TEXAS INSTRUMENTS

**TPS40170-Q1**
ZHCS826C – JANUARY 2012 – REVISED NOVEMBER 2023

# TPS40170-Q1 汽车类 4.5V 至 60V 宽输入同步 PWM 降压控制器

## 1 特性

- 可提供新款类似产品：
  - LM5145-Q1 宽输入电压和占空比范围 75V 同步降压控制器
  - LM5146-Q1 宽占空比范围 100V 同步直流/直流降压控制器
- 符合汽车应用要求
- 具有符合 AEC-Q100 标准的下列特性：
  - 器件温度等级 1：-40°C 至 125°C 的环境工作温度范围
  - 器件 HBM ESD 分级等级 1C
  - 器件 CDM ESD 分类等级 C4B
- 4.5V 至 60V 宽输入电压范围
- 600mV 基准电压，精确度 1%
- 可编程 UVLO 与迟滞功能
- 带有电压前馈的电压模式控制
- 可编程频率范围：100kHz 至 600kHz
- 可通过主要或辅助选项实现双向频率同步
- 提供基于低侧 FET 检测的过流保护与基于高侧 FET 检测的短路保护，具有集成式热补偿功能
- 可编程闭环软启动
- 支持预偏置输出
- 165°C 热关断，具有迟滞功能
- 电压跟踪
- 电源正常状态指示
- 支持 1µA 低电流关断
- 8V 与 3.3V LDO 输出
- 集成式自举二极管
- 20 引脚 4.5mm × 3.5mm VQFN (RGY) 封装

## 2 应用

- 高级驾驶员辅助系统 (ADAS)
- 汽车信息娱乐系统与仪表组

## 3 说明

TPS40170-Q1 器件是一款功能齐全的同步 PWM 降压控制器，工作输入电压介于 4.5V 至 60V 之间，并针对高功率密度、高可靠性的直流/直流转换器应用进行了优化。该控制器通过输入电压前馈补偿实施电压模式控制，可在输入电压变化时立即做出响应。开关频率可在 100kHz 至 600kHz 之间进行编程。

TPS40170-Q1 器件具有整套系统保护和监控特性，例如可编程 UVLO、通过检测低侧 FET 实现的可编程过流保护 (OCP)、通过检测高侧 FET 实现的可选短路保护 (SCP) 和热关断。启用引脚可在低电流（典型值 1µA）模式下实现系统关断。控制器支持预偏置输出，提供一个开漏电源良好 (PGOOD) 信号，并支持闭环软启动、输出电压跟踪以及自适应死区时间控制。

TPS40170-Q1 器件提供精度为 1% 的精确输出稳压。此外，该控制器还采用新型双向同步方案，即一个控制器作为主要控制器，其他下游控制器作为辅助控制器，辅助控制器可与主要控制器同相或 180° 异相同步。辅助控制器可在自由运行开关频率的 ±30% 范围内与外部时钟同步。

新款产品（LM5145-Q1 和 LM5146-Q1）具有 BOM 成本低、效率高、设计尺寸小等诸多特性。

**器件信息**

| 器件型号 | 封装[1] | 封装尺寸[2] |
|---|---|---|
| TPS40170-Q1 | RGY（VQFN，20） | 4.50mm × 3.50mm |

(1) 有关所有可选封装，请参阅 节 10。
(2) 封装尺寸（长 × 宽）为标称值，并包括引脚（如适用）。



简化版原理图

English Data Sheet: SLVSB90

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## 4 Pin Configuration and Functions
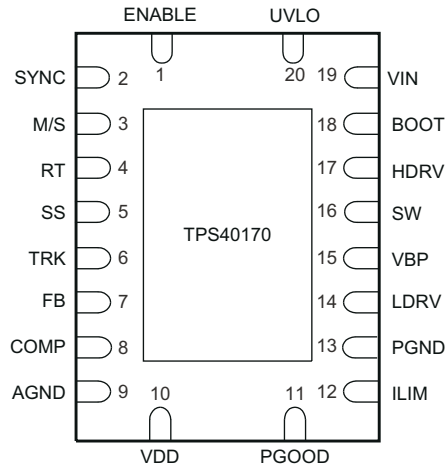
图 4-1. RGY PACKAGE QFN-20 (Top View)

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. |

提交文档反馈

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表 4-1. Pin Functions（续）

| 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

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# 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 | |
| | | Charged device model (CDM), per AEC Q100-011 — 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 | UNIT |
|---|---|---|---|
| | | RGY | |
| | | 20 PINS | |
| $R_{\theta JA}$ | 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 | |
| $\psi_{JT}$ | Junction-to-top characterization parameter | 0.5 | |
| $\psi_{JB}$ | 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 < $V_{VIN}$ ⩽ 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** | | | | | | |

Product Folder Links: *TPS40170-Q1*

## 5.5 Electrical Characteristics （续）

These specifications apply for －40°C $\leqslant$ $T_A$ $\leqslant$ +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 k$\Omega$, 4.5 V < $V_{VIN}$ $\leqslant$ 60 V | 90 | 100 | 110 | |
| | | $R_{RT}$ = 31.6 k$\Omega$, 4.5 V < $V_{VIN}$ $\leqslant$ 60 V | 270 | 300 | 330 | |
| | | $R_{RT}$ = 14.3 k$\Omega$, 4.5 V < $V_{VIN}$ $\leqslant$ 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 < $V_{VIN}$ $\leqslant$ 60 V | 14 | 15 | 16 | V/V |

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## 5.5 Electrical Characteristics （续）

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 |

Product Folder Links: *TPS40170-Q1*

English Data Sheet: *SLVSB90*

## 5.5 Electrical Characteristics （续）

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 | | |
| **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 | |
| 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 | |
| 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 | |
| V$_{PG,HYST}$ | Power good hysteresis voltage at FB pin | | 8.5 | 20 | 32 | |
| 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 |

Product Folder Links: *TPS40170-Q1*

English Data Sheet: SLVSB90

## 5.5 Electrical Characteristics （续）

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 | | M$\Omega$ |

(1)     Not production tested.

Product Folder Links: *TPS40170-Q1*

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## 5.6 Typical Characteristics
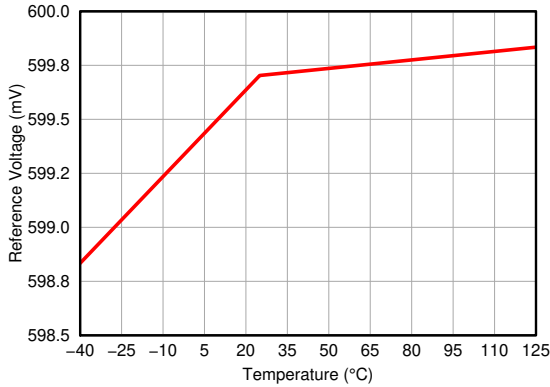


图 5-1. Reference Voltage vs Junction Temperature
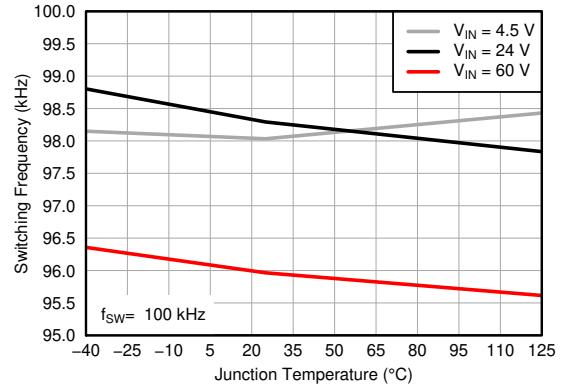


图 5-2. Switching Frequency vs Junction Temperature
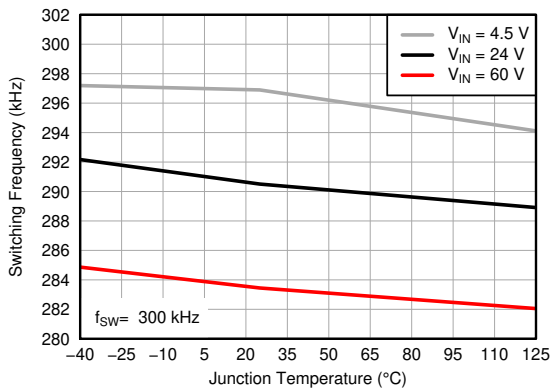


图 5-3. Switching Frequency vs Junction Temperature
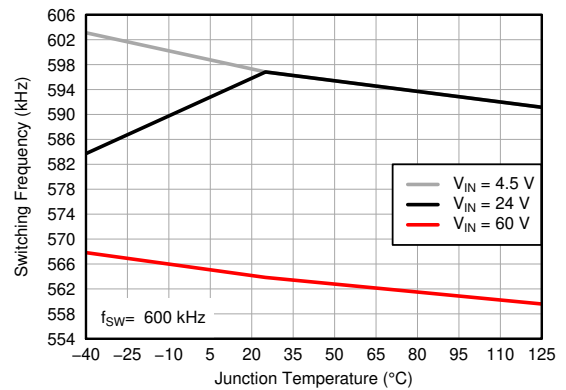


图 5-4. Switching Frequency vs Junction Temperature
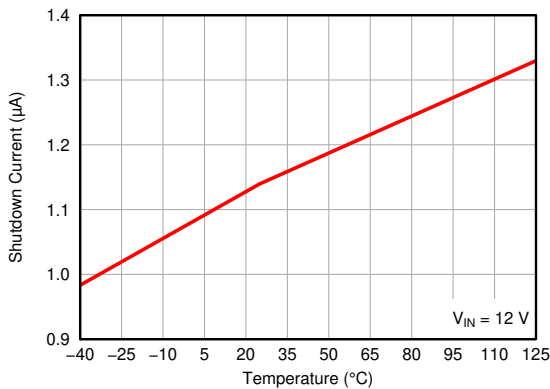


图 5-5. Shutdown Current vs Junction Temperature
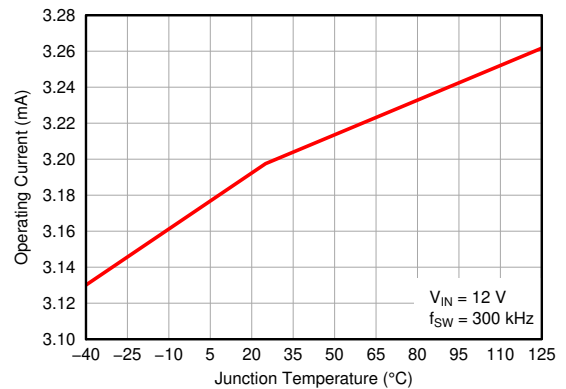


