**TEXAS INSTRUMENTS**

**TPS54302**

ZHCSF83A –MAY 2016–REVISED MAY 2016

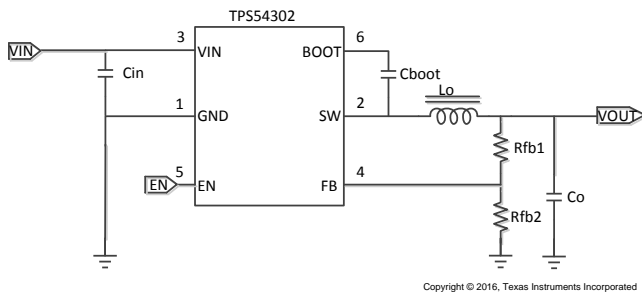# TPS54302 4.5V 至 28V 输入电压、3A 输出电流、EMI 友好型同步降压转换器

## 1 特性

- 4.5V 至 28V 宽输入电压范围
- 集成 85mΩ 和 40 mΩ 金属氧化物半导体场效应晶体管 (MOSFET)，持续输出电流为 3A
- 关断电流低至 2μA，静态电流为 45μA
- 内部 5ms 软启动
- 400kHz 固定开关频率
- 扩展频谱，可减少电磁干扰 (EMI)
- 高级 Eco-mode™脉冲跳跃
- 峰值电流模式控制
- 内部环路补偿
- 采用断续模式为两个 MOSFET 提供过流保护
- 过压保护
- 热关断
- 小外形尺寸晶体管 (SOT)-23 (6) 封装

## 2 应用

- 12V 和 24V 分布式电源总线供电
- 工业应用
  - 白色家电
- 消费类应用
  - 音频
  - 机顶盒 (STB)、数字电视 (DTV)
  - 打印机

## 3 说明

TPS54302 是一款输入电压范围为 4.5V 至 28V 的 3A 同步降压转换器。该器件包含两个集成式开关场效应晶体管 (FET) 并且具备内部回路补偿和 5ms 内部软启动功能，可降低组件数。

通过集成 MOSFET 并采用 SOT-23 封装，TPS54302 获得了高功率密度，并且在印刷电路板 (PCB) 上的占用空间非常小。

高级 Eco-mode 实现以最大限度提高了轻负载效率并降低了功率损耗。
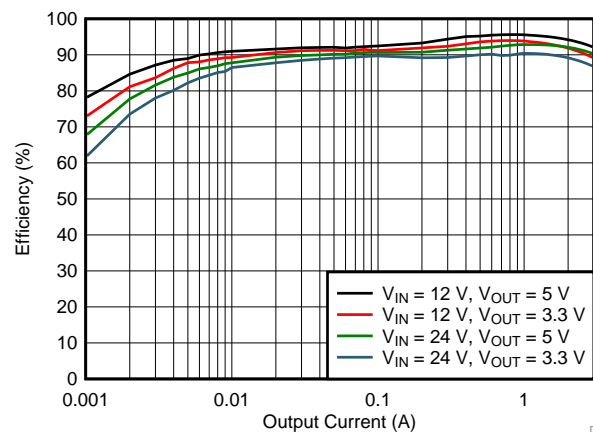
TPS54302 中引入了扩展频谱操作以降低 EMI。

高侧 MOSFET 上的逐周期电流限制功能可在过载条件下保护转换器并通过防止电流失控的低侧 MOSFET 续流电流限制功能实现功能增强。当过流持续时间超出预设时间时，将触发断续模式保护功能。

### 器件信息(1)

| 器件型号 | 封装 | 封装尺寸（标称值） |
|---|---|---|
| TPS54302 | SOT-23(6) | 1.60mm x 2.90mm |

(1) 要了解所有可用封装，请见数据表末尾的可订购产品附录。

### 简化电路原理图



Copyright © 2016, Texas Instruments Incorporated

### 效率与输出电流间的关系

English Data Sheet: SLVSDG6

## 目录

## 4 修订历史记录

**Changes from Original (May 2016) to Revision A**             **Page**

• 已更改 "产品预览"至"量产数据" .......................................................................................................................................... 1

# 5 Pin Configuration and Functions

**DDC Package**
**6-Pin SOT-23**
**Top View**
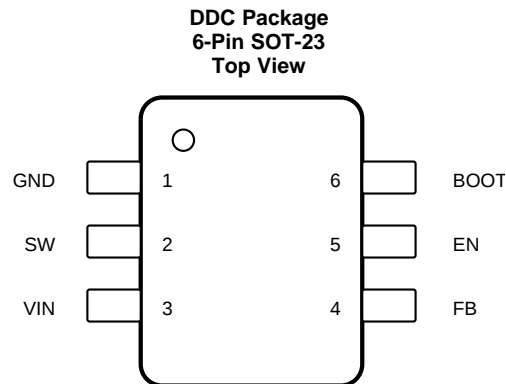


**Pin Functions**

| PIN | | TYPE[1] | DESCRIPTION |
|---|---|---|---|
| NAME | NO. | | |
| BOOT | 6 | O | Supply input for the high-side NFET gate drive circuit. Connect a 0.1-μF capacitor between BOOT and SW pins. |
| EN | 5 | I | This pin is the enable pin. Float the EN pin to enable. |
| FB | 4 | I | Converter feedback input. Connect to output voltage with feedback resistor divider. |
| GND | 1 | – | Ground pin Source terminal of low-side power NFET as well as the ground terminal for controller circuit. Connect sensitive VFB to this GND at a single point. |
| SW | 2 | O | Switch node connection between high-side NFET and low-side NFET. |
| VIN | 3 | – | Input voltage supply pin. The drain terminal of high-side power NFET. |

(1) O = Output; I = Input

# 6 Specifications

## 6.1 Absolute Maximum Ratings

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

|  |  | MIN | MAX | UNIT |
|---|---|---|---|---|
| Input voltage range, $V_I$ | VIN | –0.3 | 30 | V |
|  | EN | –0.3 | 7 | V |
|  | FB | –0.3 | 7 | V |
| Output voltage range, $V_O$ | BOOT-SW | –0.3 | 7 | V |
|  | SW | –0.3 | 30 | V |
|  | SW (20 ns transient) | –5 | 30 | V |
| Operating junction temperature, $T_J$ | | –40 | 150 | °C |
| Storage temperature range, $T_{stg}$ | | –65 | 150 | °C |

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

## 6.2 ESD Ratings

|  |  |  | VALUE | UNIT |
|---|---|---|---|---|
| $V_{(ESD)}$ | Electrostatic discharge | Human-body model (HBM), per ANSI/ESDA/JEDEC JS-001[1] | ±4000 | V |
|  |  | Charged-device model (CDM), per JEDEC specification JESD22-C101[2] | ±1500 | |

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

## 6.3 Recommended Operating Conditions

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

|  |  |  | MIN | MAX | UNIT |
|---|---|---|---|---|---|
| $V_I$ | Input voltage range | VIN | 4.5 | 28 | V |
|  |  | EN | –0.1 | 7 | V |
|  |  | FB | –0.1 | 7 | V |
| $V_O$ | Output voltage range | BOOT-SW | –0.1 | 7 | V |
|  |  | SW | –0.1 | 28 | V |
| $T_J$ | Operating junction temperature | | –40 | 125 | °C |

## 6.4 Thermal Information

|  | | TPS54302 | UNIT |
|---|---|---|---|
| **THERMAL METRIC**[1] | | **DDC (SOT-23)** | |
|  | | **6 PINS** | |
| $R_{\theta JA}$ | Junction-to-ambient thermal resistance | 87.1 | °C/W |
| $R_{\theta JC(top)}$ | Junction-to-case (top) thermal resistance | 35.5 | °C/W |
| $R_{\theta JB}$ | Junction-to-board thermal resistance | 14.4 | °C/W |
| $\psi_{JT}$ | Junction-to-top characterization parameter | 0.9 | °C/W |
| $\psi_{JB}$ | Junction-to-board characterization parameter | 14.2 | °C/W |

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

## 6.5 Electrical Characteristics

The electrical ratings specified in this section apply to all specifications in this document, unless otherwise noted. These specifications are interpreted as conditions that do not degrade the device parametric or functional specifications for the life of the product containing it. $T_J$ = –40°C to +125°C, $V_{IN}$ = 4.5 V to 28 V, (unless otherwise noted).

| PARAMETER | | TEST CONDITIONS | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| **INPUT SUPPLY** | | | | | | |
| $V_{IN}$ | Input voltage range | | 4.5 | | 28 | V |
| $I_Q$ | Non switching quiescent current | EN =5 V, VFB = 1 V | | 45 | | µA |
| $I_{OFF}$ | Shut down current | EN = GND | | 2 | | µA |
| $V_{IN(UVLO)}$ | VIN under voltage lockout | Rising $V_{IN}$ | 3.8 | 4.1 | 4.4 | V |
| | | Falling $V_{IN}$ | 3.3 | 3.6 | 3.9 | V |
| | Hysteresis | | 400 | 480 | 560 | mV |
| **ENABLE (EN PIN)** | | | | | | |
| $V_{(EN\_RISING)}$ | Enable threshold | Rising | | 1.21 | 1.28 | V |
| $V_{(EN\_FALLING)}$ | | Falling | 1.1 | 1.19 | | V |
| $I_{(EN\_INPUT)}$ | Input current | $V_{EN}$ = 1 V | | 0.7 | | µA |
| $I_{(EN\_HYS)}$ | Hysteresis current | $V_{EN}$ = 1.5 V | | 1.55 | | µA |
| **FEEDBACK AND ERROR AMPLIFIER** | | | | | | |
| $V_{FB}$ | Feedback Voltage | $V_{IN}$ = 12 V | 0.581 | 0.596 | 0.611 | V |
| **PULSE SKIP MODE** | | | | | | |
| $I_{(SKIP)}$[(1)] | Pulse skip mode peak inductor current threshold | $V_{IN}$ = 12 V, $V_{OUT}$ = 5 V, L = 10 µH | | 500 | | mA |
| **POWER STAGE** | | | | | | |
| $R_{(HSD)}$ | High-side FET on resistance | $T_A$ = 25°C, $V_{BST}$ – SW = 6 V | | 85 | | mΩ |
| $R_{(LSD)}$ | Low-side FET on resistance | $T_A$ = 25°C, $V_{IN}$ = 12 | | 40 | | mΩ |
| **CURRENT LIMIT** | | | | | | |
| $I_{(LIM\_HS)}$ | High side current limit | Maximum inductor peak current | 4 | 5 | 5.9 | A |
| $I_{(LIM\_LS)}$ | Low side source current limit | Maximum inductor valley current | 3.1 | 4 | 5.5 | A |
| **OSCILLATOR** | | | | | | |
| $F_{sw}$ | Centre switching frequency | | 290 | 400 | 510 | kHz |
| **OVER TEMPERATURE PROTECTION** | | | | | | |
| Thermal Shutdown[(1)] | Rising temperature | | | 165 | | °C |
| | Hysteresis | | | 10 | | °C |
| | Hiccup time | | | 32768 | | Cycles |

