具有 Eco-mode™ 控制的 TPS57140-EP 1.5A 42V 降压直流/直流转换器

1 特性

- 3.5V 至 42V 输入电压范围
- 200mΩ 高侧金属氧化物半导体场效应晶体 (MOSFET)
- 在具有脉冲跳跃的轻负载下实现高效率 Eco-mode ™控制机制
- 116μA 工作静态电流
- 1.5μA 关断电流
- 100kHz 至 2.5MHz 开关频率
- 同步至外部时钟
- 可调缓启动和排序
- 欠压和过压电源正常输出
- 可调的欠压闭锁 (UVLO) 电压和滞后
- 0.8V 内部电压基准
- 支持国防、航天和医疗 应用
  - 受控基线
  - 同一组装和测试场所
  - 同一制造场所
  - 支持军用（-55°C 至 125°C）温度范围
  - 延长的产品生命周期
  - 延长的产品变更通知
  - 产品可追溯性

2 应用范围

- 12V 和 24V 工业及商用低功耗系统
- 汽车售后加装配件：视频、全球卫星定位系统 (GPS)、娱乐

3 说明

TPS57140-EP 器件是一款带有集成型高侧 MOSFET 的 42V，1.5A 降压稳压器。电流模式控制提供了简单的外部补偿和灵活的组件选择。低纹波脉冲跳跃模式可将无负载的稳压输出电源电流减小至 116μA。通过使能引脚可以将关断电源电流减小至 1.5μA。

UVLO 电压在内部设定为 2.5V，但可用使能引脚将之提高。缓启动引脚控制输出电压启动斜升，该引脚还可配置用于排序或跟踪。开漏电源正常信号表示输出处于其标称电压值的 92% 至 109% 之内。
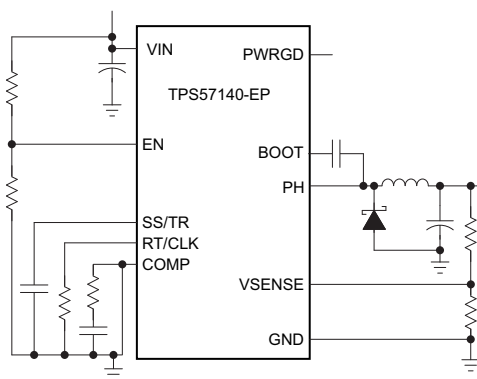
宽开关频率范围可实现效率和外部组件尺寸的优化。频率折返和热关断功能负责在过载情况下保护器件。

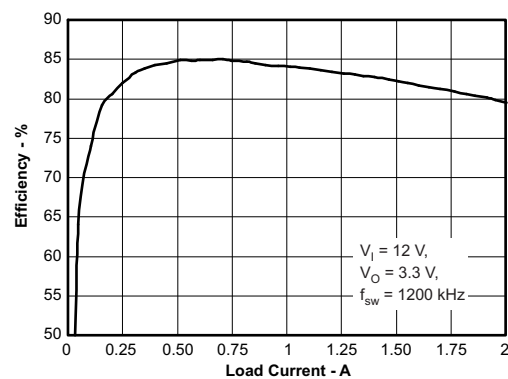TPS57140-EP 采用 10 引脚超薄型小外形尺寸无引线 (VSON) 封装 (DRC)。

器件信息[1]

器件型号	封装	封装尺寸（标称值）
TPS57140-EP	VSON (10)	3.00mm × 3.00mm

(1) 如需了解所有可用封装，请参阅数据表末尾的可订购产品附录。

简化原理图

效率与负载电流间的关系

V_I = 12 V,
V_O = 3.3 V,
f_{sw} = 1200 kHz

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4 修订历史记录

注：之前版本的页码可能与当前版本有所不同。

Changes from Revision A (September 2015) to Revision B   **Page**

• 仅有编辑更改；无技术内容更改 1

Changes from Original (September 2015) to Revision A   **Page**

• Added derating chart 7

5 Pin Configuration and Functions

DRC Package
10-Pin VSON
Top View

```
BOOT   1          10   PH
VIN    2   Exposed  9   GND
EN     3   Thermal  8   COMP
SS/TR  4    Pad     7   VSENSE
RT/CLK 5           6   PWRGD
```
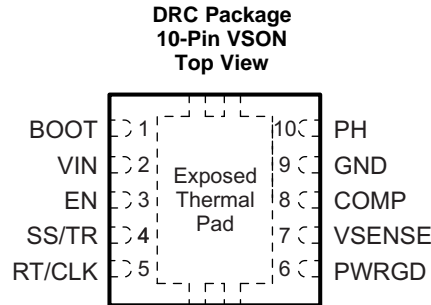
Pin Functions

