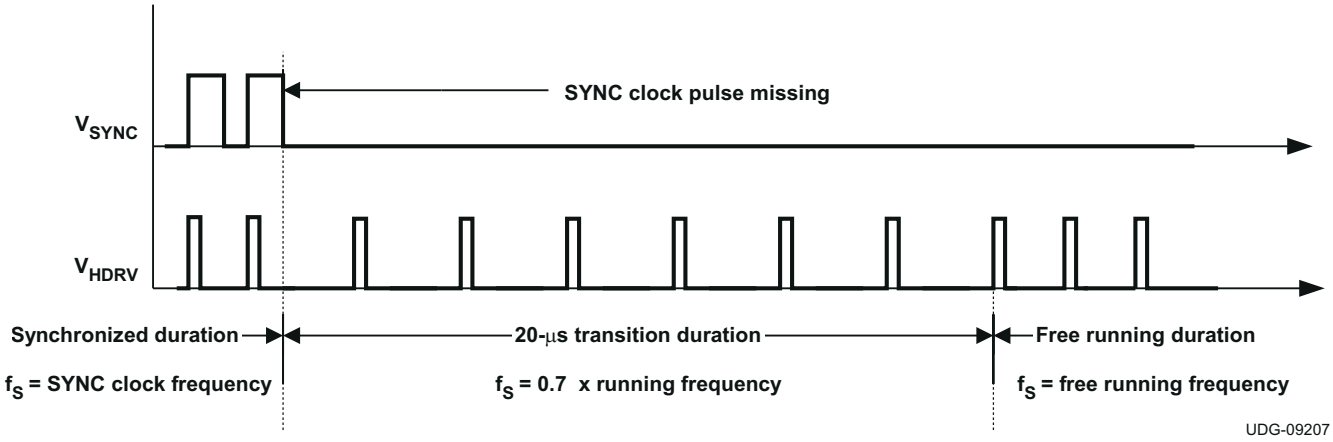


図 6-19. Transition for SYNC Clock Signal Missing (for Secondary–180° Mode)

注

When the device is operating in the primary mode with duty ratio around 50%, PWM jittering can occur. Always configure the device into the secondary mode by either connecting the M/S pin to GND or leaving it floating if primary mode is not used.

When the external SYNC clock signal is used for synchronization, limit the maximum slew rate of the clock signal to 10 V/µs to avoid potential PWM jittering,and connect the SYNC pin to the external clock signal through a 5-kΩ resistor.

7 Application and Implementation

注

以下のアプリケーション情報は、TI の製品仕様に含まれるものではなく、TI ではその正確性または完全性を保証いたしません。個々の目的に対する製品の適合性については、お客様の責任で判断していただくことになります。お客様は自身の設計実装を検証しテストすることで、システムの機能を確認する必要があります。

7.1 Application Information

The wide-input TPS40170-Q1 controller can function in a very wide range of applications. This example describes the design process for a very wide-input (10 V to 60 V) to regulated 5-V output at a load current of 6 A.

7.2 Typical Application

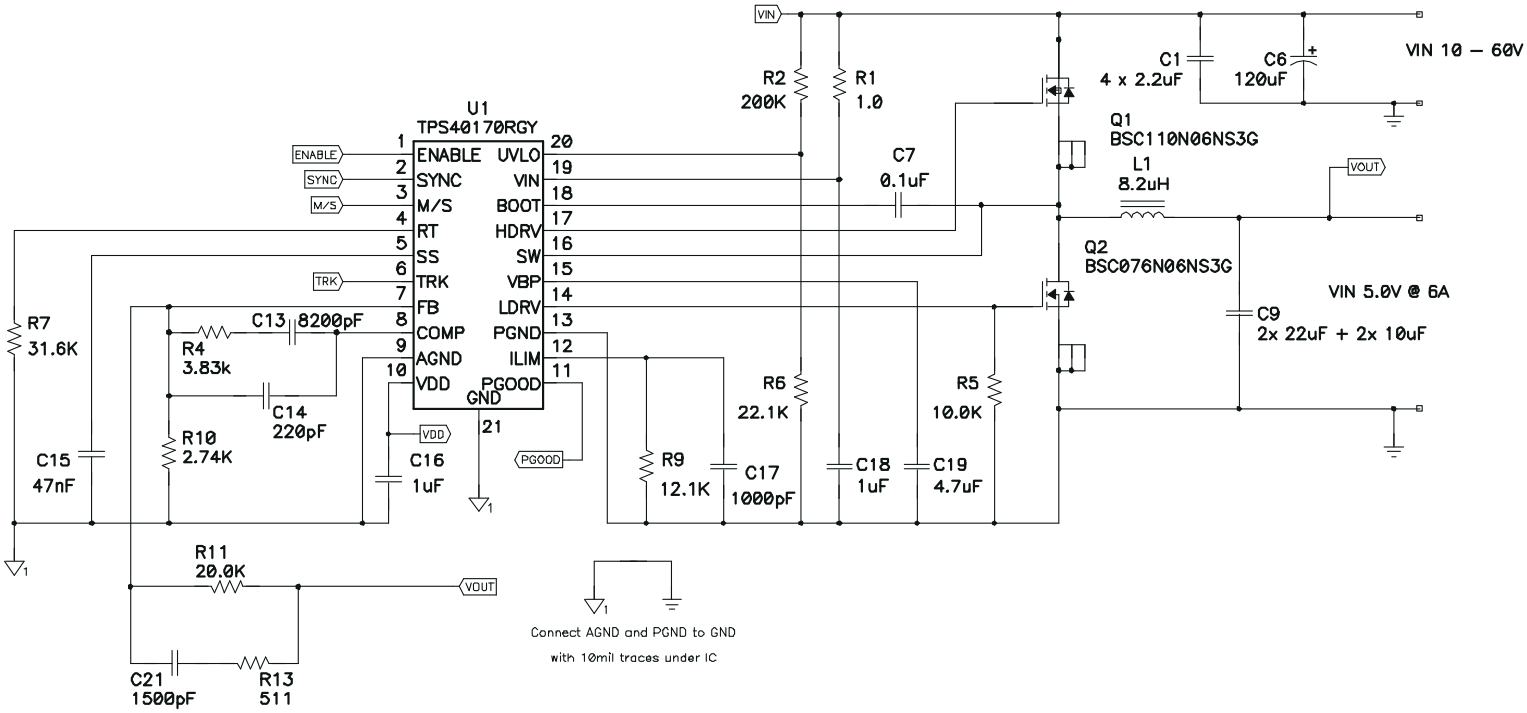


図 7-1. Design Example Application

7.2.1 Design Requirements

The design parameters are provided in 表 7-1.