(1) Not production tested

## 6.6 Timing Requirements

| | | | MIN | TYP | MAX | UNIT |
|---|---|---|---|---|---|---|
| **OVER CURRENT PROTECTION** | | | | | | |
| $t_{HIC\_WAIT}$ | Hiccup wait time | | | 512 | | Cycles |
| $t_{HIC\_RESTART}$ | Hiccup time before restart | | | 16384 | | Cycles |
| $t_{SS}$ | Soft-start time | | | 5 | | ms |
| **ON TIME CONTROL** | | | | | | |
| $t_{MIN\_ON}$[(1)] | Minimum on time, measured at 90% to 90% and 1-A loading | | | 110 | | ns |

(1) Not production tested

## 6.7 Typical Characteristics

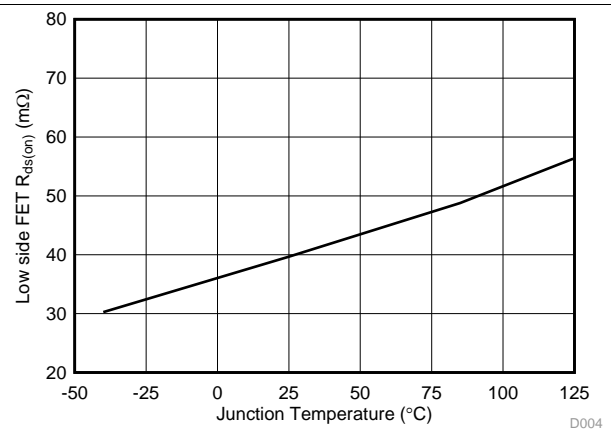$V_{IN}$ = 12, unless otherwise specified
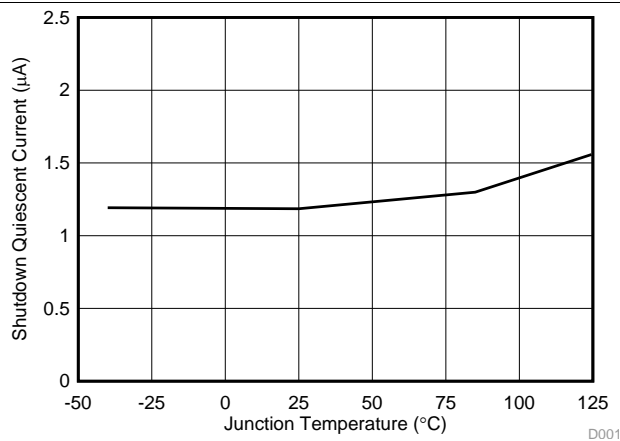


**Figure 1. Shutdown Quiescent Current vs Junction Temperature**
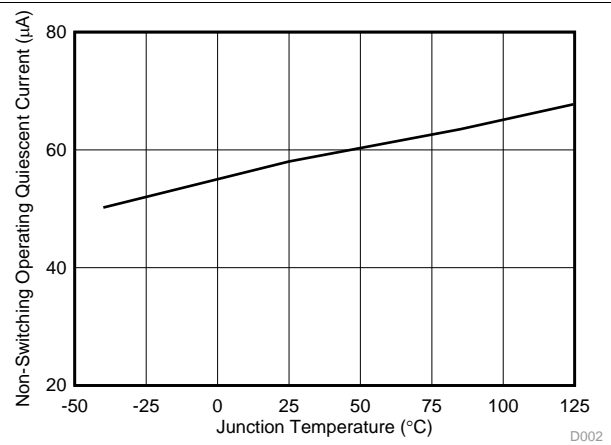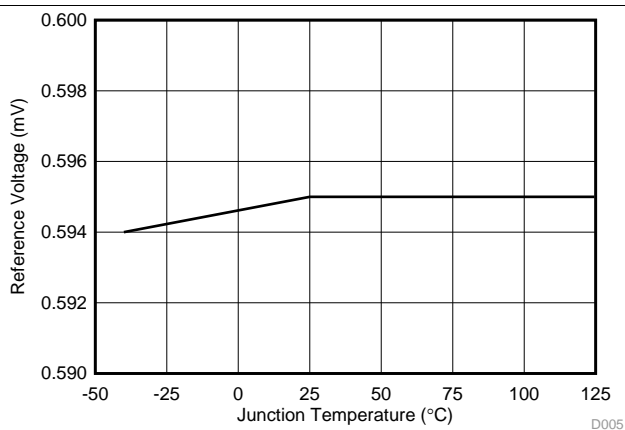


**Figure 2. Non-Switching Operating Quiescent Current vs Junction Temperature**
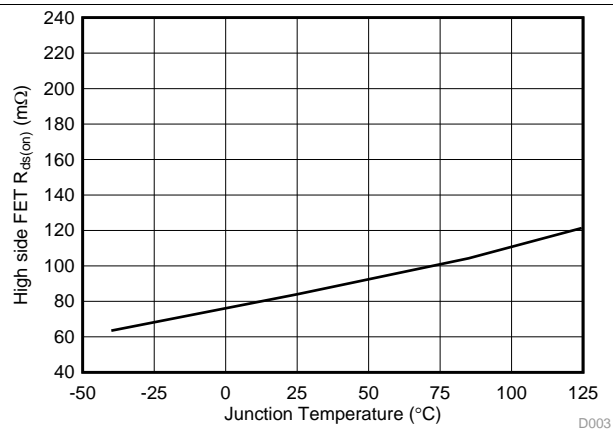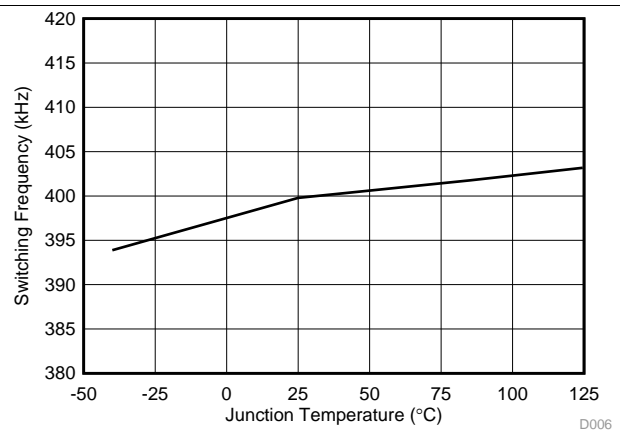


**Figure 3. High-Side Resistance vs Junction Temperature**



**Figure 4. Low-Side FET On Resistance vs Junction Temperature**
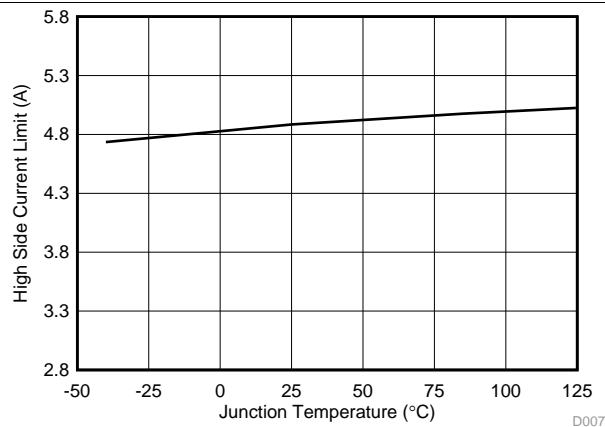


**Figure 5. Reference Voltage vs Junction Temperature**



**Figure 6. Centre Switching Frequency vs Junction Temperature**

## Typical Characteristics (continued)



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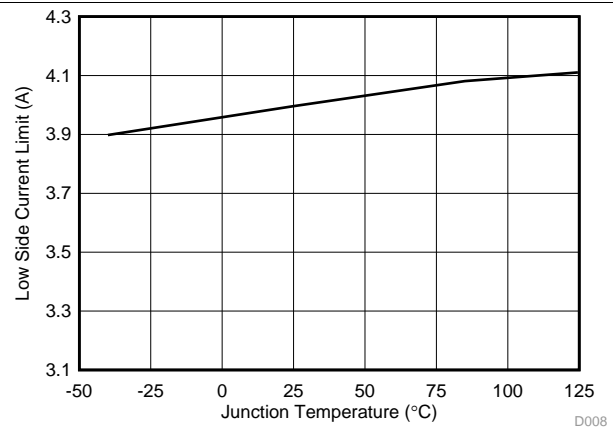