图 5-6. Operating Current vs Junction Temperature

Product Folder Links: *TPS40170-Q1*

English Data Sheet: *SLVSB90*

## 5.6 Typical Characteristics (continued)
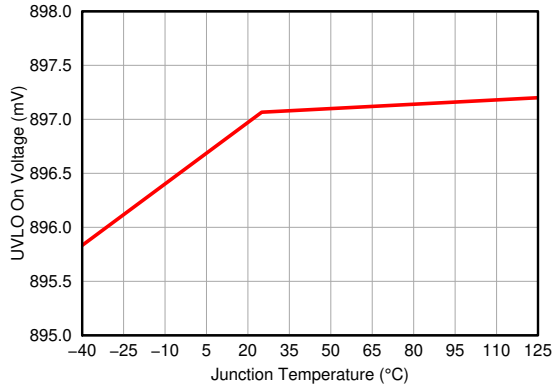


图 5-7. UVLO On Voltage vs Junction Temperature
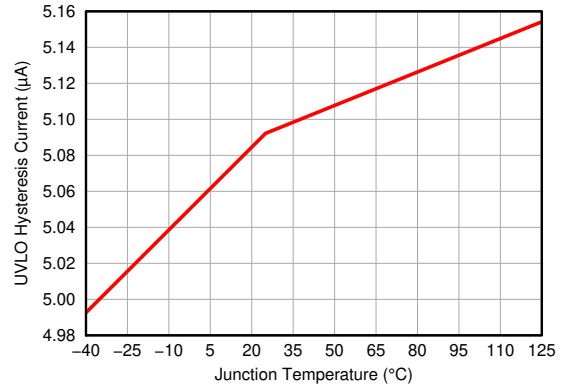
图 5-8. UVLO Hysteresis Current vs Junction Temperature
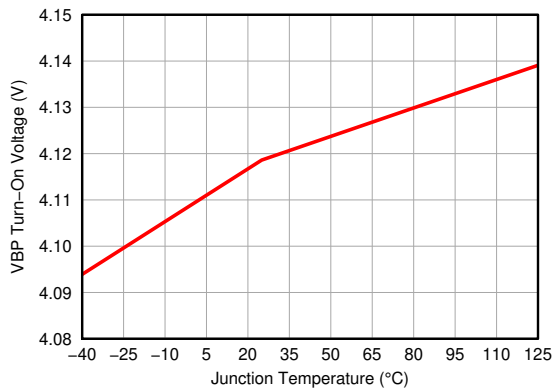
图 5-9. VBP Turnon Voltage vs Junction Temperature
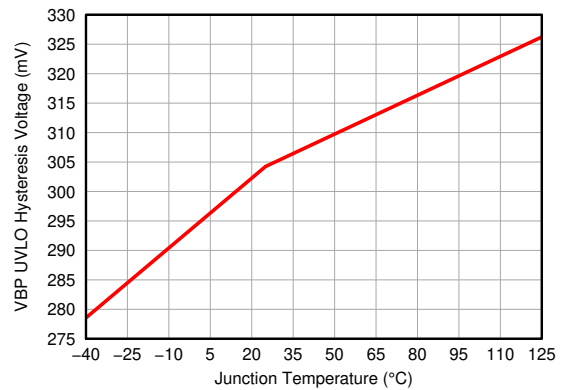
图 5-10. VBP UVLO Hysteresis Voltage
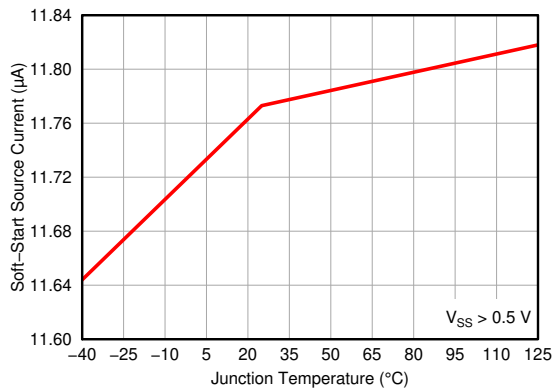
图 5-11. Soft-Start Source Current vs Junction Temperature
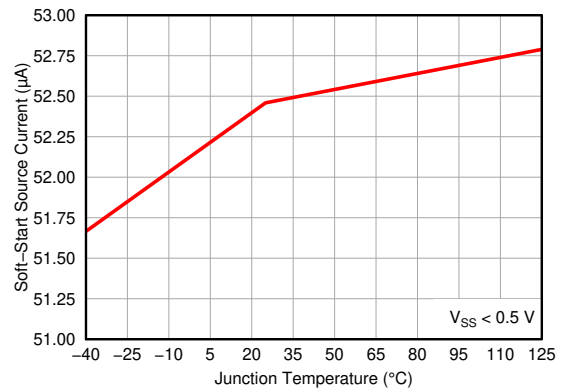
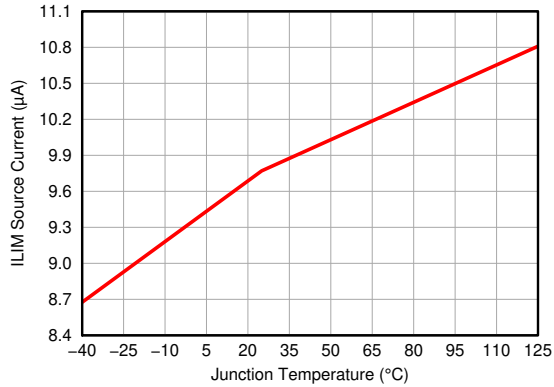图 5-12. Soft-Start Source Current vs Junction Temperature

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English Data Sheet: SLVSB90

## 5.6 Typical Characteristics (continued)



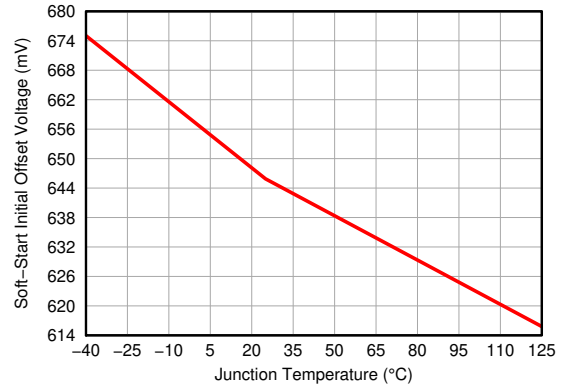图 5-13. ILIM Source Current vs Junction Temperature



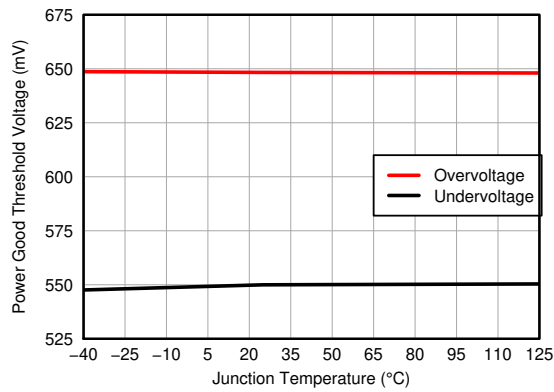图 5-14. Soft-Start Initial Offset Voltage vs Junction Temperature



图 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.

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## 6.2 Functional Block Diagram
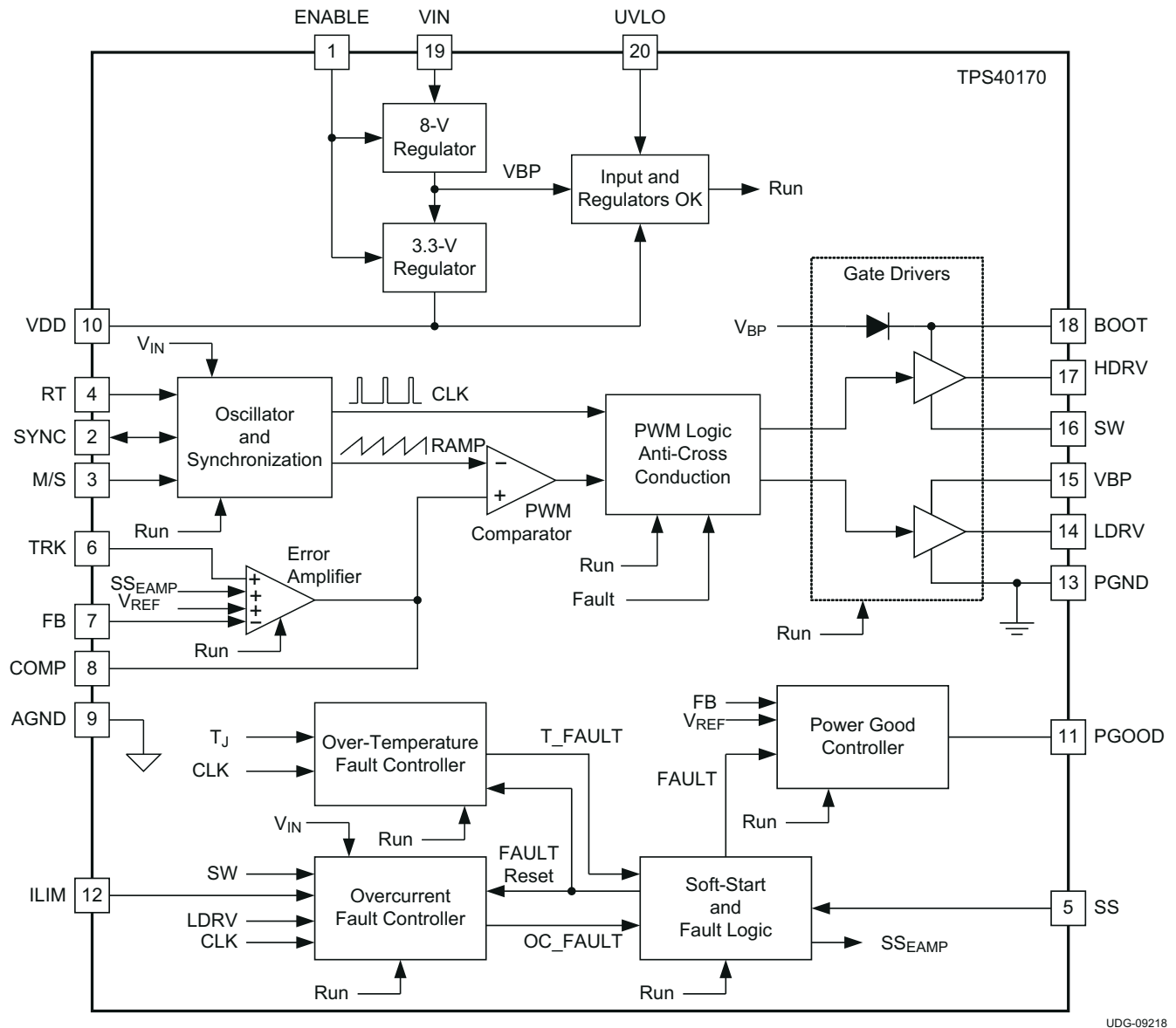