表 7-1. Design Example Parameters

PARAMETER		TEST CONDITIONS	MIN	NOM	MAX	UNIT
V_{IN}	Input voltage		10		60	V
$V_{IN(ripple)}$	Input ripple	I_{OUT} = 6 A			0.5	V
V_{OUT}	Output voltage	0 A ≤ I_{OUT} ≤ 20 A	4.8	5	5.2	V
	Line regulation	10 V ≤ V_{IN} ≤ 60 V			0.5%	
	Load regulation	0 A ≤ I_{OUT} ≤ 6 A			0.5%	
V_{RIPPLE}	Output ripple	I_{OUT} = 6 A			100	mV
V_{OVER}	Output overshoot	ΔI_{OUT} = 2.5 A		250		mV
V_{UNDER}	Output undershoot	ΔI_{OUT} = -2.5 A		250		mV
I_{OUT}	Output current	10 V ≤ V_{IN} ≤ 60 V	0		6	A
t_{SS}	Soft-start time	V_{IN} = 24 V		4		ms
I_{SCP}	Short circuit current trip point		8			A
η	Efficiency	V_{IN} = 24 V, I_{OUT} = 6 A		94%		
f_{SW}	Switching frequency			300		kHz
	Size				1.5	in^2

7.2.2 Detailed Design Procedure

表 7-2. Design Example List of Materials

REFERENCE DESIGNATOR	QTY	VALUE	DESCRIPTION	SIZE	PART NUMBER	MANUF
C1	4	2.2 µF	Capacitor, ceramic, 100-V, X7R, 15%	1210	Std	Std
C6	1	120 µF	Capacitor, aluminum, 63-V, 20%, KZE series	0.315 inch (0.8 cm)	KZE63VB121M10X16LL	Chemi-con
C7	1	0.1 µF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
C9	2 ea	22 µF 10 µF	Capacitor, ceramic, 16-V, X7R, 15%	1210	Std	Std
C13	1	8200 pF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
C14	1	220 pF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
C15	1	47 nF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
C16	1	1 µF	Capacitor, 16-V, X7R, 15%	603	Std	Std
C17	1	1000 pF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
C18	1	1 µF	Capacitor, ceramic, 100-V, X7R, 15%	1206	Std	Std
C19	1	4.7 µF	Capacitor, ceramic, 16-V, X5R, 15%	805	Std	Std
C21	1	1500 pF	Capacitor, ceramic, 50-V, X7R, 15%	603	Std	Std
L1	1	8.2 µH	Inductor, SMT, 10-A, 16-mΩ	0.51 $inch^2$ (1.3 cm^2)	IHLP5050FDER8R2M01	Vishay
Q1	1		MOSFET, N-channel, 60-V, 50-A, 11-mΩ		BSC110N06NS3G	Infineon
Q2	1		MOSFET, N-channel, 60-V, 50-A, 7.6-mΩ		BSC076N06NS3G	Infineon
R10	1	2.74 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R4	1	3.83 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R5	1	10.0 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603

表 7-2. Design Example List of Materials (続き)

REFERENCE DESIGNATOR	QTY	VALUE	DESCRIPTION	SIZE	PART NUMBER	MANUF
R9	1	12.1 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R11	1	20.0 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R6	1	22.1 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R7	1	31.6 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R2	1	200 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
R13	1	511 kΩ	Resistor, chip, 1/16W, 1%	603	Std	R603
U1			IC, 4.5 V–60 V wide input sync. PWM buck controller		TPS40170-Q1RGY	Texas Instruments

7.2.2.1 Select A Switching Frequency

To maintain acceptable efficiency and meet minimum on-time requirements, a 300-kHz switching frequency is selected.

7.2.2.2 Inductor Selection (L1)

Synchronous buck power inductors are typically sized for approximately 20%–40% of peak-to-peak ripple current (I_{RIPPLE}). Given this target ripple current, the required inductor size can be calculated in 式 19.

$$L \approx \frac{V_{IN(max)} - V_{OUT}}{0.3 \times I_{OUT}} \times \frac{V_{OUT}}{V_{IN(max)}} \times \frac{1}{f_{SW}} = \frac{60\,V - 5\,V}{0.3 \times 6\,A} \times \frac{5\,V}{60\,V} \times \frac{1}{300\,kHz} = 8.5\,\mu H \quad (19)$$

Selecting a standard 8.2-µH inductor value, solving for I_{RIPPLE} = 1.86 A.

The rms current through the inductor is approximated by 式 20.

$$I_{L(rms)} = \sqrt{\left(I_{L(avg)}\right)^2 + \tfrac{1}{12} \times \left(I_{RIPPLE}\right)^2} = \sqrt{\left(I_{OUT}\right)^2 + \tfrac{1}{12} \times \left(I_{RIPPLE}\right)^2} = \sqrt{(6)^2 + \tfrac{1}{12} \times (1.86)^2} = 6.02\,A \quad (20)$$

7.2.2.3 Output Capacitor Selection (C9)

The selection of the output capacitor is typically driven by the output transient response. 式 21 and 式 22 overestimate the voltage deviation to account for delays in the loop bandwidth and can be used to determine the required output capacitance:

$$V_{OVER} < \frac{I_{TRAN}}{C_{OUT}} \times \Delta T = \frac{I_{TRAN}}{C_{OUT}} \times \frac{I_{TRAN} \times L}{V_{OUT}} = \frac{\left(I_{TRAN}\right)^2 \times L}{V_{OUT} \times C_{OUT}} \quad (21)$$

$$V_{UNDER} < \frac{I_{TRAN}}{C_{OUT}} \times \Delta T = \frac{I_{TRAN}}{C_{OUT}} \times \frac{I_{TRAN} \times L}{\left(V_{IN} - V_{OUT}\right)} = \frac{\left(I_{TRAN}\right)^2 \times L}{\left(V_{IN} - V_{OUT}\right) \times C_{OUT}} \quad (22)$$

If $V_{IN(min)}$ > 2 × V_{OUT}, use overshoot to calculate minimum output capacitance. If $V_{IN(min)}$ < 2 × V_{OUT}, use undershoot to calculate minimum output capacitance.

$$C_{OUT(min)} = \frac{\left(I_{TRAN(max)}\right)^2 \times L}{V_{OUT} \times V_{OVER}} = \frac{(3)^2 \times 8.2\,\mu H}{5 \times 250\,mV} = 59\,\mu F \quad (23)$$

With a minimum capacitance, the maximum allowable ESR is determined by the maximum ripple voltage and is approximated by 式 24.

$$ESR_{MAX} = \frac{V_{RIPPLE(tot)} - V_{RIPPLE(cap)}}{I_{RIPPLE}} = \frac{V_{RIPPLE(tot)} - \left(\frac{I_{RIPPLE}}{8 \times C_{OUT} \times f_{SW}}\right)}{I_{RIPPLE}} = \frac{100\,mV - \left(\frac{1.86\,A}{8 \times 59\,\mu F \times 300\,kHz}\right)}{1.86\,A} = 47\,m\Omega \quad (24)$$

Two 1210, 22-µF, 16-V X7R ceramic capacitors plus two 0805 10-µF, 16-V X7R ceramic capacitors are selected to provide more than 59 µF of minimum capacitance (including tolerance and dc bias derating) and less than 47 mΩ of ESR (parallel ESR of approximately 4 mΩ).

7.2.2.4 Peak Current Rating of 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 式 25.

$$I_{CHARGE} = \frac{V_{OUT} \times C_{OUT}}{t_{SS}} = \frac{5\,V \times (2 \times 22\,\mu F + 2 \times 10\,\mu F)}{4\,ms} = 0.08\,A \quad (25)$$

$$I_{L(peak)} = I_{OUT(max)} + \left(\tfrac{1}{2} \times I_{RIPPLE}\right) + I_{CHARGE} = 6\,A + \tfrac{1}{2} \times 1.86\,A + 0.08\,A = 7.01A \quad (26)$$

An IHLP5050FDER8R2M01 8.2-µH capacitor is selected. This 10-A, 16-mΩ inductor exceeds the minimum inductor ratings in a 13-mm × 13-mm package.