**Figure 8. Low-Side Current Limit Threshold vs Junction Temperature**



**Figure 9. BOOT-SW UVLO Threshold vs Junction Temperature**



**Figure 10. VIN UVLO Threshold vs Junction Temperature**
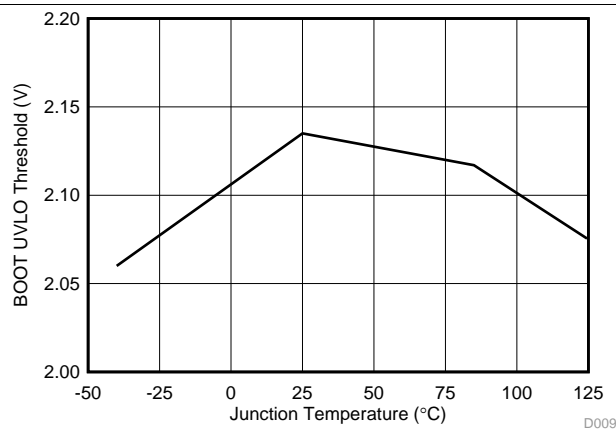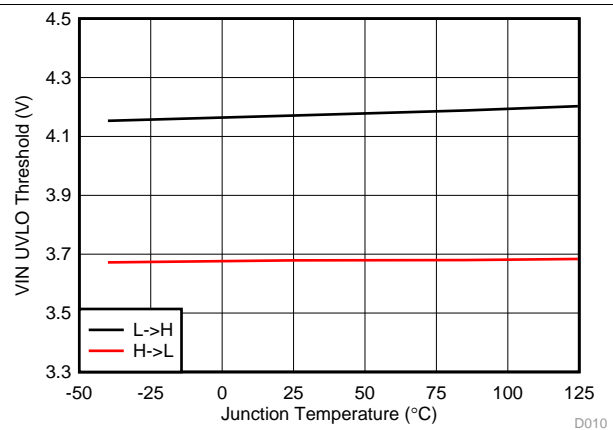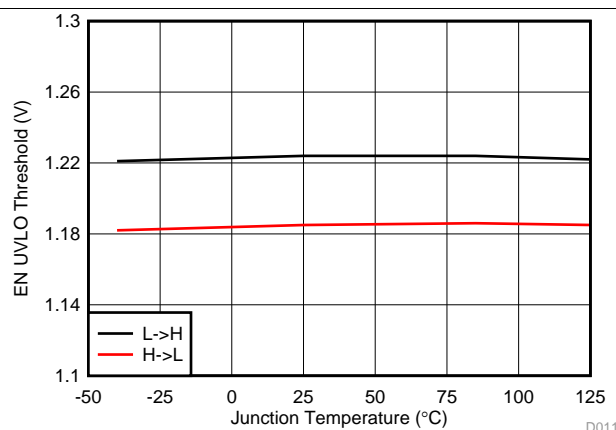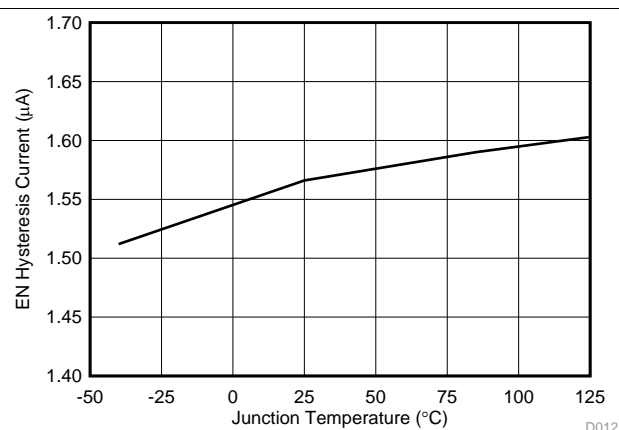


**Figure 11. EN Threshold vs Junction Temperature**



**Figure 12. EN Hysteresis Current vs Junction Temperature**

# 7 Detailed Description

## 7.1 Overview

The device is a 28-V, 3-A, synchronous step-down (buck) converter with two integrated n-channel MOSFETs. To improve performance during line and load transients the device implements a constant-frequency, peak current mode control which reduces output capacitance. The optimized internal compensation network minimizes the external component counts and simplifies the control loop design.

The TPS54302's switching frequency is fixed to 400 kHz.

The TPS54302 starts switching at VIN equal to 4.5 V. The operating current is 45 μA typically when not switching and under no load. When the device is disabled, the supply current is 2 μA typically.

The integrated 85-mΩ high-side MOSFET and 40-mΩ low-side MOSFET allow for high efficiency power supply designs with continuous output currents up to 3 A.

The TPS54302 reduces the external component count by integrating the boot recharge diode. The bias voltage for the integrated high-side MOSFET is supplied by an external capacitor on the BOOT to PH pins. The boot capacitor voltage is monitored by an UVLO circuit and will turn the high-side MOSFET off when the voltage falls below a preset threshold of 2.1 V typically.

The device minimizes excessive output overvoltage transients by taking advantage of the overvoltage comparator. When the regulated output voltage is greater than 108% of the nominal voltage, the overvoltage comparator is activated, and the high-side MOSFET is turned off and masked from turning on until the output voltage is lower than 104%.

The TPS54302 device has internal 5-ms soft-start time to minimize inrush currents.

## 7.2 Functional Block Diagram
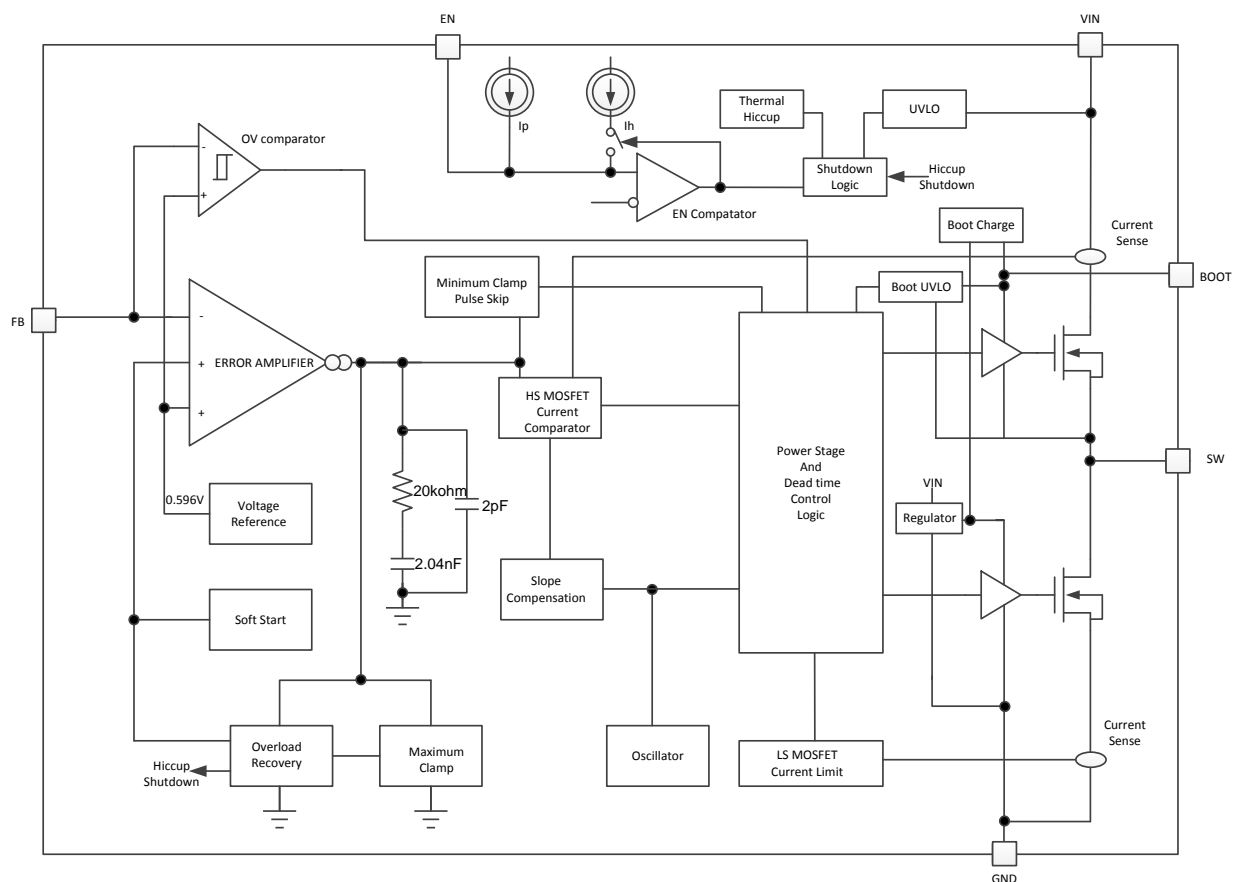


Copyright © 2016, Texas Instruments Incorporated

## 7.3 Feature Description

### 7.3.1 Fixed-Frequency PWM Control

The device uses a fixed-frequency, peak current-mode control. The output voltage is compared through external resistors on the FB pin to an internal voltage reference by an error amplifier. An internal oscillator initiates the turn on of the high-side power switch. The error amplifier output is compared to the current of the high-side power switch. When the power-switch current reaches the error amplifier output voltage level, the high side power switch is turned off and the low-side power switch is turned on. The error amplifier output voltage increases and decreases as the output current increases and decreases. The device implements a current-limit by clamping the error amplifier voltage to a maximum level and also implements a minimum clamp for improved transient-response performance.

### 7.3.2 Pulse Skip Mode

The TPS54302 is designed to operate in pulse skipping mode at light load currents to boost light load efficiency. When the peak inductor current is lower than 500 mA typically, the device enters pulse skipping mode. When the device is in pulse skipping mode, the error amplifier output voltage is clamped which prevents the high side integrated MOSFET from switching. The peak inductor current must rise above 500 mA and exit pulse skip mode. Since the integrated current comparator catches the peak inductor current only, the average load current entering pulse skipping mode varies with the applications and external output filters.

### 7.3.3 Error Amplifier

The device has a trans-conductance amplifier as the error amplifier. The error amplifier compares the FB voltage to the lower of the internal soft-start voltage or the internal 0.596-V voltage reference. The transconductance of the error amplifier is 240 µA/V typically. The frequency compensation components are placed internal between the output of the error amplifier and ground.

### 7.3.4 Slope Compensation and Output Current

The device adds a compensating ramp to the signal of the switch current. This slope compensation prevents sub-harmonic oscillations as the duty cycle increases. The available peak inductor current remains constant over the full duty-cycle range.

### 7.3.5 Enable and Adjusting Undervoltage Lockout

The EN pin provides electrical on and off control of the device. When the EN pin voltage exceeds the threshold voltage, the device begins operation. If the EN pin voltage is pulled below the threshold voltage, the regulator stops switching and enters the low-quiescent (IQ) state.

The EN pin has an internal pullup-current source which allows the user to float the EN pin to enable the device. If an application requires control of the EN pin, use open-drain or open-collector output logic to interface with the pin.

The device implements internal undervoltage-lockout (UVLO) circuitry on the VIN pin. The device is disabled when the VIN pin voltage falls below the internal VIN UVLO threshold. The internal VIN UVLO threshold has a hysteresis of 480 mV.

If an application requires a higher UVLO threshold on the VIN pin, then the EN pin can be configured as shown in Figure 13. When using the external UVLO function, setting the hysteresis at a value greater than 500 mV is recommended.

The EN pin has a small pull-up current, Ip, which sets the default state of the pin to enable when no external components are connected. The pull-up current is also used to control the voltage hysteresis for the UVLO function because it increases by Ih when the EN pin crosses the enable threshold. Use Equation 1 and Equation 2 to calculate the values of R4 and R5 for a specified UVLO threshold.