UDG-09218

## 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 图 6-1.
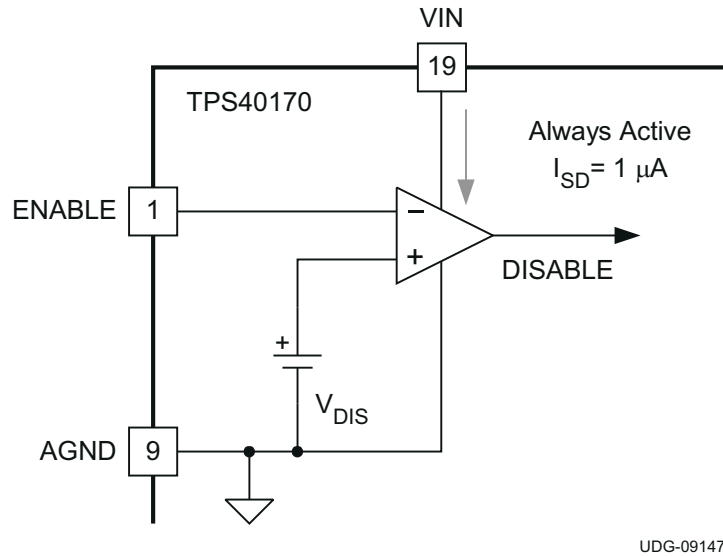


**图 6-1. TPS40170-Q1 ENABLE Functional Block**

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 图 6-2.
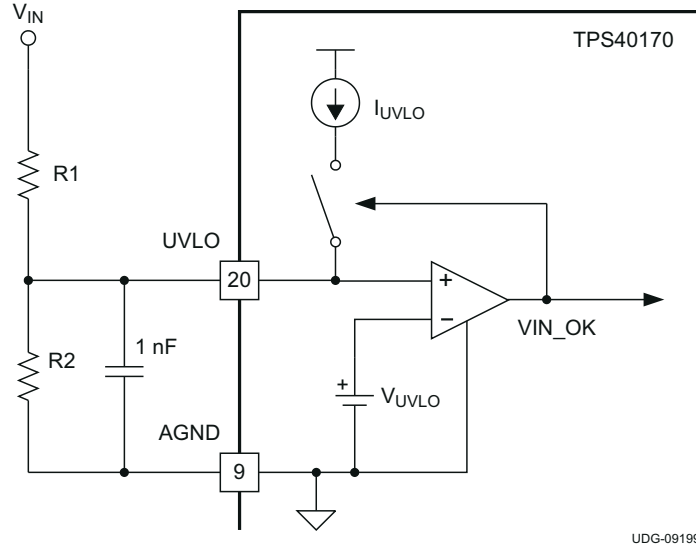
Product Folder Links: *TPS40170-Q1*

English Data Sheet: SLVSB90

图 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 方程式 1 and 方程式 2, respectively.

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

(1)

$$R_2 = R_1 \times \frac{V_{UVLO}}{\left(V_{ON} - V_{UVLO}\right)}$$

(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).

---

备注

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 方程式 3.

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

(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 方程式 4.

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

(4)

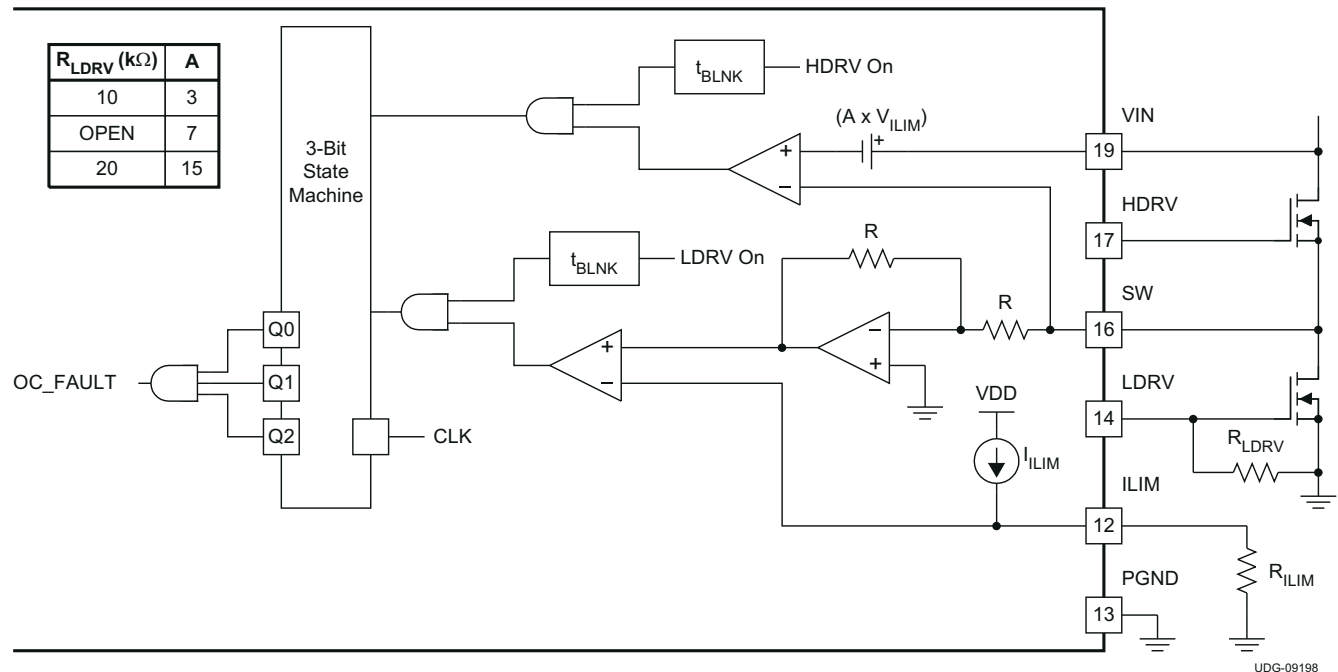图 6-3 is a functional block diagram of the two-level overcurrent protection.



**图 6-3. OCP and SCP Protection Functional Block Diagram**

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备注

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. 图 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 方程式 5).

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

(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.
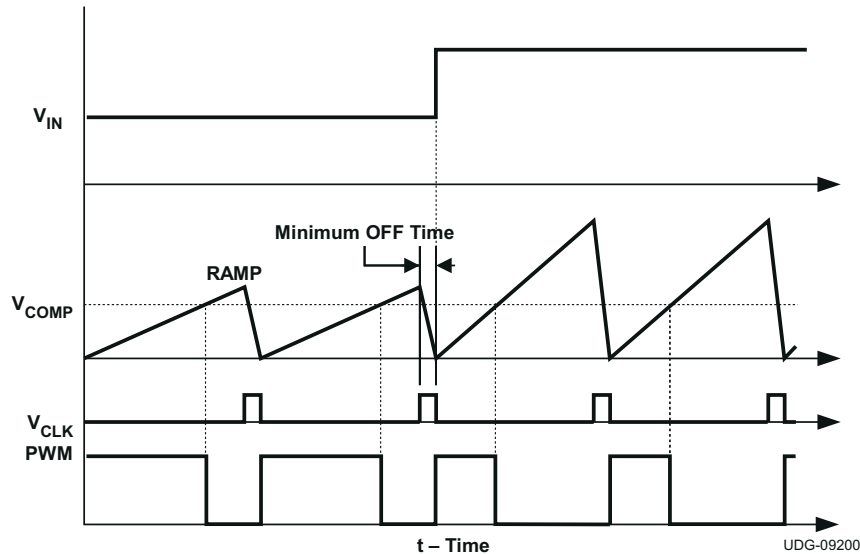


图 6-4. Feed-Forward Oscillator Timing Diagram

#### 6.3.5.1 Calculating the Timing Resistance ($R_{RT}$)

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

(6)

where

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

### 6.3.6 Feed-Forward Oscillator Timing Diagram

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备注

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 方程式 7 to determine the maximum switching frequency. For further details, see application note SLYT293.

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

(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. 图 6-5 shows the soft-start function block.

