# TPS2584x-Q1 Automotive USB Type-A BC1.2 5-V 3.5-A Output, 36-V Input Synchronous Buck With Cable Compensation

## 1 Features

- AEC-Q100 qualified for automotive applications:
  - Temperature grade 1: –40°C to +125°C, $T_A$
  - HBM ESD Classification level H2
  - CDM ESD Classification level C5
- Functional Safety-Capable
  - Documentation available to aid functional safety system design
- Synchronous Buck DC/DC regulator
  - Input voltage range: 4.5 V to 36 V
  - Output current: 3.5 A
  - 5.1-V output voltage with ±1% accuracy
  - Current mode control
  - Adjustable frequency: 300 kHz to 2.2 MHz
  - Frequency synchronization to external clock
  - FPWM with spread-spectrum dithering
  - Internal compensation for ease of use
- Compliant to USB-IF standards
  - CDP/SDP mode per USB BC1.2
- Optimized for USB power and communication
  - User-programmable USB current limit
  - Cable droop compensation up to 1.5 V
  - High bandwidth data switches on DP and DM
  - Client mode for system update
- Integrated protection
  - $V_{BUS}$ Short-to-$V_{BAT}$ protection
  - DP_IN and DM_IN Short-to-$V_{BAT}$ (TPS25840-Q1 only)
  - DP_IN and DM_IN Short-to-$V_{BUS}$
  - DP_IN, DM_IN IEC 61000-4-2 rated
    - ±8-kV contact and ±15-kV air discharge
- Fault flag reports
- 32-pin QFN package with wettable flank

## 2 Applications

- Automotive Infotainment
- USB media hubs
- USB charger ports

## 3 Description

The TPS2584x-Q1 is a USB Type-A BC1.2 charging solution that includes a synchronous DC/DC converter. With cable droop compensation, the Vbus voltage remains constant regardless of load current, ensuring connected portable devices charge at optimal current and voltage even under heavy loads.

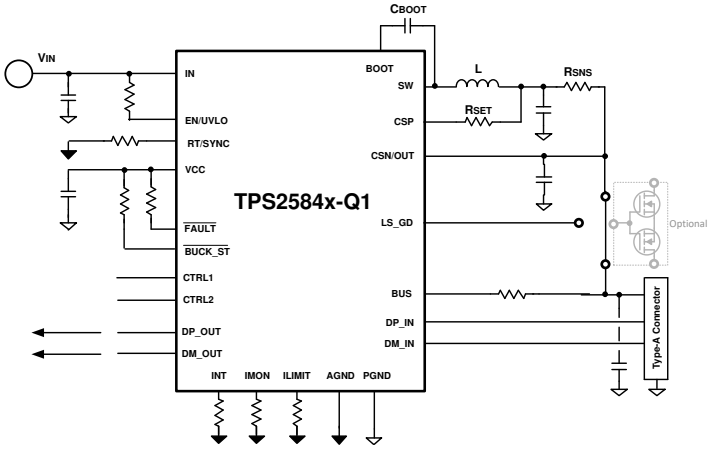
The TPS2584x-Q1 includes high bandwidth analog switches for DP and DM pass-through.

The TPS25840-Q1 also integrates short-to-battery protection on $V_{BUS}$, DM_IN and DP_IN pins. These pins can withstand voltage up to 18 V. TPS25842-Q1 does not support data line (Dx) short-to-$V_{BAT}$ protection.

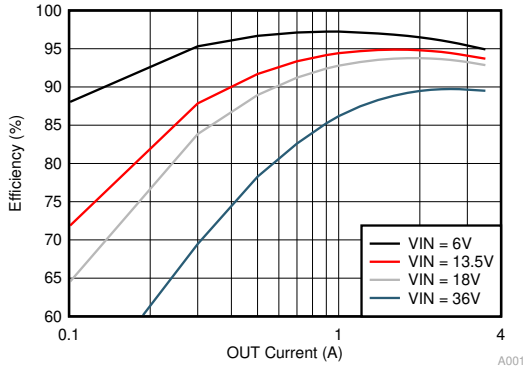
**Device Information**(1)

| PART NUMBER | PACKAGE | BODY SIZE (NOM) |
|---|---|---|
| TPS25840-Q1 | VQFN (32) | 5.00 mm x 5.00 mm |
| TPS25842-Q1 | VQFN (32) | 5.00 mm x 5.00 mm |

(1) For all available packages, see the orderable addendum at the end of the data sheet




**Simplified Schematic TPS2584x-Q1**




**Buck Efficiency vs Output Current fsw = 400 kHz**

An IMPORTANT NOTICE at the end of this data sheet addresses availability, warranty, changes, use in safety-critical applications, intellectual property matters and other important disclaimers. PRODUCTION DATA.

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## 4 Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

| Changes from Revision D (February 2022) to Revision E (March 2022) | Page |
|---|---|
| • Added the thermal information for RHB0032AA package | 8 |

| Changes from Revision C (December 2021) to Revision D (February 2022) | Page |
|---|---|
| • Added RHB0032AA package to the data sheet | 4 |

| Changes from Revision B (August 2020) to Revision C (December 2021) | Page |
|---|---|
| • Added additional description for Fault pin, for better signal quality under some sensitive application | 4 |

| Changes from Revision A (May 2020) to Revision B (August 2020) | Page |
|---|---|
| • Updated the numbering format for tables, figures and cross-references throughout the document | 1 |
| • Added functional safety link to the *Features* section | 1 |

| Changes from Revision * (September 2019) to Revision A (May 2020) | Page |
|---|---|
| • Changed *Layout* description for clarity | 48 |


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## 5 Description (Continued)

The synchronous buck regulator operates with current mode control and is internally compensated to simplify design. A resistor on the RT pin sets the switching frequency between 300 kHz and 2.2 MHz. Operating below 400 kHz results in better system efficiency. Operation above 2.1 MHz avoids the AM radio bands and allows for use of a smaller inductor.

The TPS2584x-Q1 integrates electrical signatures necessary for legacy devices which use USB data lines to determine charging configuration.

A precision current sense amplifier is included for user programmable cable droop compensation and current limit tuning. Cable compensation aids portable devices in charging at optimum current and voltage under heavy loads by changing the buck regulator output voltage linearly with load current to counteract the voltage drop due to wire resistance in automotive cabling. The VBUS voltage measured at a connected portable device remains approximately constant, regardless of load current, allowing the portable device's battery charger to work optimally.

The USB specifications require current limiting of USB charging ports, but give system designers reasonable flexibility to choose overcurrent protection levels based on system requirements. The TPS2584x-Q1 uses a novel two-threshold current limit circuit allowing system designers to either program average current limit protection of the buck regulator, or optionally, current limit using an external NMOS between the CSN/OUT and BUS pins. The NFET implementation enables the TPS2584x-Q1 buck regulator to supply a 5-V output for other loads during an overcurrent condition on the USB port.

Protection features include cycle-by-cycle current limit, hiccup short-circuit protection, undervoltage lockout, VBUS overvoltage and overcurrent, data line (Dx) short to VBUS, and die overtemperature protection.

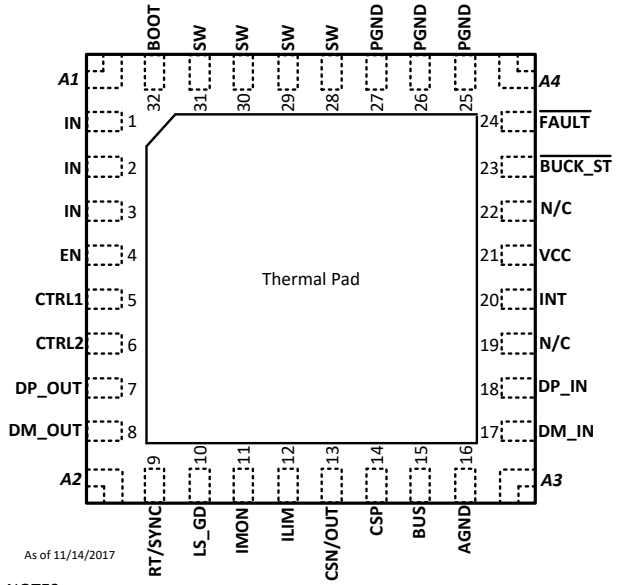
The TPS25840-Q1 includes high bandwidth analog switches for DP and DM pass-through, and support data line (Dx) short to VBAT protection. TPS25842-Q1 don't support data line (Dx) short to VBAT protection.

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## 6 Device Comparison Table

| PART NUMBER | PACKAGE | DCP AUTO | DP AND DM SWITCHES | NTC INPUT | DP/DM Short to BAT |
|---|---|---|---|---|---|
| TPS25840-Q1 | VQFN (32) | No | Yes | No | Yes |
| TPS25842-Q1 | VQFN (32) | No | Yes | No | No |

## 7 Pin Configuration and Functions




NOTES:
1) A1, A2, A3, and A4 are corner anchors for enhanced package stress performance.
2) A1, A2, A3, and A4 are electrically connected to the thermal pad.
3) A1, A2, A3, and A4 PCB lands should be electrically isolated or electrically connected to thermal pad and PGND.

**Figure 7-1. TPS25840QWRHBRQ1, TPS25842QWRHBRQ1 Package 32-Pin (VQFN) Top View (1)**

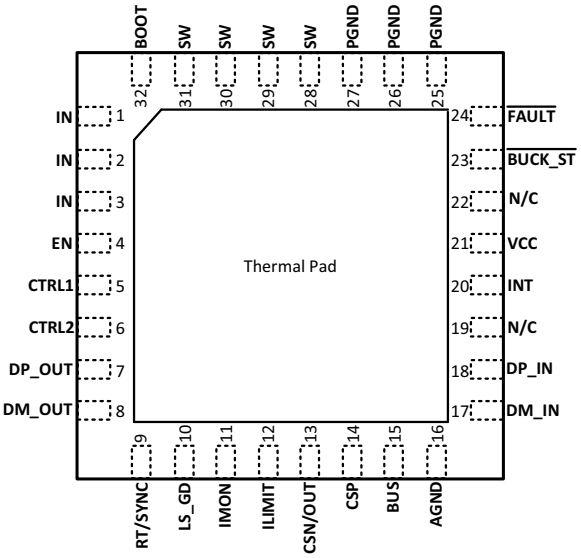



**Figure 7-2. TPS25840QCWRHBRQ1, TPS25842QCWRHBRQ1 Package 32-Pin (VQFN) Top View (2)**

**Table 7-1. Pin Functions**

| PIN | | TYPE(3) | I/O | DESCRIPTION |
|---|---|---|---|---|
| NAME | NO. | | | |
| AGND | 16 | G | - | Analog ground terminal. Ground reference for internal references and logic. All electrical parameters are measured with respect to this pin. Connect to system ground on PCB. |
| BOOT | 32 | P | | Boot-strap capacitor connection for HS FET driver. Connect a high quality 100-nF capacitor from this pin to the SW pin. |
| BUS | 15 | A | I | VBUS discharge input. Connect to VBUS on USB Connector. |
| CSN/OUT | 13 | P | I | Negative input of current sense amplifier, also buck output for internal voltage regulation. |
| CSP | 14 | P | I | Positive input of current sense amplifier. |
| CTRL1 | 5 | A | I | Logic-level control inputs for device and system configuration (see Table 10-6). |
| CTRL2 | 6 | A | I | Logic-level control inputs for device and system configuration (see Table 10-6). |
| DM_IN | 17 | A | | DM data line. Connect to USB connector. |
| DM_OUT | 8 | A | | DM data line. Connect to USB host controller. |
| DP_IN | 18 | A | | DP data line. Connect to USB connector. |
| DP_OUT | 7 | A | | DP data line. Connect to USB host controller. |


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**Table 7-1. Pin Functions (continued)**

| PIN | | TYPE(3) | I/O | DESCRIPTION |
|---|---|---|---|---|
| NAME | NO. | | | |
| EN/UVLO | 4 | A | | Enable pin. Do not float. High = on, Low = off. Can be tied to VIN. Precision enable input allows adjustable UVLO by external resistor divider. |
| $\overline{\text{FAULT}}$ | 24 | A | O | Active LOW open-drain output. Asserted during fault conditions (see Table 10-4). TI recommends series about 1-k ohm damping resistor for better signal quality. |
| ILIMIT | 12 | A | | External resistor used to set the current-limit threshold (see Table 10-2). |
| IMON | 11 | A | | External resistor used to set the max cable comp voltage at full load current. |
| IN | 1, 2, 3 | P | I | Input Supply to regulator. Connect high-quality bypass capacitors directly to this pin and PGND. |
| $\overline{\text{BUCK_ST}}$ | 23 | A | O | Active Low open-drain output. After $\overline{\text{BUCK_ST}}$ assert, the Buck converter begins to start up. At the same time, DP and DM data switch turn on accordingly. |
| LS_GD | 10 | A | | External NMOS gate driver. If TPS2584x-Q1 configured under average current limit mode, LS_GD pin must be pulled up through a 2.2-kΩ resistor (see *Current Limit Sensing using $R_{ILIMIT}$*). |
| PGND | 25, 26, 27 | G | | Power ground terminal. Connect to system ground and AGND. Connect to bypass capacitor with short wide traces. |
| N/C | 19, 22 | - | | Make no electrical connection. |
| RT/SYNC | 9 | A | | Resistor Timing or External Clock input. An internal amplifier holds this terminal at a fixed voltage when using an external resistor to ground to set the switching frequency. If the terminal is pulled above the PLL upper threshold, a mode change occurs and the terminal becomes a synchronization input. The internal amplifier is disabled and the terminal is a high impedance clock input to the internal PLL. If clocking edges stop, the internal amplifier is re-enabled and the operating mode returns to resistor frequency programming. |
| SW | 28, 29, 30, 31 | P | | Switching output of the regulator. Internally connected to source of the HS FET and drain of the LS FET. Connect to power inductor. |
| INT | 20 | A | | For internal circuit, must connect a 5.1-K resistor to AGND. |
| VCC | 21 | P | | Output of internal bias supply. Used as supply to internal control circuits. Connect a high quality 2.2-µF capacitor from this pin to GND. |

(1) For package drawing please refer to RHB0032R at the end of the data sheet.
(2) For package drawing please refer to RHB0032AA at the end of the data sheet.
(3) A = Analog, P = Power, G = Ground.

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# 8 Specifications

## 8.1 Absolute Maximum Ratings

Voltages are with respect to GND (unless otherwise noted)(1)

| PARAMETER | | MIN | MAX | UNIT |
|---|---|---|---|---|
| Input voltage | IN to PGND | –0.3 | 40 | V |
| | OUT to PGND | –0.3 | 20 | |
| | EN to AGND | –0.3 | VIN + 0.3 | |
| | CSP to AGND | –0.3 | 20 | |
| | CSN to AGND | –0.3 | 20 | |
| | BUS to AGND | –0.3 | 18 | |
| | RT/SYNC to AGND | –0.3 | 6 | |
| | CTRL1 or CTRL2 to AGND | –0.3 | 6 | |
| | AGND to PGND | –0.3 | 0.3 | |
| Output voltage | SW to PGND | –0.3 | VIN + 0.3 | V |
| | SW to PGND (less than 10 ns transients) | –3.5 | 40 | |
| | BOOT to SW | –0.3 | 6 | |
| | VCC to AGND | –0.3 | 6 | |
| | LS_GD | –0.3 | 18 | |
| Voltage range | **TPS25840-Q1**: DP_IN, DM_IN to AGND | –0.3 | 18 | V |
| | **TPS25842-Q1**: DP_IN, DM_IN to AGND | –0.3 | 7 | |
| | DP_OUT, DM_OUT to AGND | –0.3 | 6 | |
| | $\overline{\text{FAULT}}$, $\overline{\text{BUCK_ST}}$, INT to AGND | –0.3 | 6 | |
| | ILIMIT or IMON to AGND | –0.3 | 6 | |
| Pin positive source current, $I_{VCC}$ | VCC Source Current | | 5 | mA |
| Pin positive sink current, $I_{SNK}$ | $\overline{\text{FAULT}}$, $\overline{\text{BUCK_ST}}$ | | Internally Limited | A |
| I/O current | DP_IN to DP_OUT, or DM_IN to DM_OUT in SDP, CDP, or Client Mode | –100 | 100 | mA |
| $T_J$ | Junction temperature | -40 | 150 | °C |
| $T_{stg}$ | Storage temperature | –65 | 150 | °C |

(1) Stresses beyond those listed under *Absolute Maximum Rating* can 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 Condition*. Exposure to absolute-maximum-rated conditions for extended periods can affect device reliability.

## 8.2 ESD Ratings

| | | | | VALUE | UNIT |
|---|---|---|---|---|---|
| $V_{(ESD)}$ | Electrostatic discharge | Human body model (HBM), per AEC Q100-002(1) | | ±2000(2) | V |
| | | Charged device model (CDM), per AEC Q100-011 | Corner pins (1, 8, 9, 17, 25 and 32) | ±750(3) | |
| | | | Other pins | ±750(3) | |
| | | IEC 61000-4-2 contact discharge | DP_IN, DM_IN pins | ±8000(4) | |
| | | IEC 61000-4-2 air-gap discharge | DP_IN, DM_IN pins | ±15000(4) | |

(1) AEC Q100-002 indicates that HBM stressing shall be in accordance with the ANSI/ESDA/JEDEC JS-001 specification.
(2) The passing level per AEC-Q100 Classification H2.
(3) The passing level per AEC-Q100 Classification C5.
(4) Surges per IEC61000-4-2, 1999 applied between DP_IN, DM_IN and output ground of the TPS2584x-Q1 evaluation module.


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## 8.3 Recommended Operating Conditions

Voltages are with respect to GND (unless otherwise noted)

| | | | MIN | NOM | MAX | UNIT |
|---|---|---|---|---|---|---|
| $V_I$ | Input voltage | IN to PGND | 4.5 | | 36 | V |
| | | EN | 0 | | VIN | |
| | | VCC when driven from external regulator | 0 | | 5.5 | |
| | | DP_IN, DM_IN | 0 | | 3.6 | |
| | | DP_OUT, DM_OUT | 0 | | 3.6 | |
| | | CTRL1, CTRL2 | 0 | | VCC | |
| | | RT/SYNC when driven by external clock | 0 | | VCC | |
| $V_{PU}$ | Pull up voltage | $\overline{FAULT}$, $\overline{BUCK_ST}$ | 0 | | VCC | |
| $V_O$ | Output voltage | CSN/OUT | 0 | | 6.5 | |
| $I_O$ | Output current | Buck regulator output current | 0 | | 3.5 | A |
| | | DP_IN to DP_OUT or DM_IN to DM_OUT Continuous current in SDP, CDP or Client Mode | –30 | | 30 | mA |
| $I_{SNK}$ | Sink current | $\overline{FAULT}$, $\overline{BUCK_ST}$ | | | 10 | |
| $I_I$ | Input current | Continuous current into the CSP pin | | | 200 | µA |
| $R_{EXT}$ | External resistnace | $R_{IMON}$, $R_{ILIMIT}$ | 0 | | 100 | kΩ |
| $T_J$ | | Operating junction temperature | –40 | | 125(1) | °C |

(1) Operating at junction temperatures greater than 125°C is possible, however lifetime will be degraded.

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## 8.4 Thermal Information

| THERMAL METRIC(1) | | TPS2584x-Q1 | | UNIT |
|---|---|---|---|---|
| | | RHB0032R (VQFN) | RHB0032AA (VQFN) | |
| | | 32 PINS | 32 PINS | |
| $R_{\theta JA}$ | Junction-to-ambient thermal resistance | 28.7 | 29.4 | °C/W |
| $R_{\theta JC(top)}$ | Junction-to-case (top) thermal resistance | 17.6 | 18.6 | °C/W |
| $R_{\theta JB}$ | Junction-to-board thermal resistance | 7.2 | 9.7 | °C/W |
| $\Psi_{JT}$ | Junction-to-top characterization parameter | 0.2 | 0.2 | °C/W |
| $\Psi_{JB}$ | Junction-to-board characterization parameter | 7.2 | 9.7 | °C/W |
| $R_{\theta JC(bot)}$ | Junction-to-case (bottom) thermal resistance | 1 | 2.3 | °C/W |

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

## 8.5 Electrical Characteristics

Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| **SUPPLY VOLTATE (IN PIN)** | | | | | | |
| $V_{IN}$ | Operating input voltage range | | 4.5 | | 36 | V |
| $I_Q$ | Operating quiescent current (non switching) | $V_{EN/UVLO}$ = $V_{IN}$, CTRL1 = CTRL2 = $V_{CC}$, $V_{CSN}$ = 8 V, INT pull down resistance = 5.1 kΩ | | 700 | 990 | µA |
| $I_{SD}$ | Shutdown quiescent current; measured at IN pin. | EN= 0 | | 10 | 16 | µA |
| **ENABLE and UVLO (EN/UVLO PIN)** | | | | | | |
| $V_{EN/UVLO_VCC_H}$ | EN/UVLO input level required to turn on internal LDO | $V_{EN/UVLO}$ rising threshold | | | 1.14 | V |
| $V_{EN/UVLO_VCC_L}$ | EN/UVLO input level required to turn off internal LDO | $V_{EN/UVLO}$ falling threshold | 0.3 | | | V |
| $V_{EN/UVLO_H}$ | EN/UVLO input level required to turn on state machine | $V_{EN/UVLO}$ rising threshold | 1.140 | 1.200 | 1.260 | V |
| $V_{EN/UVLO_HYS}$ | Hysteresis | $V_{EN/UVLO}$ falling threshold | | 90 | | mV |
| $I_{LKG_EN/UVLO}$ | Enable input leakage current | $V_{EN/UVLO}$ = 3.3 V | | 0.5 | | uA |
| **INTERNAL LDO** | | | | | | |
| $V_{BOOT_UVLO}$ | Bootstrap voltage UVLO threshold | | | 2.2 | | V |
| $V_{CC}$ | Internal LDO output voltage appearing on VCC pin | 6 V ≤ $V_{IN}$ ≤ 36 V | 4.75 | 5 | 5.25 | V |
| $V_{CC_UVLO_R}$ | Rising UVLO threshold | | 3.4 | 3.6 | 3.8 | V |
| $V_{CC_UVLO_HYS}$ | Hysteresis | | | 600 | | mV |
| **CURRENT LIMIT VOLTAGE (CSP - CSN/OUT PINS) TO ACTIVATE BUCK AVG CURRENT LIMITING** | | | | | | |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage buck regulator control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 125°C | 43.5 | 46 | 48.5 | mV |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage buck regulator control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 150°C | 42.5 | 46 | 49.5 | mV |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage buck regulator control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 26.1 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 125°C | 20 | 22.5 | 25 | mV |


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## 8.5 Electrical Characteristics (continued)

Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage buck regulator control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 26.1 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 150°C | 19 | 22.5 | 26 | mV |
| **CURRENT LIMIT VOLTAGE (CSP - CSN/OUT PINS) TO ACTIVATE EXTERNAL NFET CURRENT LIMITING** | | | | | | |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage NFET control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 6.8 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 125°C | 40 | 43 | 46 | mV |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage NFET control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 6.8 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 150°C | 38.5 | 43 | 47.5 | mV |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage NFET control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13.7 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 125°C | 18 | 21 | 24 | mV |
| ($V_{CSP}$ – $V_{CSN/OUT}$) | Current limit voltage NFET control loop | $V_{CSN}$ = 5 V, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13.7 kΩ, $R_{IMON}$ = 13 kΩ, -40°C ≤ $T_J$ ≤ 150°C | 17 | 21 | 25 | mV |
| **CURRENT LIMIT - BUCK REGULATOR PEAK CURRENT LIMIT** | | | | | | |
| $I_{L-SC-HS}$ | High-side current limit | | 4.6 | 5.4 | 6.2 | A |
| $I_{L-SC-LS}$ | Low-side current limit | | 3.5 | 4 | 4.5 | A |
| $I_{L-NEG-LS}$ | Low-side negative current limit | | –3.1 | –2.1 | –1.3 | A |
| **CABLE COMPENSATION VOLTAGE** | | | | | | |
| $V_{IMON}$ | Cable compensation voltage | ($V_{CSP}$ – $V_{CSN}$) = 46 mV, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13 kΩ, $R_{IMON}$ = 13 kΩ | 0.935 | 1 | 1.065 | V |
| $V_{IMON}$ | Cable compensation voltage | ($V_{CSP}$ – $V_{CSN}$) = 23 mV, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13 kΩ, $R_{IMON}$ = 13 kΩ | 0.435 | 0.5 | 0.565 | V |
| $V_{IMON}$ | Cable compensation voltage (internal clamp) | ($V_{CSP}$ –$V_{CSN}$) = 46 mV, $R_{SET}$ = 300 Ω, $R_{ILIMIT}$ = 13 kΩ, $R_{IMON}$ = open | | 1.8 | | V |
| **BUCK OUTPUT VOLTAGE (CSN/OUT PIN)** | | | | | | |
| $V_{CSN/OUT}$ | Output voltage | INT pulldown resistance = 5.1kΩ, $R_{IMON}$ = 0 Ω, $R_{ILIMIT}$ = 0 Ω | 5.05 | 5.10 | 5.15 | V |
| $V_{CSN/OUT}$ | Output voltage accuracy | INT pulldown resistance = 5.1kΩ, $R_{IMON}$ = 0 Ω, $R_{ILIMIT}$ = 0 Ω | –1 | | 1 | % |
| $V_{CSN/OUT_OV}$ | Overvoltage level on CSN/OUT pin which buck regulator stops switching | $V_{CSN/OUT}$ rising | 7.1 | 7.5 | 7.9 | V |
| $V_{CSN/OUT_OV_HYS}$ | Hysteresis | | | 500 | | mV |
| $V_{HC}$ | CSN / OUT pin voltage required to trigger short circuit hiccup mode | | | 2 | | V |
| $V_{DROP}$ | Dropout voltage ( $V_{IN}$-$V_{OUT}$ ) | $V_{IN}$ = $V_{OUT}$ + $V_{DROP}$, $V_{OUT}$ = 5.1V, $I_{OUT}$ = 3A | | 150 | | mV |
| **BUCK REGULATOR INTERNAL RESISTANCE** | | | | | | |
| $R_{DS-ON-HS}$ | High-side MOSFET ON-resistance | Load = 3 A, $T_J$ = 25°C | | 40 | 45 | mΩ |
| $R_{DS-ON-HS}$ | High-side MOSFET ON-resistance | Load = 3 A, -40°C ≤ $T_J$ ≤ 125°C | | 40 | 68 | mΩ |
| $R_{DS-ON-HS}$ | High-side MOSFET ON-resistance | Load = 3 A, -40°C ≤ $T_J$ ≤ 150°C | | 40 | 75 | mΩ |
| $R_{DS-ON-LS}$ | Low-side MOSFET ON-resistance | Load = 3 A, $T_J$ = 25C | | 35 | 41 | mΩ |
| $R_{DS-ON-LS}$ | Low-side MOSFET ON-resistance | Load = 3 A, -40°C ≤ $T_J$ ≤ 125°C | | 35 | 60 | mΩ |
| $R_{DS-ON-LS}$ | Low-side MOSFET ON-resistance | Load = 3 A, -40°C ≤ $T_J$ ≤ 150°C | | 35 | 68 | mΩ |
| **NFET GATE DRIVE (LS_GD PIN)** | | | | | | |

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## 8.5 Electrical Characteristics (continued)

Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| $V_{LS_GD}$ | NFET gate drive output voltage | $V_{CSN/OUT}$ = 5.1 V, $C_G$ = 1000 pF (see Figure 9-2) | 9.5 | 11 | 12.5 | V |
| $I_{LS_DR_SRC}$ | NFET gate drive output source current | $V_{CSN/OUT}$ = 5.1 V, $C_G$ = 1000 pF | 2 | 3 | 4 | µA |
| $I_{LS_DR_SNK}$ | NFET gate drive output sink current | $V_{CSN/OUT}$ = 5.1 V, $C_G$ = 1000 pF | 20 | 35 | 50 | µA |
| $V_{LS_GD_UVLO_R}$ | $V_{CSN/OUT}$ rising threshold for LS_GD operation | $V_{CSN/OUT}$ rising | 2.85 | 3 | 3.15 | V |
| $V_{LS_GD_UVLO_HYS}$ | Hysteresis | | | 80 | | mV |
| **BUS DISCHARGE (BUS PIN)** | | | | | | |
| $V_{BUS_OV}$ | Rising threshold for BUS pin overvoltage protection | $V_{BUS}$ rising | 6.6 | 7 | 7.3 | V |
| $V_{BUS_OV_HYS}$ | Hysteresis | | | 180 | | mV |
| $R_{BUS_DCHG_18V}$ | Discharge resistance for BUS | $V_{BUS}$ = 18V, measure leakage current | | 29 | | kΩ |
| $R_{BUS_DCHG_8V}$ | Discharge resistance for BUS | $V_{BUS}$ = 8V, measure leakage current | | 35 | | kΩ |
| **$\overline{\text{FAULT}}$, $\overline{\text{BUCK_ST}}$** | | | | | | |
| $V_{OL}$ | FAULT Output low voltage | $I_{SNK_PIN}$ = 0.5 mA | | | 250 | mV |
| $I_{OFF}$ | FAULT Off-state leakage | $V_{PIN}$ = 5.5 V | | | 1 | µA |
| $V_{OL}$ | $\overline{\text{BUCK_ST}}$ Output low voltage | $I_{SNK_PIN}$ = 0.5 mA | | | 250 | mV |
| $I_{OFF}$ | $\overline{\text{BUCK_ST}}$ Off-state leakage | $V_{PIN}$ = 5.5 V | | | 1 | µA |
| **CTRL1, CTRL2 - LOGIC INPUTS** | | | | | | |
| $V_{IH}$ | Rising threshold voltage | | | 1.48 | 2 | V |
| $V_{IL}$ | Falling threshold voltage | | 0.85 | 1.30 | | V |
| $V_{HYS}$ | Hysteresis | | | 180 | | mV |
| $I_{IN}$ | Input current | | –1 | | 1 | µA |
| **DP_IN AND DM_IN OVERVOLTAGE PROTECTION** | | | | | | |
| $V_{Dx_IN_OV}$ | Rising threshold for Dx_IN overvoltage protection | DP_IN or DM_IN rising | 3.7 | 3.9 | 4.15 | V |
| | Hysteresis | | | 100 | | mV |
| $R_{Dx_IN_DCHG_18V}$ | Discharge resistance for Dx_IN | VDx_IN = 18V, measure leakage current | | 94 | | kΩ |
| $R_{Dx_IN_DCHG_5V}$ | Discharge resistance for Dx_IN | VDx_IN = 5V, measure leakage current | | 416 | | kΩ |
| **HIGH-BANDWIDTH ANALOG SWITCH** | | | | | | |
| $R_{DS_ON}$ | DP and DM switch on-resistance | $V_{DP_OUT}$ = $V_{DM_OUT}$ = 0 V, $I_{DP_IN}$ = $I_{DM_IN}$ = 30 mA | | 3.4 | 6.3 | Ω |
| $R_{DS_ON}$ | DP and DM switch on-resistance | $V_{DP_OUT}$ = $V_{DM_OUT}$ = 2.4 V, $I_{DP_IN}$ = $I_{DM_IN}$ = –15 mA | | 4.3 | 7.7 | Ω |
| $\lvert\Delta R_{DS_ON}\rvert$ | Switch resistance mismatch between DP and DM channels | $V_{DP_OUT}$ = $V_{DM_OUT}$ = 0 V, $I_{DP_IN}$ = $I_{DM_IN}$ = 30 mA | | 0.05 | 0.15 | Ω |
| $\lvert\Delta R_{DS_ON}\rvert$ | Switch resistance mismatch between DP and DM channels | $V_{DP_OUT}$ = $V_{DM_OUT}$ = 2.4 V, $I_{DP_IN}$ = $I_{DM_IN}$ = –15 mA | | 0.05 | 0.15 | Ω |
| $C_{IO_OFF}$ | DP/DM switch off-state capacitance | $V_{EN}$ = 0 V, $V_{DP_IN}$ = $V_{DM_IN}$ = 0.3 V, Vac = 0.03 $V_{PP}$ , f = 1 MHz | | 6.7 | | pF |
| $C_{IO_ON}$ | DP/DM switch on-state capacitance | $V_{DP_IN}$ = $V_{DM_IN}$ = 0.3 V, Vac = 0.03 $V_{PP}$, f = 1 MHz | | 10 | | pF |
| $O_{IRR}$ | Off-state isolation | $V_{EN}$ = 0 V, f = 250 MHz | | 9 | | dB |
| $X_{TALK}$ | On-state cross-channel isolation | f = 250 MHz | | 29 | | dB |


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## 8.5 Electrical Characteristics (continued)

Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| $I_{lkg(OFF)}$ | Off-state leakage current, DP_OUT and DM_OUT | $V_{EN}$ = 0 V, $V_{DP_IN}$ = $V_{DM_IN}$ = 3.6 V, $V_{DP_OUT}$ = $V_{DM_OUT}$ = 0 V, measure $I_{DP_OUT}$ and $I_{DM_OUT}$ | | 0.1 | 1.5 | µA |
| BW | Bandwidth (–3 dB) | $R_L$ = 50 Ω | | 800 | | MHz |
| **CHARGING DOWNSTREAM PORT (CDP) DETECT** | | | | | | |
| $V_{DM_SRC}$ | DM_IN CDP output voltage | $V_{DP_IN}$ = 0.6 V, –250 µA < $I_{DM_IN}$ < 0 µA | 0.5 | 0.6 | 0.7 | V |
| $V_{DAT_REF}$ | DP_IN rising lower window threshold for $V_{DM_SRC}$ activation | | 0.36 | 0.38 | 0.4 | V |
| $V_{DAT_REF}$ | Hysteresis | | | 50 | | mV |
| $V_{LGC_SRC}$ | DP_IN rising upper window threshold for VDM_SRC deactivation | | 0.8 | 0.84 | 0.88 | V |
| $V_{LGC_SRC_HYS}$ | Hysteresis | | | 100 | | mV |
| $I_{DP_SINK}$ | DP_IN sink current | $V_{DP_IN}$ = 0.6 V | 40 | 70 | 100 | µA |
| **RT/SYNC THRESHOLD (RT/SYNC PIN)** | | | | | | |
| $V_{IH_RT/SYNC}$ | RT/SYNC high threshold for external clock synchronization | Amplitude of SYNC clock AC signal (measured at SYNC pin) | 3.5 | | | V |
| $V_{IL_RT/SYNC}$ | RT/SYNC low threshold for external clock synchronization | Amplitude of SYNC clock AC signal (measured at SYNC pin) | | | 0.8 | V |
| **THERMAL SHUTDOWN** | | | | | | |
| $T_{SD}$ | Thermal shutdown | Shutdown threshold | | 160 | | °C |
| | | Recovery threshold | | 140 | | °C |

## 8.6 Timing Requirements

Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| | | | MIN | NOM | MAX | UNIT |
|---|---|---|---|---|---|---|
| **SYNC (RT/SYNC PIN) WITH EXTERNAL CLOCK** | | | | | | |
| $f_{SYNC}$ | Switching frequency using external clock on RT/SYNC pin | | 300 | | 2300 | kHz |
| $T_{SYNC_MIN}$ | Minimum SYNC input pulse width | $f_{SYNC}$ = 400 kHz, $V_{RT/SYNC}$ > $V_{IH_RT/SYNC}$, $V_{RT/SYNC}$ < $V_{IL_RT/SYNC}$ | | 100 | | ns |
| $T_{LOCK_IN}$ | PLL lock time | | | 100 | | µs |

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## 8.7 Switching Characteristics

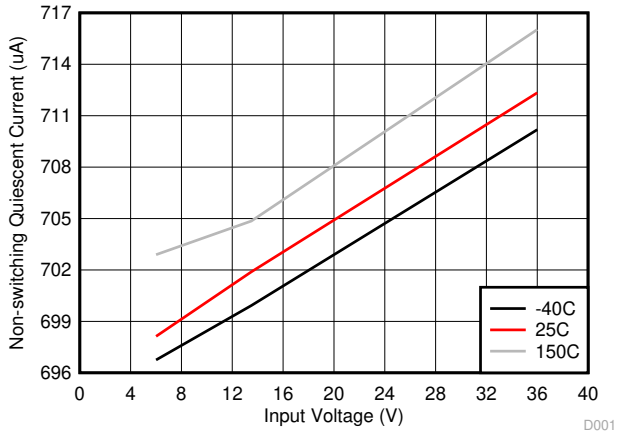
Limits apply over the junction temperature ($T_J$) range of -40°C to +150°C; $V_{IN}$ = 13.5 V, $f_{SW}$ =400 kHz, $C_{VCC}$ = 2.2 µF, $R_{SNS}$ = 15 mΩ, $R_{IMON}$ = 13 kΩ, $R_{ILIMIT}$= 13 kΩ, $R_{SET}$= 300 Ω unless otherwise stated. Minimum and maximum limits are specified through test, design or statistical correlation. Typical values represent the most likely parametric norm at $T_J$ = 25°C, and are provided for reference purposes only.

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| **SOFT START** | | | | | | |
| $T_{SS}$ | Internal soft-start time | The time of internal reference to increase from 0 V to 1.0 V | 3 | 5 | 7 | ms |
| **HICCUP MODE** | | | | | | |
| $N_{OC}$ | Number of cycles that LS current limit is tripped to enter Hiccup mode | | | 128 | | Cycles |
| $T_{OC}$ | Hiccup retry delay time | | | 118 | | ms |
| **SW (SW PIN)** | | | | | | |
| $T_{ON_MIN}$ | Minimum turnon-time | | | 105 | | ns |
| $T_{ON_MAX}$ | Maximum turnon-time, HS timeout in dropout | | | 7.5 | | µs |
| $T_{OFF_MIN}$ | Minimum turnoff time | | | 80 | | ns |
| $D_{max}$ | Maximum switch duty cycle | | | 98 | | % |
| **TIMING RESISTOR AND INTERNAL CLOCK** | | | | | | |
| $f_{SW_RANGE}$ | Switching frequency range using RT mode | | 300 | | 2300 | kHz |
| $f_{SW}$ | Switching frequency | $R_T$ = 49.9 kΩ | 360 | 400 | 440 | kHz |
| | Switching frequency | $R_T$ = 8.87 kΩ | 1953 | 2100 | 2247 | kHz |
| $FS_{SS}$ | Frequency span of spread spectrum operation | | | ±6 | | % |
| **NFET DRIVER** | | | | | | |
| $t_r$ | $V_{LS_DR}$ rise time | VOUT = 5.1 V, NFET = CSD87502Q2, time from LS_GD 10% to 90% | | 1000 | | µs |
| $t_f$ | $V_{LS_DR}$ fall time | VOUT = 5.1 V, NFET = CSD87502Q2, time from LS_GD time 90% to 10% | | 100 | | µs |
| **CURRENT LIMIT - EXTERNAL NFET CONNECTED BETWEEN CSN/OUT AND BUS, LS_GD CONNECTED TO FET GATE** | | | | | | |
| $t_{OC_HIC_ON}$ | ON-time during hiccup mode | | | 2 | | ms |
| $t_{OC_HIC_OFF}$ | OFF-time during hiccup mode | | | 263 | | ms |
| **$\overline{\text{FAULT}}$ DUE TO VBUS OC, VBUS OV, DP OV, DM OV** | | | | | | |
| $t_{DEGLA}$ | Asserting deglitch time | | 5.5 | 8.2 | 11.5 | ms |
| $t_{DEGLD}$ | De-asserting deglitch time | | 5.5 | 8.2 | 11.5 | ms |
| **$\overline{\text{BUCK_ST}}$** | | | | | | |
| $t_{DEGLA}$ | Asserting deglitch time | | 88 | 150 | 220 | ms |
| **HIGH-BANDWIDTH ANALOG SWITCH** | | | | | | |
| $t_{pd}$ | Analog switch propagation delay | | | 0.14 | | ns |
| $t_{SK}$ | Analog switch skew between opposite transitions of the same port ($t_{PHL}$ – $t_{PLH}$) | | | 0.02 | | ns |
| $t_{OV_Dn}$ | DP_IN and DM_IN overvoltage protection response time | | | 2 | | µs |
| $t_{ST_DEG_Dn}$ | Deglitch time from Vcc > 4V to DP / DM data switch turn on | | 88 | 150 | 220 | ms |


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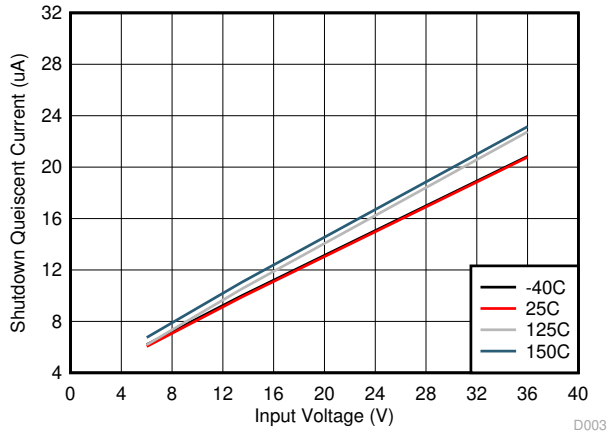
## 8.8 Typical Characteristics

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.




$V_{CSN}$ = 8 V INT = 5.1 kΩ

**Figure 8-1. Non-Switching Quiescent Current**




EN = 0 V

**Figure 8-2. Shutdown Quiescent Current**

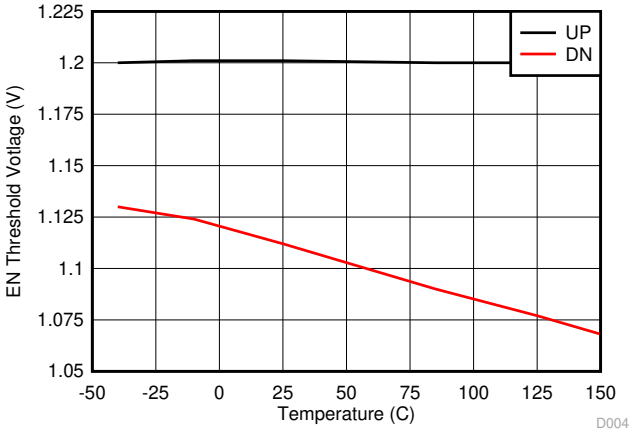



**Figure 8-3. Precision Enable Threshold**

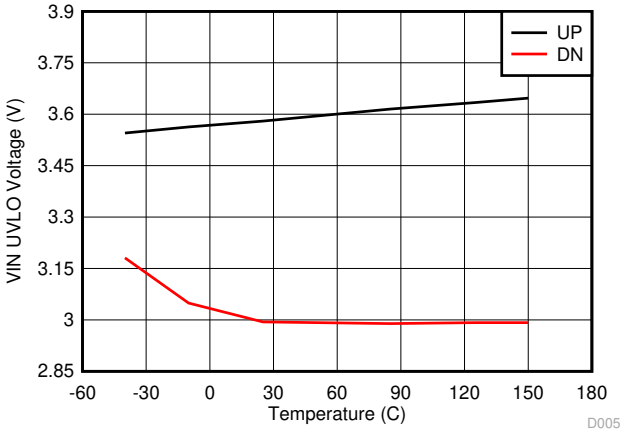



**Figure 8-4. VIN UVLO Threshold**

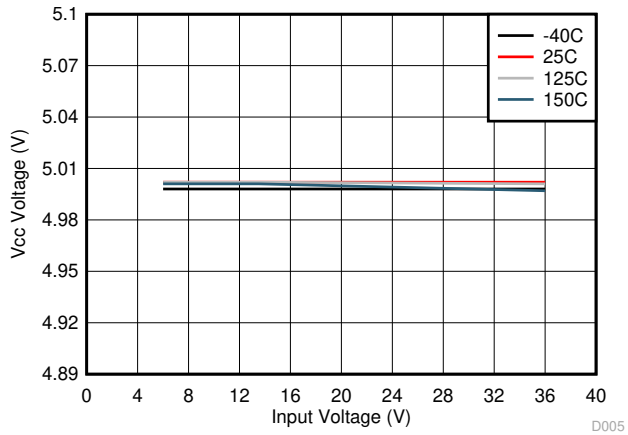
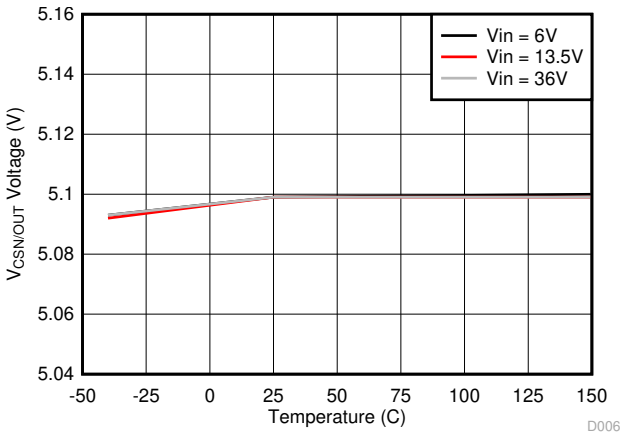



**Figure 8-5. VCC vs Input Voltage**




$R_{IMON}$ = 0 Ω

**Figure 8-6. $V_{CSN/OUT}$ Voltage vs Junction Temperature**

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## 8.8 Typical Characteristics (continued)

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.

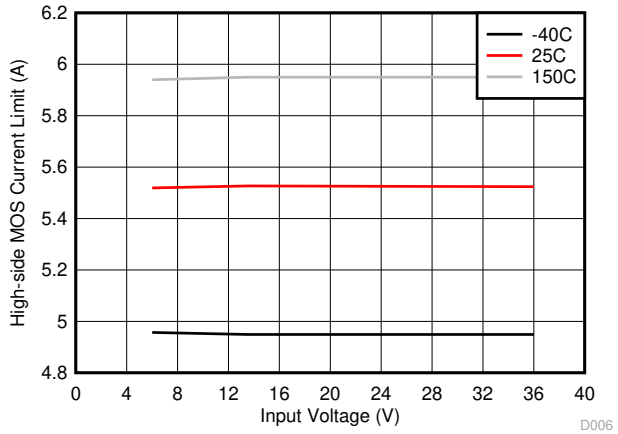



**Figure 8-7. High-side Current Limit vs Input Voltage**

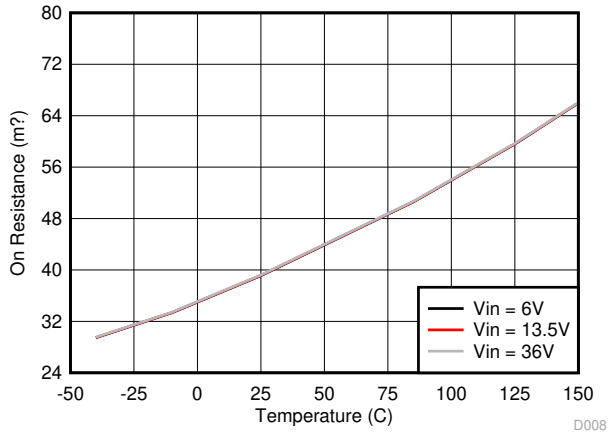



**Figure 8-8. High-side MOSFET on Resistance vs Junction Temperature**

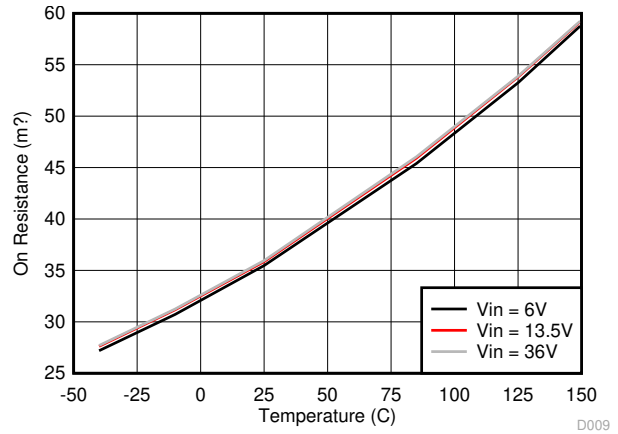
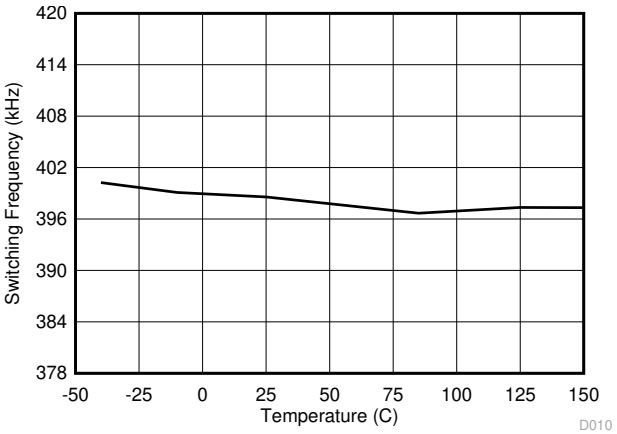
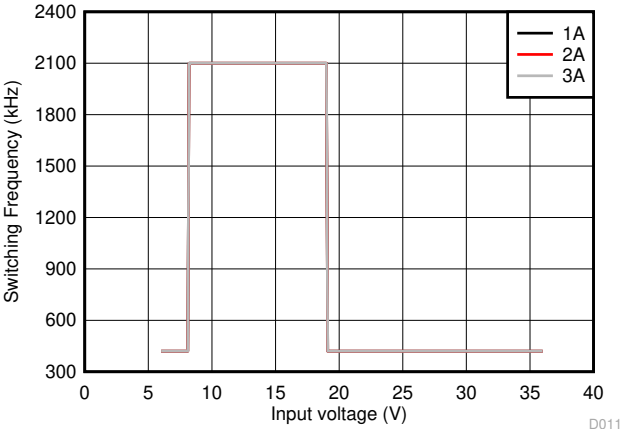



**Figure 8-9. Low-side MOSFET on Resistance vs Junction Temperature**



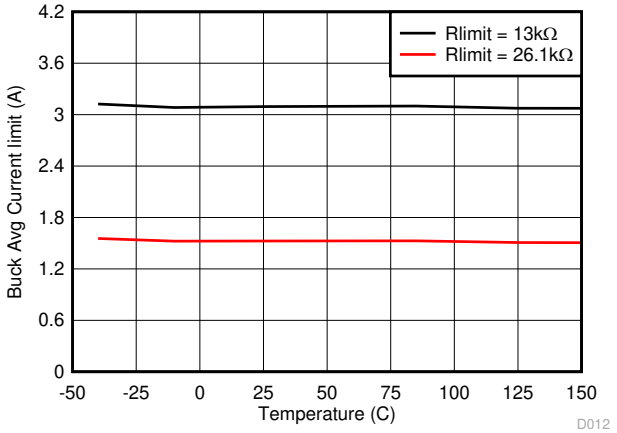

$R_T$ = 49.9 kΩ

**Figure 8-10. Switching Frequency vs Junction Temperature**




$R_T$ = 8.87 kΩ

**Figure 8-11. Switching Frequency vs VIN Voltage**



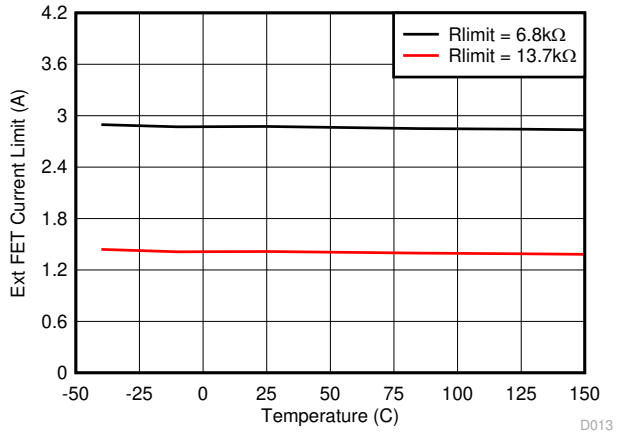

$R_{SNS}$ = 15 mΩ $R_{SET}$ = 300 Ω

**Figure 8-12. Buck Average Current Limit vs Junction Temperature**


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## 8.8 Typical Characteristics (continued)

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.




$R_{SNS}$ = 15 mΩ $R_{SET}$ = 300 Ω

**Figure 8-13. External FET Current Limit vs Junction Temperature**

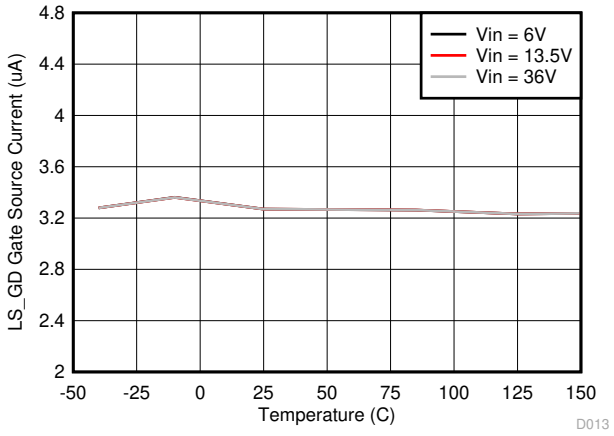
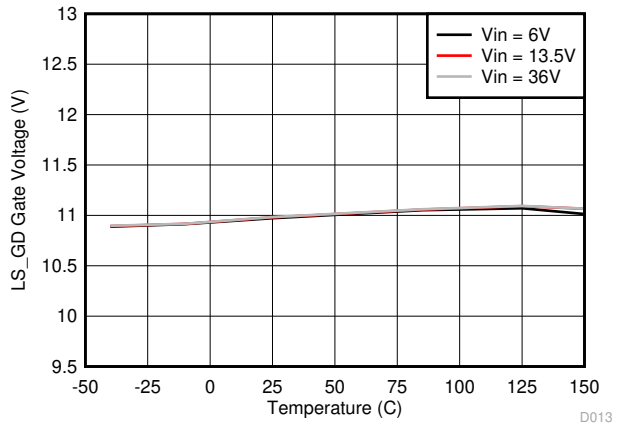
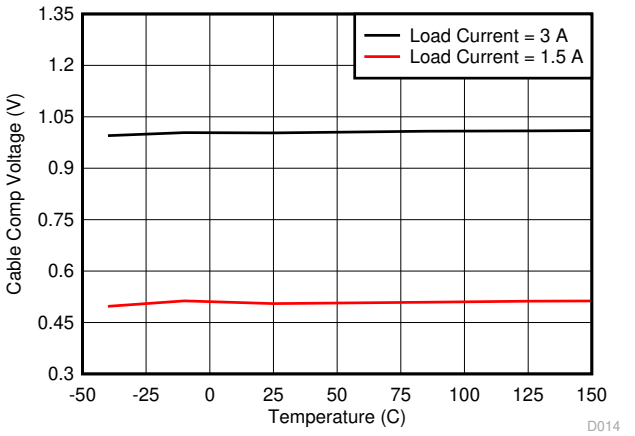



**Figure 8-14. LS_GD Gate Source Current vs Junction Temperature**



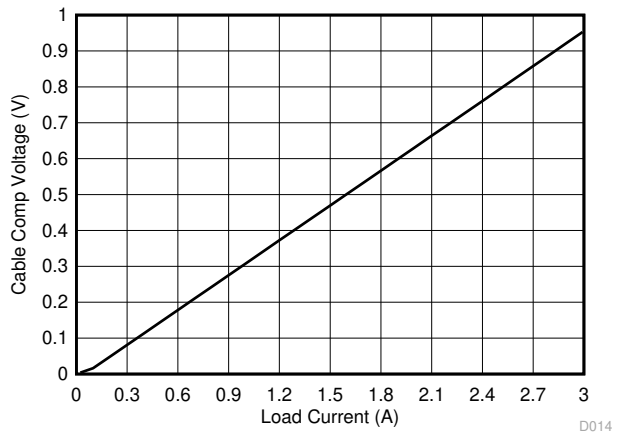

$V_{CSN/OUT}$ = 5.1 V $R_{IMON}$ = 0 kΩ

**Figure 8-15. LS_GD Gate Voltage vs Junction Temperature**




$R_{SNS}$ = 15 mΩ $R_{SET}$ = 300 Ω $R_{IMON}$ = 13 kΩ

**Figure 8-16. Cable Compensation Voltage vs Junction Temperature**




$R_{SNS}$ = 15 mΩ $R_{SET}$ = 300 Ω $R_{IMON}$ = 13 kΩ

**Figure 8-17. Cable Compensation Voltage vs Load Current**

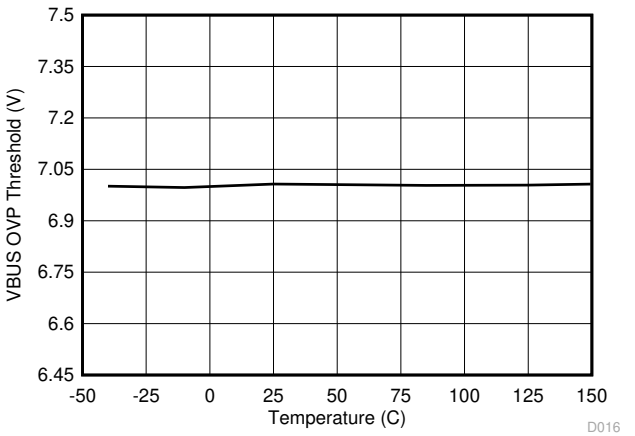



**Figure 8-18. $V_{BUS}$ Overvoltage Protection Threshold vs Junction Temperature**

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## 8.8 Typical Characteristics (continued)

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.