PIN		I/O	DESCRIPTION
NAME	NO.		
BOOT	1	O	The device requires a bootstrap capacitor between BOOT and PH. A voltage on this capacitor below the minimum required by the output device forces the output to switch off pending a refresh of the capacitor.
COMP	8	O	Error-amplifier output and input to the output-switch current comparator. Connect frequency-compensation components to this pin.
EN	3	I	Enable pin, internal pullup current source. Pull below 1.2 V to disable. Float to enable. Adjust the input UVLO with two resistors.
GND	9	—	Ground
PH	10	I	The source of the internal high-side power MOSFET
PWRGD	6	O	An open-drain output, asserts low if output voltage is low due to thermal shutdown, dropout, overvoltage, or EN shutdown.
RT/CLK	5	I	Resistor timing and external clock. An internal amplifier holds this pin at a fixed voltage when using an external resistor to ground to set the switching frequency. Pulling the pin above the PLL upper threshold causes a mode change whereby the pin becomes a synchronization input. Disabling of the internal amplifier occurs, and the pin is a high-impedance clock input to the internal PLL. If clocking edges stop, re-enabling of the internal amplifier occurs, and the mode returns to a resistor-set function.
SS/TR	4	I	Slow-start and tracking. An external capacitor connected to this pin sets the output rise time. The voltage on this pin overrides the internal reference, allowing use of the pin for tracking and sequencing.
VIN	2	I	Input supply voltage, 3.5 to 42 V
VSENSE	7	I	Inverting node of the transconductance (gm) error amplifier
Thermal pad	—	—	GND pin must have an electrical connection to the exposed pad on the printed-circuit board for proper operation.

6 Specifications

6.1 Absolute Maximum Ratings

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

				MIN	MAX	UNIT
V_{IN}	Input voltage	VIN		−0.3	47	V
		EN[2]		−0.3	5	
		BOOT			55	
		VSENSE		−0.3	3	
		COMP		−0.3	3	
		PWRGD		−0.3	6	
		SS/TR		−0.3	3	
		RT/CLK		−0.3	3.6	
V_{OUT}	Output voltage	PH to BOOT			8	
		PH		−0.6	47	
			200 ns	−1	47	
			30 ns	−2	47	
			Maximum DC voltage, T_J = −40°C	−0.85		
V_{DIFF}	Differential voltage	PAD to GND		−200	200	mV
I_{SOURCE}	Source current	EN			100	µA
		BOOT			100	mA
		VSENSE			10	µA
		PH			Current limit	
		RT/CLK			100	µA
I_{SINK}	Sink current	VIN			Current limit	
		COMP			100	µA
		PWRGD			10	mA
		SS/TR			200	µA
T_J	Operating junction temperature			−55	160	°C
T_{stg}	Storage temperature			−65	150	

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

6.2 ESD Ratings

			VALUE	UNIT
$V_{(ESD)}$	Electrostatic discharge	Human-body model (HBM), per ANSI/ESDA/JEDEC JS-001[1]	±2000	V
		Charged-device model (CDM), per JEDEC specification JESD22-C101[2]	±750	

(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 junction temperature range (unless otherwise noted)

	MIN	NOM	MAX	UNIT
VIN	3.5		42	V

6.4 Thermal Information

THERMAL METRIC[1][2]		TPS57140-EP	UNIT
		DRC (VSON)	
		10 PINS	
$R_{\theta JA}$ Junction-to-ambient thermal resistance	Standard board	56.5	°C/W
	Custom board[3]	61.5	
$R_{\theta JC(top)}$ Junction-to-case (top) thermal resistance		52.1	°C/W
$R_{\theta JB}$ Junction-to-board thermal resistance		20.8	°C/W
ψ_{JT} Junction-to-top characterization parameter		0.9	°C/W
ψ_{JB} Junction-to-board characterization parameter		20.8	°C/W
$R_{\theta JC(bot)}$ Junction-to-case (bottom) thermal resistance		5.2	°C/W