7.2.2.5 Input Capacitor Selection (C1, C6)

The input voltage ripple is divided between capacitance and ESR. For this design, VRIPPLE(cap) = 400 mV and VRIPPLE(ESR) = 100 mV. The minimum capacitance and maximum ESR are estimated by:

$$C_{IN(min)} = \frac{I_{LOAD} \times V_{OUT}}{V_{RIPPLE(cap)} \times V_{IN} \times f_{SW}} = \frac{6\,A \times 5\,V}{400\,mV \times 10\,V \times 300\,kHz} = 25\,\mu F \quad (27)$$

$$ESR_{MAX} = \frac{V_{RIPPLE(esr)}}{I_{LOAD} + \tfrac{1}{2} \times I_{RIPPLE}} = \frac{100\,mV}{6.93A} = 14.4\,m\Omega \quad (28)$$

The rms current in the input capacitors is estimated in 式 29.

$$I_{RMS(cin)} = I_{LOAD} \times \sqrt{D \times (1 - D)} = 6\,A \times \sqrt{0.5 \times (1 - 0.5)} = 3.0\,A \quad (29)$$

To achieve these values, four 1210, 2.2-µF, 100-V, X7R ceramic capacitors plus a 120-µF electrolytic capacitor are combined at the input. This provides a smaller size and overall cost than 10 ceramic input capacitors or an electrolytic capacitor with the ESR required.

表 7-3. Inductor Summary

PARAMETER		VALUE	UNIT
L	Inductance	8.2	µH
$I_{L(rms)}$	RMS current (thermal rating)	6.02	A
$I_{L(peak)}$	Peak current (saturation rating)	7.01	A

7.2.2.6 MOSFET Switch Selection (Q1, Q2)

Using the J/K method for MOSFET optimization, apply 式 30 through 式 33.

High-side gate (Q1):

$$J = (10)^{-9} \times \left(\frac{V_{IN} \times I_{OUT}}{I_{DRIVE}} + \frac{Q_G}{Q_{SW}} \times V_{DRIVE} \right) \times f_{SW} \left(\frac{W}{nC} \right) \quad (30)$$

$$K = (10)^{-3} \left((I_{OUT})^2 + \frac{1}{12} \times (I_{P-P})^2 \right) \times \left(\frac{V_{OUT}}{V_{IN}} \right) \left(\frac{W}{m\Omega} \right) \quad (31)$$

Low-side gate (Q2):

$$K = (10)^{-3} \left((I_{OUT})^2 + \frac{1}{12} \times (I_{P-P})^2 \right) \times \left(1 - \frac{V_{OUT}}{V_{IN}} \right) \left(\frac{W}{m\Omega} \right) \quad (32)$$

$$J = 10^{-9} \left(\frac{V_{FD} \times I_{OUT}}{I_{DRIVE}} + \frac{Q_G}{Q_{SW}} \times V_{DRIVE} \right) \times f_{SW} \left(\frac{W}{nC} \right) \quad (33)$$

Optimizing for 300 kHz, 24-V input, 5-V output at 6 A, calculate ratios of 5.9 mΩ/nC and 0.5 mΩ/nC for the high-side and low-side FETS, respectively. BSC110N06NS2 (ratio 1.2) and BSC076N06NS3 (ratio 0.69) MOSFETS are selected.

7.2.2.7 Timing Resistor (R7)

The switching frequency is programmed by the current through R_{RT} to GND. The R_{RT} value is calculated using 式 34.

$$R_{RT} = \frac{(10)^4}{f_{SW}} - 2k\Omega = \frac{(10)^4}{300\,kHz} - 2 = 31.3k\Omega \approx 31.6k\Omega \quad (34)$$

7.2.2.8 UVLO Programming Resistors (R2, R6)

The UVLO hysteresis level is programmed by R2 using 式 35.

$$R_{UVLO(hys)} = \frac{V_{UVLO(on)} - V_{UVLO(off)}}{I_{UVLO}} = \frac{9\,V - 8\,V}{5.0\,\mu A} = 200k\Omega \quad (35)$$

$$R_{UVLO(set)} > R_{UVLO(hys)} \frac{V_{UVLO(max)}}{\left(V_{UVLO\_ON(min)} - V_{UVLO(max)} \right)} = 200k\Omega \frac{0.919\,V}{(9.0\,V - 0.919\,V)} = 22.7k\Omega \approx 22.1k\Omega \quad (36)$$

7.2.2.9 Bootstrap Capacitor (C7)

To ensure proper charging of the high-side FET gate, limit the ripple voltage on the boost capacitor to less than 250 mV.

$$C_{BOOST} = \frac{Q_{G1}}{V_{BOOT(ripple)}} = \frac{25\,nC}{250\,mV} = 100\,nF \quad (37)$$

7.2.2.10 VIN Bypass Capacitor (C18)

Place a capacitor with a value of 1 µF. Select a capacitor with a value between 0.1 µF and 1.0 µF, X5R or better ceramic bypass capacitor for VIN as specified in 表 7-2 . For this design, a 1.0-µF, 100 V, X7R capacitor has been selected.

7.2.2.11 VBP Bypass Capacitor (C19)

Select a capacitor with a value between 1 µF and 10 µF, X5R or better ceramic bypass capacitor for BP as specified in 表 7-2. For this design a 4.7-µF, 16 V capacitor has been selected.

7.2.2.12 SS Timing Capacitor (C15)

The soft-start capacitor provides a smooth ramp of the error-amplifier reference voltage for controlled start-up. The soft-start capacitor is selected by using 式 38.

$$C_{SS} = \frac{t_{SS}}{0.09} = \frac{4\,ms}{0.09} = 44\,nF \approx 47\,nF \tag{38}$$

7.2.2.13 ILIM Resistor (R19, C17)

The TPS40170-Q1 controller uses the negative drop across the low-side FET at the end of the OFF time to measure the inductor current. Allowing for 30% over the minimum current limit for transient recovery and a 20% rise in $R_{DS(on)Q2}$ for self-heating of the MOSFET, the voltage drop across the low-side FET at the current limit is given by 式 39.

$$V_{OC} = \left(\left(1.3 \times I_{OCP(min)}\right) + \left(\frac{1}{2} \times I_{RIPPLE}\right)\right) \times 1.25 \times R_{DS(on)G2} = \left(1.3 \times 8\,A + \frac{1}{2} \times 1.86\,A\right) \times 1.25 \times 7.6\,m\Omega = 107.6\,mV \tag{39}$$

The internal current-limit temperature coefficient helps compensate for the MOSFET $R_{DS(on)}$ temperature coefficient, so the current-limit programming resistor is selected by 式 40.

$$R_{ILIM} = \frac{V_{OC}}{I_{OCSET(min)}} = \frac{107.6\,mV}{9.0\,\mu A} = 12.0\,k\Omega \approx 12.1\,k\Omega \tag{40}$$

A 1000-pF capacitor is placed in parallel to improve noise immunity of the current-limit set-point.

7.2.2.14 SCP Multiplier Selection (R5)

The TPS40170-Q1 controller uses a multiplier (A_{OC}) to translate the low-side overcurrent protection into a high-side $R_{DS(on)}$ pulse-by-pulse short-circuit protection. Ensure that 式 41 is true.

$$A_{OC} > \frac{I_{OCP(min)} + \left(\frac{1}{2} \times I_{RIPPLE}\right)}{I_{OCP(min)} + \left(\frac{1}{2} \times I_{RIPPLE}\right)} \times \frac{R_{DS(on)Q1}}{R_{DS(on)Q2}} = \frac{8\,A + \frac{1}{2} \times 1.86\,A}{8\,A + \frac{1}{2} \times 1.86\,A} \times \frac{11\,m\Omega}{7.6\,m\Omega} = 1.45 \tag{41}$$

A_{OC} = 3 is selected as the next-greater A_{OC}. The value of R5 is set to 10 kΩ.