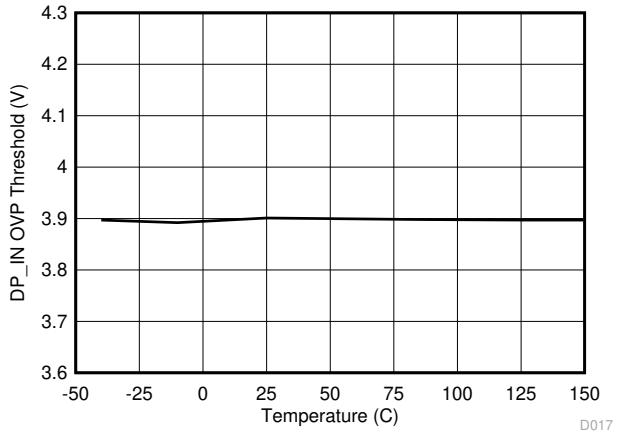



**Figure 8-19. DP_IN Overvoltage Protection Threshold vs Junction Temperature**

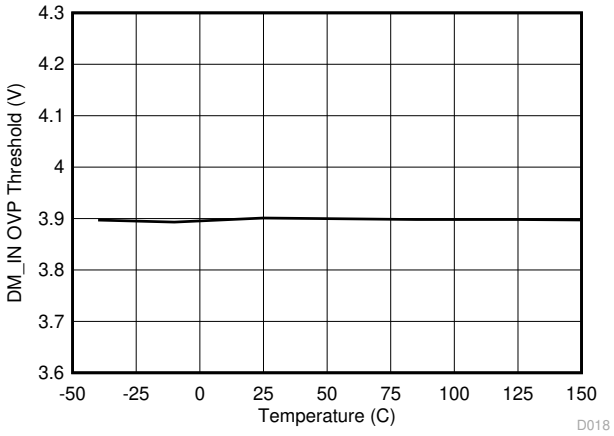
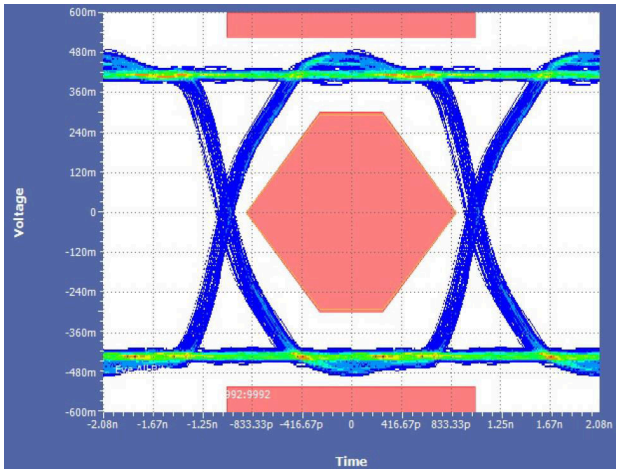
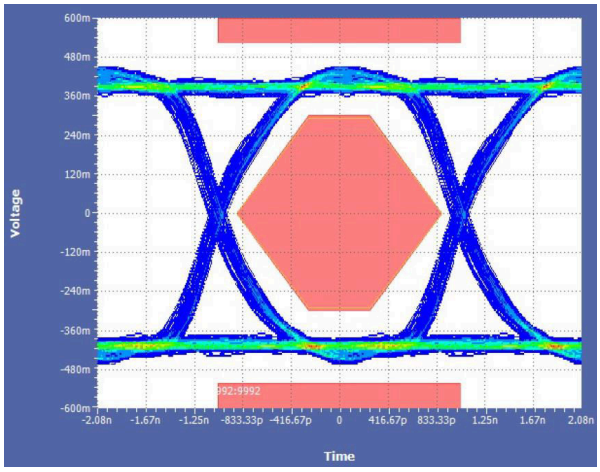



**Figure 8-20. DM_IN Overvoltage Protection Threshold vs Junction Temperature**




Measured source with 10-cm cable

**Figure 8-21. Bypassing the TPS2584x-Q1 Data Switch**




Measured on TPS25830-Q1 EVM with 10-cm cable

**Figure 8-22. Through the TPS2584x-Q1 Data Switch**




**Figure 8-23. Data Transmission Characteristics vs Frequency**




**Figure 8-24. Off-state Data-Switch Isolation vs Frequency**


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## 8.8 Typical Characteristics (continued)

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.

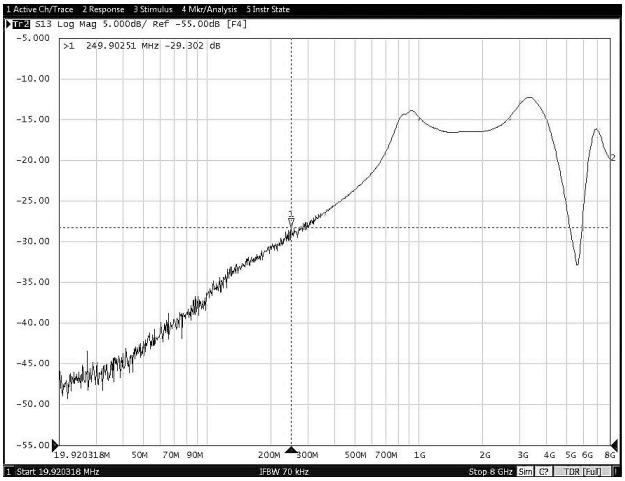



**Figure 8-25. On-state Cross-channel Isolation vs Frequency**

# 9 Parameter Measurement Information

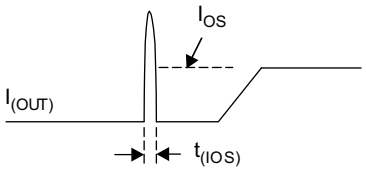



Figure 9-1. Short-Circuit Parameters

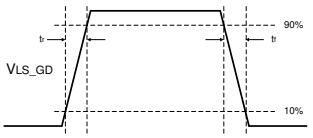



Figure 9-2. NFET Gate Drive Rise and Fall Time

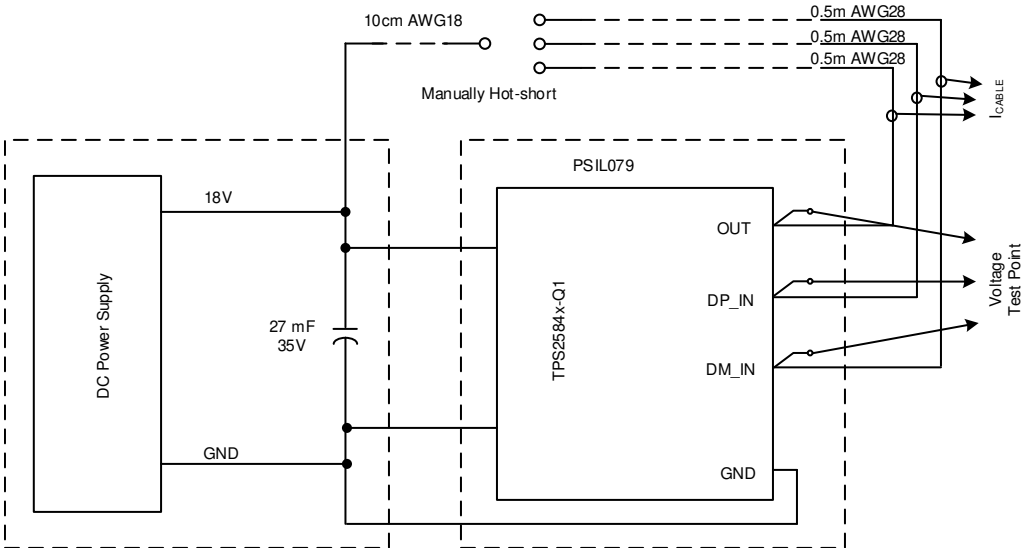



Figure 9-3. Short-to-Battery System Test Setup


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# 10 Detailed Description

## 10.1 Overview

The TPS2584x-Q1 devices are full-featured solutions for implementing a compact USB charging port with support for Type-A BC1.2 standards. Both devices contain an efficient 36-V buck regulator power source capable of providing up to 3.5 A of output current at 5.10 V (nominal). System designers can optimize efficiency or solution size through careful selection of switching frequency over the range of 300 to 2200 kHz with sufficient margin to operate above or below the AM radio frequency band. In all versions the buck regulator operates in forced PWM mode ensuring fixed switching frequency regardless of load current. Spread-spectrum feature aid reducing harmonic peaks of the switching frequency potentially simplifying EMI filter design and easing compliance.

Current sensing via a precision high-side current sense amplifier enables an accurate, user programmable overcurrent limit setting; and programmable linear cable compensation to overcome IR losses when powering remote USB ports.

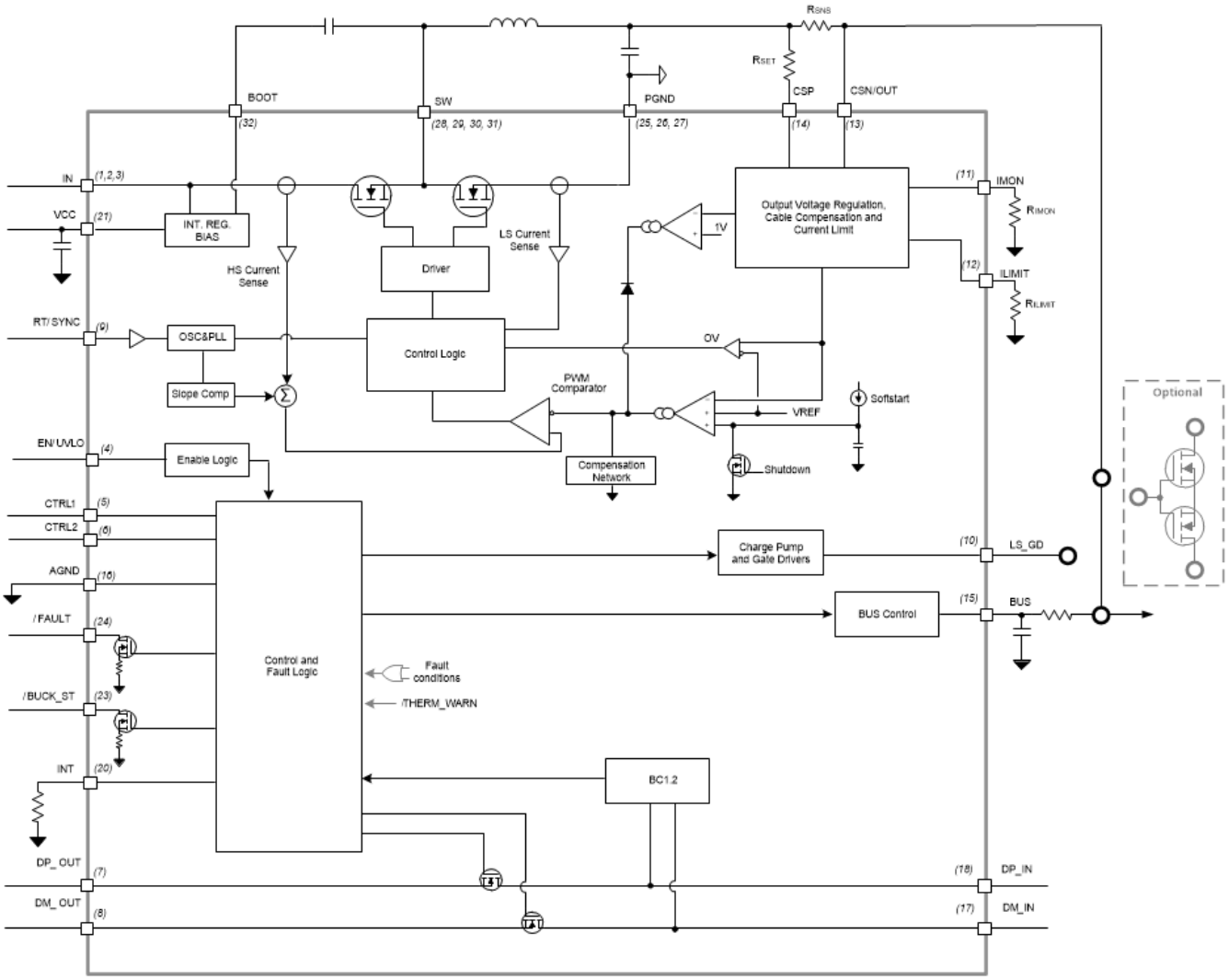
The CTRL1 and CTRL2 pins set the operating mode for the TPS2584x-Q1 device. The device can support CDP, SDP or Client configurations.

The TPS25840-Q1 integrates high band-width (800 MHz) USB switches, includes short to $V_{BAT}$ and short to $V_{BUS}$ protection as well as IEC61000-4-2 electrostatic discharge clamps to protect the host from potentially damaging overvoltage conditions.

The TPS25842-Q1 integrates high band-width (800 MHz) USB switches, includes short to $V_{BUS}$ protection as well as IEC61000-4-2 electrostatic discharge clamps, but does not support short to $V_{BAT}$ protection.

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## 10.2 Functional Block Diagram




## 10.3 Feature Description

### 10.3.1 Buck Regulator

The following operating description of the TPS2584x-Q1 refers to the *Functional Block Diagram* and the waveforms in Figure 10-1. TPS2584x-Q1 is a step-down synchronous buck regulator with integrated high-side (HS) and low-side (LS) switches (synchronous rectifier). The TPS2584x-Q1 supplies a regulated output voltage by turning on the HS and LS NMOS switches with controlled duty cycle. During high-side switch ON time, the SW pin voltage swings up to approximately $V_{IN}$, and the inductor current $i_L$ increase with linear slope ($V_{IN}$ – $V_{OUT}$) / L. When the HS switch is turned off by the control logic, the LS switch is turned on after an anti-shoot-through dead time. Inductor current discharges through the LS switch with a slope of –$V_{OUT}$ / L. The control parameter of a buck converter is defined as Duty Cycle D = $t_{ON}$ / $T_{SW}$, where $t_{ON}$ is the high-side switch ON time and $T_{SW}$ is the switching period. The regulator control loop maintains a constant output voltage by adjusting the duty cycle D. In an ideal buck converter, where losses are ignored, D is proportional to the output voltage and inversely proportional to the input voltage: D = $V_{OUT}$ / $V_{IN}$.


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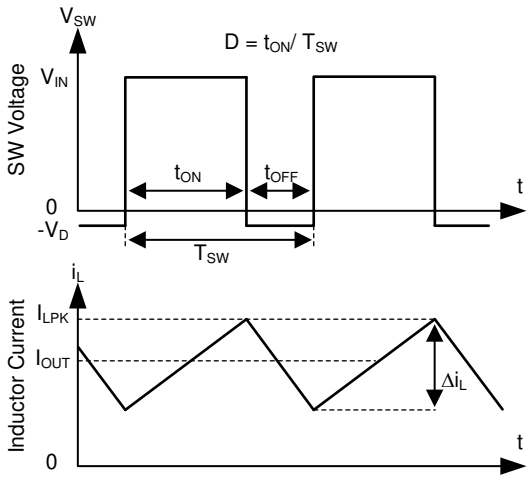



**Figure 10-1. SW Node and Inductor Current Waveforms in Continuous Conduction Mode (CCM)**

The TPS2584x-Q1 employs fixed frequency peak current mode control. A voltage feedback loop is used to get accurate DC voltage regulation by adjusting the peak current command based on voltage offset. The peak inductor current is sensed from the high-side switch and compared to the peak current threshold to control the ON time of the high-side switch. The voltage feedback loop is internally compensated, which allows for fewer external components, makes it easy to design, and provides stable operation with almost any combination of output capacitors. TPS2584x-Q1 operates in FPWM mode for low output voltage ripple, tight output voltage regulation, and constant switching frequency.

### 10.3.2 Enable/UVLO

The voltage on the EN/UVLO pin controls the ON or OFF operation of TPS2584x-Q1. An EN/UVLO pin voltage higher than $V_{EN/UVLO-VOUT-H}$ is required to start the internal regulator (Assume 5.1-k pull down resister on INT pin). The EN/UVLO pin is an input and can not be left open or floating. The simplest way to enable the operation of the TPS2584x-Q1 is to connect the EN to $V_{IN}$. This action allows self-start-up of the TPS2584x-Q1 when $V_{IN}$ is within the operation range.

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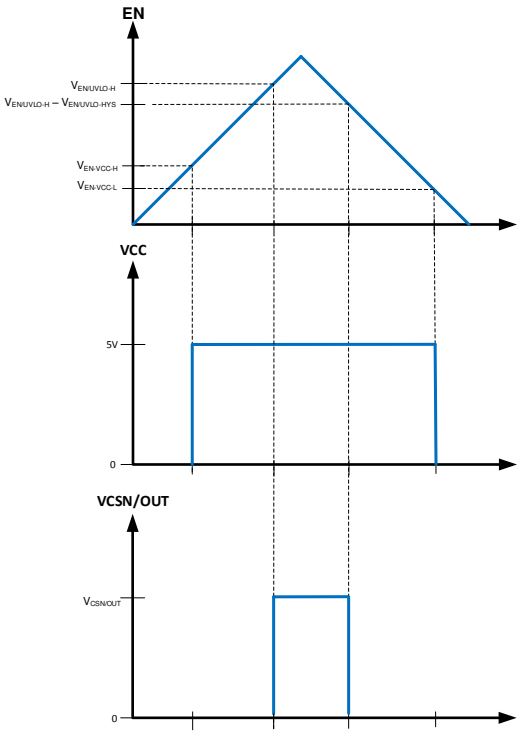



**Figure 10-2. Precision Enable Behavior**

Many applications benefit from the employment of an enable divider $R_{ENT}$ and $R_{ENB}$ (Figure 10-3) to establish a precision system UVLO level for the TPS2584x-Q1. System UVLO can be used for sequencing, ensuring reliable operation, or supply protection, such as a battery discharge level. To ensure the USB port $V_{BUS}$ is within the 5-V operating range as required for USB compliance (for the latest USB specifications and requirements, refer to *USB.org*), TI suggests that the $R_{ENT}$ and $R_{ENB}$ resistors be chosen such that the TPS2584x-Q1 enables when $V_{IN}$ is approximately 6 V. Considering the drop out voltage of the buck regulator and IR loses in the system, 6 V provides adequate margin to maintain $V_{BUS}$ within USB specifications. If system requirements such as a warm crank (start) automotive scenario require operation with $V_{IN}$ < 6 V, the values of $R_{ENT}$ and $R_{ENB}$ can be calculated assuming a lower $V_{IN}$. An external logic signal can also be used to drive EN/UVLO input when a microcontroller is present and it is desirable to enable or disable the USB port remotely for other reasons.

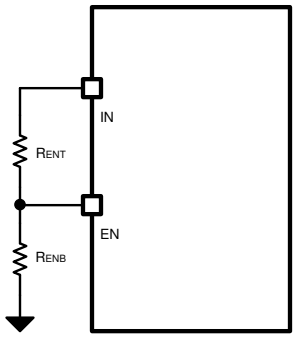



**Figure 10-3. System UVLO by Enable Divider**

UVLO configuration using external resistors is governed by the following equations:

$$R_{ENT} = (\frac{V_{IN(ON)}}{V_{EN/UVLO_H}} - 1) \times R_{ENB} \quad (1)$$


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$$V_{IN(OFF)} = V_{IN(ON)} \times (1 - \frac{V_{EN/UVLO_HYS}}{V_{EN/UVLO_H}}) \quad (2)$$

Example:

$V_{IN(ON)}$ = 6 V (user choice)

$R_{ENB}$ = 5 kΩ (user choice)

$R_{ENT}$ = [($V_{IN(ON)}$ / $V_{EN/UVLO_H}$) – 1] × $R_{ENB}$= 19.6 kΩ. Choose standard 20 kΩ.

Therefore, $V_{IN(OFF)}$ = 6 V × [1 – (0.09 V / 1.2 V)] = 5.55 V

A typical start-up waveform is shown in Figure 10-4. The rise time of DCDC VBUS voltage is about 5 ms.

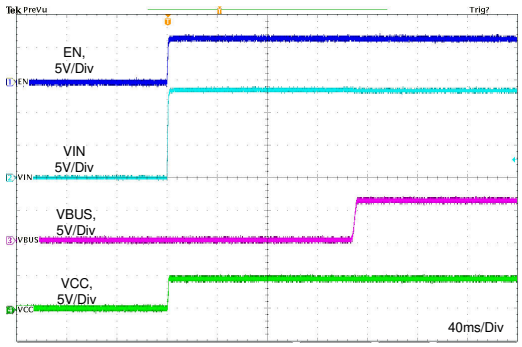



**Figure 10-4. Typical Start-up Behavior, $V_{IN}$ = 13.5 V, $R_{IMON}$ = 12.6 kΩ**

For TPS2584x-Q1, the pin voltage must meet the requirement below during startup. See Figure 10-5.

- $V_{BUS}$ < 0.8 V (typical)
- $V_{DX_OUT}$ < 2.2 V (typical)
- $V_{DX_IN}$ < 1.5 V (typical)

After the 150-ms deglitch time, no additional requirement on these pins. In real application, $\overline{\text{BUCK_ST}}$ pin can be used to configure the timing sequence.

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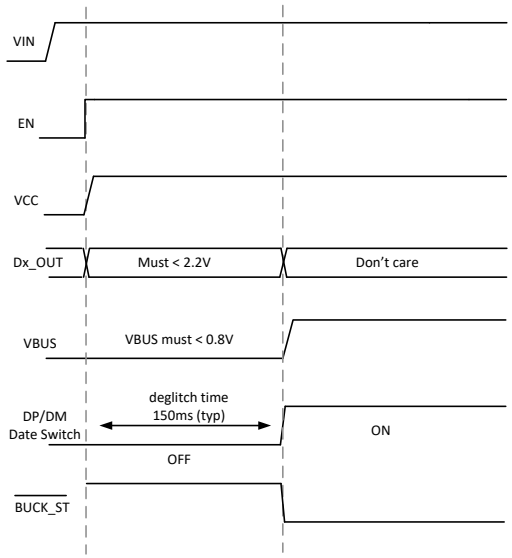



**Figure 10-5. TPS2584x-Q1 Pin Voltage Requirement During Startup**

### 10.3.3 Switching Frequency and Synchronization (RT/SYNC)

The switching frequency of the TPS2584x-Q1 can be programmed by the resistor $R_T$ from the RT/SYNC pin and GND pin. Use Equation 3 to determine the RT resistance for a given switching frequency.

$$R_{FREQ}\left(k\Omega\right) = 26660 \times f_{SW}{}^{-1.0483}\left(kHz\right) \quad (3)$$

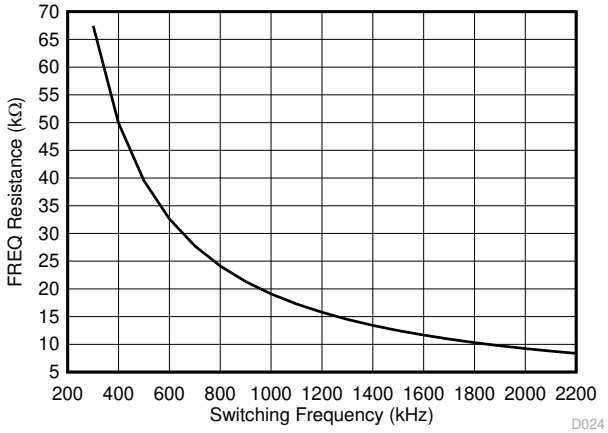



**Figure 10-6. RT Set Resistor vs Switching Frequency**

Table 10-1 lists typical RT resistors values.

**Table 10-1. Setting the Switching Frequency With RT**

| RT (kΩ) | SWITCHING FREQUENCY (kHz) |
|---|---|
| 68.1 | 300 |
| 49.9 | 400 |
| 39.2 | 500 |
| 19.1 | 1000 |
| 12.4 | 1500 |
| 9.31 | 2000 |
| 8.87 | 2100 |


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**Table 10-1. Setting the Switching Frequency With RT (continued)**

| RT (kΩ) | SWITCHING FREQUENCY (kHz) |
|---|---|
| 8.45 | 2200 |

TPS2584x-Q1 switching action can be synchronized to an external clock from 300 kHz to 2.3 MHz. The RT/SYNC pin can be used to synchronize the internal oscillator to an external clock. The internal oscillator can be synchronized by AC coupling a positive edge into the RT/SYNC pin. The AC coupled peak-to-peak voltage at the RT/SYNC pin must exceed the SYNC amplitude threshold of 3.5 V (typical) to trip the internal synchronization pulse detector, and the minimum SYNC clock ON and OFF time must be longer than 100 ns (typical). When using a low impedance signal source, the frequency setting resistor, $R_T$, is connected in parallel with an AC coupling capacitor, $C_{COUP}$, to a termination resistor, $R_{TERM}$ (for example: 50 Ω). The two resistors in series provide the default frequency setting resistance when the signal source is turned off. A 10-pF ceramic capacitor can be used for $C_{COUP}$. Figure 10-7 show the device synchronized to an external clock.