## Feature Description (continued)
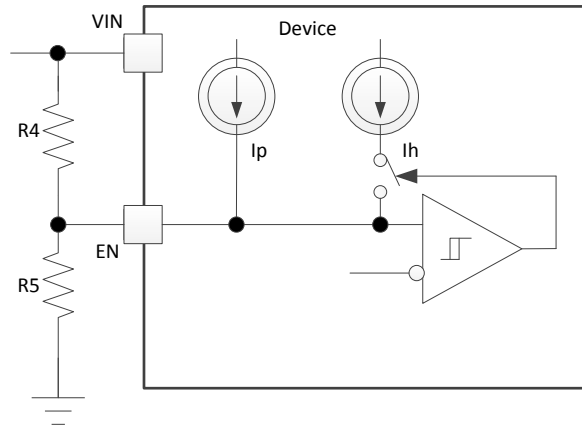


**Figure 13. Adjustable VIN Undervoltage Lockout**

$$R4 = \frac{V_{START}\left(\dfrac{V_{ENfalling}}{V_{ENrising}}\right) - V_{STOP}}{I_p\left(1 - \dfrac{V_{ENfalling}}{V_{ENrising}}\right) + I_h}$$

(1)

Where:

$I_p$ = 0.7 µA
$I_h$ = 1.55 µA
$V_{ENfalling}$ = 1.19 V
$V_{ENrising}$ = 1.22 V

$$R5 = \frac{R4 \times V_{ENfalling}}{V_{STOP} - V_{ENfalling} + R4\left(I_p + I_h\right)}$$

(2)

### 7.3.6 Safe Startup into Pre-Biased Outputs

The device has been designed to prevent the low-side MOSFET from discharging a pre-biased output. During monotonic pre-biased startup, both high-side and low-side MOSFETs are not allowed to be turned on until the internal soft-start voltage is higher than FB pin voltage.

### 7.3.7 Voltage Reference

The voltage reference system produces a precise ±2.5% voltage-reference over temperature by scaling the output of a temperature stable bandgap circuit. The typical voltage reference is designed at 0.596 V.

### 7.3.8 Adjusting Output Voltage

The output voltage is set with a resistor divider from the output node to the FB pin. It is recommended to use divider resistors with 1% tolerance or better. Start with 100 kΩ for the upper resistor divider, use Equation 3 to calculate the output voltage. To improve efficiency at light loads consider using larger value resistors. If the values are too high the regulator is more susceptible to noise and voltage errors from the FB input current are noticeable.

$$V_{out} = V_{ref} \times \left[\frac{R2}{R3} + 1\right]$$

(3)

### 7.3.9 Internal Soft-Start

The TPS54302 device uses the internal soft-start function. The internal soft start time is set to 5 ms typically.

## Feature Description (continued)

### 7.3.10 Bootstrap Voltage (BOOT)

The TPS54302 has an integrated boot regulator and requires a 0.1-µF ceramic capacitor between the BOOT and SW pins to provide the gate drive voltage for the high-side MOSFET. A ceramic capacitor with an X7R or X5R grade dielectric is recommended because of the stable characteristics over temperature and voltage. To improve drop out, the TPS54302 is designed to operate at 100% duty cycle as long as the BOOT to SW pin voltage is greater than 2.1 V typically.
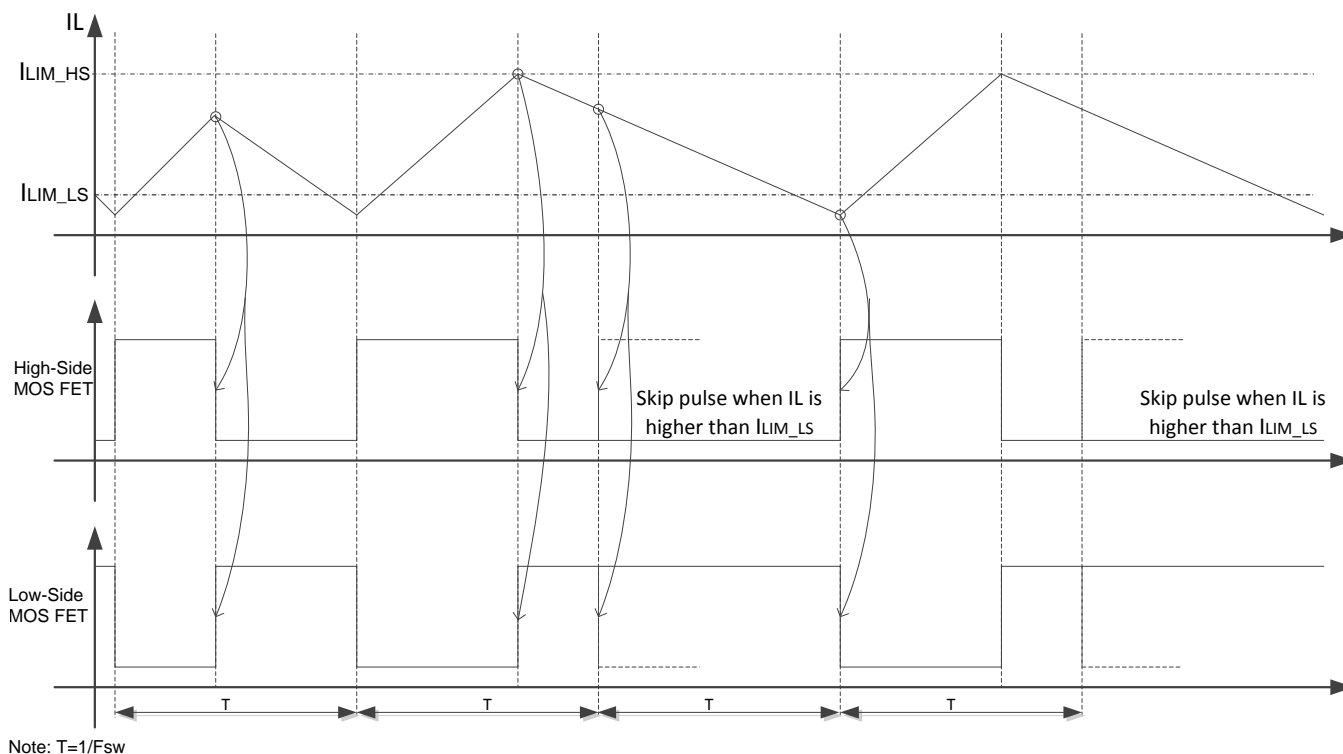
### 7.3.11 Overcurrent Protection

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

#### 7.3.11.1 High-Side MOSFET Overcurrent Protection

The device implements current mode control which uses the internal COMP voltage to control the turn off of the high-side MOSFET and the turn on of the low-side MOSFET on a cycle-by-cycle basis. During each cycle, the switch current and the current reference generated by the internal COMP voltage are compared. When the peak switch current intersects the current reference the high-side switch turns off.

#### 7.3.11.2 Low-Side MOSFET Overcurrent Protection

While the low-side MOSFET is turned on, the conduction current is monitored by the internal circuitry. During normal operation the low-side MOSFET sources current to the load. At the end of every clock cycle, the low-side MOSFET sourcing current is compared to the internally set low-side sourcing current-limit. The inductor valley current is exceeded the low-side source current limit, the high-side MOSFET does not turn on and the low-side MOSFET stays on for the next cycle. The high-side MOSFET turns on again when the inductor valley current is below the low-side sourcing current-limit at the start of a cycle as shown in Figure 14.



**Figure 14. Overcurrent Protection for Both MOSFETs**

Furthermore, if an output overload condition occurs for more than the hiccup wait time, which is programmed for 512 switching cycles, the device shuts down and restarts after the hiccup time of 16384 cycles. The hiccup mode helps to reduce the device power dissipation under severe overcurrent conditions.

## Feature Description (continued)

### 7.3.12 Spread Spectrum

In order to reduce EMI, the TPS54302 introduces frequency spread spectrum. The jittering span is ±6% of the switching frequency with 1/512 swing frequency.

### 7.3.13 Output Overvoltage Protection (OVP)

The TPS54302 incorporates an overvoltage transient protection (OVTP) circuit to minimize output voltage overshoot when recovering from output fault conditions or strong unload transients. The OVTP circuit includes an overvoltage comparator to compare the FB pin voltage and internal thresholds. When the FB pin voltage goes above 108% × Vref, the high-side MOSFET is forced off. When the FB pin voltage falls below 104% × Vref, the high-side MOSFET is enabled again.

### 7.3.14 Thermal Shutdown

The internal thermal-shutdown circuitry forces the device to stop switching if the junction temperature exceeds 165°C typically. When the junction temperature drops below 155°C typically, the internal thermal-hiccup timer begins to count. The device reinitiates the power-up sequence after the built-in thermal-shutdown hiccup time (32768 cycles) is over.

## 7.4 Device Functional Modes

### 7.4.1 Normal Operation

When the input voltage is above the UVLO threshold, the TPS54302 can operate in their normal switching modes. Normal continuous conduction mode (CCM) occurs when inductor peak current is above 0 A. In CCM, the TPS54302 operates at a fixed frequency.

### 7.4.2 Eco-mode™ Operation

The device is designed to operate in high-efficiency pulse-skipping mode under light load conditions. Pulse skipping initiates when the switch current falls to 500 mA typically. During pulse skipping, the low-side FET turns off when the switch current falls to 0 A. The switching node (the SW pin) waveform takes on the characteristics of discontinuous conduction mode (DCM) operation and the apparent switching frequency decreases. As the output current decreases, the perceived time between switching pulses increases.

# 8 Application and Implementation

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

## 8.1 Application Information

The TPS54302 is typically used as a step down converter, which converts an input voltage from 8 V - 28 V to a fixed output voltage of 5 V.

## 8.2 Typical Application

### 8.2.1 TPS54302 8-V to 28-V Input, 5-V Output Converter
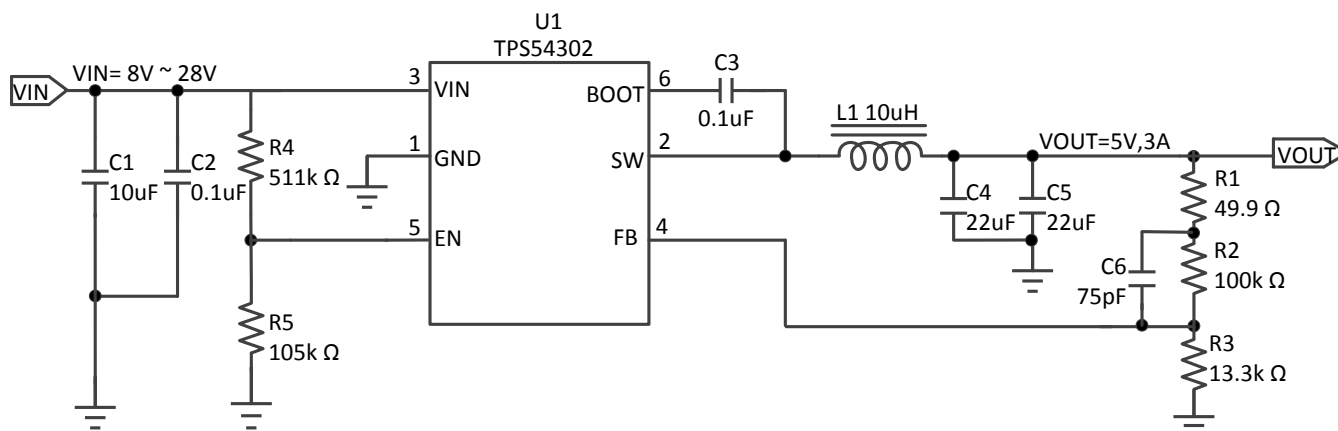


Copyright © 2016, Texas Instruments Incorporated

**Figure 15. 5-V, 3-A Reference Design**

### 8.2.2 Design Requirements

For this design example, use the parameters in Table 1.