7.2.2.15 Feedback Divider (R10, R11)

The TPS40170-Q1 controller uses a full operational amplifier with an internally fixed 0.6-V reference. The value of R11 is selected between 10 kΩ and 50 kΩ for a balance of feedback current and noise immunity. With the value of R11 set to 20 kΩ, the output voltage is programmed with a resistor divider given by 式 42.

$$R10 = \frac{V_{FB} \times R11}{\left(V_{OUT} - V_{FB}\right)} = \frac{0.600\,V \times 20.0\,k\Omega}{\left(5.0\,V - 0.600\,V\right)} = 2.73\,k\Omega \approx 2.74\,k\Omega \tag{42}$$

7.2.2.16 Compensation: (R4, R13, C13, C14, C21)

Using the TPS40k Loop Stability Tool for a 60-kHz bandwidth and a 50° phase margin with an R10 value of 20 kΩ, the following values are obtained. The tool is available from the TI Web site, Literature Number SLUC263.

- C21 = C1 = 1500 pF
- C13 = C2 = 8200 pF
- C14 = C3 = 220 pF
- R13 = R2 = 511 Ω
- R4 = R3 = 3.83 kΩ

7.2.3 Application Curves

図 7-2 shows an efficiency graph for this design with 10-V to 60-V input and 5-V at 6-A output. 図 7-3 shows a 24-V to 5-V at 6-A loop response, where V_{IN} = 24 V and I_{OUT} = 6 A, yielding 58-kHz bandwidth, 51° phase margin. 図 7-4 shows the output ripple 20 mV/div, 2 µs/div, 20 MHz bandwidth.

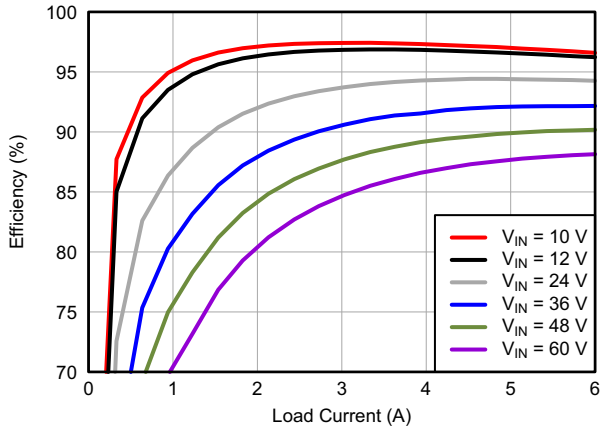


図 7-2. Efficiency vs Load Current

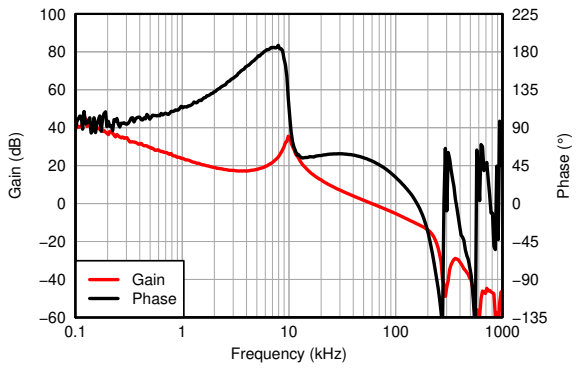


図 7-3. Loop Response

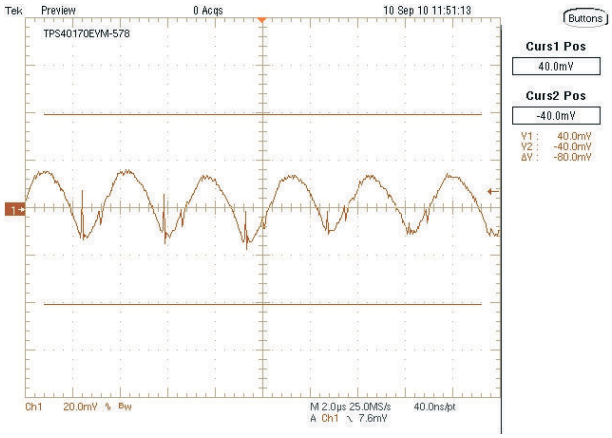


図 7-4. Output Ripple Waveform

7.3 Power Supply Recommendations

7.3.1 Bootstrap Resistor

A small resistor in series with the bootstrap capacitor reduces the turnon time of the internal MOSFET, thereby reducing the rising edge ringing of the SW node and reducing shoot-through induced by dv/dt. A bootstrap resistor value that is too large delays the turnon time of the high-side switch and can trigger an apparent SCP fault. See the *Design Examples* section.

7.3.2 SW-Node Snubber Capacitor

Observable voltage ringing at the SW node is caused by fast switching edges and parasitic inductance and capacitance. If the ringing results in excessive voltage on the SW node, or erratic operation of the converter, an R-C snubber can be used to dampen the ringing and ensure proper operation over the full load range. See the *Design Examples* section.

7.3.3 Input Resistor

The TPS40170-Q1 device has a wide input-voltage range, which allows for the device input to share a power source with the power-stage input. Power-stage switching noise can pollute the device power source if the layout is not adequate in minimizing noise. Power-stage switching noise can trigger a short-circuit fault. If so, adding a small resistor between the device input and power-stage input is recommended. This resistor, together with the device input capacitor, composes an R-C filter that filters out the switching noise from power stage. See the *Design Examples* section.

7.3.4 LDRV Gate Capacitor

Power-device selection is important for proper switching operation. If the low-side MOSFET has low gate capacitance Cgs (if Cgs < Cgd), there is a risk of short-through induced by high dv/dt at the switching node during high-side turnon. If this happens, add a small capacitance between LDRV and GND. See the *Design Examples* section.

7.4 Layout

7.4.1 Layout Guidelines

図 7-5 illustrates an example layout. For the controller, it is important to connect carefully noise-sensitive signals such as RT, SS, FB, and COMP as close to the IC device as possible and connect to AGND as shown. The thermal pad must be connected to any internal PCB ground planes using multiple vias directly underneath the IC device. AGND and PGND must be connected at a single point.

High-performance FETs such as NexFET™ power MOSFETs from Texas Instruments, require careful attention to the layout. Minimize the distance between the positive node of the input ceramic capacitor and the drain pin of the control (high-side) FET. Minimize the distance between the negative node of the input ceramic capacitor and the source pin of the synchronize (low-side) FET. Because of the large gate drive, smaller gate charge, and faster turn-on times of the high-performance FETs, use a minimum of four 10-µF ceramic input capacitors such as TDK #C3216X5R1A106M. Ensure the layout allows a continuous flow of the power planes.

The layout of the HPA578 EVM is shown in 図 7-5 through 図 7-8 for reference.