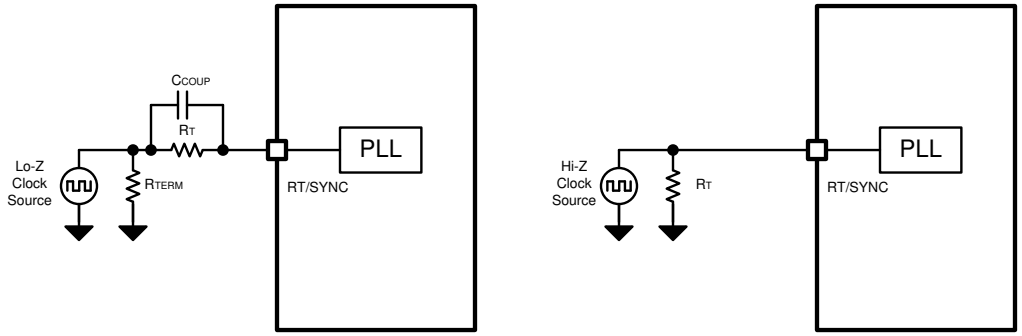



**Figure 10-7. Synchronize to External Clock**

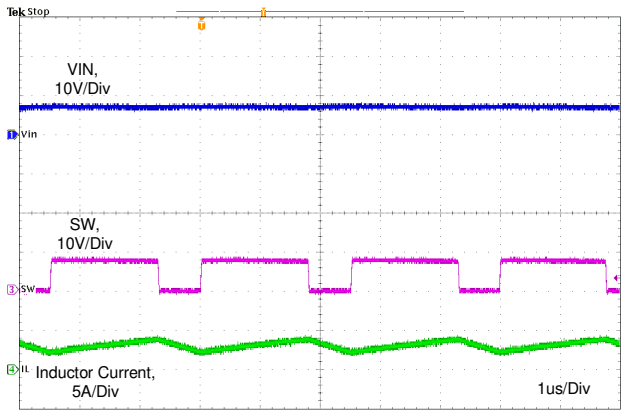
To avoid AM radio frequency brand and maintain proper regulation when minimum ON-time or minimum OFF-time is reached, the TPS2584x-Q1 implement frequency foldback scheme depends on VIN voltage. Refer to Figure 8-10.

- When 8 V < VIN ≤ 19 V, the switching frequency of TPS2584x-Q1 is determined by $R_T$ resistor or external sync clock.
- When VIN ≤ 8 V, the switching frequency of TPS2584x-Q1 is set to default 420 kHz, regardless of $R_T$ resistor setting or external sync clock.
- When VIN > 19 V, the switching frequency of TPS2584x-Q1 is set to default 420 kHz, regardless of $R_T$ resistor setting or external sync clock.

figure 10-8, figure 10-9 and Figure 10-10 show the device switching frequency and behavior under different VIN voltage and $R_T$ = 8.87 kΩ.

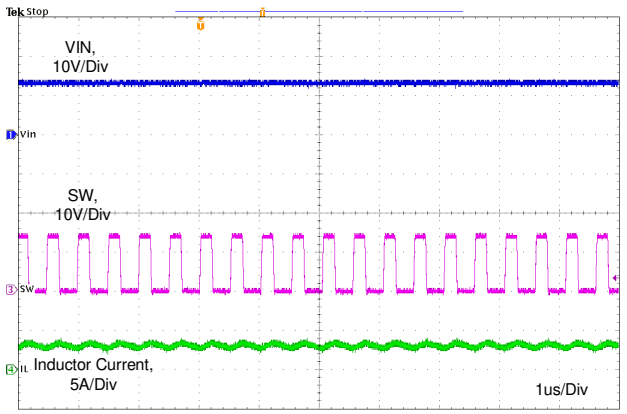
Figure 10-11, Figure 10-12 and Figure 10-13 show the device switching frequency and behavior under different VIN voltage and synchronized to an external 2.1-M system clock.

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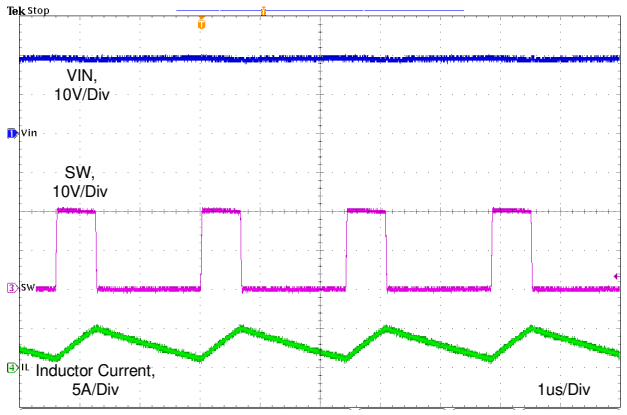

VIN = 7.5 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-8. Switching Frequency When $R_T$ = 8.87 kΩ**



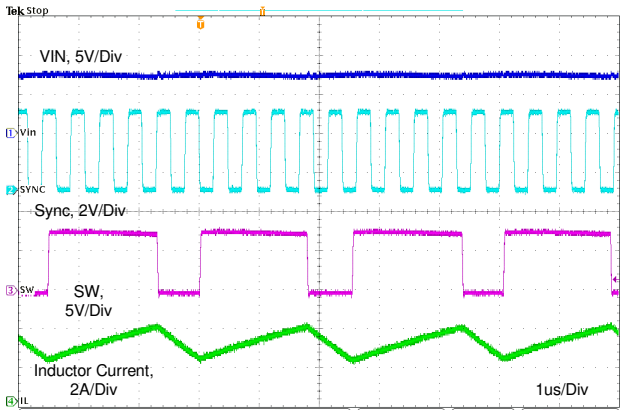

VIN = 13.5 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-9. Switching Frequency When $R_T$ = 8.87 kΩ**



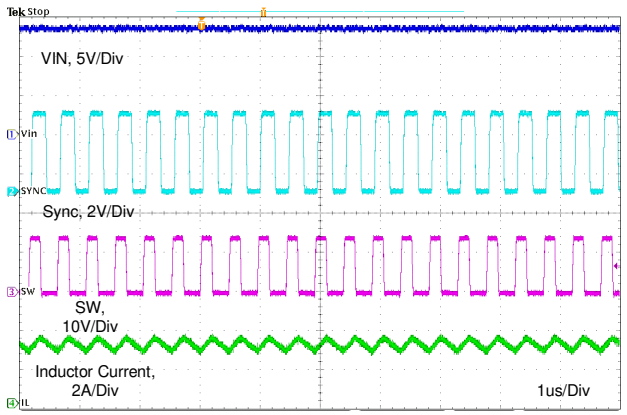

VIN = 20 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-10. Switching Frequency When $R_T$ = 8.87 kΩ**



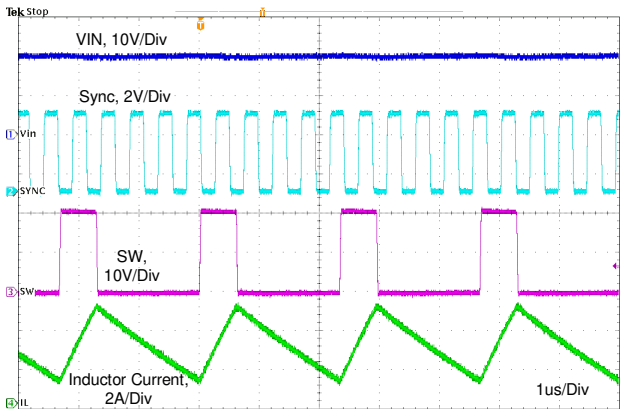

VIN = 7.5 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-11. Synchronizing to External 2.1-MHz Clock**




VIN = 13.5 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-12. Synchronizing to External 2.1-MHz Clock**




VIN = 20 V L = 2.2 uH $I_{LOAD}$ = 3 A

**Figure 10-13. Synchronizing to External 2.1-MHz Clock**


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### 10.3.4 Spread-Spectrum Operation

To reduce EMI, the TPS2584x-Q1 introduce frequency spread spectrum. The spread spectrum is used to eliminate peak emissions at specific frequencies by spreading emissions across a wider range of frequencies than a part with fixed frequency operation. In most systems, low frequency conducted emissions from the first few harmonics of the switching frequency can be easily filtered. A more difficult design criterion is reduction of emissions at higher harmonics which fall in the FM band. These harmonics often couple to the environment through electric fields around the switch node. The TPS2584x-Q1 devices use ±6% spread of switching frequencies with 1/256 swing frequency.

The spread spectrum function is only available when using the TPS2584x-Q1 internal oscillator. If the RT/SYNC pin is synchronized to an external clock, the spread spectrum function turns off.

### 10.3.5 VCC, VCC_UVLO

The TPS2584x-Q1 integrates an internal LDO to generate $V_{CC}$ for control circuitry and MOSFET drivers. The nominal voltage for $V_{CC}$ is 5 V. The $V_{CC}$ pin is the output of an LDO and must be properly bypassed. A high quality ceramic capacitor with a value of 2.2 µF to 4.7 µF, 10 V or higher rated voltage must be placed as close as possible to VCC and grounded to the PGND ground pin. The $V_{CC}$ output pin must not be loaded with more than 5 mA, or shorted to ground during operation. Shorting $V_{CC}$ to ground during operation can cause damage to the TPS2584x-Q1.

### 10.3.6 Minimum ON-time, Minimum OFF-time

Minimum ON-time, $T_{ON_MIN}$, is the smallest duration of time that the HS switch can be on. $T_{ON_MIN}$ is typically 105 ns in the TPS2584x-Q1. Minimum OFF-time, $T_{OFF_MIN}$, is the smallest duration that the HS switch can be off. $T_{OFF_MIN}$ is typically 80 ns in the TPS2584x-Q1. In CCM (FPWM) operation, $T_{ON_MIN}$ and $T_{OFF_MIN}$ limit the voltage conversion range given a selected switching frequency.

The minimum duty cycle allowed is:

$$D_{MIN} = T_{ON_MIN} \times f_{sw} \tag{4}$$

And the maximum duty cycle allowed is:

$$D_{MAX} = 1 - T_{OFF_MIN} \times f_{sw} \tag{5}$$

Given fixed $T_{ON_MIN}$ and $T_{OFF_MIN}$, the higher the switching frequency the narrower the range of the allowed duty cycle.

### 10.3.7 Internal Compensation

The TPS2584x-Q1 is internally compensated as shown in Figure 10-14. The internal compensation is designed such that the loop response is stable over the specified operating frequency and output voltage range. The TPS2584x-Q1 is optimized for transient response over the range 300 kHz ≤ fsw ≤ 2300 kHz.

### 10.3.8 Bootstrap Voltage (BOOT)

The TPS2584x-Q1 provides an integrated bootstrap voltage regulator. A small capacitor between the BOOT and SW pins provides the gate drive voltage for the high-side MOSFET. The BOOT capacitor is refreshed when the high-side MOSFET is off and the low-side switch conducts. The recommended value of the BOOT capacitor is 0.1 µF. TI recommends a ceramic capacitor with an X7R or X5R grade dielectric with a voltage rating of 10 V or higher for stable performance overtemperature and voltage.

### 10.3.9 $R_{SNS}$, $R_{SET}$, $R_{ILIMIT}$ and $R_{IMON}$

The programmable current limit threshold and full-scale cable compensation voltage are determined by the values of the $R_{SNS}$, $R_{SET}$, $R_{ILIMIT}$ and $R_{IMON}$ resistors. Refer to Figure 10-14.

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- $R_{SNS}$ is the current sense resistor. The recommended voltage across $R_{SNS}$ under current limit must be approximately 50 mV as a compromise between accuracy and power dissipation. For example, if current limiting is desired for $I_{OUT(MAX)} \geq 3.3$ A, then $R_{SNS} = 0.05$ V / 3.3 A = 0.01515 Ω. Choose a standard value of 15 mΩ.
- $R_{SET}$ determines the input current to the transconductance amplifier and current mirror. The amplifier balances the voltage to be equal to that across $R_{SNS}$. Choose a $R_{SET}$ value to produce an $I_{SET}$ current between 75–180 µA at the desired $I_{OUT(MAX)}$. Considering 50 mV across $R_{SET}$, a value of 300 Ω provides approximately 166 µA of $I_{SET}$ current to the amplifier and mirror circuit. Care must be taken to limit the $I_{SET}$ current below 200 µA to avoid saturating the internal amplifier circuit.
- $R_{ILIMIT}$ in conjuction with the 0.5 × $I_{SET}$ current produces a voltage on the ILIMIT pin which is proportional to the load current flowing in $R_{SNS}$. For details on setting the current limit, see *Current Limit Sensing Using $R_{ILIMIT}$*.
- $R_{IMON}$ in conjuction with the 0.5 × $I_{SET}$ current produces a voltage on the IMON pin which is proportional to the load current flowing in $R_{SNS}$. For details on setting the current limit, see *Cable Compensation*.

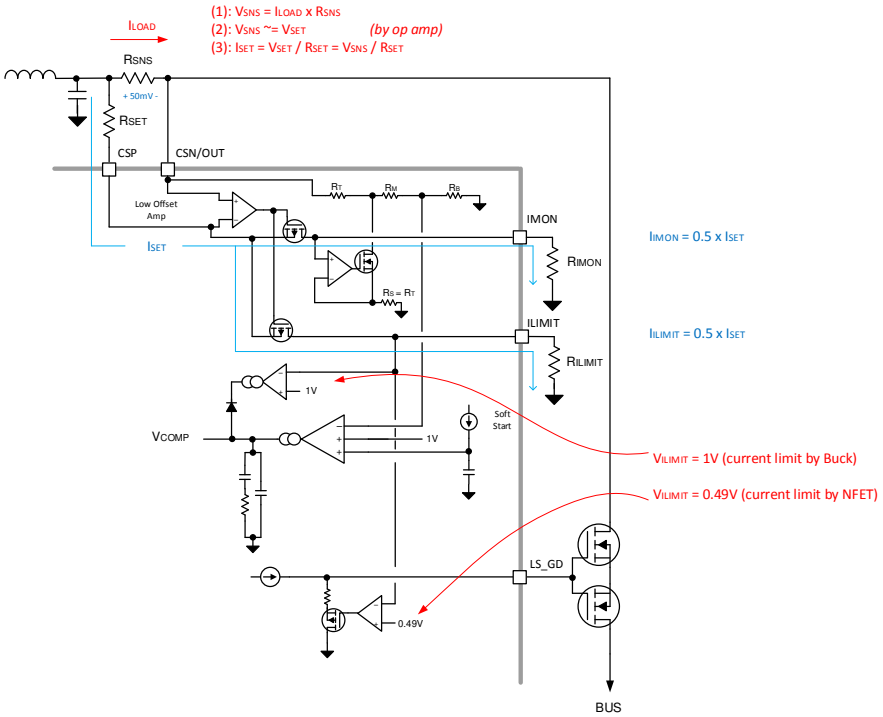



**Figure 10-14. Current Limit and Cable Compensation Circuit**

### 10.3.10 Overcurrent and Short Circuit Protection

For maximum versatility, TPS2584x-Q1 includes both a precision, programmable current limit as well cycle-by-cycle current limit to protect the USB port from extreme overload conditions. In most applications, the $R_{ILIMIT}$ resistor in conjunction with the selection of $R_{SNS}$ and $R_{SET}$ determines the overload threshold. The cycle-by-cycle current limit serves as a backup means of protection in the event $R_{ILMIT}$ is shorted to ground, disabling the programmable current limit function.

In some applications, the setting of TPS2584x-Q1 over-current need meets MFi requirement. For more details, please refer to the *How to Pass MFi Overcurrent Protection Test With USB Charger and Switch Device application report*.

#### *10.3.10.1 Current Limit Setting using $R_{ILIMIT}$*

Refer to Figure 10-14. The TPS2584x-Q1 can establish current limit by two methods.

- Using external a single or back-to-back N-Channel MOFETs between CSN/OUT and BUS: a voltage of 0.49 V on the ILIMIT pin initiates current limiting using the external MOSFET by decreasing the LS_GD voltage


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causing the FET to operate in the saturation region. To protect the MOSFETs from damage a hiccup timer limits the duty cycle to prevent thermal runaway. Refer to the *Specifications* for MOSFET hiccup timing.

- Buck average current limit: no MOSFET, CSN/OUT connected to BUS. The LS_GD must be pulled up through a 2.2-kΩ resistor. In this configuration, a voltage of 1 V across $R_{ILIMIT}$ on the ILIMIT pin initiates average current limiting of the buck regulator.

The 2-level current limit is described below:

- With external MOSFET Figure 10-15:
  - Isolating a fault on the USB port from other loads connected to the CSP output of the TPS2584x-Q1. In some applications, it can be useful to power additional circuitry (for example: USB HUB) from the output of the TPS2584x-Q1 and maintain operation of these circuits in the event of a short circuit downstream of the BUS pin. To prevent triggering the MOSFET current limit below the programmed ILIMIT threshold, external circuits must be supplied after the inductor and before the current sense resistor, $R_{SNS}$.
  - After $R_{SNS}$ and $R_{SET}$ are determined and the full load $I_{SET}$ current is known, the resistor value, $R_{ILIMIT}$, can be determined by:

$$R_{ILIMIT} = \frac{0.49 \times R_{SET}}{0.5 \times (I_{LIMIT} \times R_{SNS} + 0.0007)} \quad (6)$$

  - In most cases, the recommended voltage across $R_{SNS}$ under current limit must be approximately 50 mV as a compromise between accuracy and power dissipation. While in some application, $R_{ILIMIT}$ is the only resistor that can be changed to achieve different current limit. Typical $R_{ILIMIT}$ resistors value are listed in Table 10-2 given the condition $R_{SNS}$ = 15 mΩ and $R_{SET}$ = 300 Ω

**Table 10-2. Setting the Current Limit with $R_{ILIMIT}$**

| Current-Limit Threshold (mA) | $R_{ILIMIT}$ (kΩ) | |
|---|---|---|
| | With External MOSFET | Without External MOSFET |
| 700 | 26.1 | 53.6 |
| 1500 | 12.7 | 26.1 |
| 1700 | 11.3 | 22.6 |
| 2700 | 7.15 | 14.7 |
| 3000 | 6.49 | 13 |
| 3400 | 5.62 | 11.5 |
| 3800 | 5.11 | 10.5 |

- Buck Average Current Limit Figure 10-16:
  1. CSN/OUT connected directly to BUS, LS_GD must be pulled up through 2.2-kΩ resistor. The TPS2584x-Q1 can operate as a stand-alone USB charging port. In this configuration, the internal buck regulator operates with average current limiting as programmed by the ILIMIT pin, potentially producing less heat compared to N-channel MOSFET current limiting
  2. After $R_{SNS}$ and $R_{SET}$ are determined and the full load $I_{SET}$ current is known, the resistor value $R_{ILIMIT}$ can be determined by:

$$R_{ILIMIT} = \frac{1 \times R_{SET}}{0.5 \times (I_{LIMIT} \times R_{SNS} + 0.0007)} \quad (7)$$

  3. Table 10-2 lists Typical $R_{ILIMIT}$ resistors values given the condition $R_{SNS}$ = 15 mΩ and $R_{SET}$ = 300 Ω.

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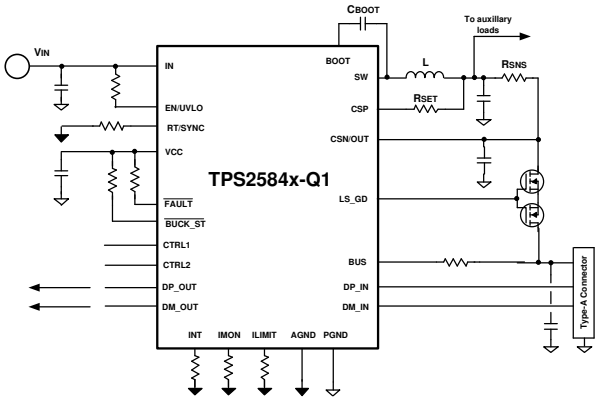



**Figure 10-15. Current Limit With External MOSFET**

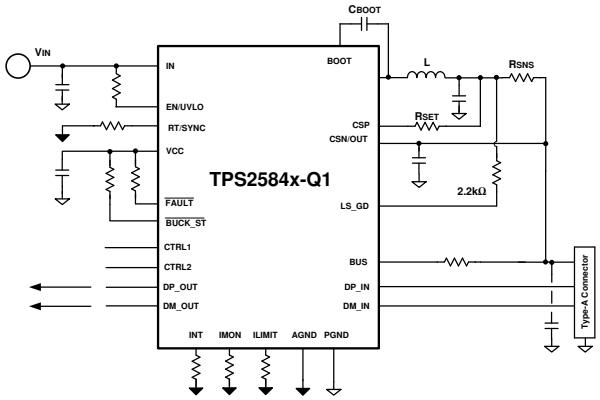



**Figure 10-16. Buck Average Current Limit**

#### *10.3.10.2 Buck Average Current Limit Design Example*

To start the procedure, the $I_{LOAD(MAX)}$, $R_{SNS}$ and $R_{SET}$, must be known.

1. Determine $I_{LIMIT}$, usually chose $I_{LIMIT}$= $I_{LOAD(MAX)}$ / (1 – 10%).
2. Determine $R_{SNS}$ to achieve 50 mV at current limit. For 3-A load current, choose $I_{LIMIT}$ = 3.3A. $R_{SNS}$ = (0.05 V / 3.3 A) = 15 mΩ.
3. Choose $R_{SET}$ = 300 Ω
4. According to Equation 7, $R_{LIMIT}$ = 300 / (0.5 × ( 3.3 X 0.015 + 0.0007)) = 11.95 kΩ.
5. Choose standard 11.8 kΩ.

#### *10.3.10.3 External MOSFET Gate Drivers*

The TPS2584x-Q1 has integrated NFET gate drivers, and can support current limit with external NFET. Refer to Figure 10-15.

The LS_GD pin of TPS2584x-Q1 can source 3-uA (typical) current to enhance the external MOSFET. A 6.2-V clamp between LS_GD and CSN/OUT pin limits the gate-to-source voltage. During DCDC start up, the LS_GD gate drivers begin to source current after $V_{CSN/OUT}$ reach 3 V. If the $V_{CSN/OUT}$ > 7.5 V or $V_{BUS}$ > 7 V is under overvoltage condition, the LS_GD turns off immediately with 35-uA (typical) sink current.

If load current above NFET current limit threshold, LS_GD also turns off the NFET after 2 ms (typical) and enters hiccup mode to protect NFET from thermal issue. Refer to Figure 11-24 for application waveform.

In real application, if $V_{BUS}$ short to $V_{BAT}$ function is needed, 20 V back-to-back NFET is suggested in circuit design.

#### *10.3.10.4 Cycle-by-Cycle Buck Current Limit*

The buck regulator cycle-by-cycle current limit on both the peak and valley of the inductor current. Hiccup mode is activated if a fault condition persists to prevent over-heating.

High-side MOSFET overcurrent protection is implemented by the nature of the Peak Current Mode control. The HS switch current is sensed when the HS is turned on after a set blanking time. The HS switch current is compared to the output of the Error Amplifier (EA) minus slope compensation every switching cycle. for more details, refer to the *Functional Block Diagram*. The peak current of HS switch is limited by a clamped maximum peak current threshold $I_{HS_LIMIT}$ which is constant. So the peak current limit of the high-side switch is not affected by the slope compensation and remains constant over the full duty cycle range.

The current going through LS MOSFET is also sensed and monitored. When the LS switch turns on, the inductor current begins to ramp down. The LS switch is not turned OFF at the end of a switching cycle if its current is above the LS current limit $I_{LS_LIMIT}$. The LS switch is kept ON so that the inductor current keeps ramping down, until the inductor current ramps below the LS current limit $I_{LS_LIMIT}$. Then the LS switch is turned OFF and the HS switch is turned on after a dead time. This is somewhat different than the more typical peak current limit, and results in Equation 8 for the maximum load current.


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$$I_{OUT_MAX} = 0.5 \times (I_{LS_LIMIT} + I_{HS_LIMIT}) \quad (8)$$

If $V_{CSN/OUT}$ < 2-V typical due to a short circuit for 128 consecutive cycles, hiccup current protection mode is activated. In hiccup mode, the regulator is shut down and kept off for 118 ms typically, then TPS2583x-Q1 go through a normal re-start with soft start again. If the short-circuit condition remains, hiccup repeats until the fault condition is removed. Hiccup mode reduces power dissipation under severe overcurrent conditions, prevents over-heating and potential damage to the device and serves as a backup to the programmable current limit. See *Current Limit Setting Using $R_{ILIMIT}$*. After the output short is removed, the hiccup delay is passed and the output voltage recovers normally as shown in Figure 11-21.

### 10.3.11 Overvoltage, IEC and Short-to-Battery Protection

The TPS25840-Q1 integrates OVP and short to battery protection on VBUS, DM_IN and DP_IN pins. These pins can withstand voltage up to 18 V, and can protect upstream processor or Hub data line when overvoltage or short to battery condition occurs. Refer to Figure 9-3 for the short-to-battery test setup.

The TPS2584x-Q1 also integrates IEC ESD cell on DP_IN and DM_IN pins.

For more detailed TPS2584x-Q1 short-to-battery consideration and test report, please refer to the *TPS2583x-Q1 and TPS2584x-Q1 Short-to- Battery Application application report*.

#### *10.3.11.1 $V_{BUS}$ and $V_{CSN/OUT}$ Overvoltage Protection*

The TPS25840-Q1 integrates overvoltage protection on both BUS and CSN/OUT pin to meet different application requirement.

BUS pin can withstand up to 18 V, and the OVP threshold is 7-V typical. After overvoltage is detected on BUS pin, the LS_GD turns off immediately. Also, $\overline{\text{FAULT}}$ asserts after 8-ms deglitch time. After the excessive voltage is removed, the LS_GD turns on again and $\overline{\text{FAULT}}$ deasserts.

CSN/OUT pin can withstand up to 20 V, and the OVP threshold is 7.5-V typical. After overvoltage is detected on CSN/OUT pin, the buck converter stops regulation. Also, LS_GD turns off immediately. After the excessive voltage is removed, the buck converter resumes and the LS_GD turns on again.

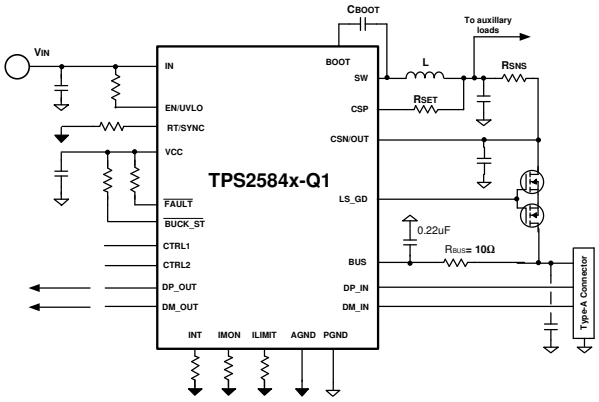



**Figure 10-17. Current Limit With External MOSFET**

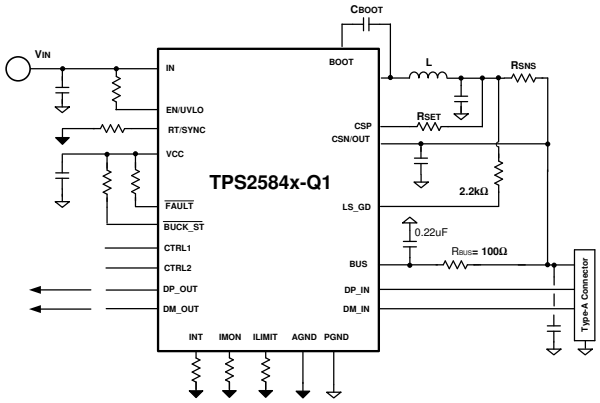



**Figure 10-18. Buck Average Current Limit**

As shown in Figure 10-17, TPS25840-Q1 is configured in external FET current limit mode. When short-to-battery occurs on BUS_Connector, the external MOSFET turns off immediately after BUS pin detect over voltage. The $\overline{\text{FAULT}}$ signal assertd after 8-ms deglitch time. See Figure 11-28. With Back-to-back FET, the TPS2583x-Q1 can withstand short-to-battery event even when Vin is off. TI recommends a 10-Ω 0805 resistor between BUS pin and BUS_Connector.

As shown in Figure 10-18, TPS25840-Q1 is configured in buck average current limit mode. When short-to-battery occurs on BUS_Connector, the buck regulator stops switching after CSN/OUT pin detect overvoltage.

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The $\overline{\text{FAULT}}$ signal also asserts after 8-ms deglitch time. TI recommends a 100-Ω 0805 resistor between BUS pin and BUS_Connector in buck average current limit mode.

#### *10.3.11.2 DP_IN and DM_IN Protection*

DP_IN and DM_IN protection consists of IEC ESD and overvoltage protection.