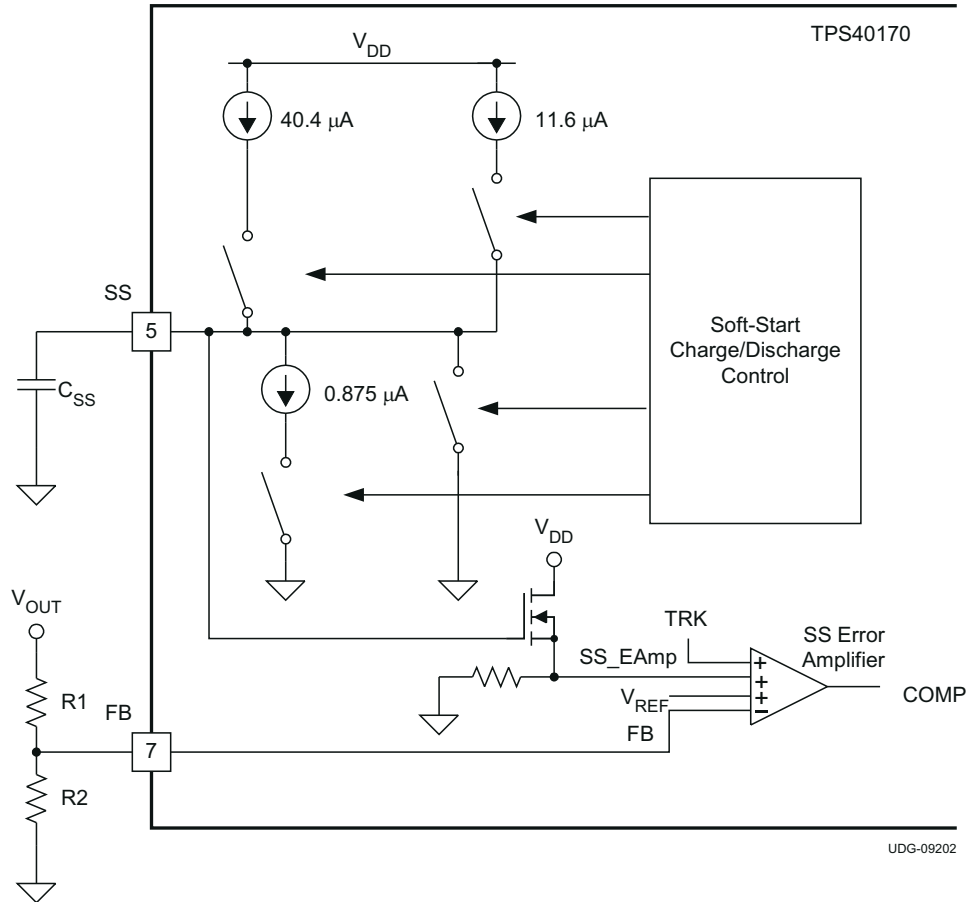Product Folder Links: *TPS40170-Q1*

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图 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 图 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 节 6.3.7.1 section.
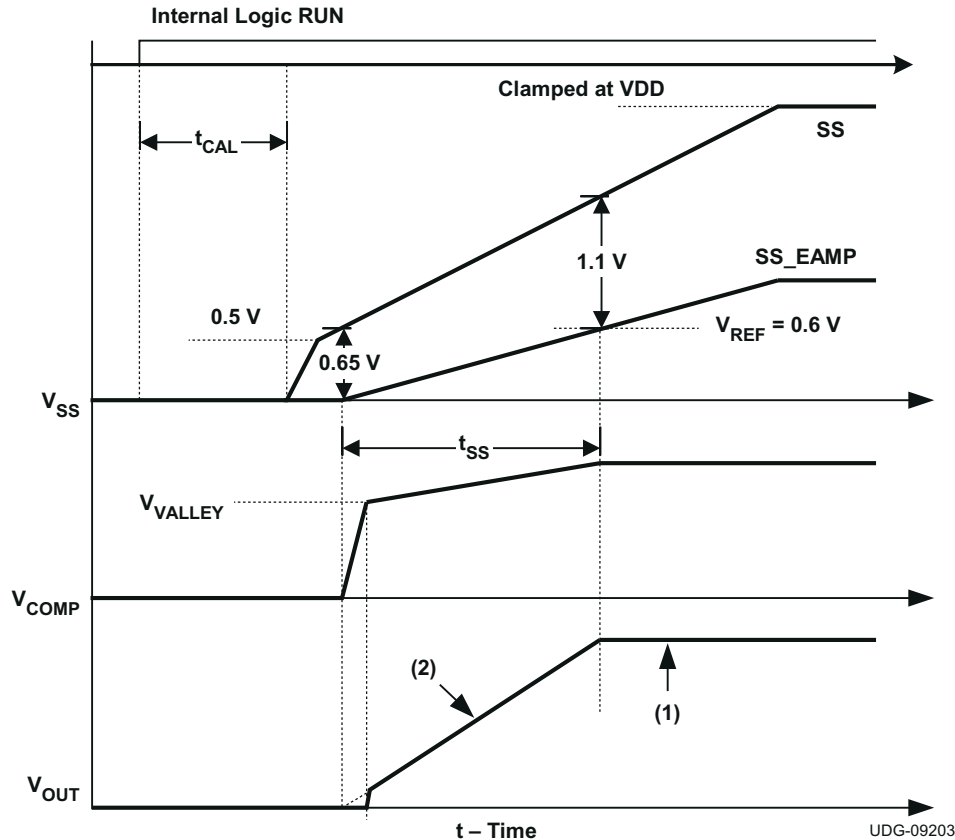
图 **6-6. Soft-Start Waveforms**

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备注

Referring to 图 6-6:

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

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

---

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

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

(8)

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

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

(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

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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. 图 6-7 shows the restart timing.
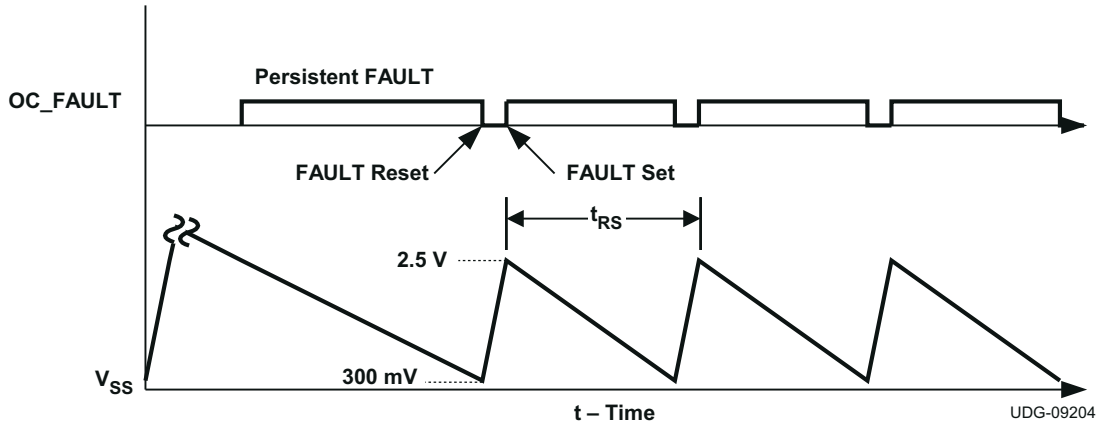


**图 6-7. Overcurrent Fault Restart Timing**

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**备注**

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.

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### 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. 图 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 方程式 10).

$$C_{SS} = \frac{t_{SS}}{0.09} \qquad (10)$$

The restart time ($t_{RS}$) is defined in 方程式 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.

$$t_{RS} \approx 2.28 \times C_{SS} \qquad (11)$$

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

---

**备注**

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

图 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.
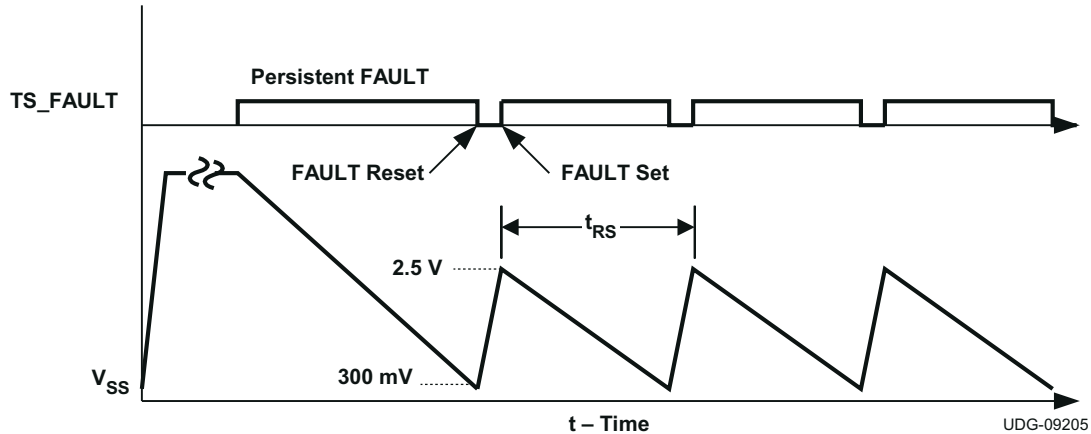


图 **6-8. Overtemperature Fault Restart Timing**

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 图 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.

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

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

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

(12)

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

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

(13)

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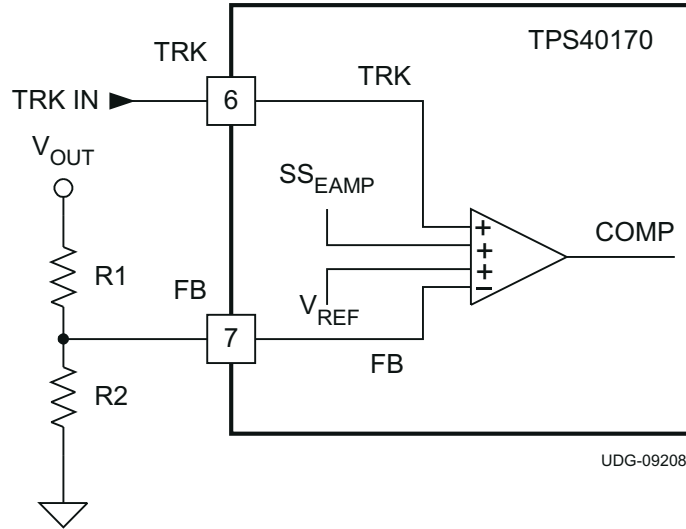
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**图 6-9. Tracking Functional Block**

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 图 6-10, 图 6-11, 图 6-12, 图 6-13, 图 6-14, and 图 6-15 respectively.

In simultaneous voltage tracking, shown in 图 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), 方程式 12 is used. As a result, components selection must meet 方程式 14.

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

(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, 方程式 13 is used. As a result, the resistor settings must meet 方程式 15 and 方程式 16.

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

(15)

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

(16)

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

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

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

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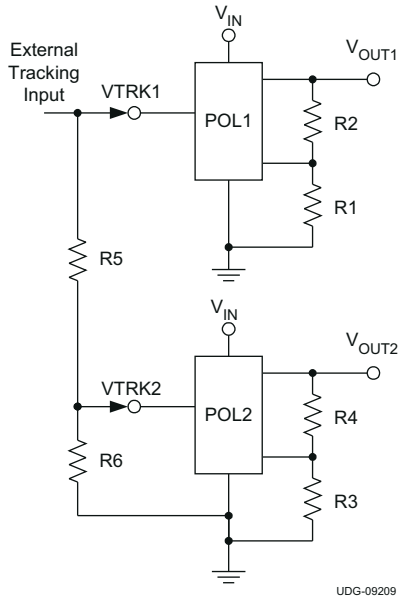- $R_1 = 3.16\ k\Omega$
- $R_3 = 1.37\ k\Omega$



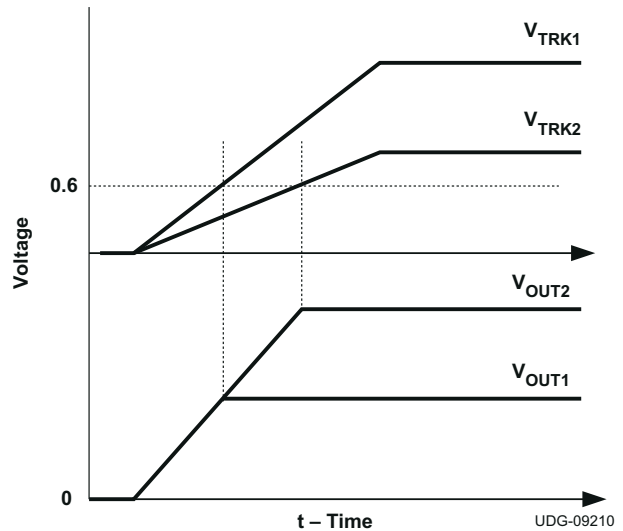图 **6-10. Simultaneous Voltage-Tracking Schematic**



图 **6-11. Simultaneous Voltage-Tracking Waveform**

In ratiometric voltage tracking as shown in 图 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.