7.4.2 Layout Example

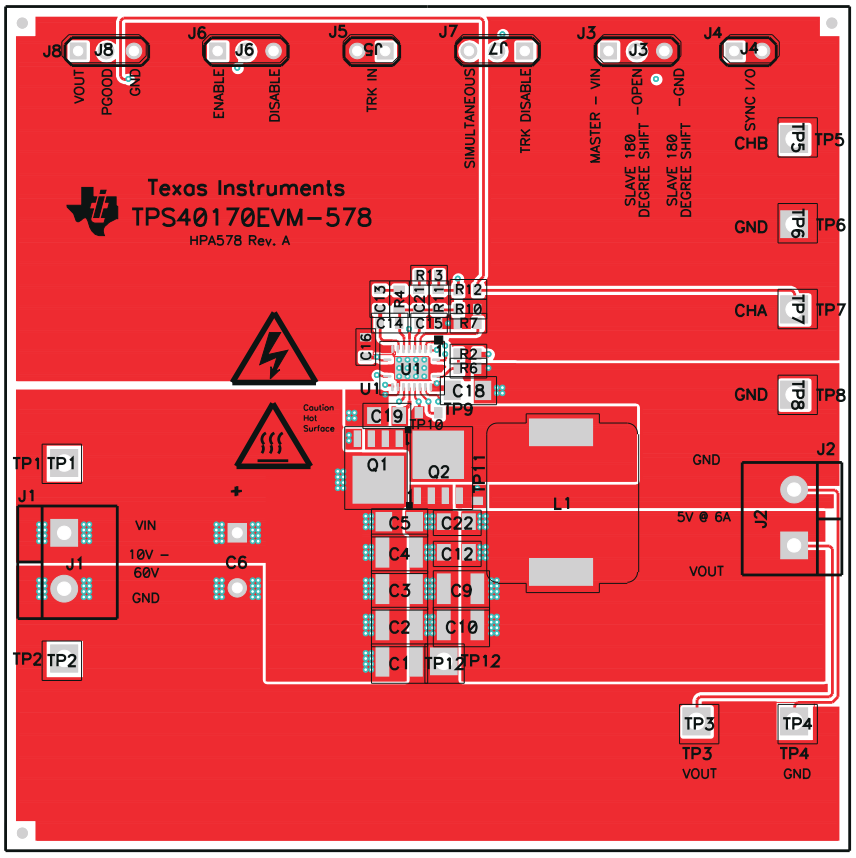


図 7-5. Top Copper, Viewed From Top

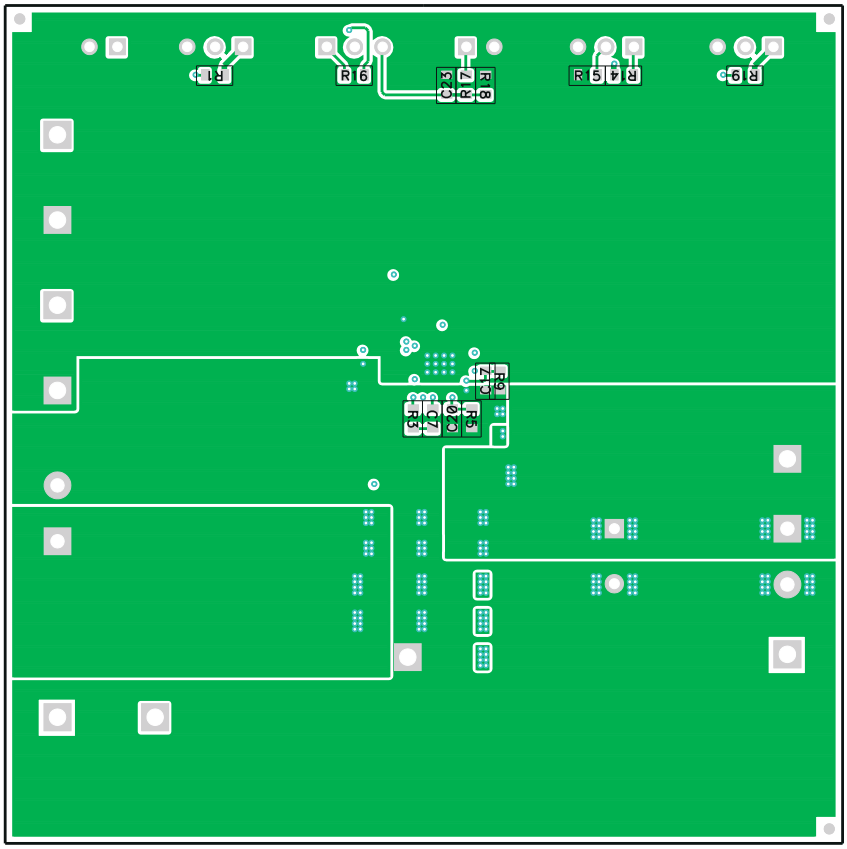


図 7-6. Bottom Copper, Viewed From Bottom

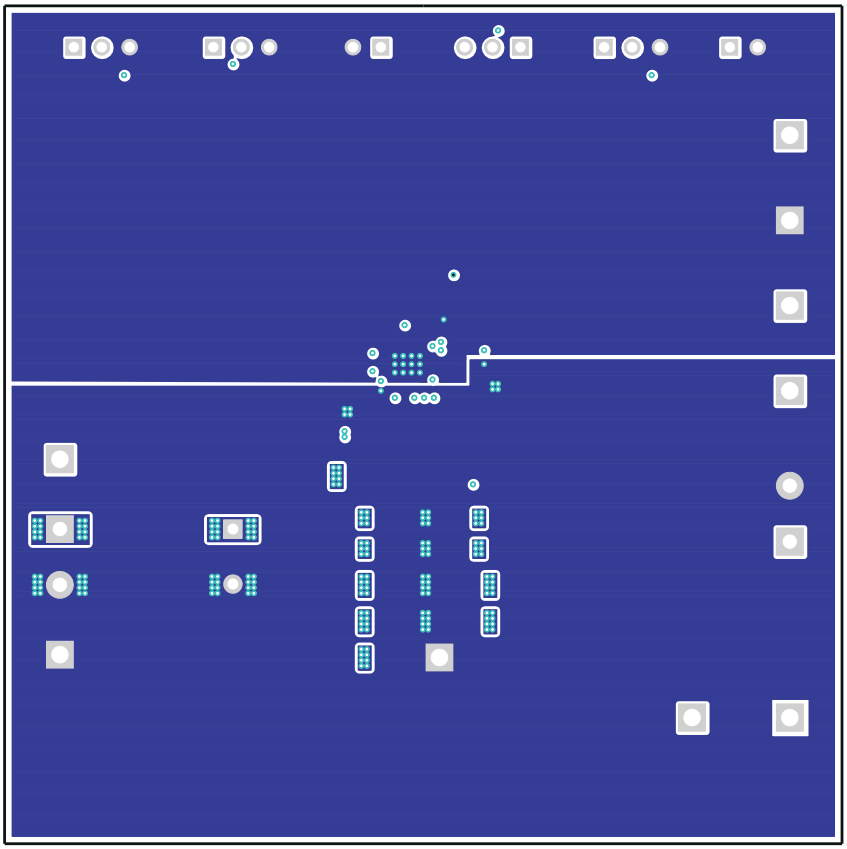


図 7-7. Internal Layer 1, Viewed From Top

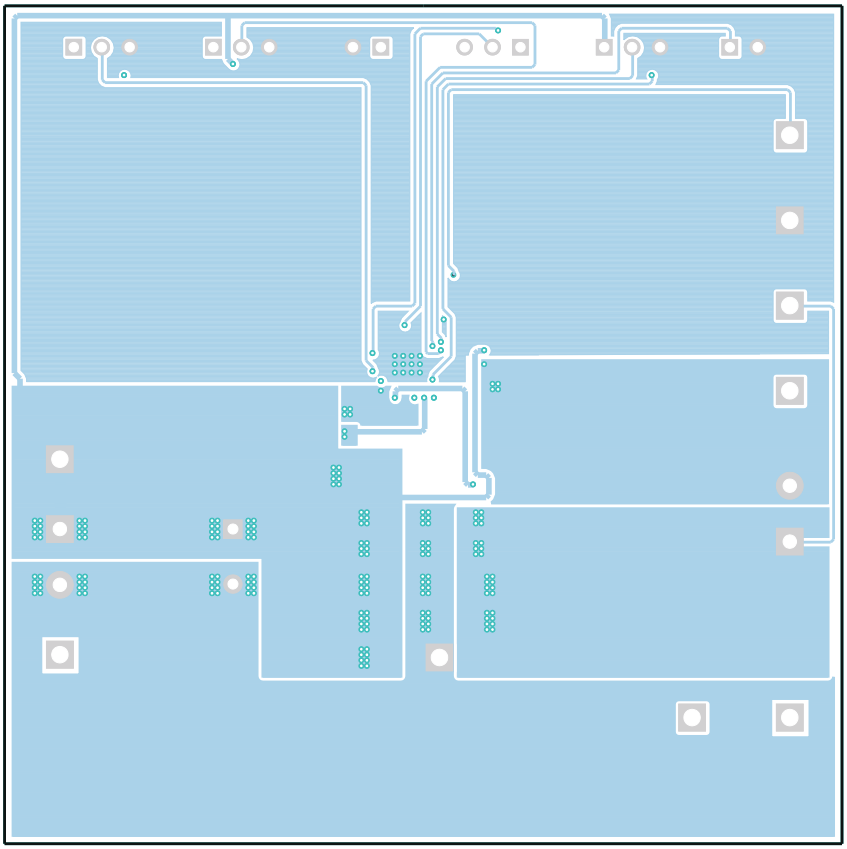


図 7-8. Internal Layer 2, Viewed From Top

8 Device and Documentation Support

8.1 Device Support

8.1.1 サード・パーティ製品に関する免責事項

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8.2 Documentation Support

8.2.1 Related Documentation

For related documentation see the following:

- Steve Mappus, *DV/DT Immunity Improved in Synchronous Buck Converters*. July, 2005, Power Electronics Technology.
- Texas Instruments, *TPS4005x Wide-Input Synchronous Buck Controller* data sheet
- Texas Instruments, *TPS40k Loop Stability Tool*

8.3 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、www.tij.co.jp のデバイス製品フォルダを開いてください。[通知] 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ることができます。 変更の詳細については、改訂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。

8.4 サポート・リソース

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ESD による破損は、わずかな性能低下からデバイスの完全な故障まで多岐にわたります。精密な IC の場合、パラメータがわずかに変化するだけで公表されている仕様から外れる可能性があるため、破損が発生しやすくなっています。

8.7 用語集

テキサス・インスツルメンツ用語集　　この用語集には、用語や略語の一覧および定義が記載されています。

9 Revision History

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。

Changes from Revision B (December 2014) to Revision C (November 2023)	Page
• 「特長」セクションに新しい類似製品の導入を追加	1
• データシートのタイトルに「車載用」を追加	1
• ドキュメント全体にわたって表、図、相互参照の採番方法を更新	1
• 「概要」セクションに新しい類似製品の導入を追加	1
• 本体サイズからパッケージ・サイズに変更し、「パッケージ情報」表に表の注を追加	1