The DP_IN and DM_IN pins integrate an IEC ESD cell to provide ESD protection up to ±15-kV air discharge and ±8-kV contact discharge per IEC 61000-4-2 (for test conditions, see the *ESD Ratings* section). The IEC ESD performance of the TPS2584x-Q1 device depends on the capacitance connected from BUS pin to GND. TI recommends placing a A 0.22-µF capacitor close to the BUS pin

The ESD stress seen at DP_IN and DM_IN is impacted by many external factors like the parasitic resistance and inductance between ESD test points and the DP_IN and DM_IN pins. For air discharge, the temperature and humidity of the environment can cause some difference, so the IEC performance must always be verified in the end-application circuit.

Overvoltage protection (OVP) is provided for short-to-$V_{BUS}$ or short-to-battery conditions in the vehicle harness, preventing damage to the upstream USB transceiver or hub. When the voltage on DP_IN or DM_IN exceeds 3.9 V (typical), the TPS25840-Q1 device immediately turns off DP/DM switch and responds to block the high-voltage reverse connection to DP_OUT and DM_OUT. $\overline{\text{FAULT}}$ signal asserts after 8-ms deglitch time. See Figure 11-30.

For DP_IN and DM_IN, when OVP is triggered, the device turns on an internal discharge path with 416-kΩ resistance to ground. On removal of the overvoltage condition, the pin automatically turns off this discharge path and returns to normal operation by turning on the previously affected analog switch.

### 10.3.12 Cable Compensation

When a load draws current through a long or thin wire, there is an IR drop that reduces the voltage delivered to the load. Cable droop compensation linearly increases the voltage at the CSN/OUT pin of TPS2584x-Q1 as load current increases with the objective of maintaining $V_{BUS_CON}$ (the bus voltage at the USB connector) at 5 V, regardless of load conditions. Most portable devices charge at maximum current when 5 V is present at the USB connector. Figure 10-19 provides an example of resistor drops encountered when designing an automotive USB system with a remote USB connector location.

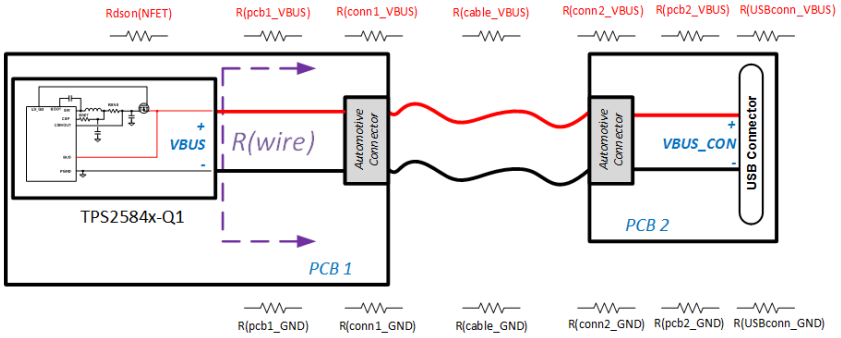



**Figure 10-19. Automotive USB Resistances**


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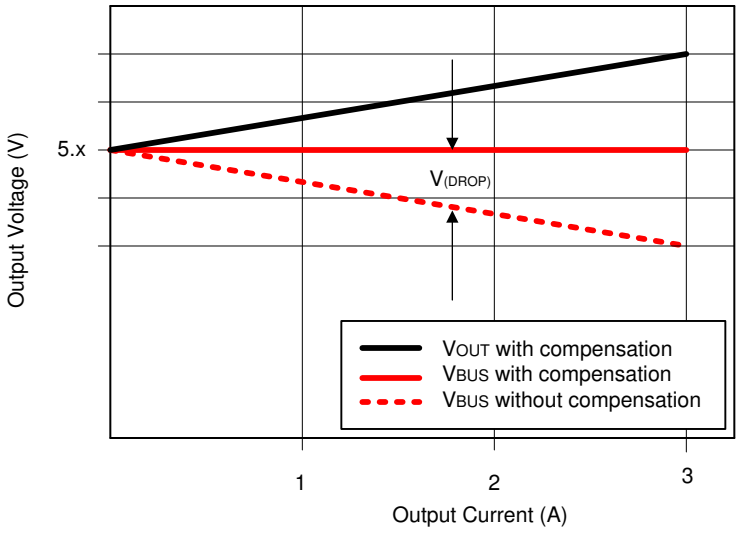



**Figure 10-20. Voltage Drop**

The TPS2584x-Q1 detects the load current and increases the voltage at the CSN/OUT pin to compensate the IR drop in the charging path according to the gain set by the $R_{SNS}$, $R_{SET}$, and $R_{IMON}$ resistors as described in *$R_{SNS}$, $R_{SET}$, $R_{ILIMIT}$, and $R_{IMON}$*.

The amount of cable droop compensation required can be estimated by the following equation: $\Delta V_{OUT} = (R_{SNS} + R_{DSON_NFET} + R_{WIRE}) I_{BUS} \times R_{IMON}$ is then chosen by $R_{IMON} = (\Delta V_{OUT} \times R_{SET} \times 2) / (I_{BUS} \times R_{SNS})$, Where $\Delta V_{OUT}$ is the desired cable droop compensation voltage at full load.

In most cases, the recommended voltage across $R_{SNS}$ must be 50 mV. See the *$R_{SNS}$, $R_{SET}$, $R_{ILIMIT}$, and $R_{IMON}$* section. In type-C application, typical $R_{IMON}$ resistors value are listed in Table 10-3 given the condition full load current = 3 A, $R_{SNS}$ = 15 mΩ and $R_{SET}$ = 300 Ω.

**Table 10-3. Setting the Cable Compensation Voltage with $R_{IMON}$**

| Cable Compensation Voltage at 3-A Full Load (V) | $R_{IMON}$ (kΩ) |
|---|---|
| 0.3 | 4.02 |
| 0.6 | 8.06 |
| 0.9 | 12.1 |
| 1.2 | 16.2 |
| 1.5 | 20 |

**Note**

The maximum cable compensation voltage in TPS2584x-Q1 is 1.5 V.

### *10.3.12.1 Cable Compensation Design Example*

To start the procedure, the $R_{SNS}$, $R_{DSON_NFET}$ and wire resistance $R_{WIRE}$ must be known.

1. Determine $R_{SNS}$ to achieve 50 mV at full current. For 3.3 A (3-A load current plus at approximately 10% for overcurrent threshold). $R_{SNS}$ = (0.05 V / 3.3 A) = 15 mΩ.
2. $R_{DSON_NFET}$ = 50 mΩ
3. $R_{WIRE}$ = 200 mΩ
4. $\Delta V_{OUT} = (R_{SNS} + R_{DSON_NFET} + R_{WIRE}) \times I_{BUS} = (0.015 + 0.05 + 0.2) \times 3 = 0.795$ V
5. Choose $R_{SET}$ = 300 Ω
6. $R_{IMON} = (\Delta V_{OUT} * R_{SET} \times 2) / (I_{BUS} \times R_{SNS}) = (0.795 \times 300 \times 2) / (3 \times 0.015) = 10.6$ kΩ

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### 10.3.13 USB Port Control

The TPS25840-Q1 and TPS25842-Q1 include DP_IN, DM_IN pins for automatic or host facilitated USB port power management of a Type-A downstream facing connector. For details on configuring the TPS2584x-Q1, see *Device Functional Modes*.

### 10.3.14 FAULT Response

The device features an active-low, open-drain fault output. Connect a 100-kΩ pullup resistor from $\overline{\text{FAULT}}$ to VCC or other suitable I/O voltage. $\overline{\text{FAULT}}$ can be left open or tied to GND when not used.

Table 10-4 summarizes the conditions that generate a fault and actions taken by the device.

**Table 10-4. Fault and Warning Conditions**

| EVENT | CONDITION | ACTION |
|---|---|---|
| Overcurrent on OUT | NFET or Buck average current limit implemented. See *Current Limit Sensing using $R_{ILIMIT}$*.<br>$I_{CSN/OUT}$ > programmed $I_{SNS}$. | The device regulates current at $I_{SNS}$ either by external NFET or by the buck regulator control loop.<br>When current limiting by external NFET, there is NO fault indicator assertion under minor overload conditions.<br>When current limiting by buck average current, there is NO fault indicator assertion under minor overload conditions. Hard shorts during average buck current limiting can trigger buck hiccup operation. The $\overline{\text{FAULT}}$ indicator asserts immediately after $N_{OC}$ cycles in and persists for $T_{OC}$ as specified in *Cycle-by-Cycle Buck Current Limit*. |
| Overvoltage on BUS | $V_{BUS}$ > $V_{BUS_OV}$ | The device turns on the BUS discharge path in the event of an overvoltage conditions and turns off the LS_GD immediately. The $\overline{\text{FAULT}}$ indicator asserts and de-asserts with an 8-ms deglitch. |
| Overvoltage on the data lines | DP_IN or DM_IN > $V_{Dx_IN_OV}$ | The device immediately shuts off the USB data switches. The $\overline{\text{FAULT}}$ indicator asserts and de-asserts with an 8-ms deglitch. |

### 10.3.15 USB Specification Overview

Universal Serial Bus specifications provide critical physical and electrical requirements to electronics manufacturers of USB capable equipment. Adherence to these specifications during product development coupled with standardized compliance testing assures very high degrees of interoperability amongst USB products in the market. Since its inception in the mid 1990s, USB has undergone a number of revisions to enhance utility and extend functionality. For the most up to date standards, please consult the USB Implementers Forum (USB-IF).

All USB ports are capable of providing a 5-V output making them a convenient power source for operating and charging portable devices. USB specification documents outline specific power requirements to ensure interoperability. In general, a USB 2.0 port host port is required to provide up to 500 mA; a USB 3.0 or USB 3.1 port is required to provide up to 900 mA; ports adhering to the USB Battery Charging 1.2 Specification provide up to 1500 mA; and newer Type-C ports can provide up to 3000 mA. Though USB standards governing power requirements exist, some manufacturers of popular portable devices created their own proprietary mechanisms to extend allowed available current beyond the 1500 mA maximum per BC 1.2. While not officially part of the standards maintained by the USB-IF, these proprietary mechanisms are recognized and implemented by manufacturers of USB charging ports.

The TPS2584x-Q1 device supports the most-common USB-charging schemes BC1.2 in popular hand-held media and cellular devices.

The BC1.2 specification includes three different port types:

- Standard downstream port (SDP, supported)
- Charging downstream port (CDP, supported)
- Dedicated charging port (DCP, NOT supported)

BC1.2 defines a charging port as a downstream-facing USB port that provides power for charging portable equipment. Under this definition, CDP and DCP are defined as charging ports.


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Table 10-5 lists the difference between these port types.

**Table 10-5. Operating Modes Table**

| PORT TYPE | SUPPORTS USB2.0 COMMUNICATION | MAXIMUM ALLOWABLE CURRENT DRAWN BY PORTABLE EQUIPMENT (A) |
|---|---|---|
| SDP (USB 2.0) | YES | 0.5 |
| SDP (USB 3.0 and 3.1) | YES | 0.9 |
| CDP | YES | 1.5 |
| DCP | NO | 1.5 |

### 10.3.16 Device Power Pins (IN, CSN/OUT, and PGND)

The IN pins are the input power path to the TPS2584x-Q1 devices. The internal LDO and buck regulator high side switch are supplied from the IN pins. The CSN/OUT pin connects to the negative terminal of the current sense amplifier and the internal voltage feedback network. This pin must be connected to the output LC filter for proper operation. PGND is the power ground return. For optimum performance, ensure the IN pin is properly bypassed to PGND with adequate bulk and high-frequency bypass capacitance located as close to these pins as possible.

### 10.3.17 Thermal Shutdown

The device has an internal overtemperature shutdown threshold, $T_{SD}$ to protect the device from damage and overall safety of the system. When device temperature exceeds $T_{SD}$, the LD_GD pin is pulled low, and the buck regulator stops switching. The device attempts to power-up when die temperature decreases by approximately 20°C.

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## 10.4 Device Functional Modes

### 10.4.1 Shutdown Mode

The EN pin provides electrical ON and OFF control for the TPS2584x-Q1. When $V_{EN}$ is below 1.2 V (typical), the device is in shutdown mode. The TPS2584x-Q1 also employs VIN and VCC undervoltage lock out protection. If $V_{IN}$ or $V_{CC}$ voltage is below their respective UVLO level, the regulator is turned off.

### 10.4.2 Active Mode

The TPS2584x-Q1 is in Active Mode when $V_{EN}$ is above the precision enable threshold, $V_{IN}$ and $V_{CC}$ are above their respective UVLO levels. The simplest way to enable the TPS2584x-Q1 is to connect the EN pin to VIN pin. This connection allows self startup when the input voltage is in the operating range of 3.8 V to 36 V and a UFP detection is made. For details on setting these operating levels, refer to *VCC, VCC_UVLO* and *Enable/UVLO*.

In Active Mode, the TPS2584x-Q1 buck regulator operates with Forced Pulse Width Modulation (FPWM), also referred to as Forced Continuous Conduction Mode (FCCM). This operation ensures the buck regulator switching frequency remains constant under all load conditions. FPWM operation provides low output voltage ripple, tight output voltage regulation, and constant switching frequency. Built-in spread-spectrum modulation aids in distributing spectral energy across a narrow band around the switching frequency programmed by the RT/SYNC pin. Under light load conditions the inductor current is allowed to go negative. A negative current limit of $I_{L_NEG}$ is imposed to prevent damage to the regulator's low side FET. During operation the TPS2584x-Q1 synchronized to any valid clock signal on the RT/SYNC input.

### 10.4.3 Device Truth Table (TT)

The device truth table (Table 10-6) lists all valid combinations for the two control pins (CTRL1 and CTRL2). The TPS2584x-Q1 devices monitor the CTRL inputs and transitions to whichever charging mode it is commanded.

**Table 10-6. Truth Table**

| DEVICE(S) | CTRL1 | CTRL2 | CURRENT LIMIT SETTING | USB MODES | BUCK REGULATOR | LS_GD |
|---|---|---|---|---|---|---|
| TPS2584x-Q1 | 0 | 0 | Buck Hiccup Only | Client Mode(1) | ON | OFF |
| | 0 | 1 | Buck Hiccup Only | Client Mode(1) | ON | OFF |
| | 1 | 0 | See *Current Limit Sensing using $R_{ILIMIT}$* | SDP Mode | ON | |
| | 1 | 1 | | CDP Mode | ON | |

(1) TPS2584x-Q1: USB data switches ON during client mode.

### 10.4.4 USB Port Operating Modes

#### *10.4.4.1 Standard Downstream Port (SDP) Mode — USB 2.0, USB 3.0, and USB 3.1*

An SDP is a traditional USB port that follows USB 2.0, USB 3.0 or USB 3.1 protocol. A USB 2.0 SDP supplies a minimum of 500 mA per port and supports USB 2.0 communications. A USB 3.x SDP supplies a minimum of 900 mA per port and supports USB 3.0 or USB 3.1 communications. For both types, the host controller must be active to allow charging.

#### *10.4.4.2 Charging Downstream Port (CDP) Mode*

A CDP is a USB port that follows USB BC1.2 and supplies a minimum of 1.5 A per port. A CDP provides power and meets the USB 2.0 requirements for device enumeration. USB-2.0 communication is supported, and the host controller must be active to allow charging. The difference between CDP and SDP is the host-charge handshaking logic that identifies this port as a CDP. A CDP is identifiable by a compliant BC1.2 client device and allows for additional current draw by the client device.

The CDP handshaking process occurs in two steps. During step one, the portable equipment outputs a nominal 0.6-V output on the D+ line and reads the voltage input on the D– line. The portable device detects the connection to an SDP if the voltage is less than the nominal data-detect voltage of 0.3 V. The portable device detects the connection to a CDP if the D– voltage is greater than the nominal data detect voltage of 0.3 V and optionally less than 0.8 V.


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The second step is necessary for portable equipment to determine whether the equipment is connected to a CDP or a DCP. The portable device outputs a nominal 0.6-V output on the D– line and reads the voltage input on the D+ line. The portable device concludes the equipment is connected to a CDP if the data line being read remains less than the nominal data detects voltage of 0.3 V. The portable device concludes it is connected to a DCP if the data line being read is greater than the nominal data detect voltage of 0.3 V.

#### *10.4.4.3 Client Mode*

The TPS2584x-Q1 device integrates client mode as shown in Figure 10-21. The external MOSFET power switch is OFF and only the data analog switch is ON. This mode can be used by automotive USB system manufacturers and OEMs for factory-only software programming with the USB port.

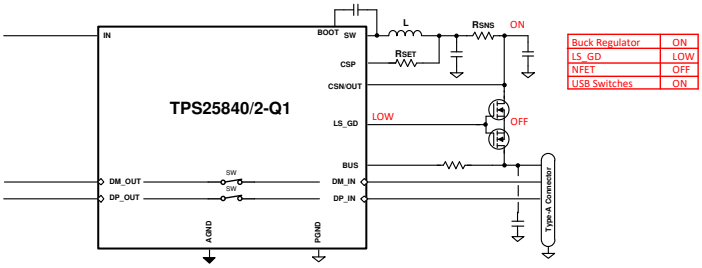



**Figure 10-21. Client-mode Equivalent Circuit**

### 10.4.5 High-bandwidth Data-line Switches

The TPS2584x-Q1 device passes the D+ and D– data lines through the device to enable monitoring and handshaking while supporting the charging operation. A wide-bandwidth signal switch allows data to pass through the device without corrupting signal integrity. The data-line switches are turned on in any of the CDP, SDP, or client operating modes. The EN input must be at logic high for the data line switches to be enabled.

For more detailed USB2.0 data line consideration and eye diagram test report, please refer to the *How to Improve USB2.0 Eye Diagram Using Long USB Cable application report*.

**Note**

- The data switches are ON while in CDP mode, even during CDP handshaking.
- The data line switches are OFF if EN/UVLO is low.
- The data line switches are OFF during External FET Current limit conditions.
- The data switches are only for a USB-2.0 differential pair. In the case of a USB-3.0 host, the super-speed differential pairs must be routed directly to the USB connector without passing through the TPS2584x device.

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# 11 Application and Implementation

**Note**

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

## 11.1 Application Information

The TPS2584x-Q1 is a step down DC-to-DC regulator and USB charge port controller. TheTPS2584x-Q1 is typically used in automotive systems to convert a DC voltage from the vehicle battery to 5-V DC with a maximum output current of 3 A. The following design procedure can be used to select components for the TPS2584x-Q1.

## 11.2 Typical Application

The TPS2584x-Q1 only requires a few external components to convert from a wide voltage range supply to a 5-V output for powering USB devices. Figure 11-1 shows a basic schematic.

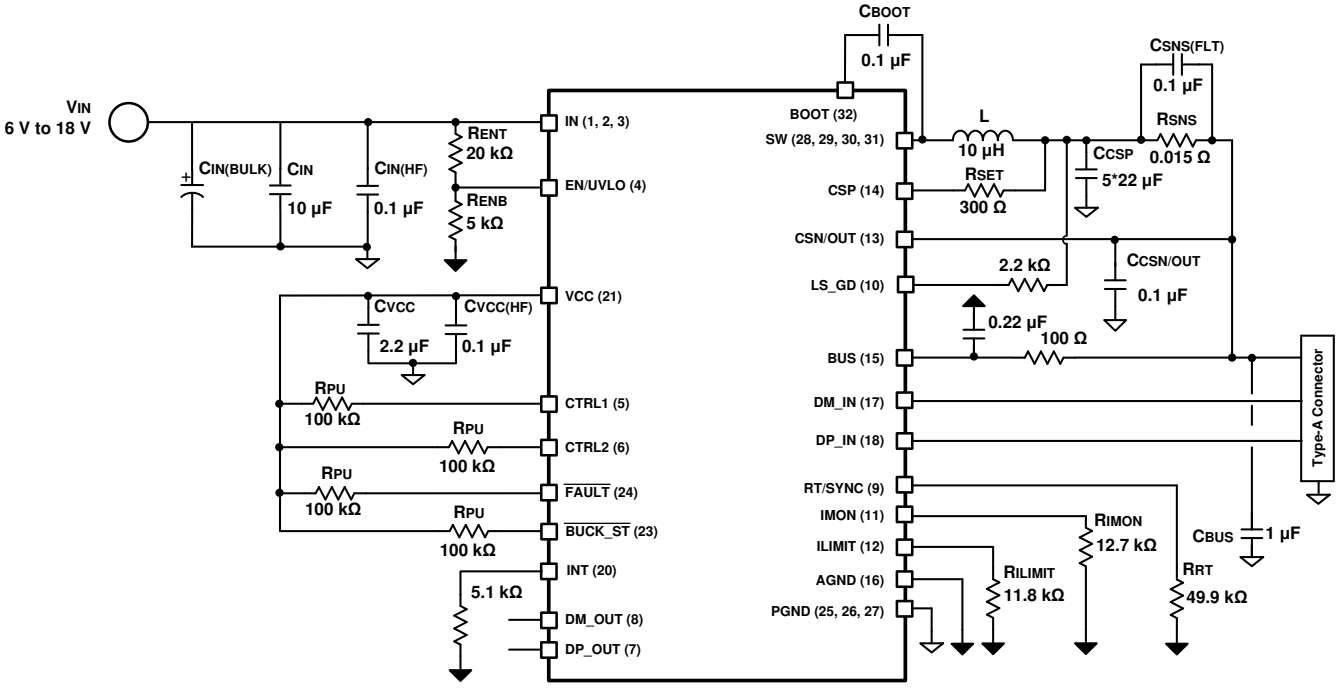



**Figure 11-1. Application Circuit**

The integrated buck regulator of TPS2584x-Q1 is internally compensated and optimized for a reasonable selection of external inductance and capacitance. The external components have to fulfill the needs of the application, but also the stability criteria of the device's control loop. Table 11-2 can be used to simplify the output filter component selection.

### 11.2.1 Design Requirements

To begin the design process, a few parameters must be known:

- Cable compensation: Total resistance including cable resistance, contact resistance of connectors, TPS2584x-Q1 current sense resistor and external NFET $r_{DS(on)}$ (if used). Refer to Figure 10-19 for examples of resistance in an automotive application.
- The maximum continuous output current for the charging port. The minimum current-limit setting of TPS2584x-Q1 device must be higher than this current.


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For this example, use the parameters listed in Table 11-1 as the input parameters.

**Table 11-1. Design Example Parameters**

| PARAMETER | VALUE |
|---|---|
| Input Voltage, $V_{IN}$ | 13.5-V typical, range from 6 V to 18 V |
| Output Voltage, $V_{OUT}$ | 5.1 V |
| Maximum Output Current $I_{OUT(MAX)}$ | 3.0 A |
| Transient Response 0.3 A to 3 A | 5% |
| Output Voltage Ripple | 50 mV |
| Input Voltage Ripple | 400 mV |
| Switching Frequency $f_{SW}$ | 400 kHz |
| Cable Resistance for Cable Compensation | 300 mΩ |
| Current Limit by Buck Average | 3.3 A |

**Table 11-2. L, and $C_{OUT}$ Typical Values**

| $f_{SW}$ | $V_{OUT}$ without Cable Compensation | $C_{IN}$ + $C_{HF}$ | L | Current Limit | $C_{CSP}$ | $C_{CSN/OUT}$ | $C_{BUS}$ |
|---|---|---|---|---|---|---|---|
| 400 kHz | 5.10 V | 1 × 10 µF + 1 × 100 nF | 10 µH | Buck Avg | 5 × 22 µF | 100 nF | 1 to 4.7 µF |
| 400 kHz | 5.10 V | 1 × 10 µF + 1 × 100 nF | 10 µH | Ext. NFET | 5 × 22 µF | 100 nF | 1 to 4.7 µF |
| 2100 kHz | 5.10 V | 1 × 10 µF + 1 × 100 nF | 2.2 µH | Buck Avg | 2 × 22 µF | 100 nF | 1 to 4.7 µF |

1. Inductance value is calculated based on $V_{IN}$ = 18 V.
2. All the $C_{OUT}$ values are after derating.

### 11.2.2 Detailed Design Procedure

#### *11.2.2.1 Output Voltage*

The output voltage of TPS2584x-Q1 is internally fixed at 5.10 V. Cable compensation can be used to increase the voltage on the CSN/OUT pin linearly with increasing load current. For more details on output voltage variation versus load current, refer to *Cable Compensation*. If cable compensation is not desired, use a 0-Ω $R_{IMON}$ resistor.

#### *11.2.2.2 Switching Frequency*

The recommended switching frequency of the TPS2584x-Q1 is in the range of 300–400 kHz for best efficiency. Choose $R_{RT}$ = 49.9 kΩ for 400-kHz operation. Refer to Table 10-1 to choose a different switching frequency.

#### *11.2.2.3 Inductor Selection*

The most critical parameters for the inductor are the inductance, saturation current and the rated current. The inductance is based on the desired peak-to-peak ripple current $\Delta i_L$. Because the ripple current increases with the input voltage, the maximum input voltage is always used to calculate the minimum inductance $L_{MIN}$. Use Equation 10 to calculate the minimum value of the output inductor. $K_{IND}$ is a coefficient that represents the amount of inductor ripple current relative to the maximum output current of the device. A reasonable value of $K_{IND}$ must be 20% to 40%. During an instantaneous short or over current operation event, the RMS and peak inductor current can be high. The inductor current rating must be higher than the current limit of the device.

$$\Delta i_L = \frac{V_{OUT} \times \left(V_{IN_MAX} - V_{OUT}\right)}{V_{IN_MAX} \times L \times f_{SW}} \tag{9}$$

$$L_{MIN} = \frac{V_{IN_MAX} - V_{OUT}}{I_{OUT} \times K_{IND}} \times \frac{V_{OUT}}{V_{IN_MAX} \times f_{SW}} \tag{10}$$

In general, choose lower inductance in switching power supplies because it usually corresponds to faster transient response, smaller DCR, and reduced size for more compact designs. But too low of an inductance

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can generate too large of an inductor current ripple such that over current protection at the full load can be falsely triggered. Too low of an inductance also generates more conduction loss and inductor core loss. Larger inductor current ripple also implies larger output voltage ripple with same output capacitors. With peak current mode control, TI recommends to have a larger inductor current ripple. A larger peak current ripple improves the comparator signal to noise ratio.

For this design example, choose $K_{IND}$ = 0.3, the minimum inductor value is calculated to be 8.7 µH. Choose the nearest standard 10 µH ferrite inductor with a capability of 4 A RMS current and 6-A saturation current.

#### *11.2.2.4 Output Capacitor Selection*

The value of the output capacitor, and its ESR, determine the output voltage ripple and load transient performance. The output capacitor bank is usually limited by the load transient requirements, rather than the output voltage ripple. Equation 11 can be used to estimate a lower bound on the total output capacitance, and an upper bound on the ESR, required to meet a specified load transient.

$$C_{OUT} \geq \frac{\Delta I_{OUT}}{f_{SW} \cdot \Delta V_{OUT} \cdot K} \cdot \left[ (1-D) \cdot (1+K) + \frac{K^2}{12} \cdot (2-D) \right]$$

$$ESR \leq \frac{(2+K) \cdot \Delta V_{OUT}}{2 \cdot \Delta I_{OUT} \left[ 1+K+\frac{K^2}{12} \cdot \left( 1+\frac{1}{(1-D)} \right) \right]}$$

$$D = \frac{V_{OUT}}{V_{IN}} \tag{11}$$

where

- $\Delta V_{OUT}$ = output voltage transient
- $\Delta I_{OUT}$ = output current transient
- K = Ripple factor from *Inductor Selection*

After the output capacitor and ESR have been calculated, Equation 12 can be used to check the peak-to-peak output voltage ripple, $V_r$.

$$V_r \cong \Delta I_L \cdot \sqrt{ESR^2 + \frac{1}{(8 \cdot f_{SW} \cdot C_{OUT})^2}} \tag{12}$$

The output capacitor and ESR can then be adjusted to meet both the load transient and output ripple requirements.