Product Folder Links: *TPS40170-Q1*
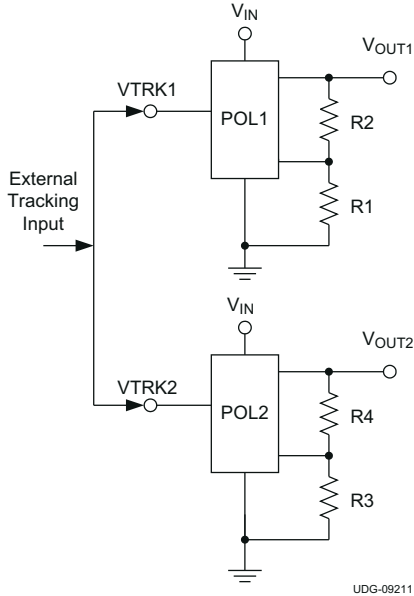
English Data Sheet: SLVSB90

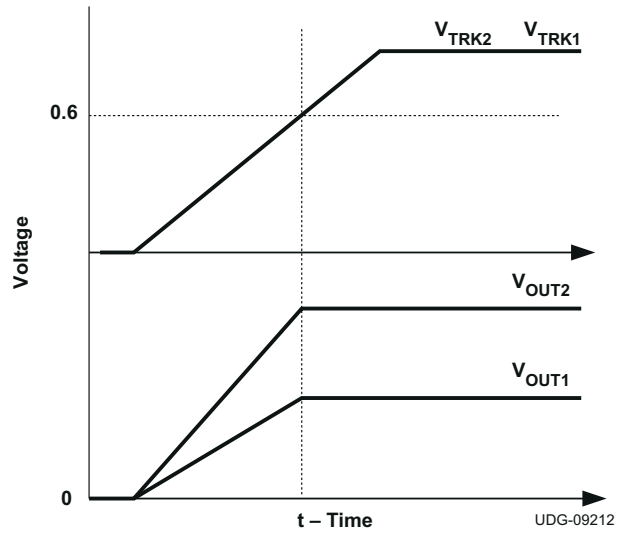图 **6-12. Ratiometric Voltage-Tracking Schematic**



图 **6-13. Ratiometric Voltage-Tracking Waveform**

Sequential start-up is shown in 图 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.
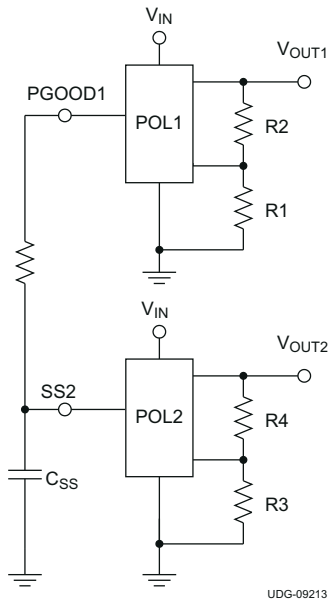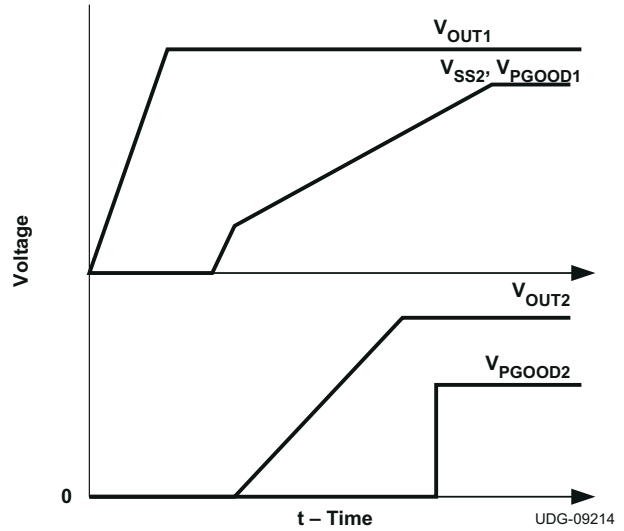


图 **6-14. Sequential Start-Up Schematic**



图 **6-15. Sequential Start-Up Waveform**

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备注

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 k$\Omega$ and 100 k$\Omega$) 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 图 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.
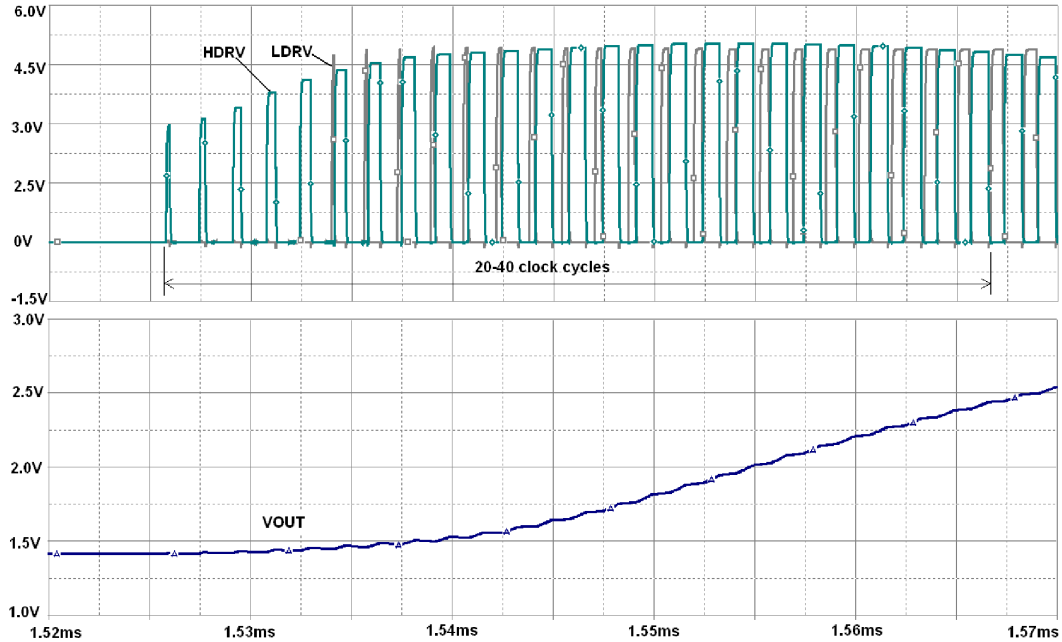
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**图 6-16. Start-Up Switching Waveform During Pre-Biased Condition**

### 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.

---

**备注**

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 方程式 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.

---

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

(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

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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.

图 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.
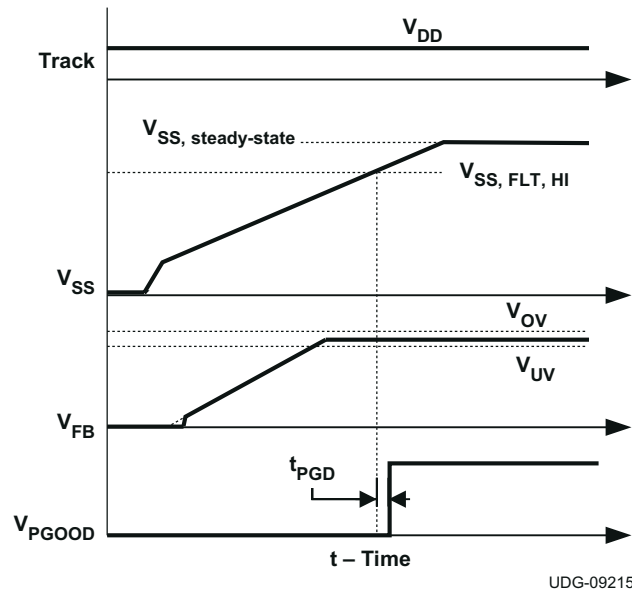


图 **6-17. PGOOD Signal**

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

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备注

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

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- 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 图 6-18. The synchronization frequency must be within ±30% of its programmed free-running frequency.

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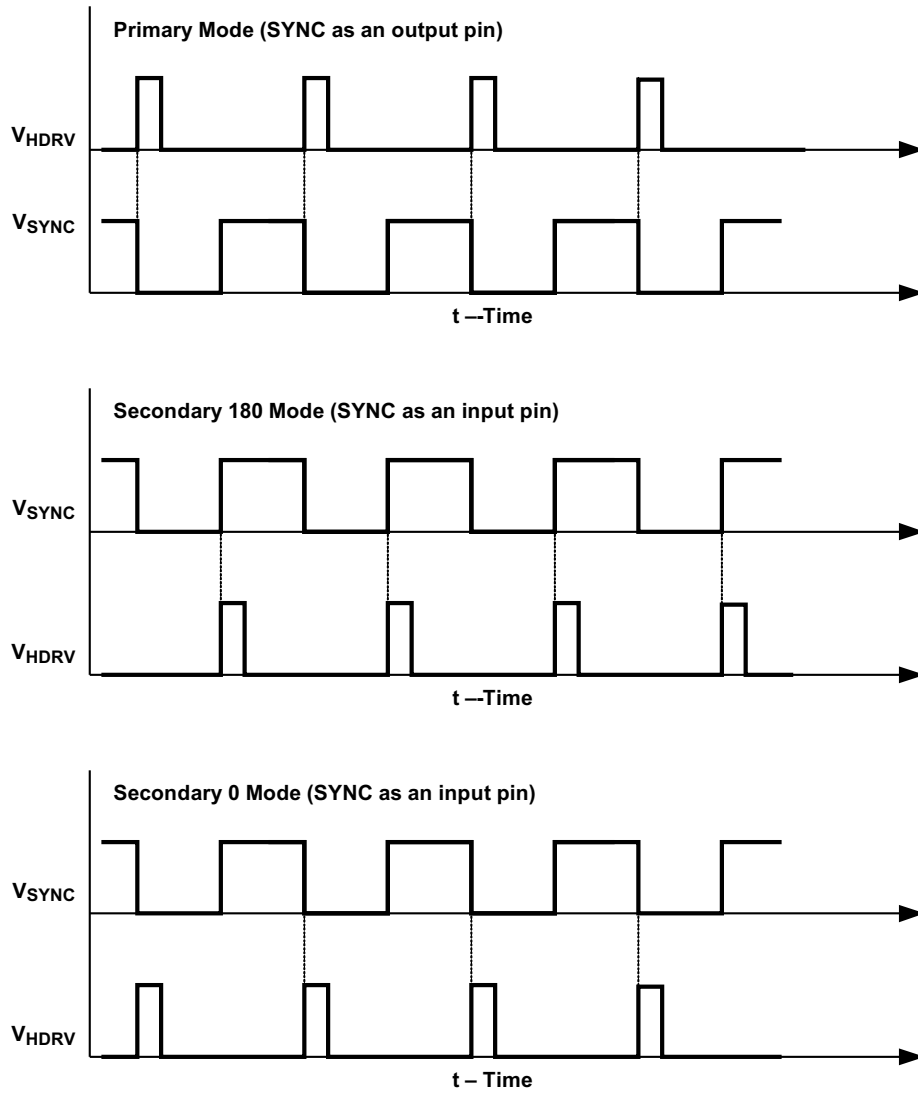
UDG-09206

**图 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. 图 6-19 shows this process.

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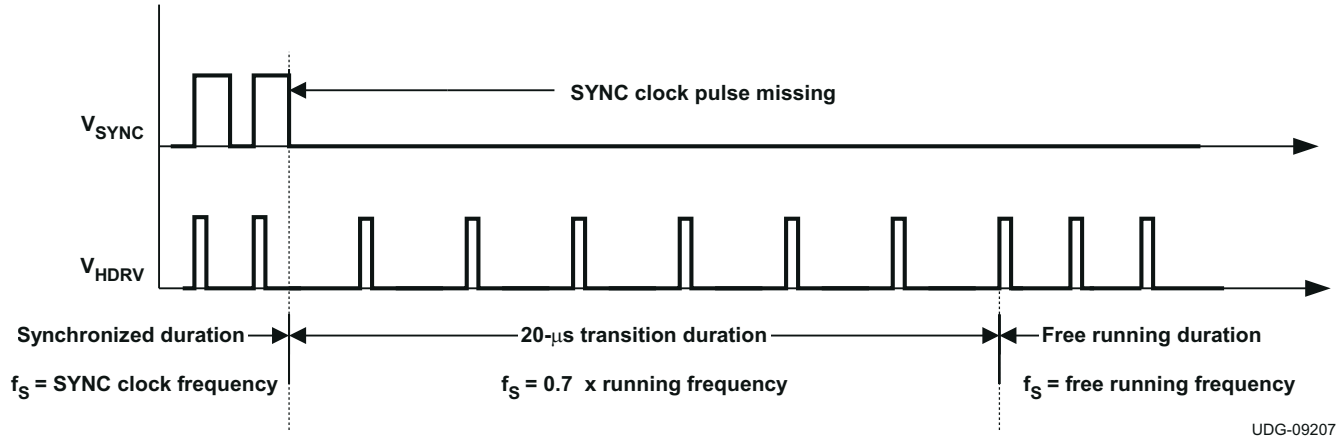
**图 6-19. Transition for SYNC Clock Signal Missing (for Secondary–180° Mode)**

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备注

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-k$\Omega$ resistor.