**Table 1. Design Parameters**

| PARAMETER | VALUE |
| --- | --- |
| Input voltage range | 8 V to 28 V |
| Output voltage | 5 V |
| Output current | 3 A |
| Transient response, 1.5-A load step | $\Delta V_{OUT}$ = ±5 % |
| Input ripple voltage | 400 mV |
| Output voltage ripple | 30 mVpp |
| Switching frequency | 400 kHz |

### 8.2.3 Detailed Design Procedure

#### 8.2.3.1 Input Capacitor Selection

The device requires an input decoupling capacitor and a bulk capacitor is needed depending on the application. A ceramic capacitor over 10 µF is recommended for the decoupling capacitor. An additional 0.1-µF capacitor (C2) from VIN to GND is optional to provide additional high frequency filtering. The capacitor voltage rating needs to be greater than the maximum input voltage.

Use Equation 4 to calculate the input ripple voltage ($\Delta V_{IN}$).

$$\Delta V_{IN} = \frac{I_{OUT(MAX)} \times 0.25}{C_{BULK} \times f_{sw}} + \left( I_{OUT(MAX)} \times ESR_{MAX} \right)$$

(4)

where:

  $C_{BULK}$ is the bulk capacitor value

  $f_{SW}$ is the switching frequency

  $I_{OUT(MAX)}$ is the maximum loading current

  $ESR_{MAX}$ is maximum series resistance of the bulk capacitor

The maximum RMS (root mean square) ripple current must also be checked. For worst case conditions, use Equation 5 to calculate $I_{CIN(RMS)}$.

$$I_{CIN(RMS)} = \frac{I_{O(MAX)}}{2}$$

(5)

The actual input-voltage ripple is greatly affected by parasitic associated with the layout and the output impedance of the voltage source. Design Requirements show the actual input voltage ripple for this circuit which is larger than the calculated value. This measured value is still below the specified input limit of 400 mV. The maximum voltage across the input capacitors is VIN (MAX) + $\Delta$VIN/2. The selected bypass capacitor is rated for 35 V and the ripple current capacity is greater than 2 A. Both values provide ample margin. The maximum ratings for voltage and current must not be exceeded under any circumstance.

#### 8.2.3.2 Bootstrap Capacitor Selection

A 0.1-µF ceramic capacitor must be connected between the BOOT to SW pin for proper operation. It is recommended to use a ceramic capacitor.

#### 8.2.3.3 Output Voltage Set Point

The output voltage of the TPS54302 device is externally adjustable using a resistor divider network. In the application circuit of Figure 15, this divider network is comprised of R2 and R3. Use Equation 6 and Equation 7 to calculate the relationship of the output voltage to the resistor divider.

$$R3 = \frac{R2 \times V_{ref}}{V_{OUT} - V_{ref}}$$

(6)

$$V_{out} = V_{ref} \times \left[ \frac{R2}{R3} + 1 \right]$$

(7)

Select a value of R2 to be approximately 100 kΩ. Slightly increasing or decreasing R3 can result in closer output voltage matching when using standard value resistors. In this design, R2 = 100 kΩ and R3 = 13.3 kΩ which results in a 5-V output voltage. The 49.9-Ω resistor, R1, is provided as a convenient location to break the control loop for stability testing.

#### 8.2.3.4 Undervoltage Lockout Set Point

The undervoltage lockout (UVLO) set point can be adjusted using the external-voltage divider network of R4 and R5. R4 is connected between the VIN and EN pins of the TPS54302 device. R5 is connected between the EN and GND pins. The UVLO has two thresholds, one for power up when he input voltage is rising and one for power down or brown outs when the input voltage is falling. For the example design, the minimum input voltage is 8 V, so the start voltage threshold is set to 6.74 V and the stop voltage threshold is set to 5.83 V. Use Equation 1 and Equation 2 to calculate the values for the upper and lower resistor values of R4 and R5.

### 8.2.3.5 *Output Filter Components*

Two components must be selected for the output filter, the output inductor ($L_O$) and $C_O$.

#### 8.2.3.5.1 Inductor Selection

Use Equation 8 to calculate the minimum value of the output inductor ($L_{MIN}$).

$$L_{MIN} = \frac{V_{OUT} \times \left( V_{IN(MAX)} - V_{OUT} \right)}{V_{IN(MAX)} \times K_{IND} \times I_{OUT} \times f_{sw}}$$

(8)

Where:

> $K_{IND}$ is a coefficient that represents the amount of inductor ripple current relative to the maximum output current.

In general, the value of $K_{IND}$ is at the discretion of the designer; however, the following guidelines may be used. For designs using low-ESR output capacitors, such as ceramics, a higher $K_{IND}$ can be used. When using higher ESR output capacitors, $K_{IND} = 0.2$ yields better results.

For this design example, use $K_{IND} = 0.35$. The minimum inductor value is calculated as 9.78 µH. For this design, a close standard value of 10 µH was selected for $L_{MIN}$.

For the output filter inductor, the RMS current and saturation current ratings must not be exceeded. Use Equation 9 to calculate the RMS inductor current ($I_{L(RMS)}$).

$$I_{L(MAX)} = \sqrt{I_{OUT(MAX)}^2 + \frac{1}{12} \times \left( \frac{V_{OUT} \times \left( V_{IN(MAX)} - V_{OUT} \right)}{V_{IN(MAX)} \times L_{OUT} \times f_{SW} \times 0.8} \right)^2}$$

(9)

Use Equation 10 to calculate the peak inductor current ($I_{L(PK)}$).

$$I_{L(PK)} = I_{OUT(MAX)} + \frac{V_{OUT} \times \left( V_{IN(MAX)} - V_{OUT} \right)}{1.6 \times V_{IN(MAX)} \times L_{OUT} \times f_{SW}}$$

(10)

Smaller or larger inductor values can be used depending on the amount of ripple current the designer wants to allow so long as the other design requirements are met. Larger value inductors have lower AC current and result in lower output voltage ripple. Smaller inductor values increase AC current and output voltage ripple.

#### 8.2.3.5.2 Output Capacitor Selection

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

The desired response to a large change in the load current is the first criterion. The output capacitor must supply the load with current when the regulator cannot. This situation occurs if the desired hold-up times are present for the regulator. In this case, the output capacitor must hold the output voltage above a certain level for a specified amount of time after the input power is removed. The regulator is also temporarily unable to supply sufficient output current if a large, fast increase occurs affecting the current requirements of the load, such as a transition from no load to full load. The regulator usually requires two or more clock cycles for the control loop to notice the change in load current and output voltage and to adjust the duty cycle to react to the change. The output capacitor must be sized to supply the extra current to the load until the control loop responds to the load change. The output capacitance must be large enough to supply the difference in current for 2 clock cycles while only allowing a tolerable amount of drop in the output voltage. Use Equation 11 to calculate the minimum required output capacitance.

$$C_O > \frac{2 \times \Delta I_{OUT}}{f_{sw} \times \Delta V_{OUT}}$$

(11)

where:

$\Delta I_{OUT}$ is the change in output current

$f_{SW}$ is the switching frequency of the regulator

$\Delta V_{(OUT )}b$ is the allowable change in the output voltage

For this example, the transient load response is specified as a 5% change in the output voltage, $V_{OUT}$, for a load step of 1.5 A. For this example, $\Delta I_{OUT} = 1.5$ A and $\Delta V_{OUT} = 0.05 \times 5 = 0.25$ V. Using these values results in a minimum capacitance of 30 μF. This value does not consider the ESR of the output capacitor in the output voltage change. For ceramic capacitors, the ESR is usually small enough to ignore in this calculation.

Equation 12 calculates the minimum output capacitance required to meet the output voltage ripple specification. In this case, the maximum output voltage ripple is 30 mV. Under this requirement, Equation 12 yields 10.7 μF.

$$C_O > \frac{1}{8 \times f_{SW}} \times \frac{1}{\dfrac{V_{OUTripple}}{I_{ripple}}}$$

(12)

where:

$f_{SW}$ is the switching frequency

$V_{(OUTripple)}$ is the maximum allowable output voltage ripple

$I_{(ripple)}$ is the inductor ripple current

Use Equation 13 to calculate the maximum ESR an output capacitor can have to meet the output-voltage ripple specification. Equation 13 indicates the ESR should be less than 29.2 mΩ. In this case, the ESR of the ceramic capacitor is much smaller than 29.2 mΩ.

$$R_{ESR} < \frac{V_{OUTripple}}{I_{ripple}}$$

(13)

The output capacitor can affect the crossover frequency $f_o$. Considering to the loop stability and effect of the internal parasitic parameters, choose the crossover frequency less than 40 kHz without considering the feed forward capacitor. A simple estimation for the crossover frequency without feed forward capacitor C6 is shown in Equation 14, assuming $C_{OUT}$ has small ESR.

$$f_o = \frac{5.1}{V_{OUT} \times C_{OUT}}$$

(14)

Additional capacitance deratings for aging, temperature, and DC bias should be considered which increases this minimum value. For this example, two 22-uF 25-V, X7R ceramic capacitors are used. Capacitors generally have limits to the amount of ripple current they can handle without failing or producing excess heat. An output capacitor that can support the inductor ripple current must be specified. Some capacitor data sheets specify the RMS value of the maximum ripple current. Use Equation 15 to calculate the RMS ripple current that the output capacitor must support. For this application, Equation 15 yields 296 mA for each capacitor.

$$I_{COUT(RMS)} = \frac{1}{\sqrt{12}} \times \left( \frac{V_{OUT} \times \left( V_{IN(MAX)} - V_{OUT} \right)}{V_{IN(MAX)} \times L_{OUT} \times f_{SW} \times N_C} \right)$$

(15)

#### 8.2.3.5.3 Feed-Forward Capacitor

The TPS54302 is internally compensated and the internal compensation network is composed of two capacitors and one resister shown on the block diagram. Depending on the $V_{OUT}$, if the output capacitor $C_{OUT}$ is dominated by low ESR (ceramic types) capacitors, it could result in low phase margin. To improve the phase boost an external feedforward capacitor C6 can be added in parallel with R2. C6 is chosen such that phase margin is boosted at the crossover frequency.