Changes from Revision A (March 2012) to Revision B (December 2014)	Page
• Added *Handling Rating* table, *Feature Description* section, *Device Functional Modes*, *Application and Implementation* section, *Power Supply Recommendations* section, *Layout* section, *Device and Documentation Support* section, and *Mechanical, Packaging, and Orderable Information* section	5

Changes from Revision * (January 2012) to Revision A (March 2012)	Page
• Changed R_{HDHI}, RHDLO and R_{LDLO} MAX values	6
• Changed I_{ILIM} and $I_{ILIM(ss)}$ values	6

10 Mechanical, Packaging, and Orderable Information

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

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)
TPS40170QRGYRQ1	Active	Production	VQFN (RGY) | 20	3000 | LARGE T&R	Yes	NIPDAU	Level-3-260C-168 HR	-40 to 125	PXXQ
TPS40170QRGYRQ1.A	Active	Production	VQFN (RGY) | 20	3000 | LARGE T&R	Yes	NIPDAU	Level-3-260C-168 HR	-40 to 125	PXXQ

(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.

OTHER QUALIFIED VERSIONS OF TPS40170-Q1 :

- Catalog : TPS40170

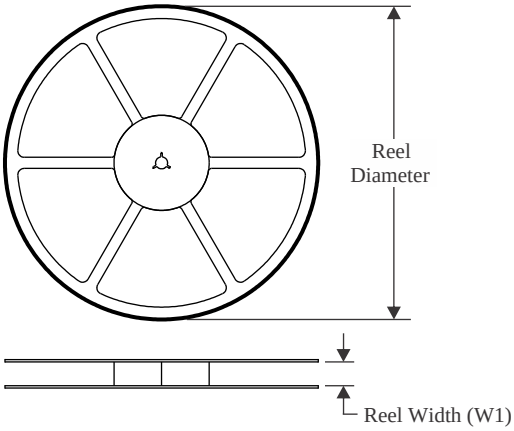
- Enhanced Product : TPS40170-EP

NOTE: Qualified Version Definitions:

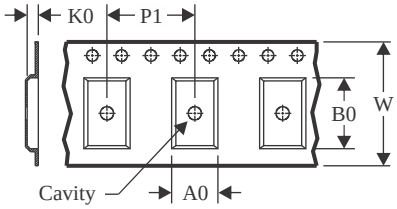
- Catalog - TI's standard catalog product
- Enhanced Product - Supports Defense, Aerospace and Medical Applications

TAPE AND REEL INFORMATION

REEL DIMENSIONS

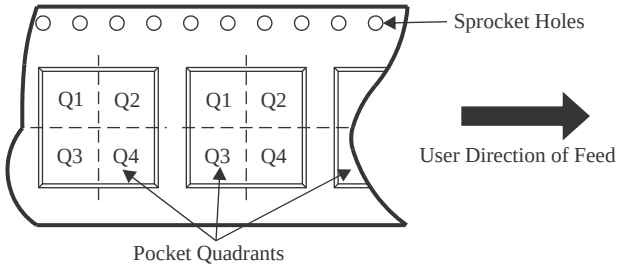


TAPE DIMENSIONS



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

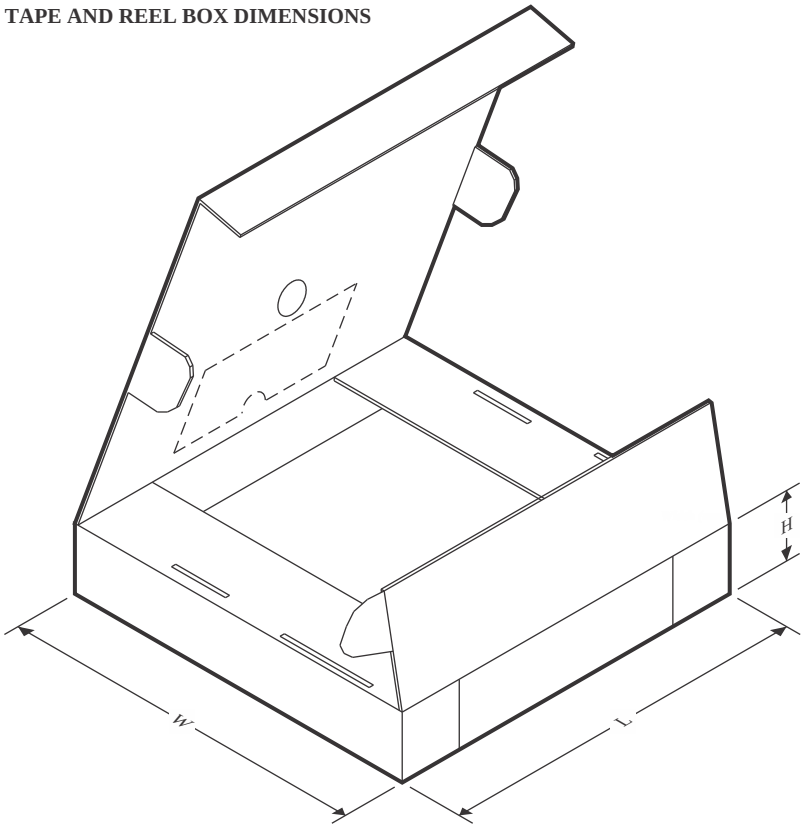
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE