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# 7 Application and Implementation

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**备注**

以下应用部分中的信息不属于 TI 器件规格的范围，TI 不担保其准确性和完整性。TI 的客 户应负责确定器件是否适用于其应用。客户应验证并测试其设计，以确保系统功能。

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## 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.
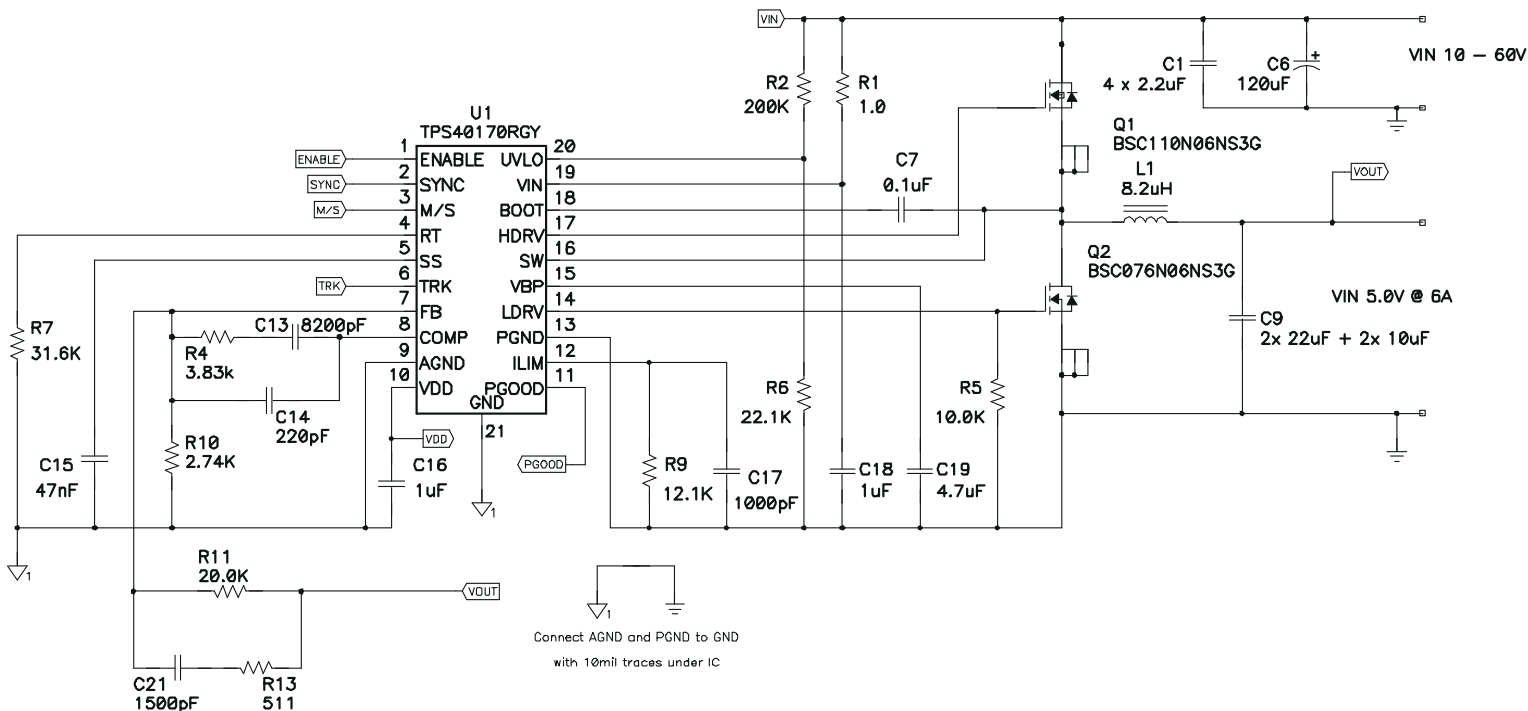
## 7.2 Typical Application



图 7-1. Design Example Application

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### 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 | $\triangle I_{OUT}$ = 2.5 A | | 250 | | mV |
| $V_{UNDER}$ | Output undershoot | $\triangle I_{OUT}$ = -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-m Ω | 0.51 inch$^2$ (1.3 cm$^2$) | IHLP5050FDER8R2M01 | Vishay |
| Q1 | 1 | | MOSFET, N-channel, 60-V, 50-A, 11-m Ω | | BSC110N06NS3G | Infineon |
| Q2 | 1 | | MOSFET, N-channel, 60-V, 50-A, 7.6-m Ω | | BSC076N06NS3G | Infineon |
| R10 | 1 | 2.74 k Ω | Resistor, chip, 1/16W, 1% | 603 | Std | R603 |
| R4 | 1 | 3.83 k Ω | Resistor, chip, 1/16W, 1% | 603 | Std | R603 |
| R5 | 1 | 10.0 k Ω | Resistor, chip, 1/16W, 1% | 603 | Std | R603 |

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

| 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 方程式 19.

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

(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 方程式 20.

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

(20)

### 7.2.2.3 Output Capacitor Selection (C9)

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

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

(21)

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

(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.

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

(23)

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

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

(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 方程式 25.

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

(25)

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

(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:

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

(27)

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

(28)

The rms current in the input capacitors is estimated in 方程式 29.

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

(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 |

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### 7.2.2.6 MOSFET Switch Selection (Q1, Q2)

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

High-side gate (Q1):

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

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

Low-side gate (Q2):

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

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

Optimizing for 300 kHz, 24-V input, 5-V output at 6 A, calculate ratios of 5.9 m$\Omega$/nC and 0.5 m$\Omega$/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 方程式 34.

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

### 7.2.2.8 UVLO Programming Resistors (R2, R6)

The UVLO hysteresis level is programmed by R2 using 方程式 35.

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

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

### 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.

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

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### 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 方程式 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 方程式 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 方程式 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 方程式 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 方程式 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

图 7-2 shows an efficiency graph for this design with 10-V to 60-V input and 5-V at 6-A output. 图 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. 图 7-4 shows the output ripple 20 mV/div, 2 μs/div, 20 MHz bandwidth.



图 7-2. Efficiency vs Load Current



图 7-3. Loop Response



图 7-4. Output Ripple Waveform
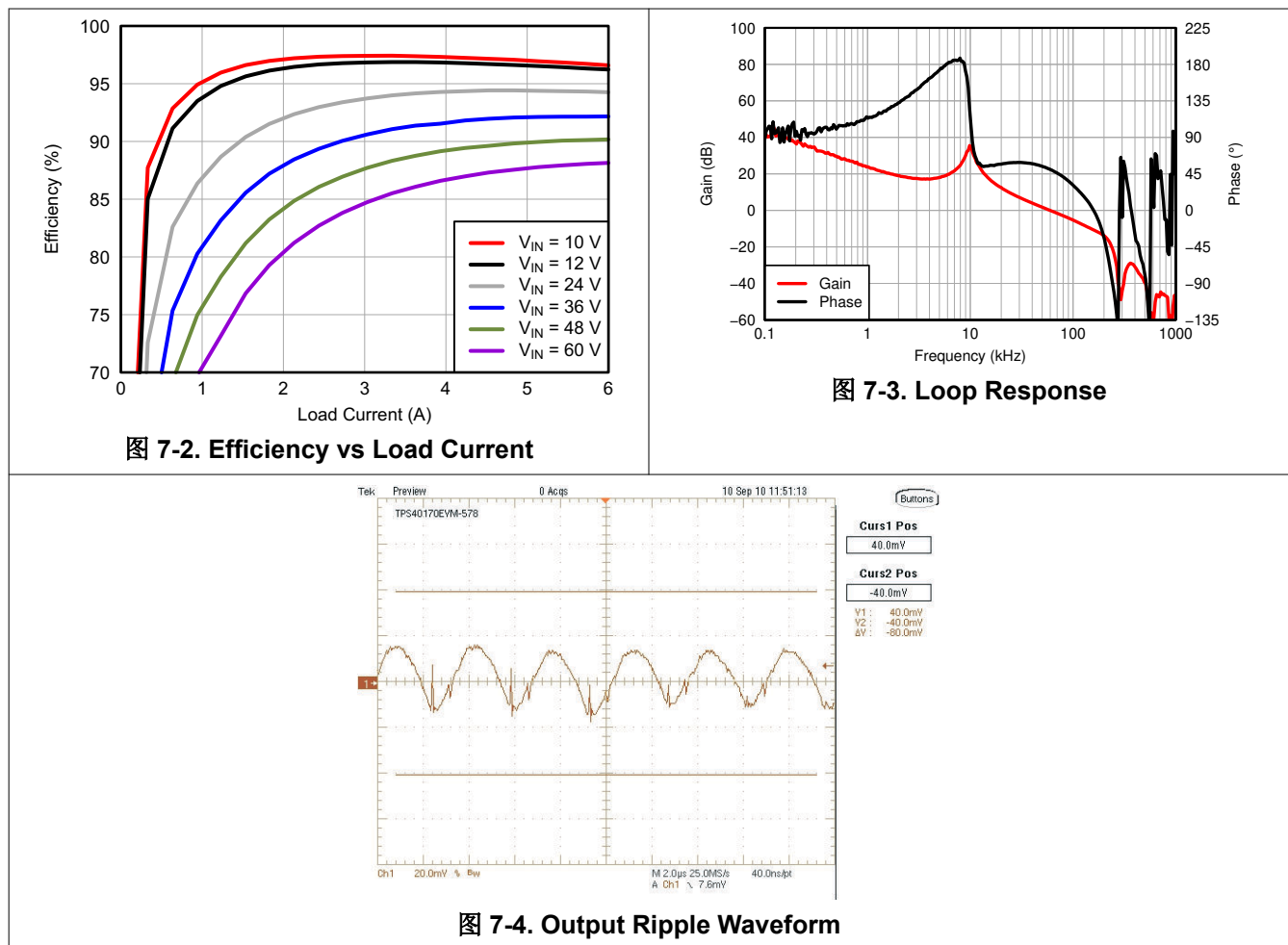
## 7.3 Power Supply Recommendations