Equation 16 for C6 was tested:

$$C6 = \frac{1}{2\pi f_o} \times \frac{1}{R_2}$$

(16)

For this design, C6 = 75 pF. C6 is not needed when $C_{OUT}$ has high ESR, and C6 calculated from Equation 16 should be reduced with medium ESR. Table 2 can be used as a starting point.

**Table 2. Recommended Component Values**

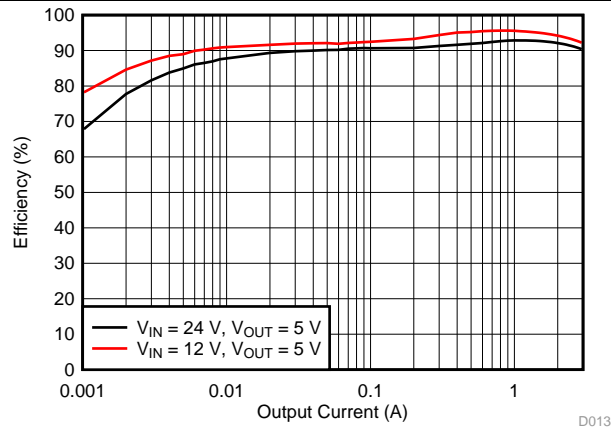| $V_{OUT}$ (V) | L (µH) | $C_{OUT}$ (µF) | R2 (kΩ) | R3 (kΩ) | C8 (pF) |
|---|---|---|---|---|---|
| 1.8 | 4.7 | 66 | 100 | 49.9 | 33 |
| 2.5 | 5.6 | 66 | 100 | 31.6 | 47 |
| 3.3 | 6.8 | 44 | 100 | 22.1 | 47 |
| 5 | 10 | 44 | 100 | 13.3 | 75 |
| 12 | 15 | 44 | 100 | 5.23 | 100 |

### 8.2.4 Application Curves
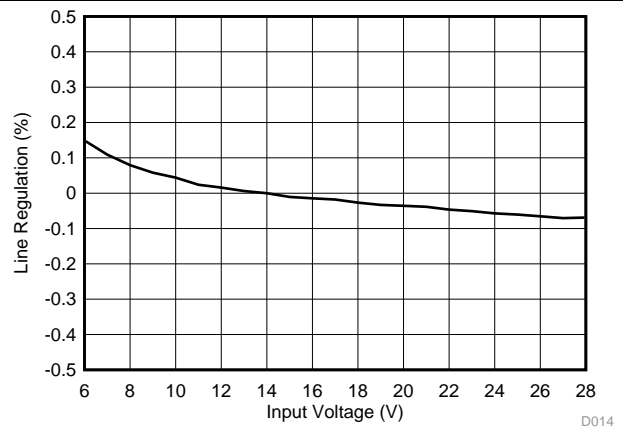


**Figure 16. Efficiency**
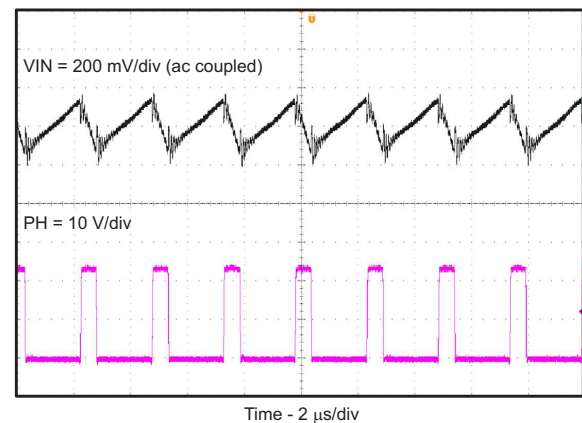


**Figure 17. Line Regulation**
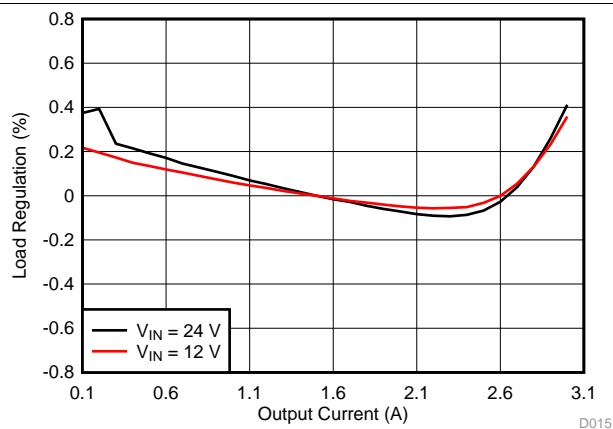


**Figure 18. Load Regulation**



**Figure 19. Input Voltage Ripple**



$I_{OUT}$ = 0 A

**Figure 20. Output Voltage Ripple**



$I_{OUT}$ = 10 mA

**Figure 21. Output Voltage Ripple**

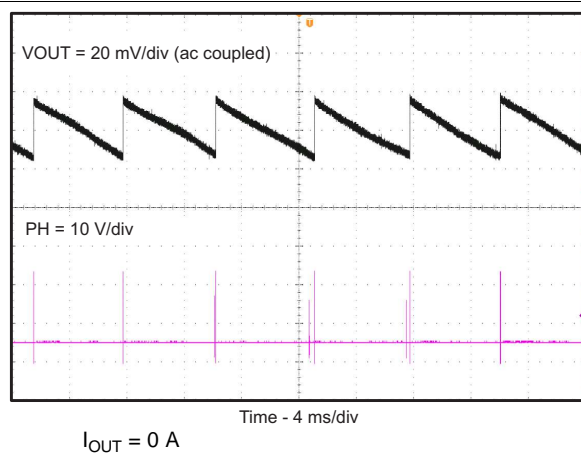VOUT = 20 mV/div (ac coupled)

PH = 10 V/div

Time - 4 µs/div

I_OUT = 100 mA

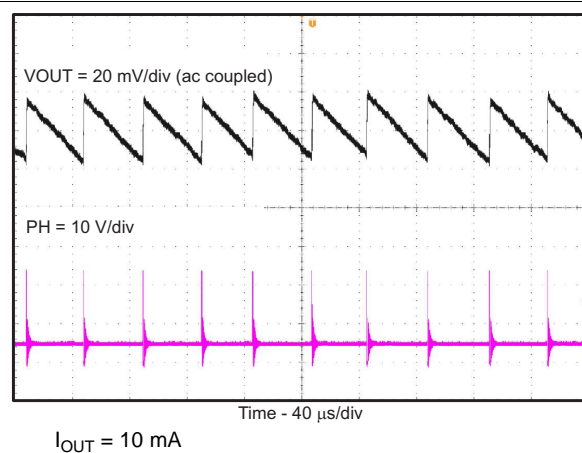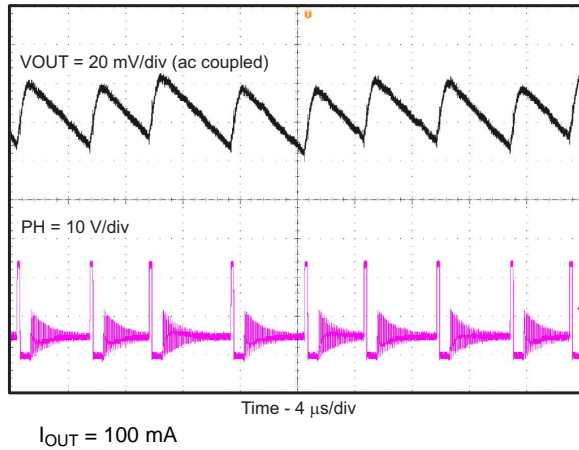**Figure 22. Output Voltage Ripple**



VOUT = 20 mV/div (ac coupled)

PH = 10 V/div

Time - 2 µs/div

I_OUT = 1.5 A
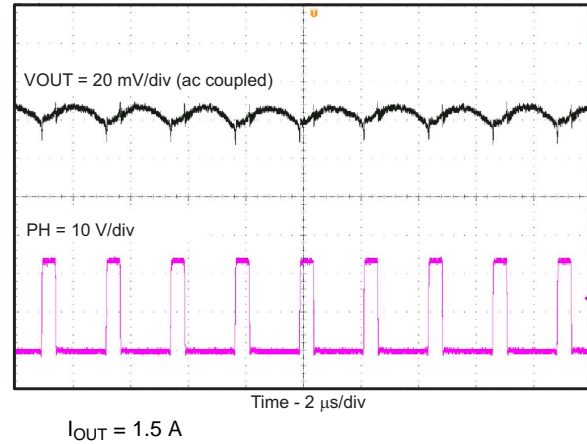
**Figure 23. Output Voltage Ripple**



VOUT - 20 mV/div (ac coupled)

PH = 10 V/div

Time - 200 µs/div

I_OUT = 3 A
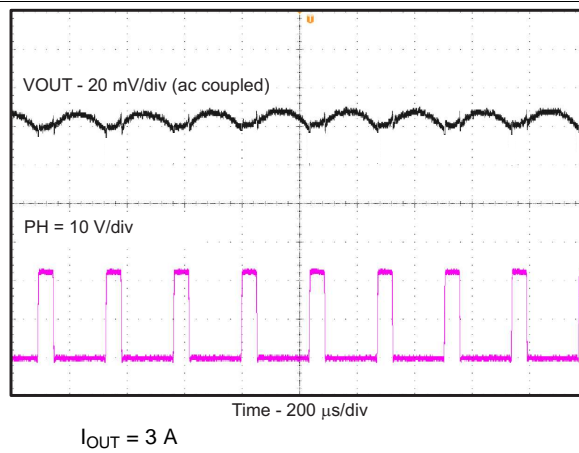
**Figure 24. Transient Response**



VOUT = 100 mV/div (ac coupled)