For this example, we require a $\Delta V_{OUT}$ of ≤ 250 mV for an output current step of $\Delta I_{OUT}$ = 2.7 A. Equation 11 gives a minimum value of 86 µF and a maximum ESR of 0.08 Ω. Assuming a 20% tolerance and a 10% bias de-rating, we arrive at a minimum capacitance of 110 µF. This can be achieved with a bank of 5 × 22-µF, 10-V, ceramic capacitors in the 1210 case size. More output capacitance can be used to improve the load transient response. Ceramic capacitors can easily meet the minimum ESR requirements. In some cases an aluminum electrolytic capacitor can be placed in parallel with the ceramics to help build up the required value of capacitance.

In practice, the output capacitor has the most influence on the transient response and loop phase margin. Load transient testing and Bode plots are the best way to validate any given design and must always be completed before the application goes into production. In addition to the required output capacitance, a small ceramic placed on the output can help to reduce high frequency noise. Small case size ceramic capacitors in the range of 1 nF to 100 nF can be very helpful in reducing voltage spikes on the output caused by inductor and board parasitics.


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The maximum value of total output capacitance must be limited to about 10 times the design value, or 1000 µF, whichever is smaller. Large values of output capacitance can adversely affect the start-up behavior of the regulator as well as the loop stability. If values larger than noted here must be used, then a careful study of start-up at full load and loop stability must be performed.

#### 11.2.2.5 Input Capacitor Selection

The TPS2584x-Q1 device requires high frequency input decoupling capacitor(s) and a bulk input capacitor, depending on the application. A high-quality ceramic capacitor type X5R or X7R with sufficient voltage ratings are recommended. To compensate the derating of ceramic capacitors, TI recommends a voltage rating of twice the maximum input voltage. The bulk capacitance selection depends upon a number of factors: long leads from the automotive battery to the IN pin of TPS2584x-Q1, cold or warm engine crank requirements, and so forth. The bulk capacitor is used to dampen voltage spike due to the lead inductance of the cable or the trace. For this design, one 10-µF, 50-V, X7R ceramic capacitors are used. Use a 0.1 µF for high-frequency filtering and place it as close as possible to the device pins. Consider adding additional bulk capacitance for operation through low $V_{IN}$ warm-crank profiles is required by the vehicle OEM.

#### 11.2.2.6 Bootstrap Capacitor Selection

Every TPS2584x-Q1 design requires a bootstrap capacitor ($C_{BOOT}$). The recommended capacitor is 0.1 µF and rated 10 V or higher. The bootstrap capacitor is located between the SW pin and the BOOT pin. The bootstrap capacitor must be a high-quality ceramic type with an X7R or X5R grade dielectric for temperature stability.

#### 11.2.2.7 VCC Capacitor Selection

The VCC pin is the output of an internal LDO for TPS2584x. The LDO supplies gate charge to the LS buck switch and is the supply to the digital state-machine and analog USB circuitry. To insure stability of the device, place a minimum of 2.2 µF, 10 V, X7R capacitor from this pin to ground. In addition a 0.1-µF high frequency decoupling capacitor is highly recommended.

#### 11.2.2.8 Enable and Under Voltage Lockout Set-Point

The system enable and undervoltage lockout (UVLO) is adjusted using the external voltage divider network of $R_{ENT}$ and $R_{ENB}$. The EN/UVLO has two thresholds, one for power up when the input voltage is rising and one for power down or brown outs when the input voltage is falling. The following equations can be used to determine the $V_{IN(ON)}$ and $V_{IN(OFF)}$ levels.

$$R_{ENT} = \left(\frac{V_{IN(ON)}}{V_{EN/UVLO_H}} - 1\right) \times R_{ENB} \tag{13}$$

$$V_{IN(OFF)} = V_{IN(ON)} \times \left(1 - \frac{V_{EN/UVLO_HYS}}{V_{EN/UVLO_H}}\right) \tag{14}$$

$V_{IN(ON)}$ = 6 V (user choice)

$R_{ENB}$ = 5 kΩ (user choice)

$R_{ENT}$ = [($V_{IN(ON)}$ / ($V_{EN/UVLO_H}$) – 1] × $R_{ENB}$

$R_{ENB}$ = [(6 V / 1.2 V) – 1] × 5 kΩ = 20 kΩ. Choose standard 20 kΩ.

Therefore $V_{IN(OFF)}$ = 6 V × [1 – (0.09 V / 1.2 V)] = 5.55 V

#### 11.2.2.9 Current Limit Set-Point

The TPS2584x-Q1 provides an accurate current limit to protect the USB port from overload based upon the values of $R_{SNS}$, $R_{SET}$ and $R_{ILIMIT}$. The design process is the same regardless of whether buck average current limiting or external NFET current limiting is chosen. The only difference is the current limit threshold voltage on the ILIMIT pin.

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- $R_{SNS}$ is the current sense resistor. The recommended voltage across $R_{SNS}$ under current limit must be approximately 50 mV as a compromise between accuracy and power dissipation. For example, if current limiting is desired for $I_{OUT(MAX)} \geq 3.3$ A, then $R_{SNS} = 0.05$ V / 3.3 A = 0.01515 Ω. Choose a standard value of 15 mΩ.
- $R_{SET}$ determines the input current to the transconductance amplifier and current mirror. The amplifier balances the voltage to be equal to that across $R_{SNS}$. Choose a $R_{SET}$ value to produce an $I_{SET}$ current between 75 –180 µA at the desired $I_{OUT(MAX)}$. Considering 50 mV across $R_{SET}$, a value of 300 Ω provides approximately 166 µA of $I_{SET}$ current to the amplifier and mirror circuit. Care must be taken to limit the $I_{SET}$ current below 200 µA to avoid saturating the internal amplifier circuit.
- Buck average current limiting occurs when $V_{ILIMIT} = 1$ V. $R_{ILIMIT}$ is calculated as 1 V × 300 Ω / [ 0.5 × (3.3 A × 15 mΩ + 0.7 mV) ] = 11.95 kΩ. A standard 11.8 kΩ value is chosen.

#### *11.2.2.10 Cable Compensation Set-Point*

From Table 11-1 the total cable resistance to be accounted for is 300 mΩ.

1. From *Current Limit Set-Point* $R_{SNS}$ and $R_{SET}$ have been determined as 15 mΩ and 300 Ω, respectively.
2. $R_{WIRE} = 300$ mΩ
3. $\Delta V_{OUT} = (R_{SNS} + R_{WIRE}) \times I_{BUS} = (0.015 + 0.3) \times 3 = 1.0395$ V
4. $R_{IMON} = (\Delta V_{OUT} \times R_{SET} \times 2) / (I_{BUS} \times R_{SNS}) = (1.0395 \times 300 \times 2) / (3.3 \times 0.015) = 12.6$ kΩ. A standard value of 12.7 kΩ is selected.

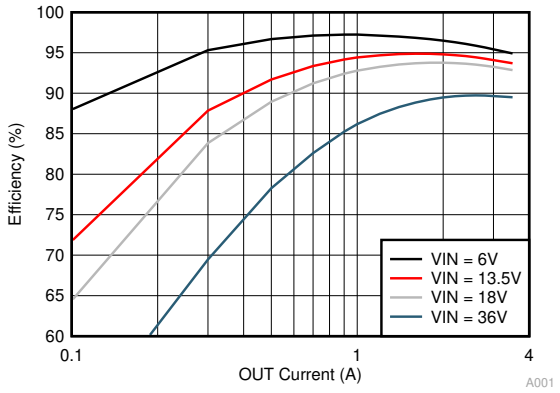
#### *11.2.2.11 $\overline{FAULT}$ Resistor Selection*

The $\overline{FAULT}$ pins are open-drain output flags. The pins can be connected to the TPS2584x-Q1 VCC pin with 100 kΩ resistors, or connected to another suitable I/O voltage supply if actively monitored by a USB HUB or MCU. The pins can be left floating if unused.


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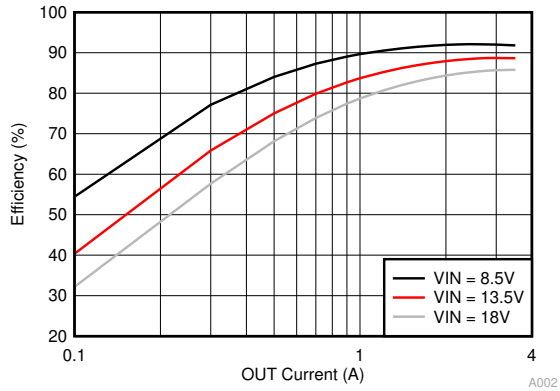
### 11.2.3 Application Curves

Unless otherwise specified the following conditions apply: $V_{IN}$ = 13.5 V, $f_{SW}$ = 400 kHz, L = 10 µH, $C_{OUT_CSP}$ = 66 µF, $C_{OUT_CSN}$ = 0.1 µF, $C_{BUS}$ = 1 µF, $T_A$ = 25 °C.



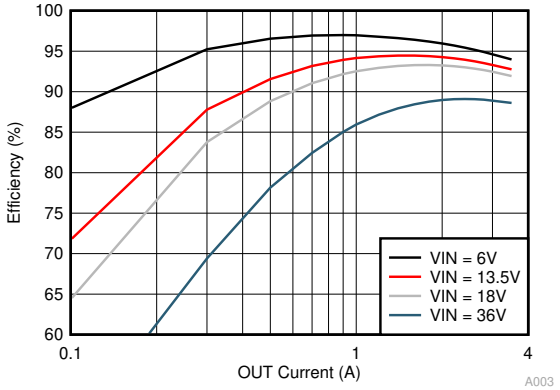

$V_{OUT}$ = 5.1 V $f_{SW}$ = 400 kHz

**Figure 11-2. Buck Only Efficiency**



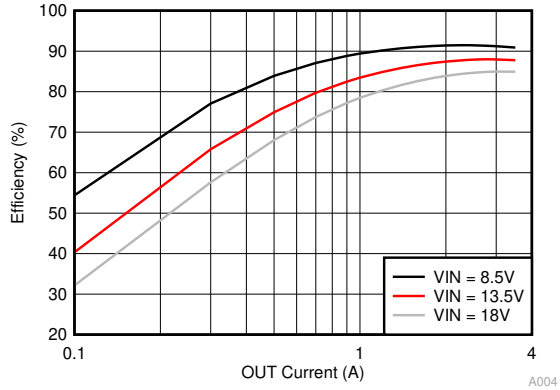

$V_{OUT}$ = 5.1 V $f_{SW}$ = 2100 kHz

**Figure 11-3. Buck Only Efficiency**



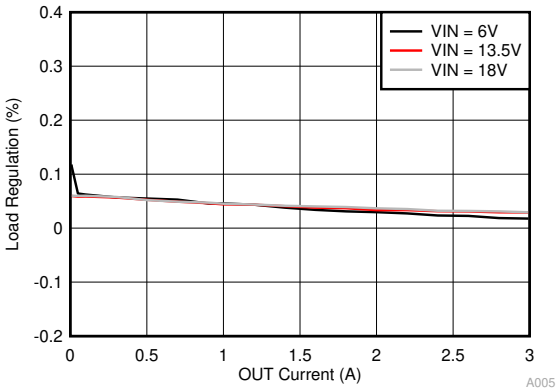

$V_{OUT}$ = 5.1 V $f_{SW}$ = 400 kHz $R_{SENS}$ = 15 mΩ

**Figure 11-4. Efficiency With Sense Resistor**



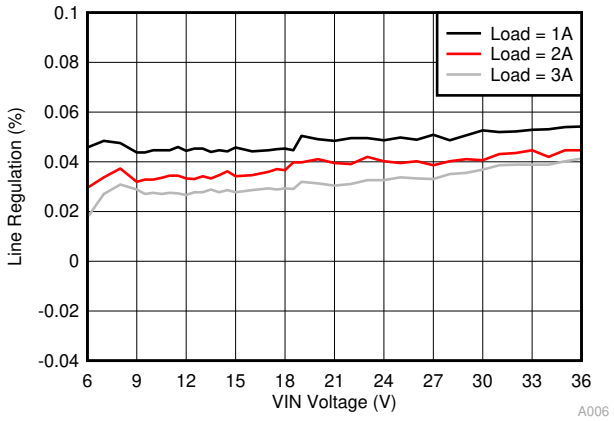

$V_{OUT}$ = 5.1 V $f_{SW}$ = 2100 kHz $R_{SENS}$ = 15 mΩ

**Figure 11-5. Efficiency With Sense Resistor**



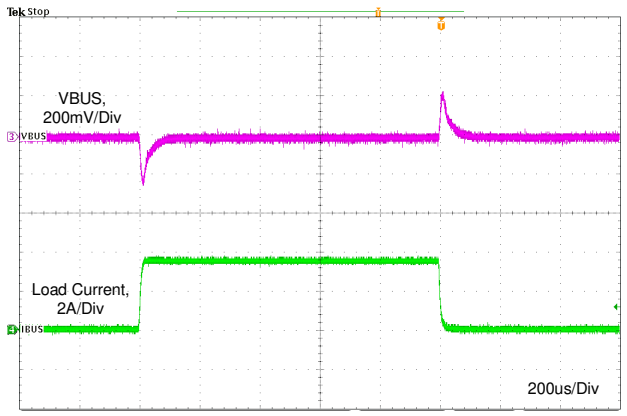

$V_{OUT}$ = 5.1 V $f_{SW}$ = 400 kHz

**Figure 11-6. Load Regulation**



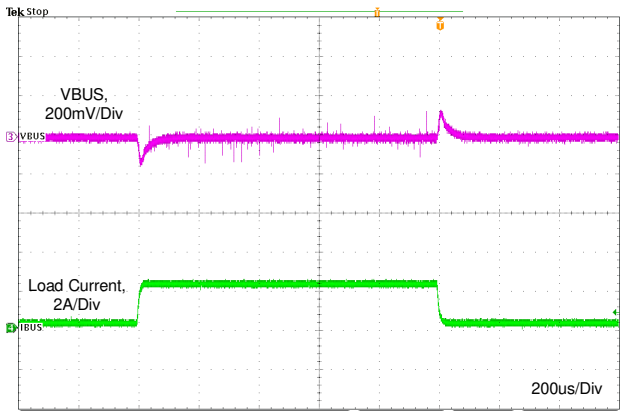

$V_{OUT}$ = 5.1 V $f_{SW}$ = 400 kHz

**Figure 11-7. Line Regulation**



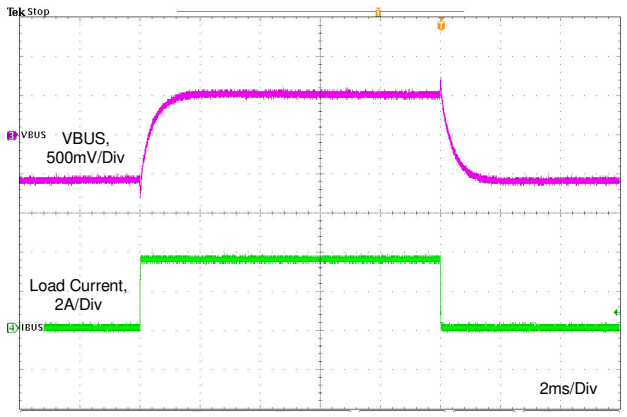

$V_{OUT}$ = 5.1 V $I_{LOAD}$ = 0 A to 3.5A $R_{IMON}$ = 0 Ω

**Figure 11-8. Load Transient Without Cable Compensation**



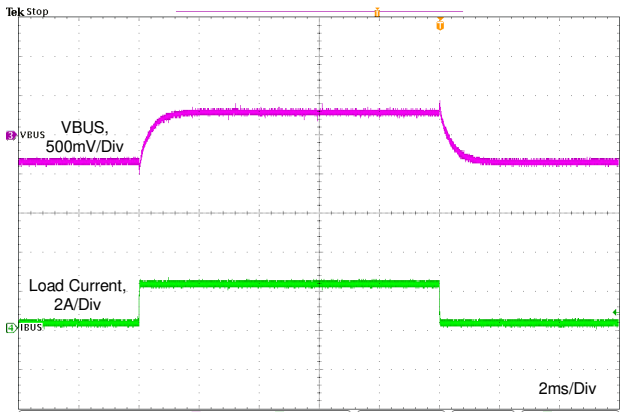

$V_{OUT}$ = 5.1 V $I_{LOAD}$ = 0.75 A to 2.25A $R_{IMON}$ = 0 Ω

**Figure 11-9. Load Transient Without Cable Compensation**




$I_{LOAD}$ = 0 A to 3.5A $R_{IMON}$ = 13 kΩ

**Figure 11-10. Load Transient With Cable Compensation**




$I_{LOAD}$ = 0.75 A to 2.25A $R_{IMON}$ = 13 kΩ

**Figure 11-11. Load Transient With Cable Compensation**

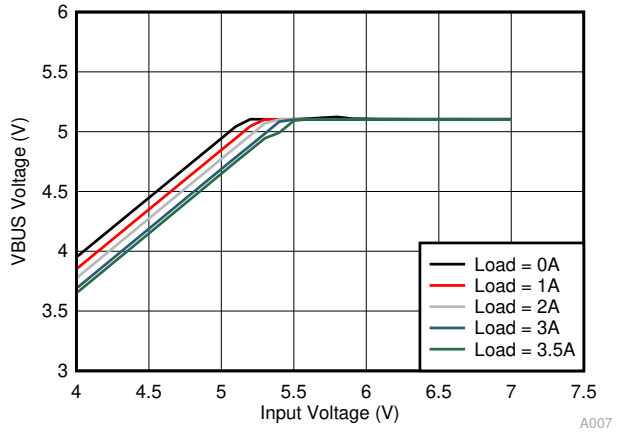
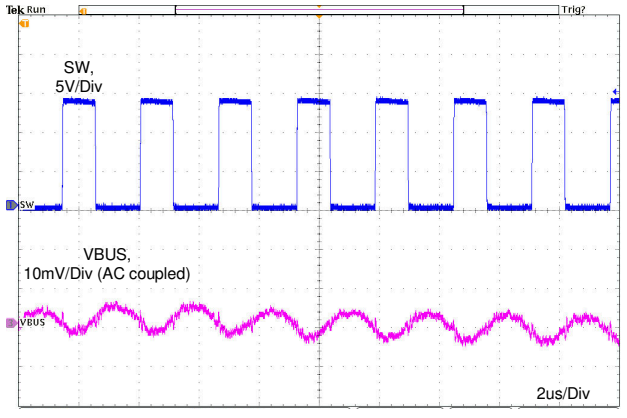


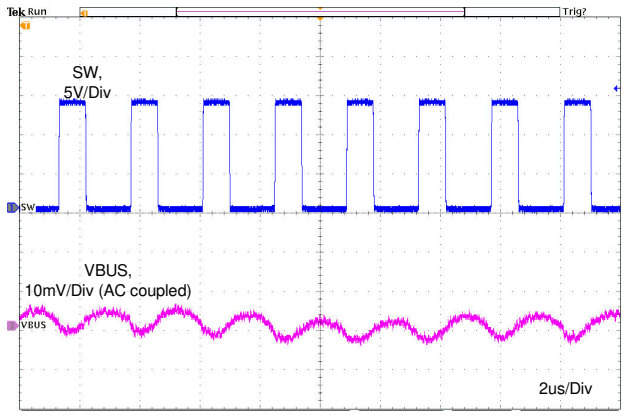

**Figure 11-12. Dropout Characteristic**




$R_{IMON}$ = 13 kΩ

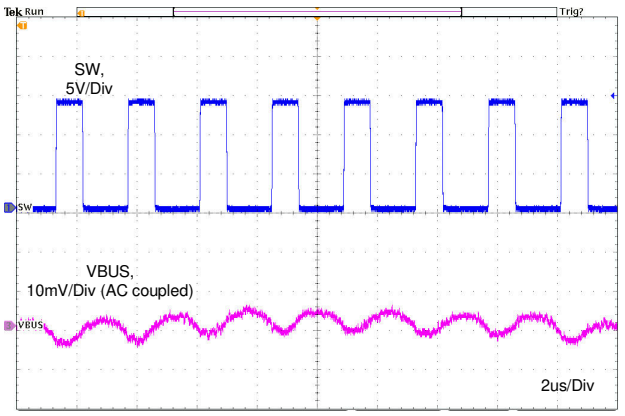
**Figure 11-13. 3.5-A Output Ripple**


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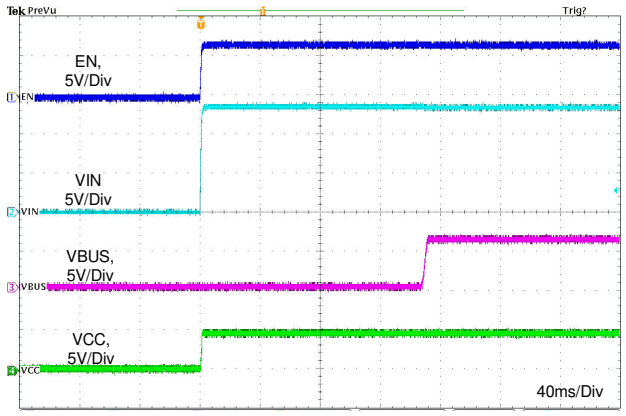

$R_{IMON}$ = 13 kΩ

**Figure 11-14. 100-mA Output Ripple**



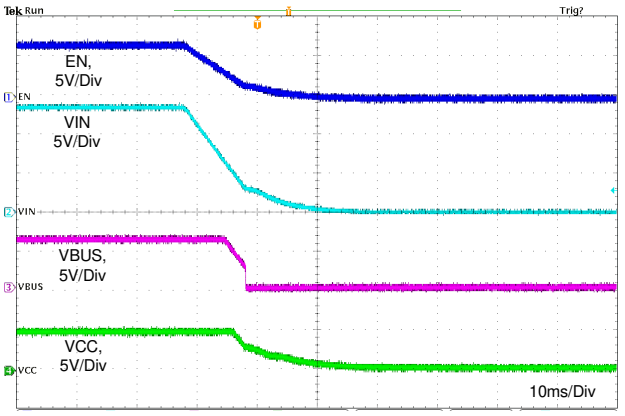

$R_{IMON}$ = 13 kΩ

**Figure 11-15. No Load Output Ripple**



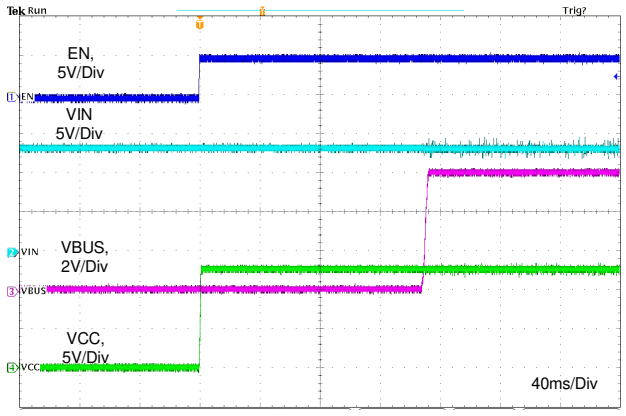

VIN = 0 V to 13.5 V INT = 5.1 kΩ $I_{LOAD}$ = 3A

**Figure 11-16. Startup Relate to VIN**



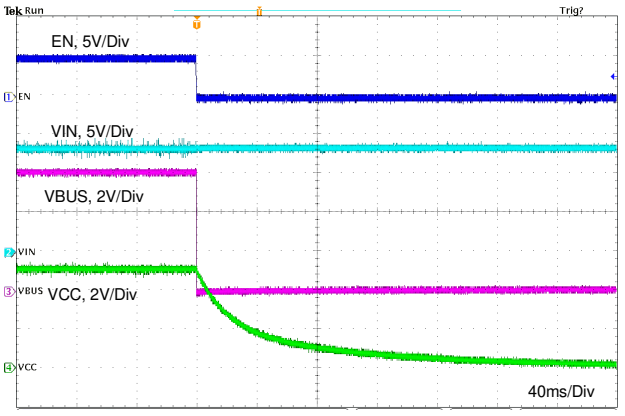

VIN = 13.5 V to 0 V INT = 5.1 kΩ $I_{LOAD}$ = 3A

**Figure 11-17. Shutdown Relate to VIN**




EN = 0V to 5V INT = 5.1 kΩ $I_{LOAD}$ = 3A

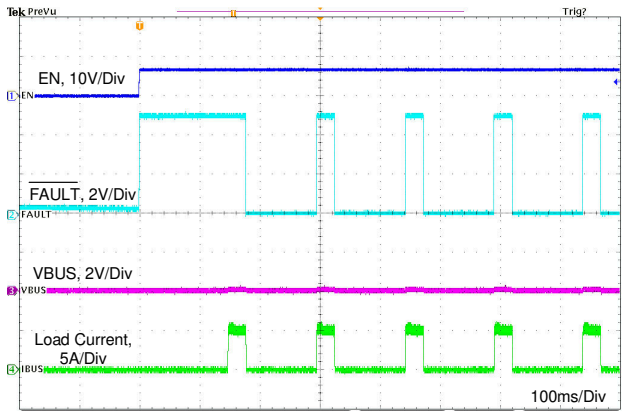
**Figure 11-18. Startup Relate to EN**




EN = 5V to 0V INT = 5.1 kΩ $I_{LOAD}$ = 3A

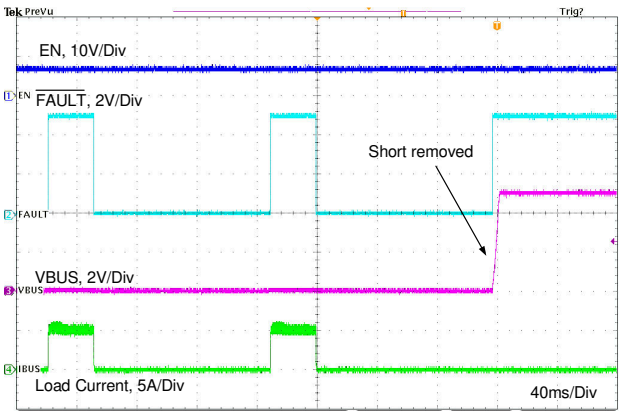
**Figure 11-19. Shutdown Relate to EN**

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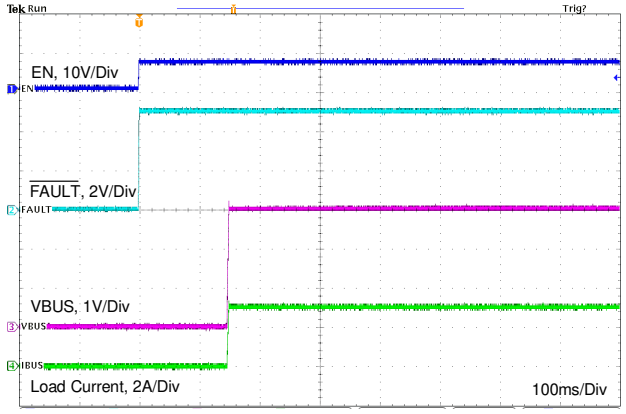