### 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.

Product Folder Links: *TPS40170-Q1*

### 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.
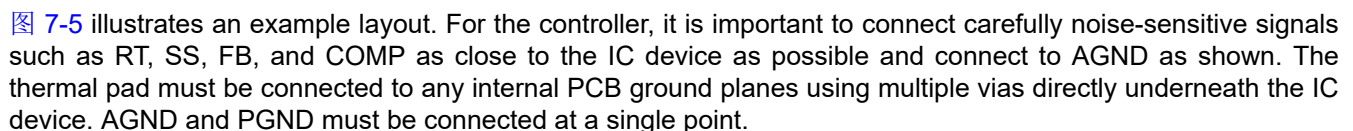
### 7.3.3 Input Resistor

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

图 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 图 7-5 through 图 7-8 for reference.
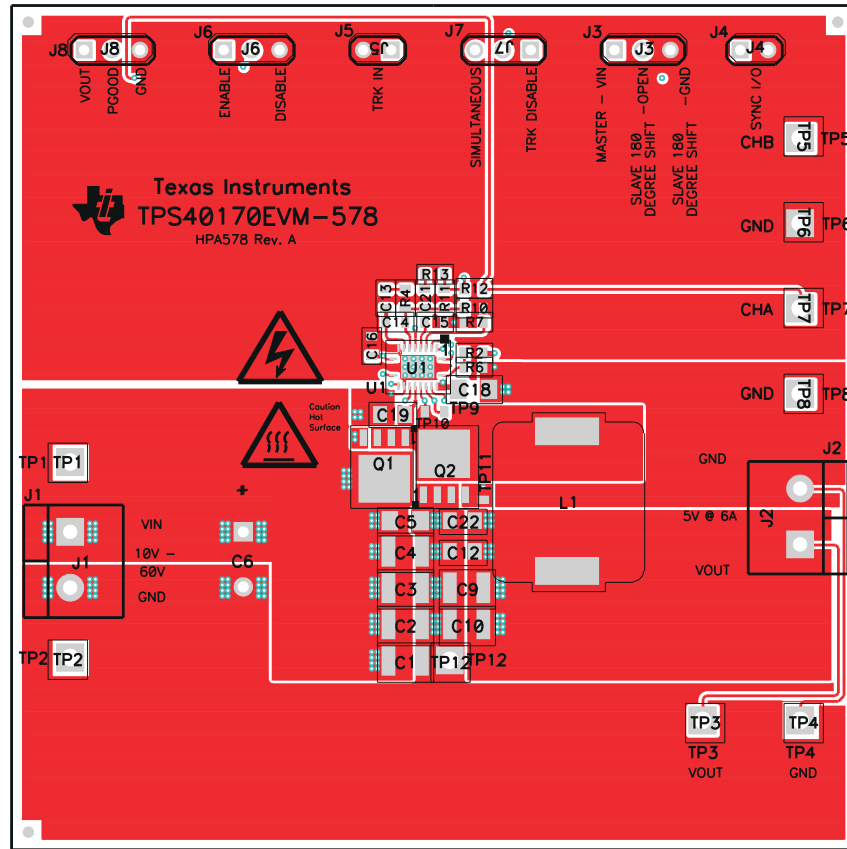
### 7.4.2 Layout Example



图 **7-5. Top Copper, Viewed From Top**

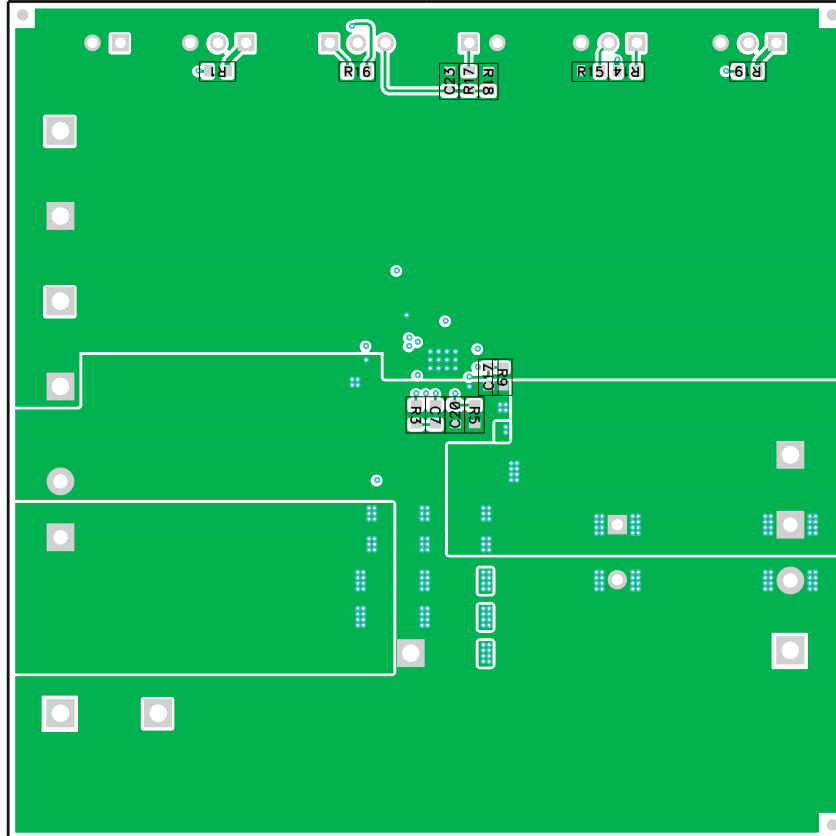Product Folder Links: *TPS40170-Q1*

English Data Sheet: SLVSB90

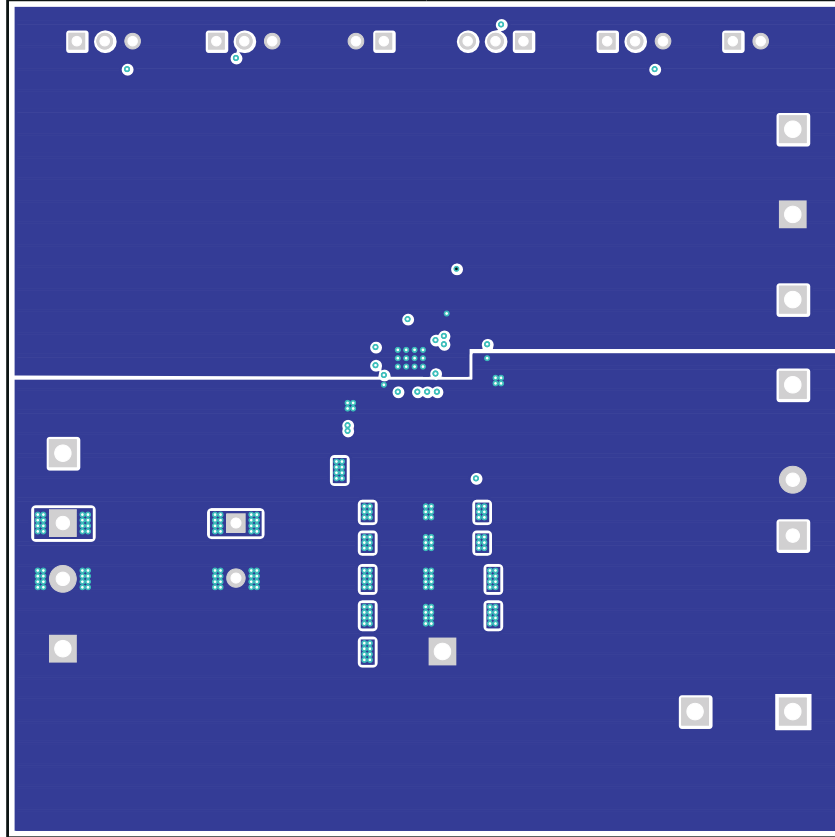图 7-6. Bottom Copper, Viewed From Bottom

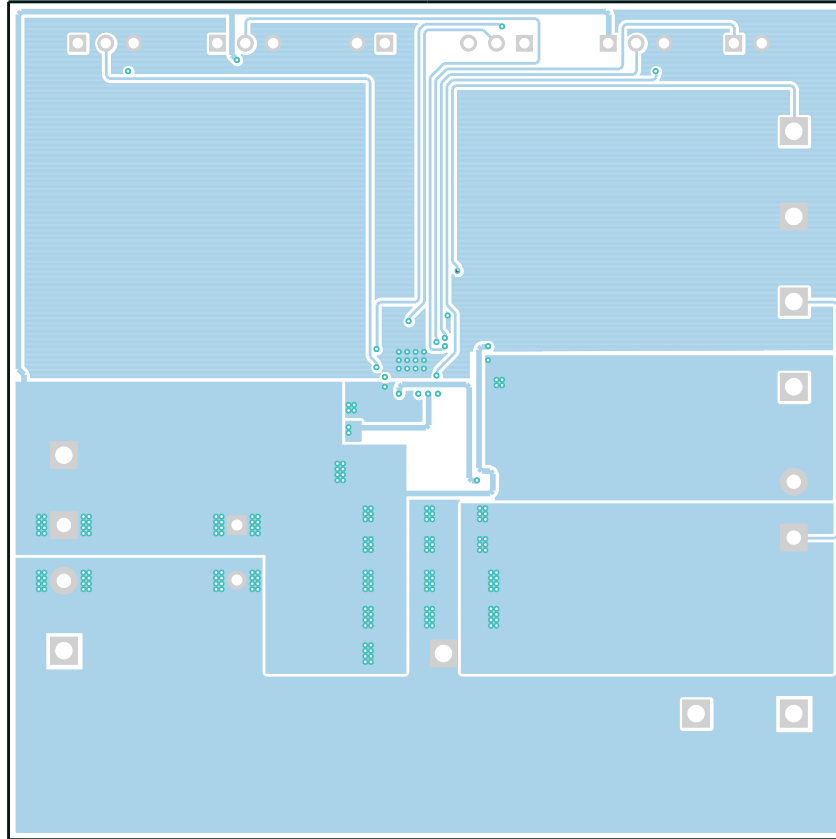图 7-7. **Internal Layer 1, Viewed From Top**

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

# 8 Device and Documentation Support

## 8.1 Device Support

### 8.1.1 第三方产品免责声明

TI 发布的与第三方产品或服务有关的信息，不能构成与此类产品或服务或保修的适用性有关的认可，不能构成此类产品或服务单独或与任何 TI 产品或服务一起的表示或认可。

## 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 接收文档更新通知

要接收文档更新通知，请导航至 ti.com 上的器件产品文件夹。点击*通知* 进行注册，即可每周接收产品信息更改摘要。有关更改的详细信息，请查看任何已修订文档中包含的修订历史记录。