IOUT = 1 A/div

Time - 200 µs/div

0.75 to 2.25-A load step, slew rate = 250 mA/µsec
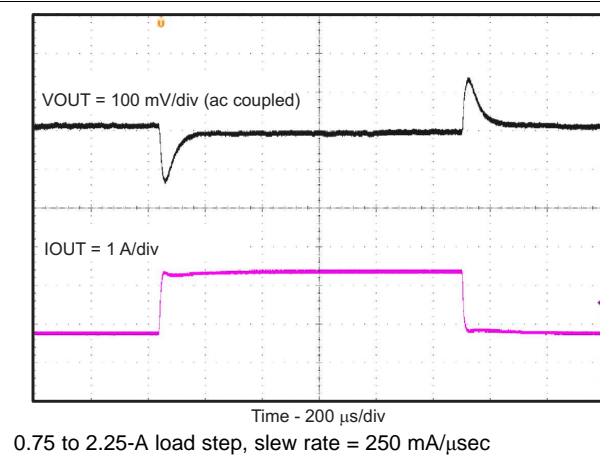
**Figure 25. Transient Response**



Vin = 10 V/div

EN = 2 V/div

Vout = 2 V/div

Time - 2 ms/div
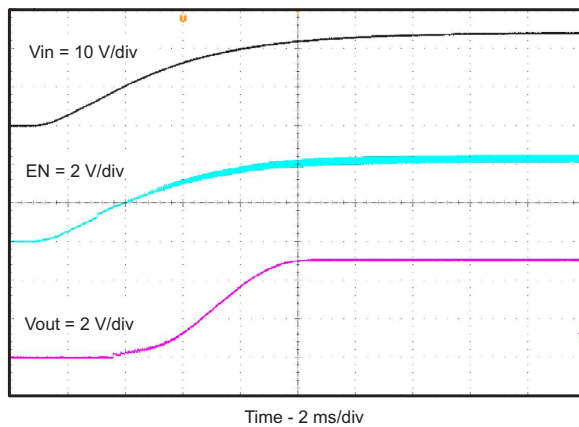
**Figure 26. Start-Up Relative to VIN**



Vin = 10 V/div

EN = 2 V/div

Vout = 2 V/div

Time - 4 ms/div
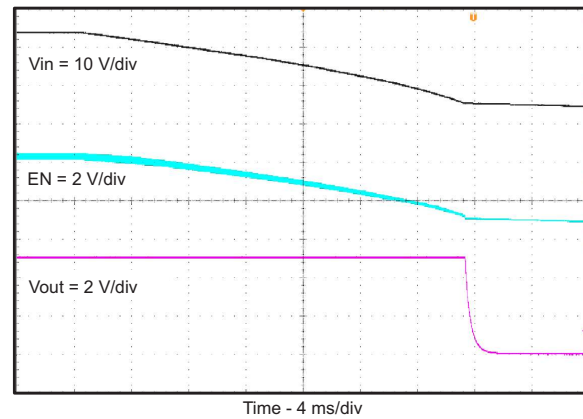
**Figure 27. Shutdown Relative to VIN**
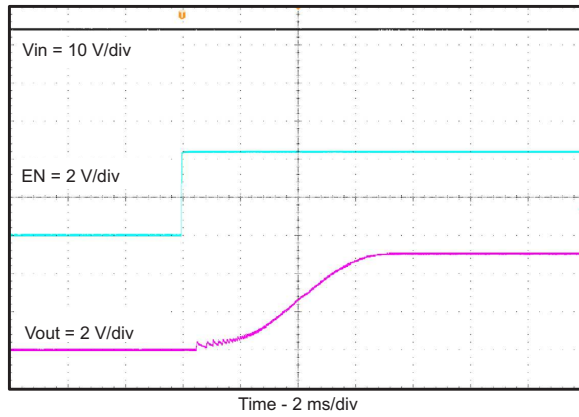
**Figure 28. Start-Up Relative to EN**
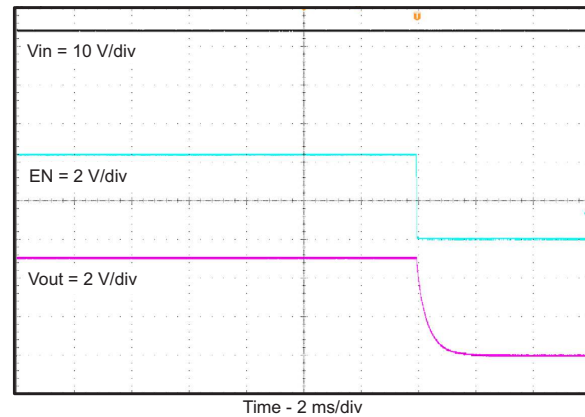


**Figure 29. Shutdown Relative to EN**

# 9 Power Supply Recommendations

The device is designed to operate from an input voltage supply range between 4.5 V and 28 V. This input supply must be well regulated. If the input supply is located more than a few inches from the device or converter, additional bulk capacitance may be required in addition to the ceramic bypass capacitors. An electrolytic capacitor with a value of 47 µF is a typical choice.

# 10 Layout

## 10.1 Layout Guidelines

- VIN and GND traces should be as wide as possible to reduce trace impedance. The wide areas are also of advantage from the view point of heat dissipation.
- The input capacitor and output capacitor should be placed as close to the device as possible to minimize trace impedance.
- Provide sufficient vias for the input capacitor and output capacitor.
- Keep the SW trace as physically short and wide as practical to minimize radiated emissions.
- Do not allow switching current to flow under the device.
- A separate VOUT path should be connected to the upper feedback resistor.
- Make a Kelvin connection to the GND pin for the feedback path.
- Voltage feedback loop should be placed away from the high-voltage switching trace, and preferably has ground shield.
- The trace of the VFB node should be as small as possible to avoid noise coupling.
- The GND trace between the output capacitor and the GND pin should be as wide as possible to minimize its trace impedance.

## 10.2 Layout Example
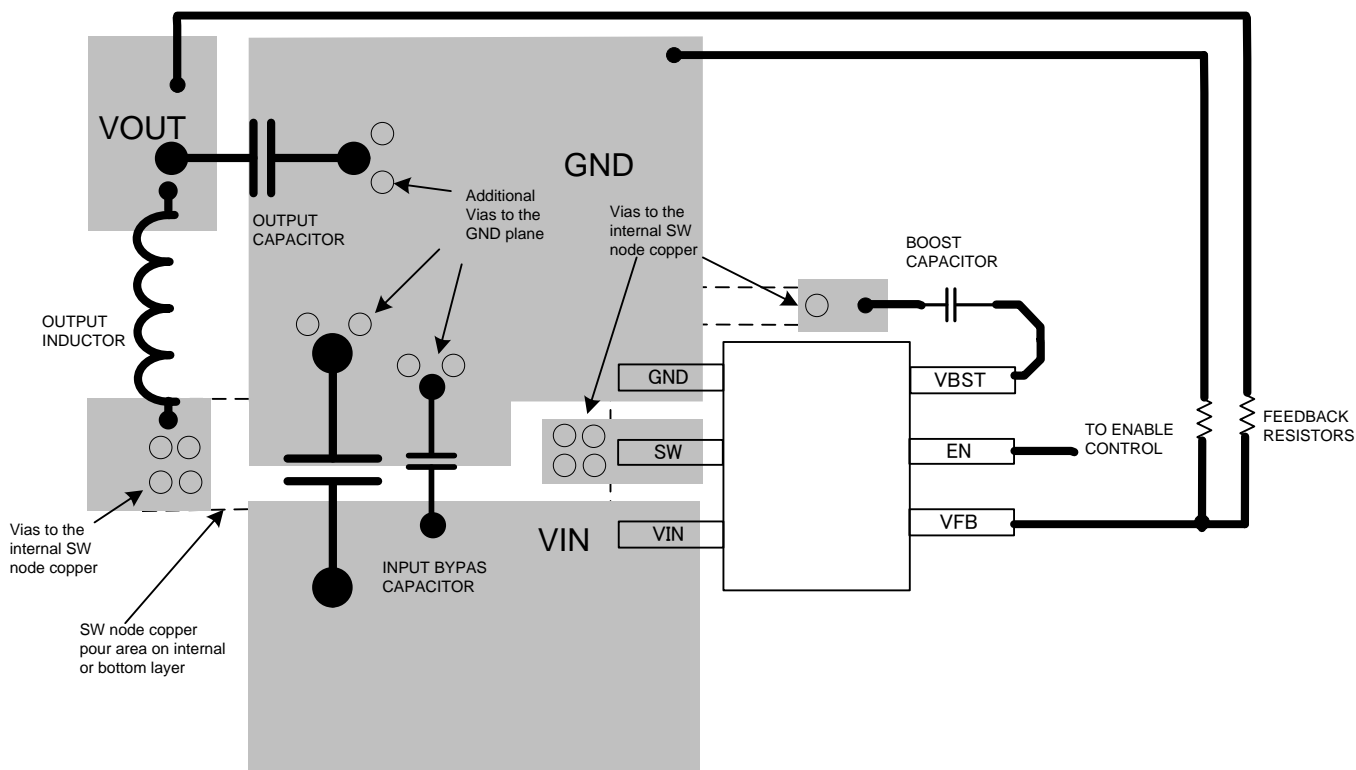


**Figure 30. Board Layout**

## 11   器件和文档支持

### 11.1   社区资源

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

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

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

### 11.2   商标

Eco-mode, E2E are trademarks of Texas Instruments.
All other trademarks are the property of their respective owners.

### 11.3   静电放电警告

这些装置包含有限的内置 ESD 保护。 存储或装卸时，应将导线一起截短或将装置放置于导电泡棉中，以防止 MOS 门极遭受静电损伤。

### 11.4   Glossary

SLYZ022 — *TI Glossary*.