*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
TPS40170QRGYRQ1	VQFN	RGY	20	3000	330.0	12.4	3.71	4.71	1.1	8.0	12.0	Q1

TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS40170QRGYRQ1	VQFN	RGY	20	3000	353.0	353.0	32.0

GENERIC PACKAGE VIEW

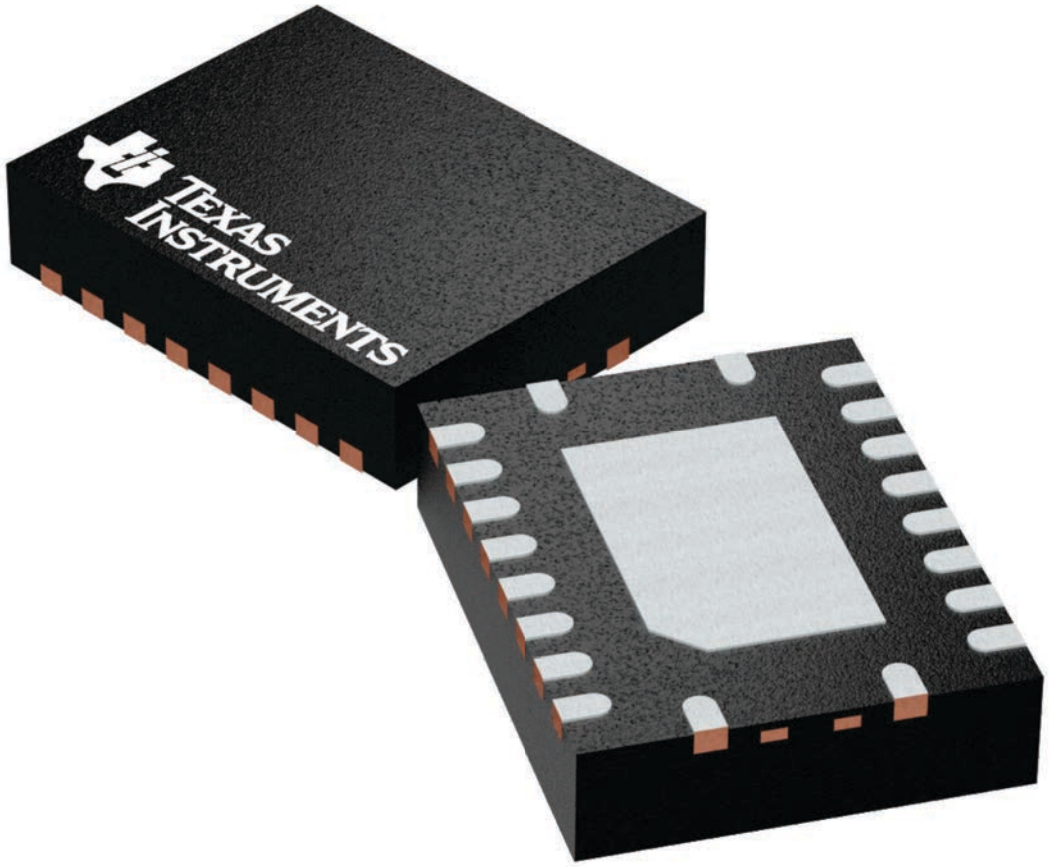
RGY 20

VQFN - 1 mm max height

3.5 x 4.5, 0.5 mm pitch

PLASTIC QUAD FGLATPACK - NO LEAD

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



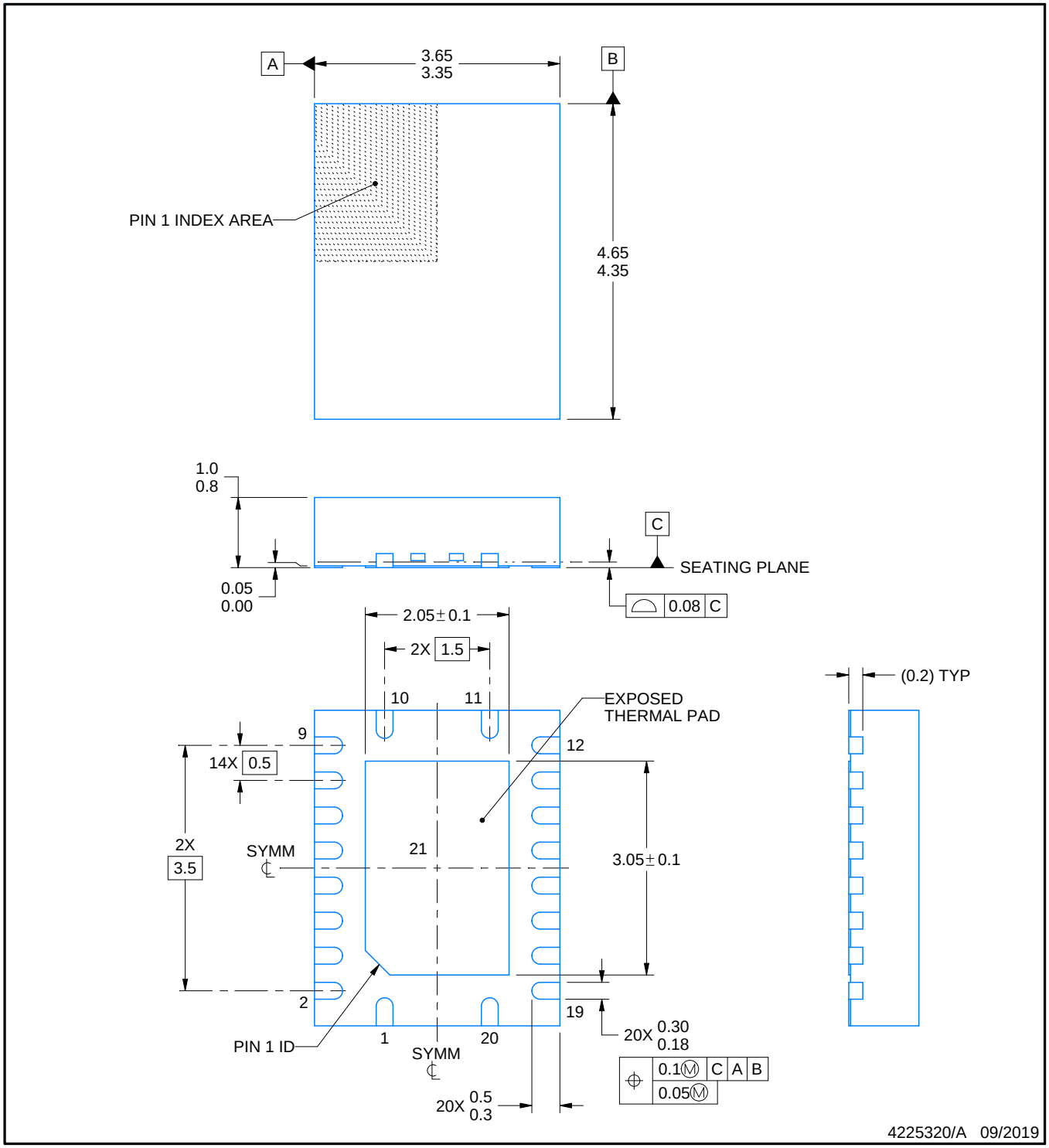
4225264/A

PACKAGE OUTLINE

RGY0020A

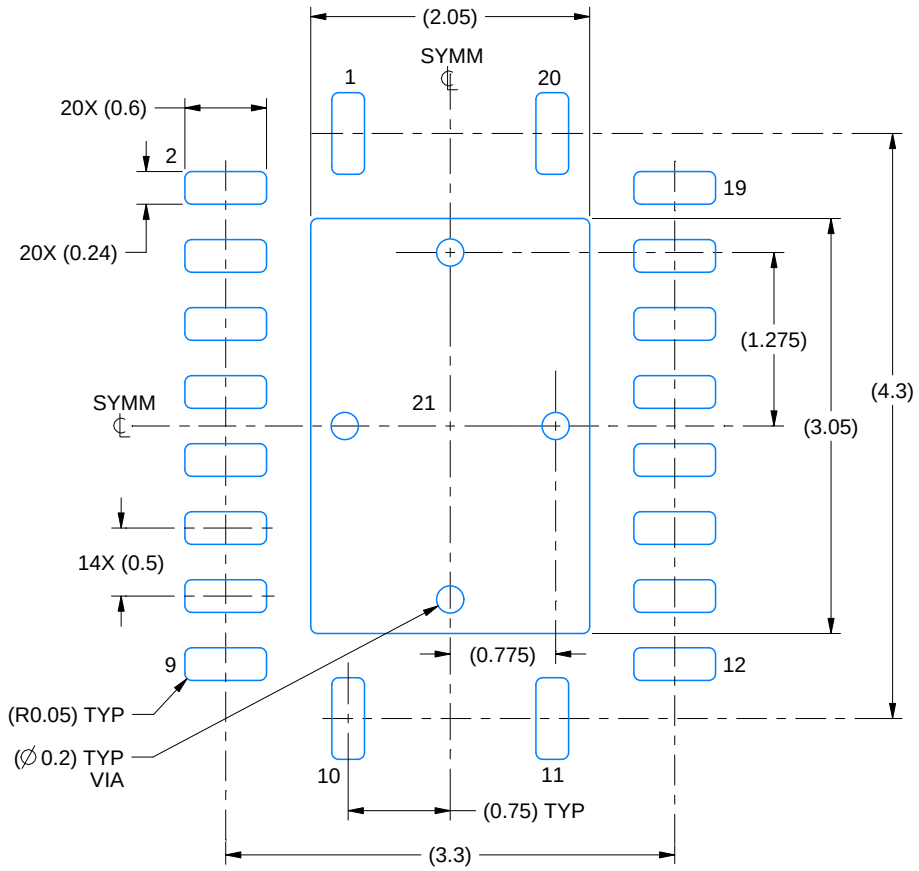