(1) For more information about traditional and new thermal metrics, see the *Semiconductor and IC Package Thermal Metrics* application report, SPRA953.
(2) Power rating at a specific ambient temperature T_A should be determined with a junction temperature of 150°C. This is the point where distortion starts to increase substantially. See *Power-Dissipation Estimate* for more information.
(3) Test-board conditions:
   (a) 3 inches (7.62 cm) × 3 inches (7.62 cm), 2 layers, thickness: 0.062 inch (1.57 mm)
   (b) 2-oz. (0.071-mm thick) copper traces located on the top of the PCB
   (c) 2-oz. (0.071-mm thick) copper ground plane, bottom layer
   (d) 6 thermal vias (13 mil (1 mil = 0.0254 mm)) located under the device package

6.5 Electrical Characteristics

T_J = –55°C to 150°C, VIN = 3.5 to 42 V (unless otherwise noted)

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
SUPPLY VOLTAGE (VIN PIN)					
Operating input voltage		3.5		42	V
Internal UVLO threshold	No voltage hysteresis, rising and falling		2.5		V
Shutdown supply current	EN = 0 V, 25°C, 3.5 V ≤ V_{IN} ≤ 60 V		1.5	4	μA
	EN = 0 V, 125°C, 3.5 V ≤ V_{IN} ≤ 60 V		1.9	6.5	
Operating: nonswitching supply current	V_{SENSE} = 0.83 V, V_{IN} = 12 V, 25°C		116	140	
ENABLE AND UVLO (EN PIN)					
Enable threshold voltage	No voltage hysteresis, rising and falling, 25°C	1.15	1.25	1.36	V
Input current	Enable threshold 50 mV		−3.8		μA
	Enable threshold −50 mV		−0.9		
Hysteresis current			−2.9		μA
VOLTAGE REFERENCE					
Voltage reference	T_J = 25°C	0.790	0.8	0.808	V
		0.780	0.8	0.819	
HIGH-SIDE MOSFET					
On-resistance	V_{IN} = 3.5 V, BOOT-PH = 3 V		300		mΩ
	V_{IN} = 12 V, BOOT-PH = 6 V		200	410	
ERROR AMPLIFIER					
Input current			50		nA
Error amplifier transconductance (gm)	−2 μA < I_{COMP} < 2 μA, V_{COMP} = 1 V		97		μS
Error amplifier transconductance (gm) during slow start	−2 μA < I_{COMP} < 2 μA, V_{COMP} = 1 V, V_{VSENSE} = 0.4 V		26		μS
Error amplifier dc gain	V_{VSENSE} = 0.8 V		10 000		V/V
Error amplifier bandwidth			2700		kHz
Error amplifier source/sink	$V_{(COMP)}$ = 1 V, 100-mV overdrive		±7		μA
COMP to switch current transconductance			6		S

Electrical Characteristics (continued)

T_J = –55°C to 150°C, VIN = 3.5 to 42 V (unless otherwise noted)

	PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
CURRENT LIMIT						
	Current-limit threshold	V_{IN} = 12 V, T_J = 25°C	1.8	2.7		A
THERMAL SHUTDOWN						
	Thermal shutdown			182		°C
TIMING RESISTOR AND EXTERNAL CLOCK (RT/CLK PIN)						
	Switching-frequency range using RT mode	V_{IN} = 12 V	100		2500	kHz
f_{SW}	Switching frequency	V_{IN} = 12 V, R_T = 200 kΩ	450	581	720	kHz
	Switching-frequency range using CLK mode	V_{IN} = 12 V	300		2200	kHz
	Minimum CLK pulse duration			40		ns
	RT/CLK high threshold	V_{IN} = 12 V		1.9	2.2	V
	RT/CLK low threshold	V_{IN} = 12 V	0.45	0.7		V
	RT/CLK falling-edge to PH rising-edge delay	Measured at 500 kHz with RT resistor in series		60		ns
	PLL lock in time	Measured at 500 kHz		100		µs
SLOW START AND TRACKING (SS/TR)						
	Charge current	$V_{SS/TR}$ = 0.4 V		2		µA
	SS/TR-to-VSENSE matching	$V_{SS/TR}$ = 0.4 V		45		mV
	SS/TR-to-reference crossover	98% nominal		1		V
	SS/TR discharge current (overload)	V_{SENSE} = 0 V, $V_{(SS/TR)}$ = 0.4 V		112		µA
	SS/TR discharge voltage	V_{SENSE} = 0 V		54		mV
POWER GOOD (PWRGD PIN)						
V_{VSENSE}	VSENSE threshold	V_{SENSE} falling		92%		
		V_{SENSE} rising		94%		
		V_{SENSE} rising		109%		
		V_{SENSE} falling		107%		
	Hysteresis	V_{SENSE} falling		2%		
	Output-high leakage	V_{SENSE} = V_{REF}, $V_{(PWRGD)}$ = 5.5 V, 25°C		10		nA
	On-resistance	$I_{(PWRGD)}$ = 3 mA, V_{SENSE} < 0.79 V		50		Ω
	Minimum VIN for defined output	$V_{(PWRGD)}$ < 0.5 V, $I_{I(PWRGD)}$ = 100 µA		0.95	1.5	V
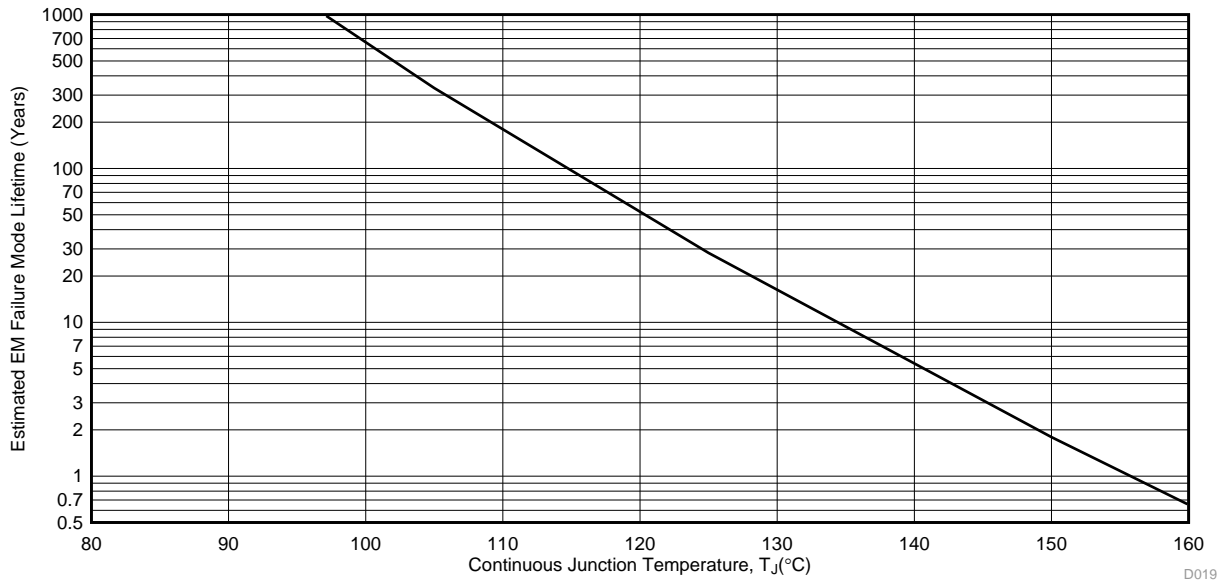
(1) Electromigration fail mode = Time at temperature with bias

(2) Silicon operating life design goal is 10 years at 105°C junction temperature (does not include package interconnect life).

(3) The predicted operating lifetime versus junction temperature is based on reliability modeling and available qualification data.

Figure 1. Predicted Lifetime Derating Chart for TPS57140-EP
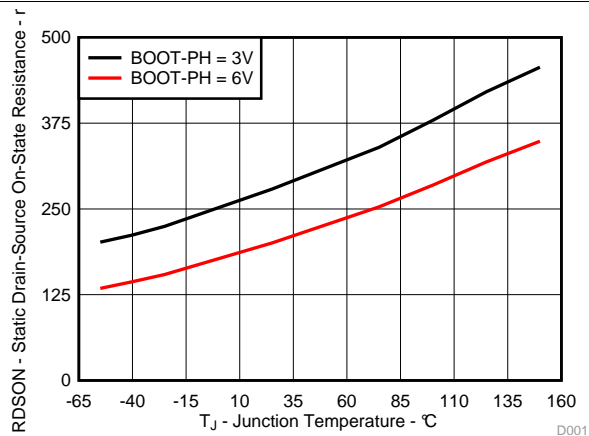
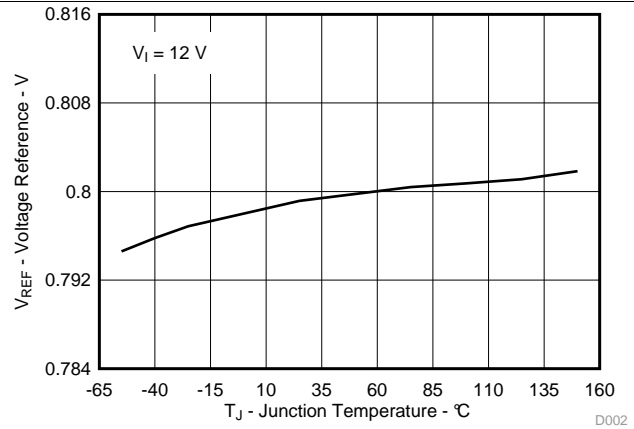## 6.6 Typical Characteristics