EN to High VBUS = GND $R_{LIMIT}$ = 13 kΩ

**Figure 11-20. Enable into Short Without External FET**



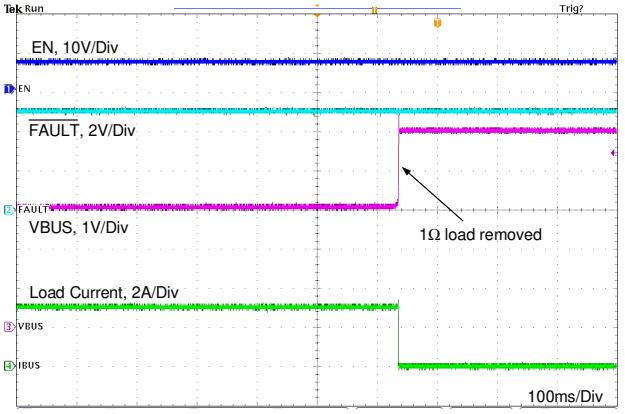

$R_{LIMIT}$ = 13 kΩ

**Figure 11-21. Short Circuit Recovery Without External FET**



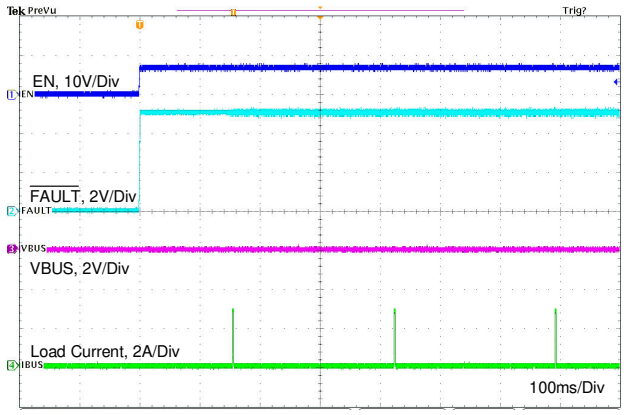

EN to High VBUS = GND $R_{LIMIT}$ = 13 kΩ

**Figure 11-22. Enable Into 1-Ω Load Without External FET**



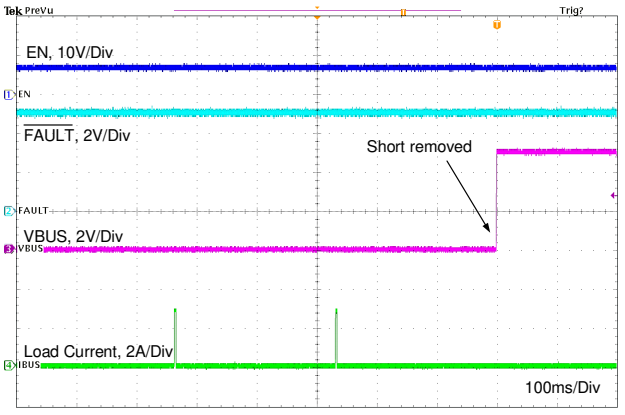

$R_{LIMIT}$ = 13 kΩ

**Figure 11-23. 1-Ω Load Recovery Without External FET**




EN to High VBUS = GND $R_{LIMIT}$ = 6.8 kΩ

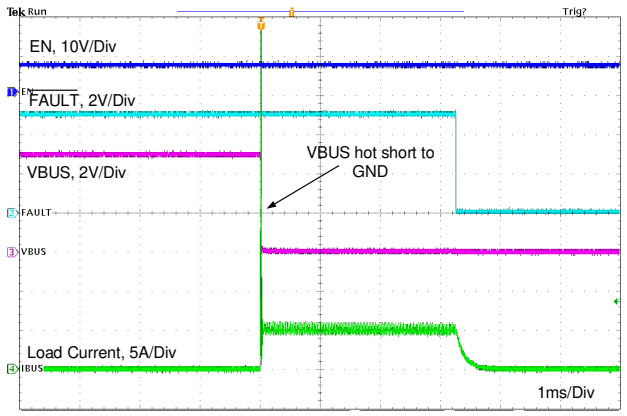
**Figure 11-24. Enable Into Short With External FET**




$R_{LIMIT}$ = 6.8 kΩ

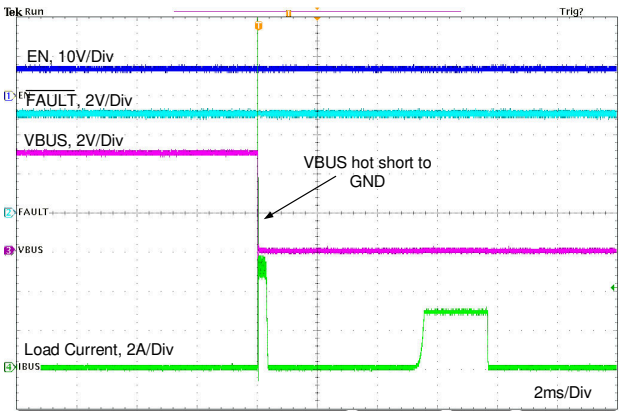
**Figure 11-25. Short Circuit Recovery With External FET**


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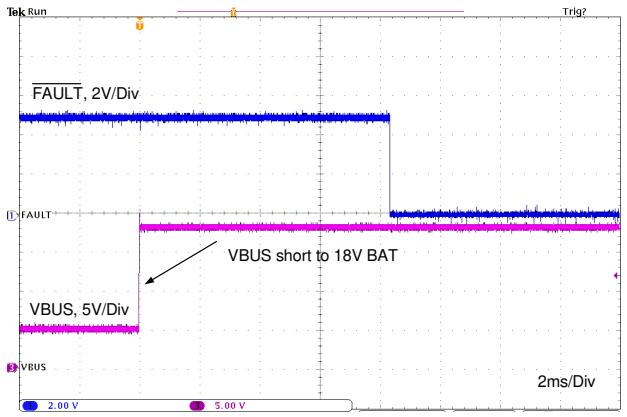

$R_{LIMIT}$ = 13 kΩ

**Figure 11-26. VBUS Hot Short to GND Without External FET**



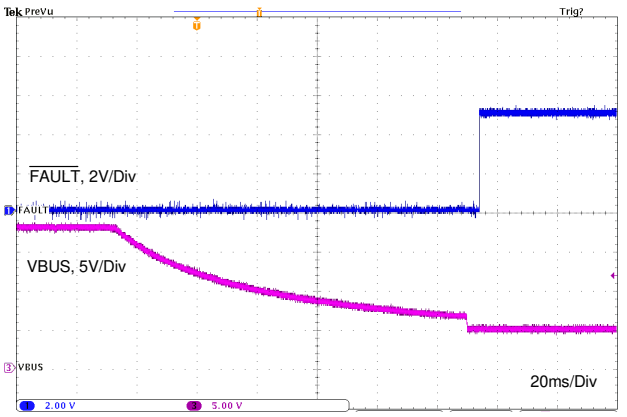

$R_{LIMIT}$ = 6.8 kΩ

**Figure 11-27. VBUS Hot Short to GND With External FET**



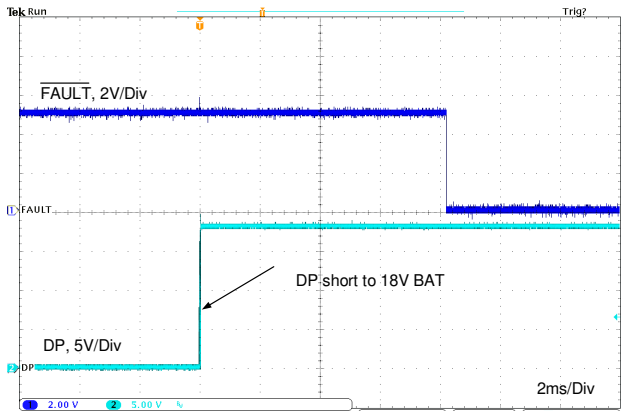

INT = 5.1 kΩ NO LOAD

**Figure 11-28. VBUS Short to BAT With External FET**



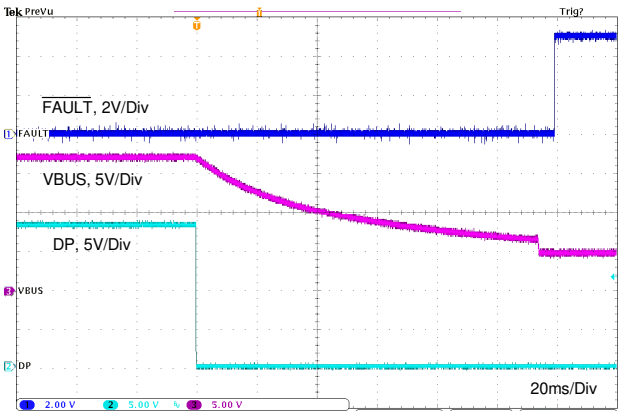

INT = 5.1 kΩ NO LOAD

**Figure 11-29. VBUS Short to BAT Recovery With External FET**




INT = 5.1 kΩ NO LOAD

**Figure 11-30. DP Short to BAT**




INT = 5.1 kΩ NO LOAD

**Figure 11-31. DP Short to BAT Recovery**

# 12 Power Supply Recommendations

The TPS2584x-Q1 is designed to operate from an input voltage supply range between 6 V and 36 V. This input supply must be able to withstand the maximum input current and maintain a stable voltage. The resistance of the input supply rail must be low enough that an input current transient does not cause a high enough drop at the TPS2584x-Q1 supply voltage that can cause a false UVLO fault triggering and system reset. If the input supply is located more than a few inches from the TPS2584x-Q1, additional bulk capacitance can be required in addition to the ceramic input capacitors. The amount of bulk capacitance is not critical, but a 47-µF or 100-µF electrolytic capacitor is a typical choice.

# 13 Layout

## 13.1 Layout Guidelines

Layout is a critical portion of good power supply design. The following guidelines help users design a PCB with the best power conversion performance, thermal performance, and minimized generation of unwanted EMI. For more detailed EMC design consideration and test report, please see the *PCB Layout and Parameters Recommendation for TPS2583X EMC Performance application report*.

1. ***Input capacitor:*** The input bypass capacitor, $C_{IN}$, must be placed as close as possible to the IN and PGND pins. Grounding for both the input and output capacitors must consist of localized top side planes that connect to the PGND pin and PAD. A combination of different values and packages of capacitors can help improve the EMC performance (for example: 10 µF + 0.1 µF + 2.2 nF). Besides, the distance between the input filter section and the output power section must be at least 15mm to prevent the output high-frequency signal from coupling into the input filter. A 10-µF cap cross $V_{IN}$ and PGND pin on top of SW is suggested for TPS2584x-Q1.
2. ***$V_{CC}$ bypass capacitor:*** Place bypass capacitors for $V_{CC}$ close to the VCC pin and ground the bypass capacitor to device ground.
3. Use a ground plane in one of the middle layers as noise shielding and heat dissipation path.
4. Connect the thermal pad to the ground plane. The QFN package has a thermal pad (PAD) connection that must be soldered down to the PCB ground plane. This pad acts as a heat-sink connection. The integrity of this solder connection has a direct bearing on the total effective $R_{\theta JA}$ of the application.
5. Make $V_{IN}$, $V_{OUT}$ and ground bus connections as wide as possible. This action reduces any voltage drops on the input or output paths of the converter and maximizes efficiency.
6. Provide enough PCB area for proper heat sinking. As stated in the section, enough copper area must be used to ensure a low $R_{\theta JA}$, commensurate with the maximum load current and ambient temperature. Make the top and bottom PCB layers with 2-ounce copper and no less than one ounce. Use an array of heat-sinking vias to connect the thermal pad (PAD) to the ground plane on the bottom PCB layer. If the PCB design uses multiple copper layers (recommended), thermal vias can also be connected to the inner layer heat-spreading ground planes.
7. The SW pin connecting to the inductor must be as short as possible, and just wide enough to carry the load current without excessive heating. Short, thick traces or copper pours (shapes) bring a high current conduction capacity to minimize parasitic resistance, but also cause a larger parasitic capacitance. Thus a balance must be found between smaller parasitic resistance and larger parasitic capacitance. And the current path must be kept straight forward to the inductor, otherwise the L-shaped or T-shaped path makes a sudden change of the impedance which causes signal reflection and impacts the performance of EMC. The output capacitors must be placed close to the $V_{OUT}$ end of the inductor and closely grounded to PGND pin and exposed PAD. Besides, do not punch vias on SW lines. Using shielded inductors or molded inductors to reduce high-frequency radiation.
8. ***Sense and Set Resistors:*** The $R_{SNS}$ and $R_{SET}$ resistors connect to the current sense amplifier inputs at the CSP and CSN/OUT pins. For best current limit and cable compensation accuracy; short, parallel traces give the best performance. If it is not possible to place $R_{SNS}$ and $R_{SET}$ near the CSP and CSN/OUT pins, TI recommends that the traces from sense resistor be routed in parallel and of similar lengths. A small filter capacitor in parallel with $R_{SNS}$ and a small filter capacitor from CSN/OUT to AGND help decouple noise.
9. $R_{ILIMIT}$ and $R_{IMON}$ resistors must be placed as close as possible to the ILIMIT and IMON pins and connected to AGND. If needed, these components can be placed on the bottom side of the PCB with signals routed through small vias.


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10. Trace routing of DP_IN, DM_IN, DP_OUT, and DM_OUT: Route these traces as micro-strips with nominal differential impedance of 90 Ω. Minimize the use of vias in the high-speed data lines. Keep the reference GND plane devoid from cuts or splits above the differential pairs to prevent impedance discontinuities.
11. $\overline{\text{FAULT}}$ are open-drain outputs. They can be connected to the VCC pin via pull-up resistors. Suggested resistor value is 100 kΩ.
12. The area enclosed by current loop of input side and output side must be as small as possible; the area enclosed by the BOOT circuit must be as small as possible.
13. Power ground PGND and the signal ground AGND must be separated in the actual PCB layout.

### 13.2 Ground Plane and Thermal Considerations

TI recommends to use one of the middle layers as a solid ground plane. Ground plane provides shielding for sensitive circuits and traces. Ground plane also provides a quiet reference potential for the control circuitry. The PGND pins must be connected to the ground plane using vias right next to the bypass capacitors. PGND pin is connected to the source of the internal LS switch. The PGND net contains noise at switching frequency and can bounce due to load variations. PGND trace, as well as VIN and SW traces, must be constrained to one side of the ground plane. The other side of the ground plane contains much less noise and must be used for sensitive routes. AGND and PGND must be connected under the QFN package PAD.

TI recommends to provide adequate device heat sinking by using the PAD of the IC as the primary thermal path. Use a minimum 2 row, 2 column "+" array of 12 mil thermal vias to connect the PAD to the system ground plane heat sink. The vias must be evenly distributed under the PAD. Use as much copper as possible, for system ground plane, on the top and bottom layers for the best heat dissipation. Use a four-layer board with the copper thickness for the four layers, starting from the top of 2 oz, 1 oz, 1 oz, 2 oz. Four layer boards with enough copper thickness provide low current conduction impedance, proper shielding and lower thermal resistance.

The thermal characteristics of the TPS2584x-Q1 are specified using the parameter, $\theta_{JA}$, which characterize the junction temperature of silicon to the ambient temperature in a specific system. Although the value of $\theta_{JA}$ is dependent on many variables, it still can be used to approximate the operating junction temperature of the device. To obtain an estimate of the device junction temperature, one can use the following relationship:

$$T_J = P_D \times \theta_{JA} + T_A \quad (15)$$

where

$T_J$ = Junction temperature is in °C

$P_D = V_{IN} \times I_{IN} \times (1 - \text{Efficiency}) - 1.1 \times I_{OUT}^{\ 2} \times \text{DCR}$ in Watt

DCR = Inductor DC parasitic resistance in Ω

$\theta_{JA}$ = Junction to ambient thermal resistance of the device in °C/W

$T_A$ = Ambient temperature in °C

$\theta_{JA}$ is highly related to PCB size and layout, as well as environmental factors such as heat sinking and air flow.

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## 13.3 Layout Example

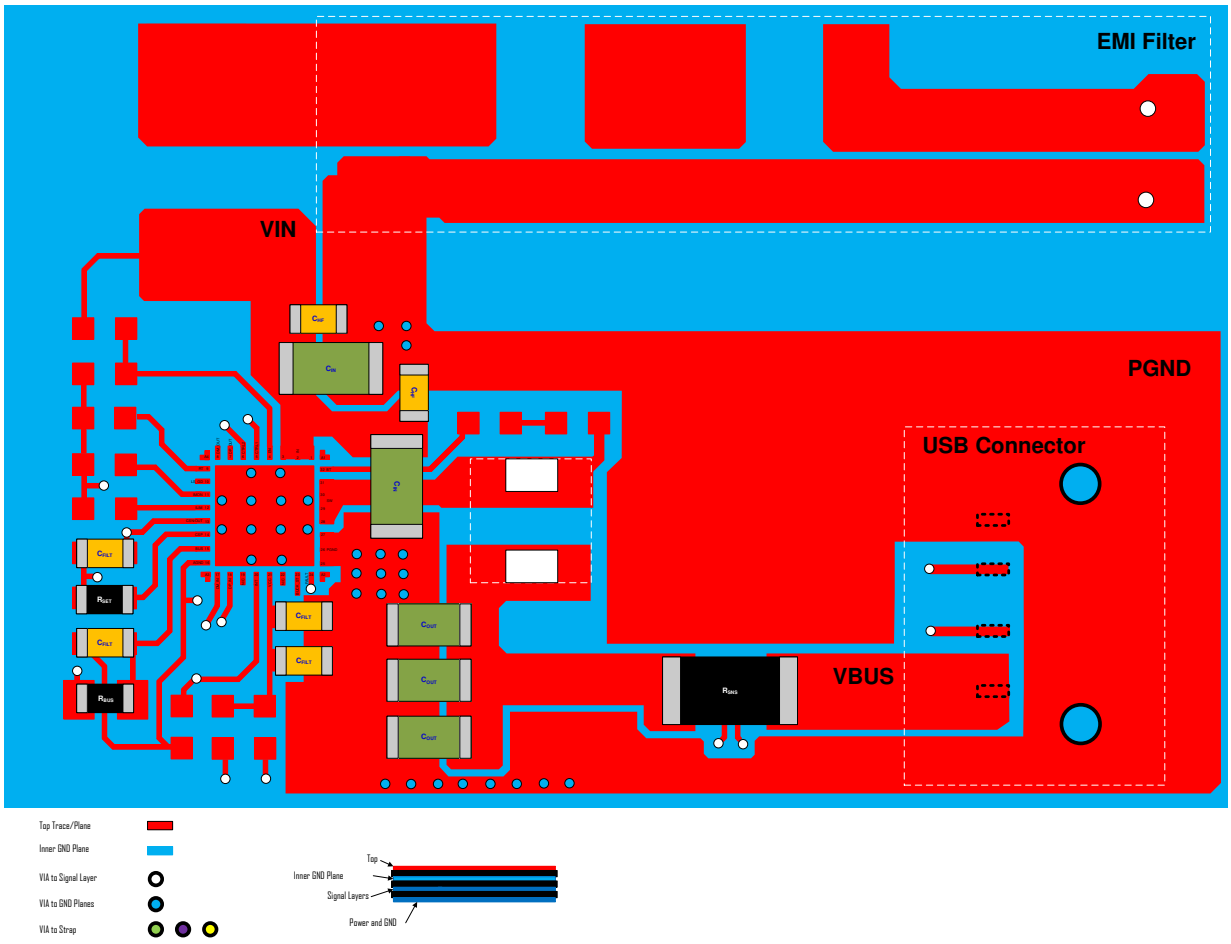



**Figure 13-1. Layout Example**


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# 14 Device and Documentation Support

## 14.1 Documentation Support

### 14.1.1 Related Documentation

- Texas Instruments, *How to Improve USB2.0 Eye Diagram Using Long USB Cable application report*
- Texas Instruments, *TPS2583x-Q1 and TPS2584x-Q1 Short-to- Battery Application application report*
- Texas Instruments, *How to Pass MFi Overcurrent Protection Test With USB Charger and Switch Device application report*
- Texas Instruments, *PCB Layout and Parameters Recommendation for TPS2583X EMC Performance application report*

## 14.2 Related Links

The table below lists quick access links. Categories include technical documents, support and community resources, tools and software, and quick access to order now.

**Table 14-1. Related Links**

| PARTS | PRODUCT FOLDER | ORDER NOW | TECHNICAL DOCUMENTS | TOOLS & SOFTWARE | SUPPORT & COMMUNITY |
|---|---|---|---|---|---|
| TPS25840-Q1 | Click here | Click here | Click here | Click here | Click here |
| TPS25842-Q1 | Click here | Click here | Click here | Click here | Click here |

## 14.3 Receiving Notification of Documentation Updates

To receive notification of documentation updates, navigate to the device product folder on ti.com. Click on *Subscribe to updates* to register and receive a weekly digest of any product information that has changed. For change details, review the revision history included in any revised document.

## 14.4 Support Resources

TI E2E™ support forums are an engineer's go-to source for fast, verified answers and design help — straight from the experts. Search existing answers or ask your own question to get the quick design help you need.

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## 14.5 Trademarks

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## 14.6 Electrostatic Discharge Caution



This integrated circuit can be damaged by ESD. Texas Instruments recommends that all integrated circuits be handled with appropriate precautions. Failure to observe proper handling and installation procedures can cause damage.

ESD damage can range from subtle performance degradation to complete device failure. Precision integrated circuits may be more susceptible to damage because very small parametric changes could cause the device not to meet its published specifications.

## 14.7 Glossary

TI Glossary — This glossary lists and explains terms, acronyms, and definitions.

# 15 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.

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## 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 |
|---|---|---|---|---|---|---|---|---|---|---|---|
| TPS25840QCWRHBRQ1 | ACTIVE | VQFN | RHB | 32 | 5000 | RoHS & Green | NIPDAU | Level-2-260C-1 YEAR | -40 to 125 | T25840 | Samples |
| TPS25840QWRHBRQ1 | ACTIVE | VQFN | RHB | 32 | 3000 | RoHS & Green | SN | Level-2-260C-1 YEAR | -40 to 125 | T25840 | Samples |
| TPS25840QWRHBTQ1 | ACTIVE | VQFN | RHB | 32 | 250 | RoHS & Green | SN | Level-2-260C-1 YEAR | -40 to 125 | T25840 | Samples |
| TPS25842QCWRHBRQ1 | ACTIVE | VQFN | RHB | 32 | 5000 | RoHS & Green | NIPDAU | Level-2-260C-1 YEAR | -40 to 125 | T25842 | Samples |
| TPS25842QWRHBRQ1 | ACTIVE | VQFN | RHB | 32 | 3000 | RoHS & Green | SN | Level-2-260C-1 YEAR | -40 to 125 | T25842 | Samples |
| TPS25842QWRHBTQ1 | ACTIVE | VQFN | RHB | 32 | 250 | RoHS & Green | SN | Level-2-260C-1 YEAR | -40 to 125 | T25842 | 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.

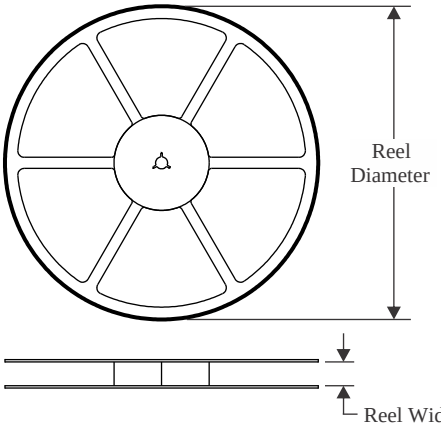
(6) Lead finish/Ball material - Orderable Devices may have multiple material finish options. Finish options are separated by a vertical ruled line. Lead finish/Ball material values may wrap to two lines if the finish value exceeds the maximum column width.

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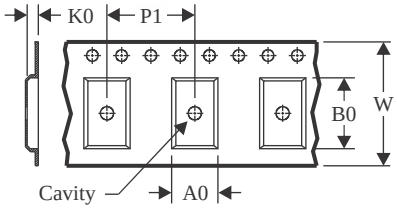
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## TAPE AND REEL INFORMATION

**REEL DIMENSIONS**

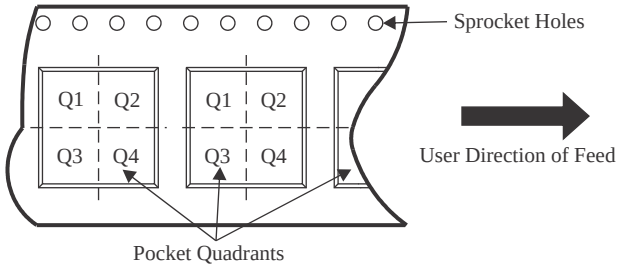



**TAPE DIMENSIONS**




| | |
|---|---|
| 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 |

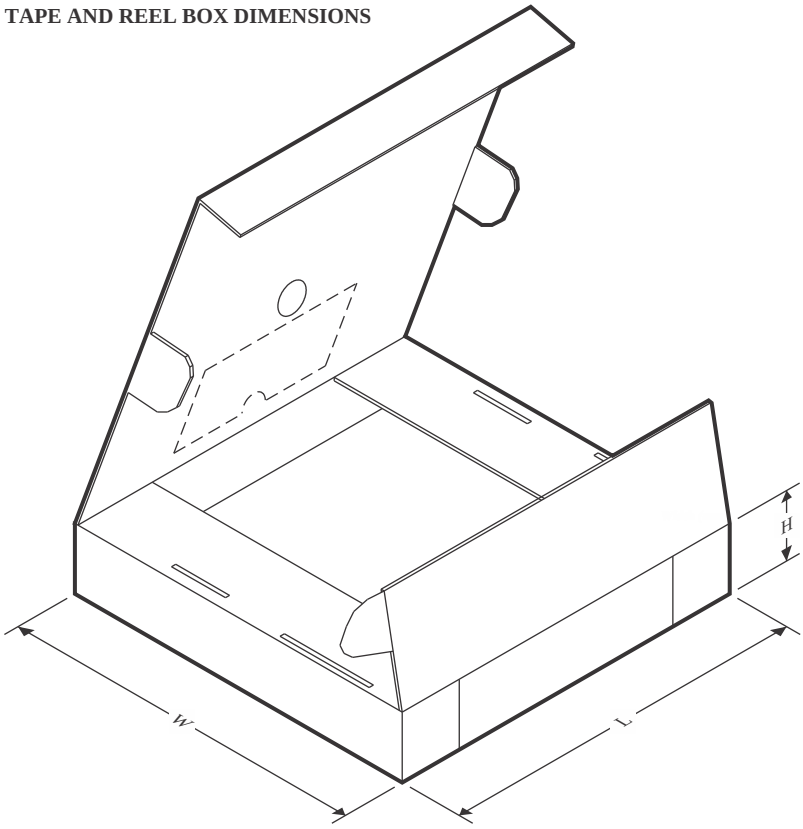
**QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE**




*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 |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| TPS25840QCWRHBRQ1 | VQFN | RHB | 32 | 5000 | 330.0 | 12.4 | 5.3 | 5.3 | 1.1 | 8.0 | 12.0 | Q2 |
| TPS25840QWRHBRQ1 | VQFN | RHB | 32 | 3000 | 330.0 | 12.4 | 5.25 | 5.25 | 1.1 | 8.0 | 12.0 | Q2 |
| TPS25840QWRHBTQ1 | VQFN | RHB | 32 | 250 | 180.0 | 12.4 | 5.25 | 5.25 | 1.1 | 8.0 | 12.0 | Q2 |
| TPS25842QCWRHBRQ1 | VQFN | RHB | 32 | 5000 | 330.0 | 12.4 | 5.3 | 5.3 | 1.1 | 8.0 | 12.0 | Q2 |
| TPS25842QWRHBRQ1 | VQFN | RHB | 32 | 3000 | 330.0 | 12.4 | 5.25 | 5.25 | 1.1 | 8.0 | 12.0 | Q2 |
| TPS25842QWRHBTQ1 | VQFN | RHB | 32 | 250 | 180.0 | 12.4 | 5.25 | 5.25 | 1.1 | 8.0 | 12.0 | Q2 |

TAPE AND REEL BOX DIMENSIONS




*All dimensions are nominal

| Device | Package Type | Package Drawing | Pins | SPQ | Length (mm) | Width (mm) | Height (mm) |
|---|---|---|---|---|---|---|---|
| TPS25840QCWRHBRQ1 | VQFN | RHB | 32 | 5000 | 367.0 | 367.0 | 35.0 |
| TPS25840QWRHBRQ1 | VQFN | RHB | 32 | 3000 | 367.0 | 367.0 | 38.0 |
| TPS25840QWRHBTQ1 | VQFN | RHB | 32 | 250 | 213.0 | 191.0 | 35.0 |
| TPS25842QCWRHBRQ1 | VQFN | RHB | 32 | 5000 | 367.0 | 367.0 | 35.0 |
| TPS25842QWRHBRQ1 | VQFN | RHB | 32 | 3000 | 367.0 | 367.0 | 38.0 |
| TPS25842QWRHBTQ1 | VQFN | RHB | 32 | 250 | 213.0 | 191.0 | 35.0 |