## 8.4 支持资源

TI E2E™ 中文支持论坛是工程师的重要参考资料，可直接从专家处获得快速、经过验证的解答和设计帮助。搜索现有解答或提出自己的问题，获得所需的快速设计帮助。

链接的内容由各个贡献者"按原样"提供。这些内容并不构成 TI 技术规范，并且不一定反映 TI 的观点；请参阅 TI 的使用条款。

## 8.5 Trademarks

NexFET™ and TI E2E™ are trademarks of Texas Instruments.
所有商标均为其各自所有者的财产。

## 8.6 静电放电警告

静电放电 (ESD) 会损坏这个集成电路。德州仪器 (TI) 建议通过适当的预防措施处理所有集成电路。如果不遵守正确的处理和安装程序，可能会损坏集成电路。

ESD 的损坏小至导致微小的性能降级，大至整个器件故障。精密的集成电路可能更容易受到损坏，这是因为非常细微的参数更改都可能会导致器件与其发布的规格不相符。

## 8.7 术语表

TI 术语表　　　　　本术语表列出并解释了术语、首字母缩略词和定义。

Product Folder Links: *TPS40170-Q1*

English Data Sheet: SLVSB90

## 9 Revision History

注：以前版本的页码可能与当前版本的页码不同

## 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.

# PACKAGE OPTION ADDENDUM

## PACKAGING INFORMATION

| Orderable Device | Status (1) | Package Type | Package Drawing | Pins | Package Qty | Eco Plan (2) | Lead finish/ Ball material (6) | MSL Peak Temp (3) | Op Temp (°C) | Device Marking (4/5) | Samples |
|---|---|---|---|---|---|---|---|---|---|---|---|
| TPS40170QRGYRQ1 | ACTIVE | VQFN | RGY | 20 | 3000 | RoHS & Green | NIPDAU | Level-3-260C-168 HR | -40 to 125 | PXXQ | Samples |

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

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

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

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

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

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

**Important Information and Disclaimer:**The information provided on this page represents TI's knowledge and belief as of the date that it is provided. TI bases its knowledge and belief on information provided by third parties, and makes no representation or warranty as to the accuracy of such information. Efforts are underway to better integrate information from third parties. TI has taken and continues to take reasonable steps to provide representative and accurate information but may not have conducted destructive testing or chemical analysis on incoming materials and chemicals. TI and TI suppliers consider certain information to be proprietary, and thus CAS numbers and other limited information may not be available for release.

In no event shall TI's liability arising out of such information exceed the total purchase price of the TI part(s) at issue in this document sold by TI to Customer on an annual basis.

**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

**PACKAGE MATERIALS INFORMATION**

## TAPE AND REEL INFORMATION

**REEL DIMENSIONS**



Reel Diameter

Reel Width (W1)

**TAPE DIMENSIONS**



Cavity

| 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**



Sprocket Holes

User Direction of Feed

Pocket Quadrants

*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 | 3.8 | 4.8 | 1.6 | 8.0 | 12.0 | Q1 |

# PACKAGE MATERIALS INFORMATION

**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 | 367.0 | 367.0 | 35.0 |

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



4225264/A

3.65
3.35

4.65
4.35

PIN 1 INDEX AREA

A

B

1.0
0.8

0.05
0.00

C

SEATING PLANE

0.08 C

2.05 ± 0.1

2X 1.5

10    11

9

14X 0.5

2X
3.5

SYMM

21

EXPOSED
THERMAL PAD

12

3.05 ± 0.1

(0.2) TYP

2

PIN 1 ID

1    SYMM    20

19

20X 0.30
0.18

0.1 Ⓜ C A B
0.05 Ⓜ

20X 0.5
0.3
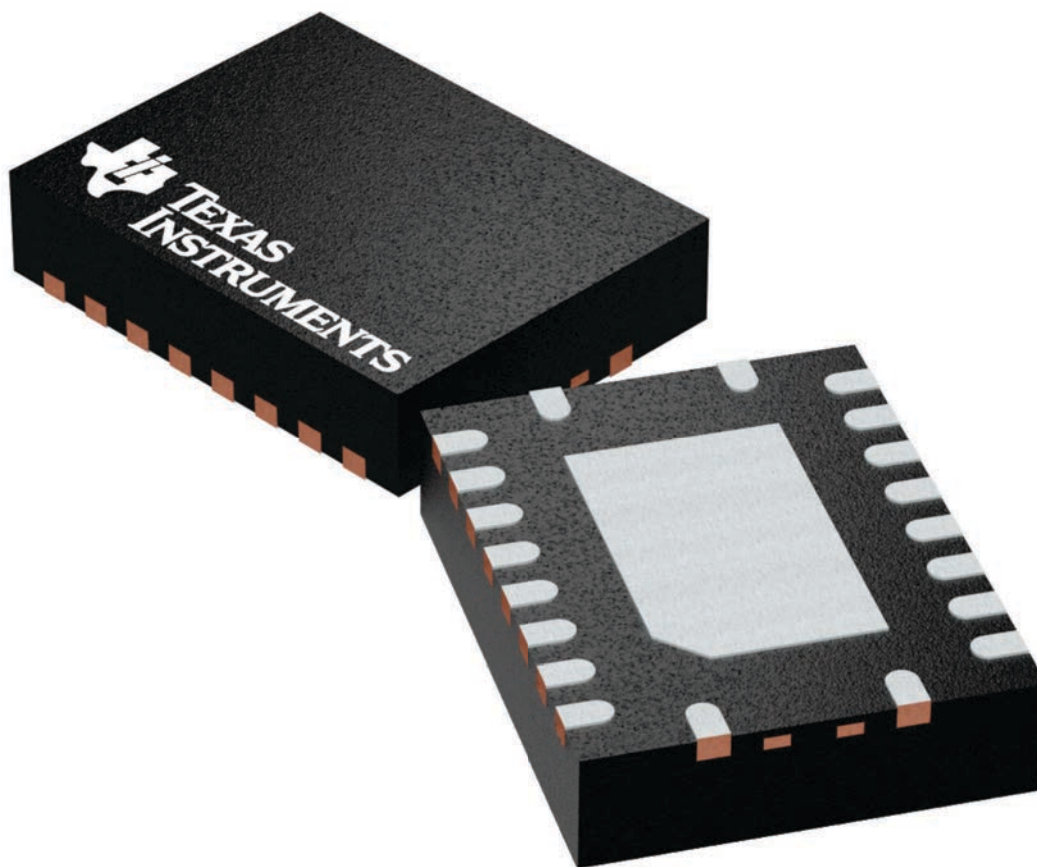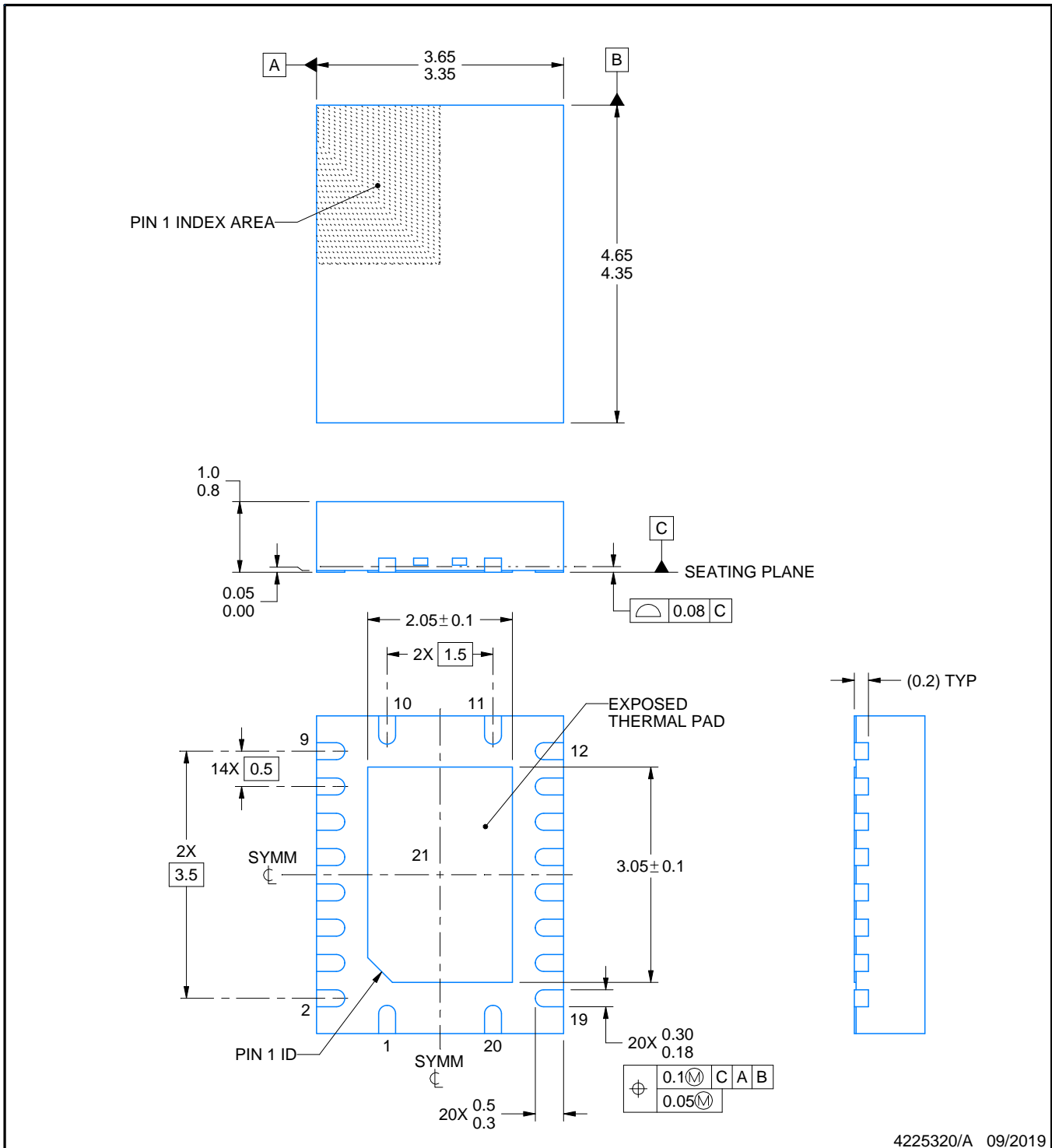
4225320/A   09/2019

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.

TEXAS
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## RGY0020A

LAND PATTERN EXAMPLE
EXPOSED METAL SHOWN
SCALE:18X

NON SOLDER MASK
DEFINED
(PREFERRED)

SOLDER MASK
DEFINED

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.

## RGY0020A

SYMM

4X (0.92)

1

20

(R0.05) TYP

20X (0.6)

2

20X (0.24)

19

4X
(1.33)

SYMM

21

(4.3)

(0.77)

14X (0.5)

(0.56)

9

12

METAL
TYP

10

11

(0.75)
TYP

(3.3)

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.

## 重要声明和免责声明