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

## 12   机械、封装和可订购信息

以下页中包括机械、封装和可订购信息。这些信息是针对指定器件可提供的最新数据。这些数据会在无通知且不对本文档进行修订的情况下发生改变。欲获得该数据表的浏览器版本，请查阅左侧的导航栏。

## 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 |
|---|---|---|---|---|---|---|---|---|---|---|---|
| TPS54302DDCR | ACTIVE | SOT-23-THIN | DDC | 6 | 3000 | RoHS & Green | SN | Level-1-260C-UNLIM | -40 to 125 | 4302 | Samples |
| TPS54302DDCT | ACTIVE | SOT-23-THIN | DDC | 6 | 250 | RoHS & Green | SN | Level-1-260C-UNLIM | -40 to 125 | 4302 | Samples |

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

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

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

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

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

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

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

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

# PACKAGE MATERIALS INFORMATION

## 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**



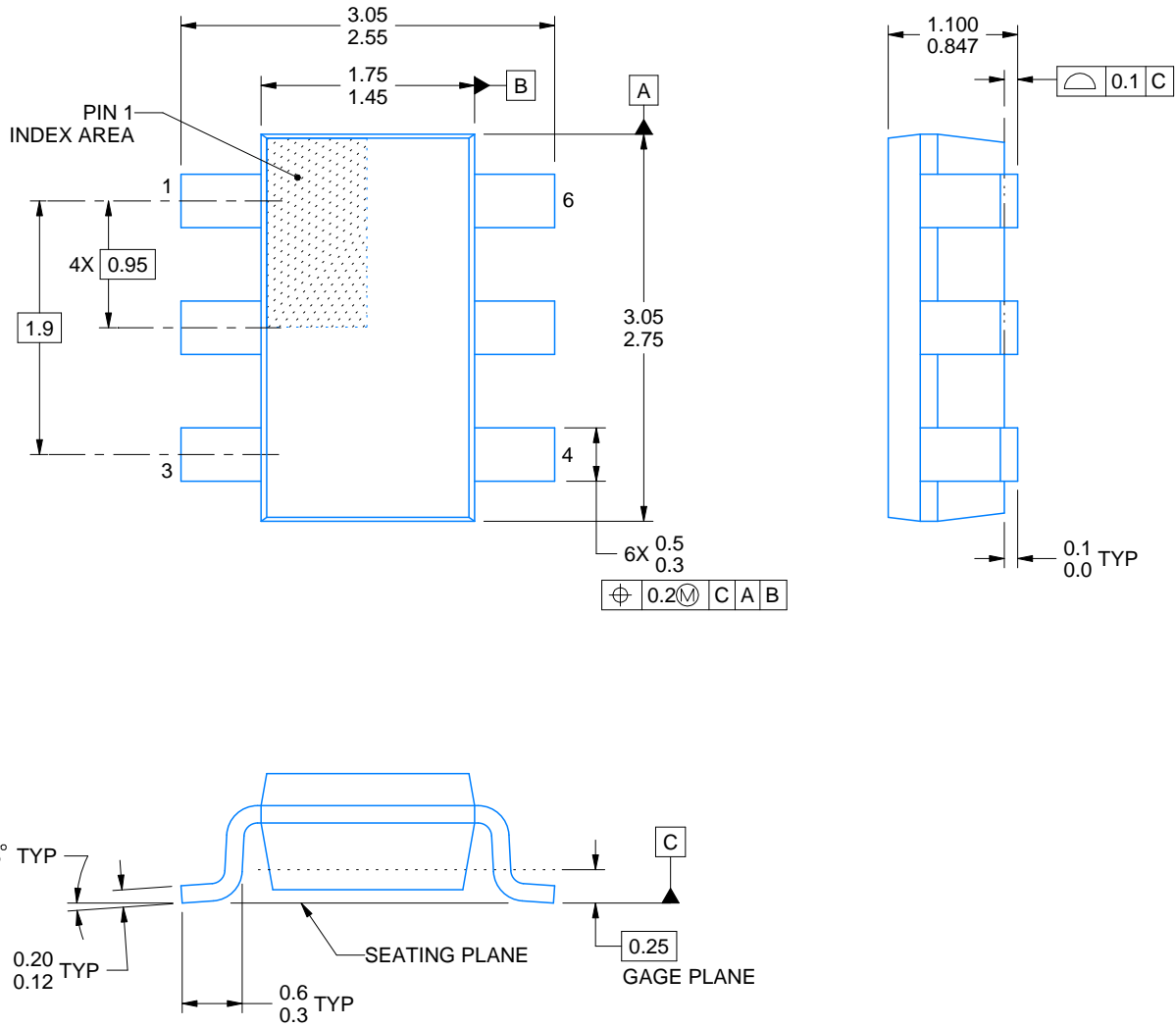*All dimensions are nominal

| 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 |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| TPS54302DDCR | SOT-23-THIN | DDC | 6 | 3000 | 180.0 | 9.5 | 3.17 | 3.1 | 1.1 | 4.0 | 8.0 | Q3 |
| TPS54302DDCR | SOT-23-THIN | DDC | 6 | 3000 | 180.0 | 9.5 | 3.17 | 3.1 | 1.1 | 4.0 | 8.0 | Q3 |
| TPS54302DDCT | SOT-23-THIN | DDC | 6 | 250 | 180.0 | 9.5 | 3.17 | 3.1 | 1.1 | 4.0 | 8.0 | Q3 |
| TPS54302DDCT | SOT-23-THIN | DDC | 6 | 250 | 180.0 | 9.5 | 3.17 | 3.1 | 1.1 | 4.0 | 8.0 | Q3 |

**TAPE AND REEL BOX DIMENSIONS**



*All dimensions are nominal

| Device | Package Type | Package Drawing | Pins | SPQ | Length (mm) | Width (mm) | Height (mm) |
|---|---|---|---|---|---|---|---|
| TPS54302DDCR | SOT-23-THIN | DDC | 6 | 3000 | 184.0 | 184.0 | 19.0 |
| TPS54302DDCR | SOT-23-THIN | DDC | 6 | 3000 | 184.0 | 184.0 | 19.0 |
| TPS54302DDCT | SOT-23-THIN | DDC | 6 | 250 | 184.0 | 184.0 | 19.0 |
| TPS54302DDCT | SOT-23-THIN | DDC | 6 | 250 | 184.0 | 184.0 | 19.0 |

3.05
2.55

1.75
1.45

B

A

PIN 1
INDEX AREA

1

6

4X 0.95

1.9

3.05
2.75

3

4

6X 0.5
0.3

⊕ 0.2Ⓜ C A B

1.100
0.847

⌒ 0.1 C

0.1 TYP
0.0

0°-8° TYP

C

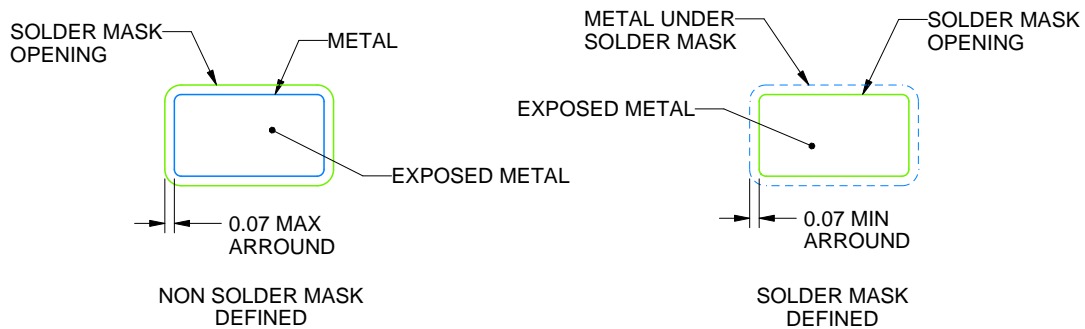0.20 TYP
0.12

SEATING PLANE

0.25

GAGE PLANE

0.6 TYP
0.3

4214841/B  11/2020

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. Reference JEDEC MO-193.

6X (1.1)

SYMM

1

6X (0.6)

6

SYMM

4X (0.95)

3

4

(R0.05) TYP

(2.7)
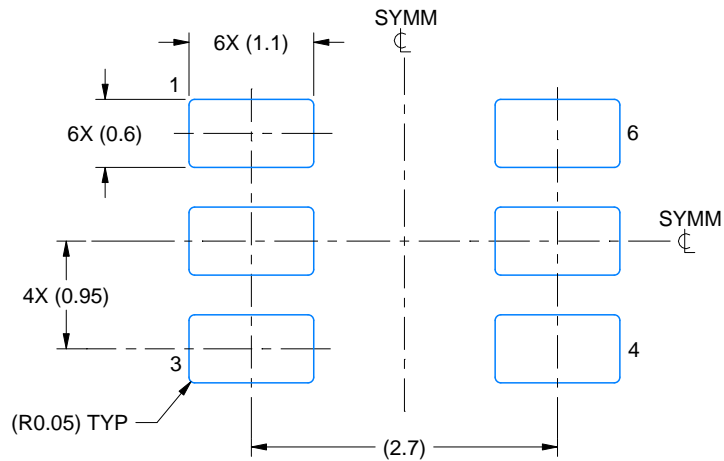
LAND PATTERN EXAMPLE
EXPLOSED METAL SHOWN
SCALE:15X

SOLDER MASK
OPENING

METAL

EXPOSED METAL

0.07 MAX
ARROUND

NON SOLDER MASK
DEFINED

METAL UNDER
SOLDER MASK

SOLDER MASK
OPENING
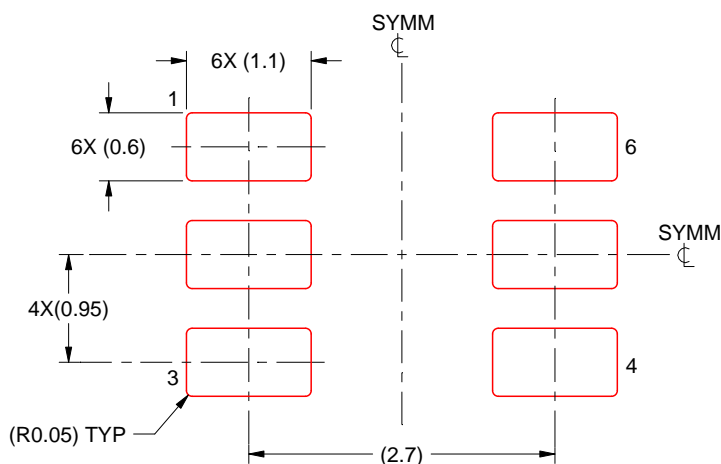
EXPOSED METAL

0.07 MIN
ARROUND

SOLDER MASK
DEFINED

SOLDERMASK DETAILS

4214841/B  11/2020

NOTES: (continued)

4. Publication IPC-7351 may have alternate designs.
5. Solder mask tolerances between and around signal pads can vary based on board fabrication site.

SYMM

6X (1.1)

1

6X (0.6)

6

SYMM

4X(0.95)

3

4

(R0.05) TYP

(2.7)

SOLDER PASTE EXAMPLE
BASED ON 0.125 THICK STENCIL
SCALE:15X

4214841/B   11/2020

NOTES: (continued)

6. Laser cutting apertures with trapezoidal walls and rounded corners may offer better paste release. IPC-7525 may have alternate design recommendations.
7. Board assembly site may have different recommendations for stencil design.

## 重要声明和免责声明