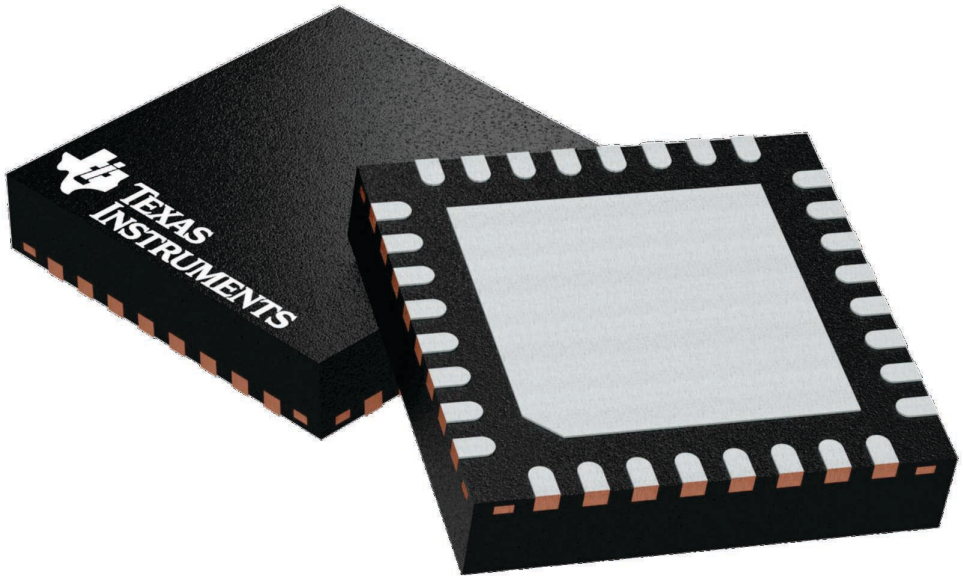
# GENERIC PACKAGE VIEW

## RHB 32

## VQFN - 1 mm max height

**5 x 5, 0.5 mm pitch**

PLASTIC QUAD FLATPACK - NO LEAD




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

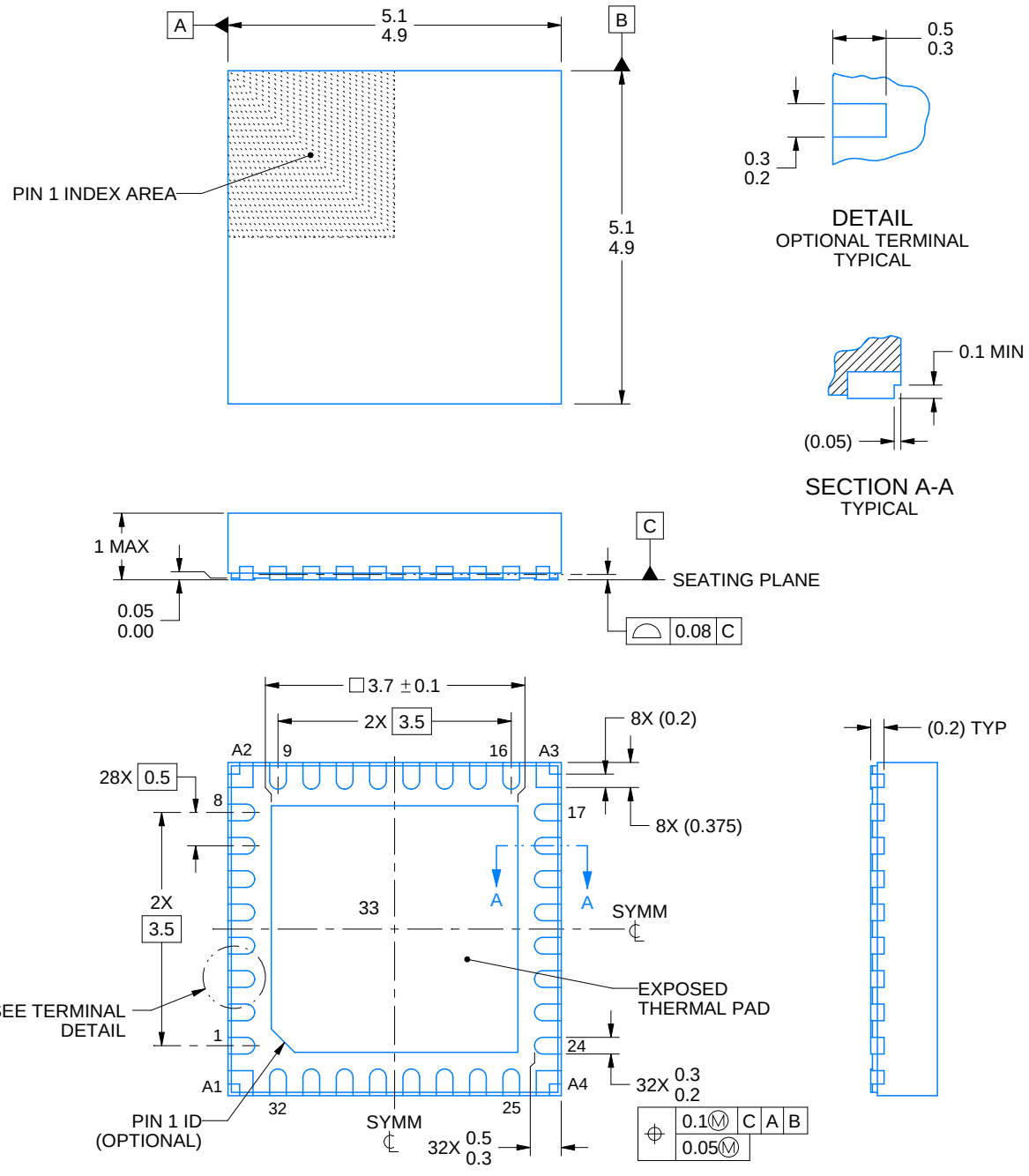
4224745/A

# PACKAGE OUTLINE

## RHB0032R

## VQFN - 1 mm max height

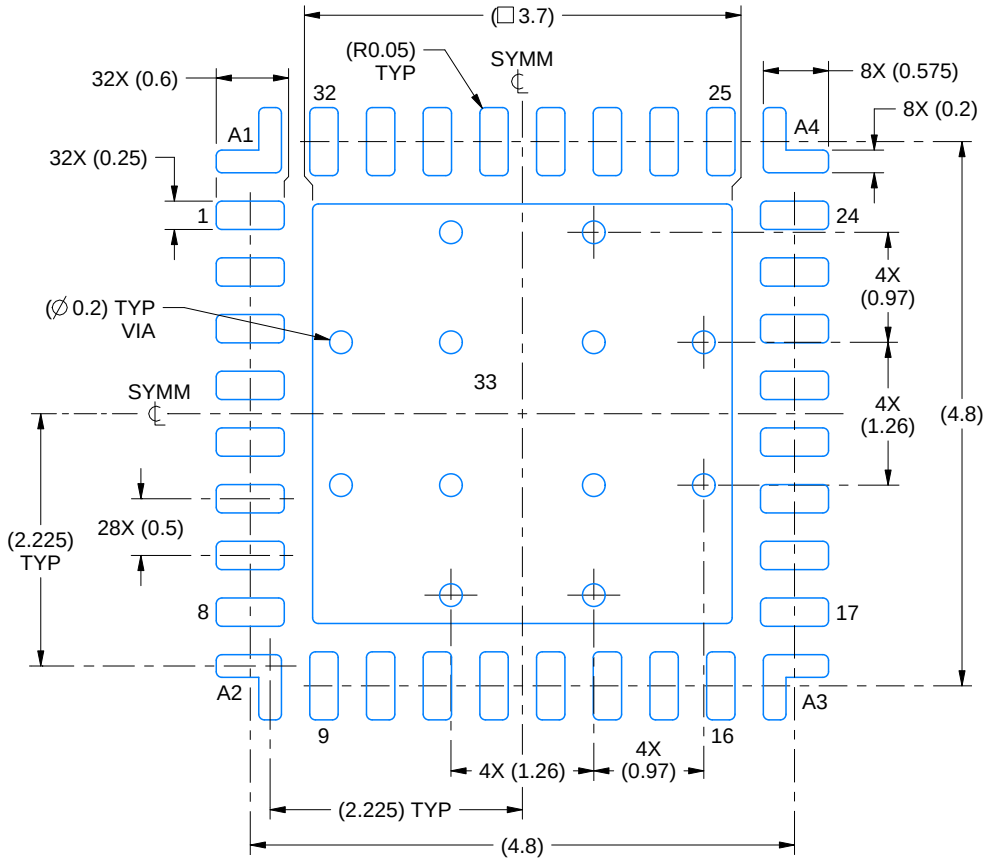
PLASTIC QUAD FLATPACK - NO LEAD



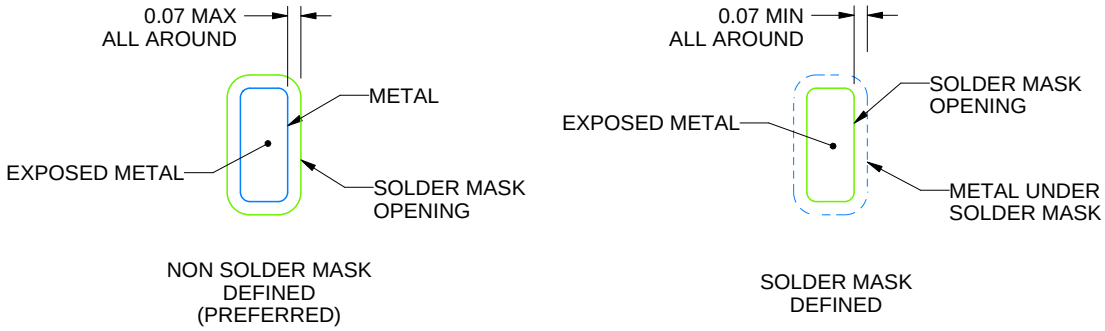

4223771/A 06/2017

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:15X

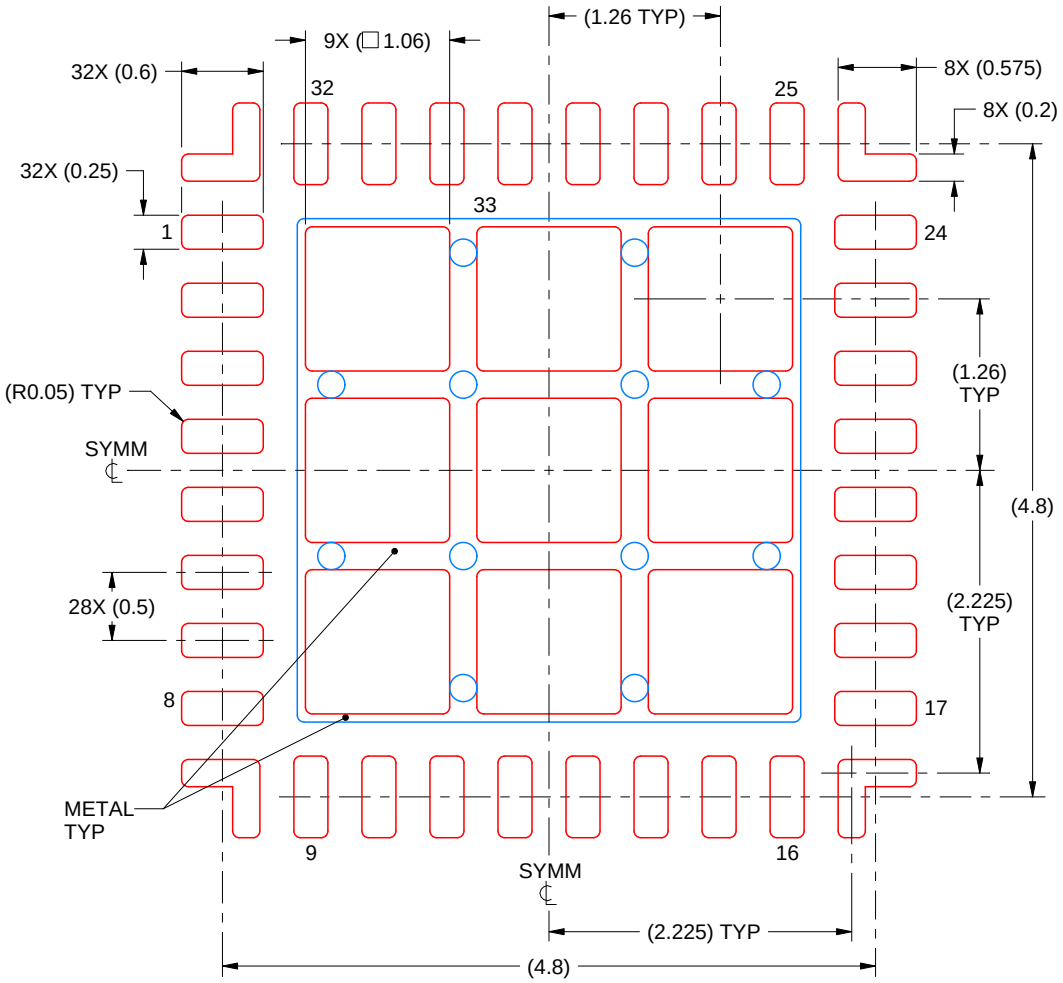



SOLDER MASK DETAILS

4223771/A 06/2017

NOTES: (continued)

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




SOLDER PASTE EXAMPLE
BASED ON 0.125 mm THICK STENCIL

EXPOSED PAD 33
74% PRINTED SOLDER COVERAGE BY AREA UNDER PACKAGE
SCALE:18X

4223771/A 06/2017

NOTES: (continued)

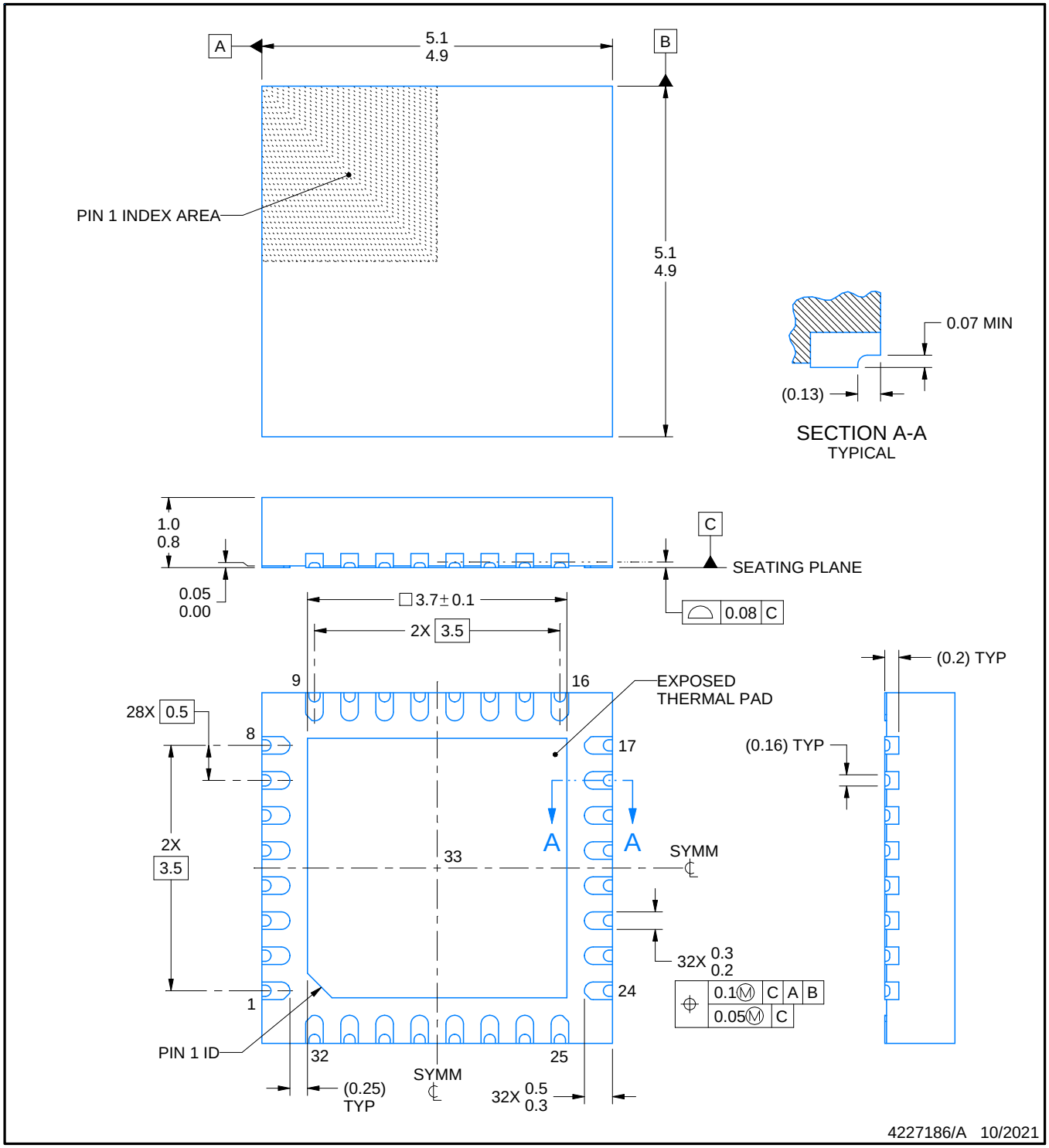
6. Laser cutting apertures with trapezoidal walls and rounded corners may offer better paste release. IPC-7525 may have alternate design recommendations.

# RHB0032AA

# PACKAGE OUTLINE

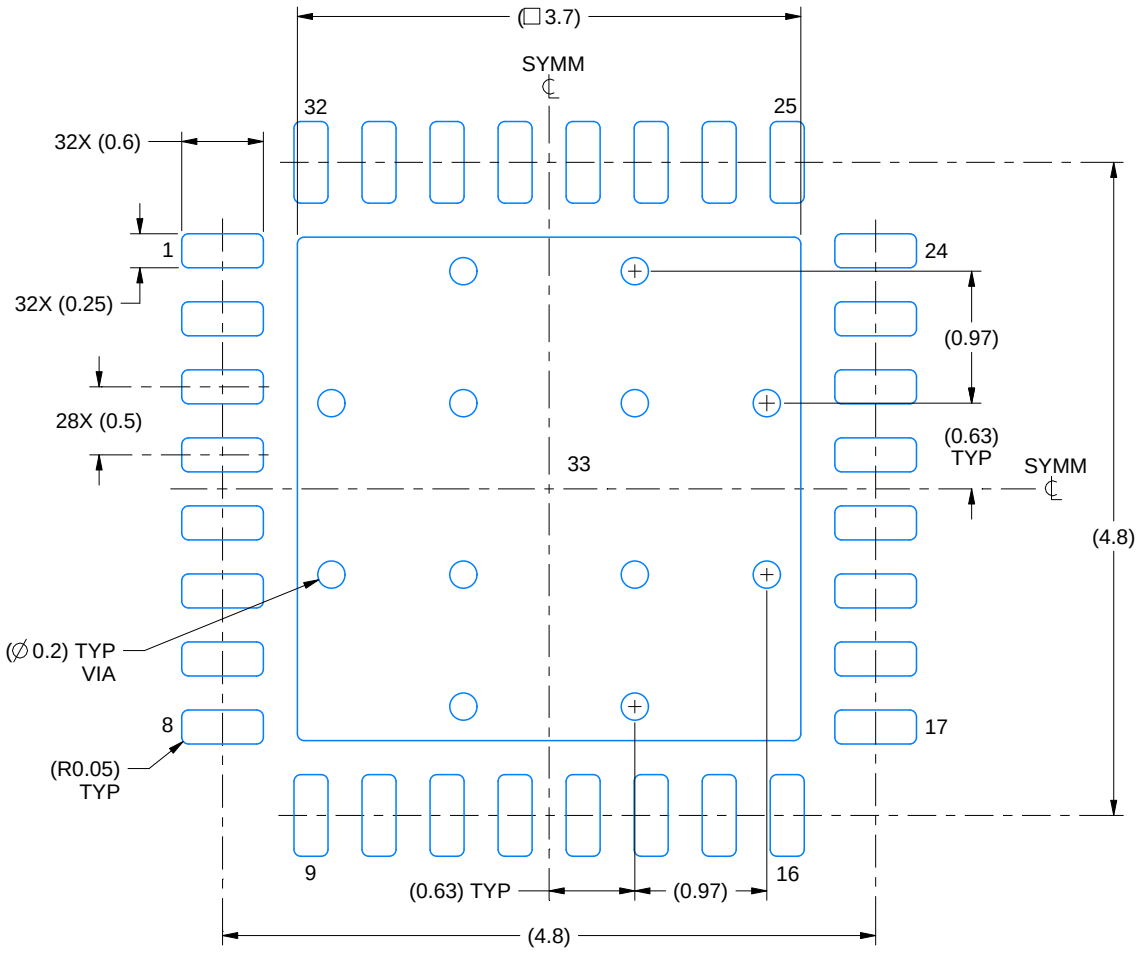
## VQFN - 1 mm max height

PLASTIC QUAD FLATPACK - NO LEAD




NOTES:

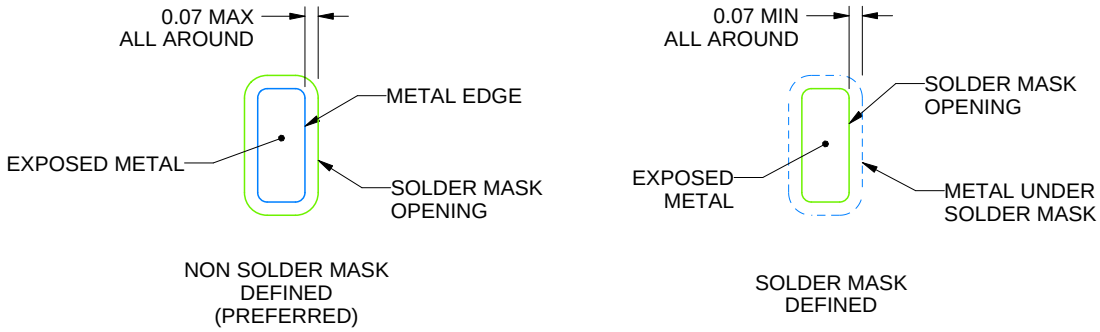
1. All linear dimensions are in millimeters. Any dimensions in parenthesis are for reference only. Dimensioning and tolerancing per ASME Y14.5M.
2. This drawing is subject to change without notice.
3. The package thermal pad must be soldered to the printed circuit board for thermal and mechanical performance.




LAND PATTERN EXAMPLE

EXPOSED METAL SHOWN

SCALE:18X

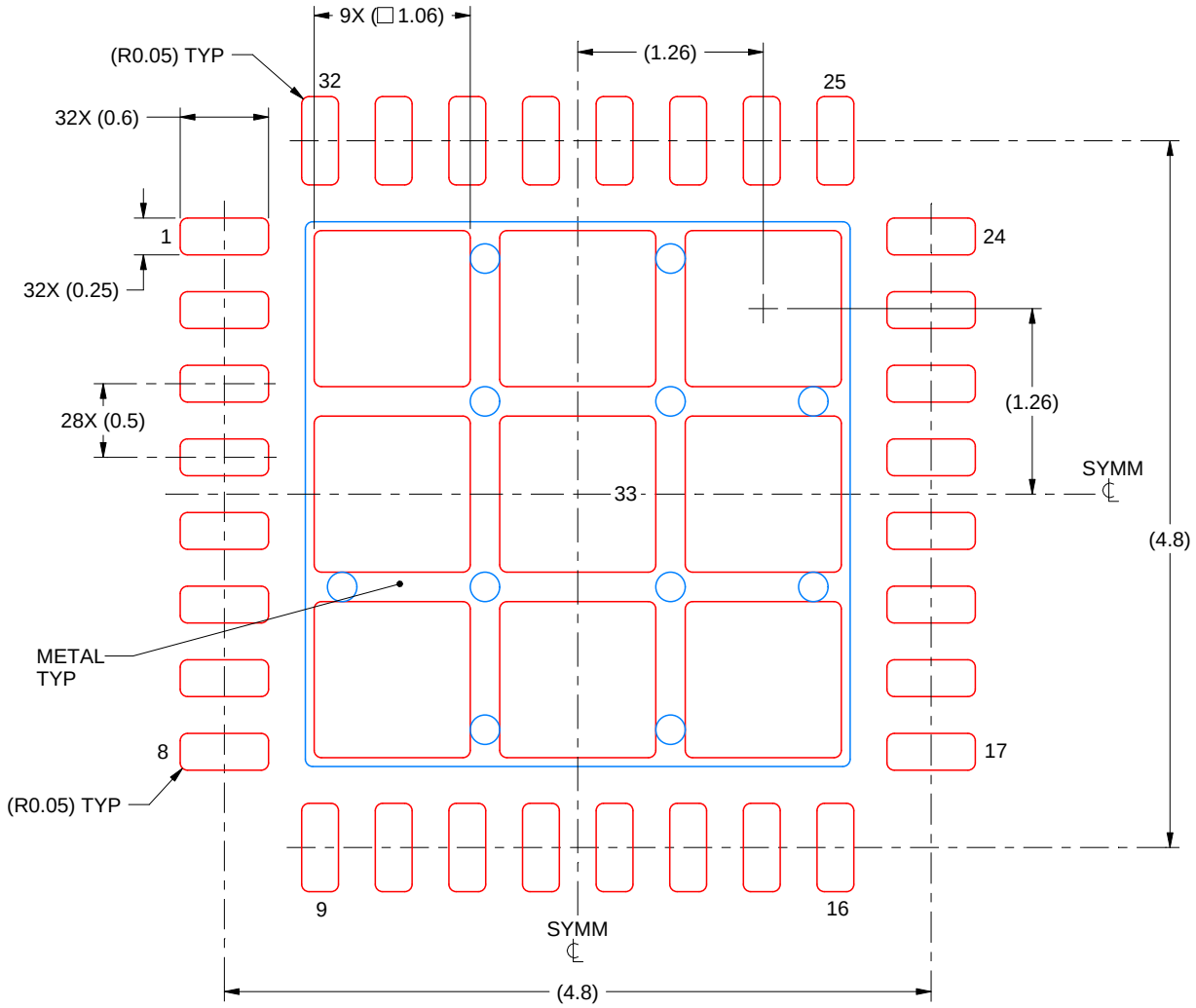



SOLDER MASK DETAILS

4227186/A 10/2021

NOTES: (continued)

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




SOLDER PASTE EXAMPLE
BASED ON 0.125 mm THICK STENCIL

EXPOSED PAD 33:
74% PRINTED SOLDER COVERAGE BY AREA UNDER PACKAGE
SCALE:20X

4227186/A 10/2021

NOTES: (continued)

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

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SCAN926260TUFX/NOPB OPA4137P TMS320DM6437ZWT6 TPS51462RGER TPS566250DDA
UC3854BDW GD75323DWR LM2619ATL/NOPB SN74CBTD3861DBR TPS2056AD CSD95410RRB
LP2989AIM-1.8/NOPB LP3990TLX-1.8/NOPB OPA4243EA/2K5 SN65LBC171DB TS5V330CDBQR
LM2831YSD/NOPB LM2852XMXAX-2.5/NOPB SN74AHC594D LM431CIM/NOPB LMP7731MA/NOPB
REG101NA-2.85/3K TPS73018DBVR DS90UB903QSQ/NOPB LM2671N-3.3/NOPB LP3882EMR-1.5/NOPB