VQFN - 1 mm max height

PLASTIC QUAD FLATPACK - NO LEAD

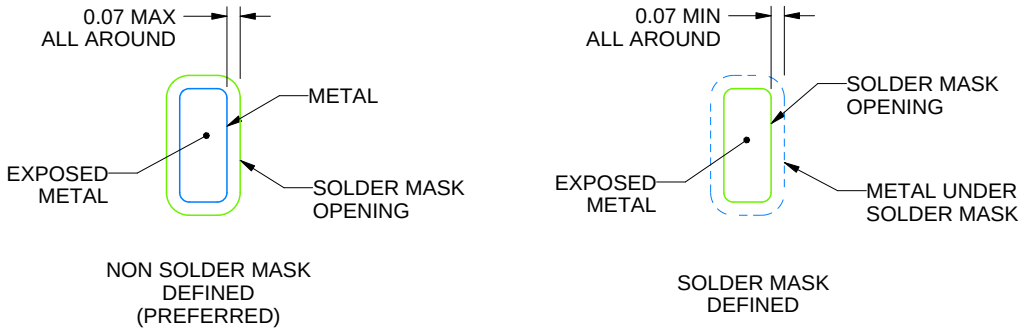


NOTES:

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
EXPOSED METAL SHOWN
SCALE:18X

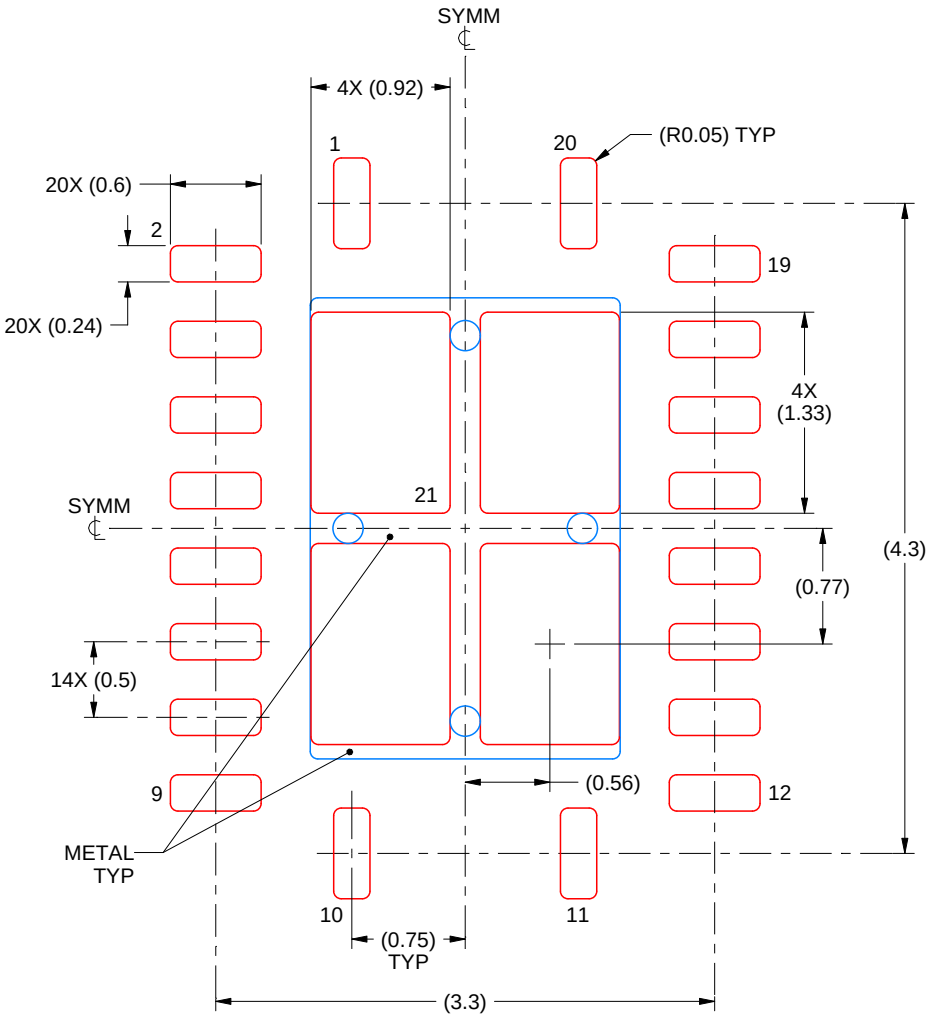


SOLDER MASK DETAILS

4225320/A 09/2019

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 21
78% PRINTED SOLDER COVERAGE BY AREA UNDER PACKAGE
SCALE:20X

4225320/A 09/2019

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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最終更新日：2025 年 10 月