Figure 2. On Resistance vs Junction Temperature

Figure 3. Voltage Reference vs Junction Temperature
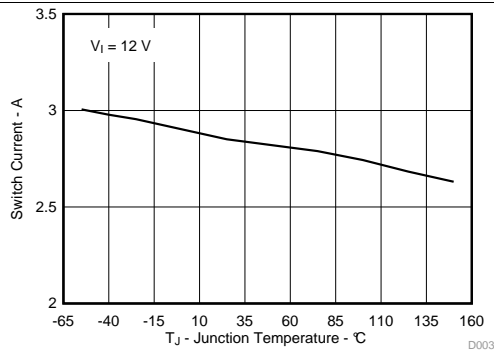
Figure 4. Switch Current Limit vs Junction Temperature
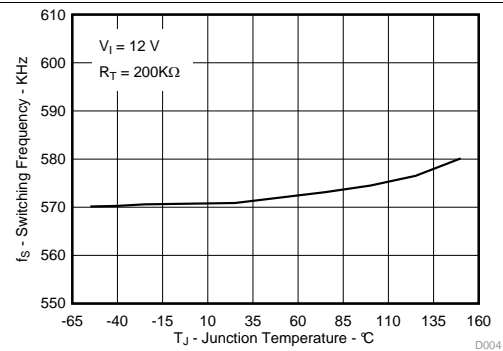
Figure 5. Switching Frequency vs Junction Temperature
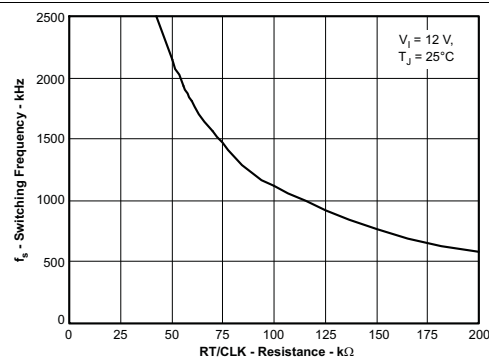
Figure 6. Switching Frequency vs RT/CLK Resistance, High-Frequency Range
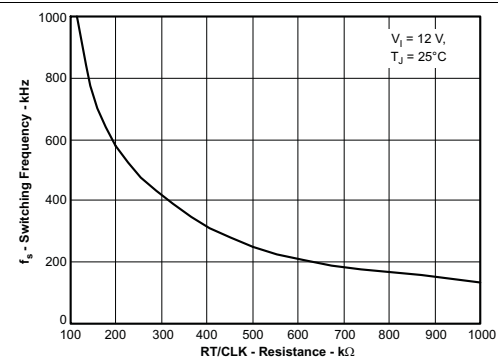
Figure 7. Switching Frequency vs RT/CLK Resistance, Low-Frequency Range

Typical Characteristics (continued)
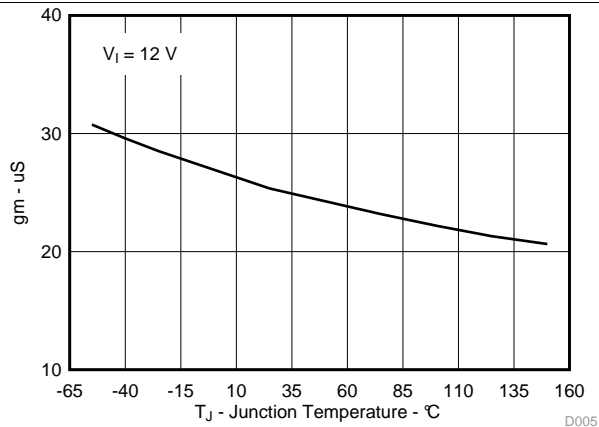
Figure 8. EA Transconductance During Slow Start vs Junction Temperature
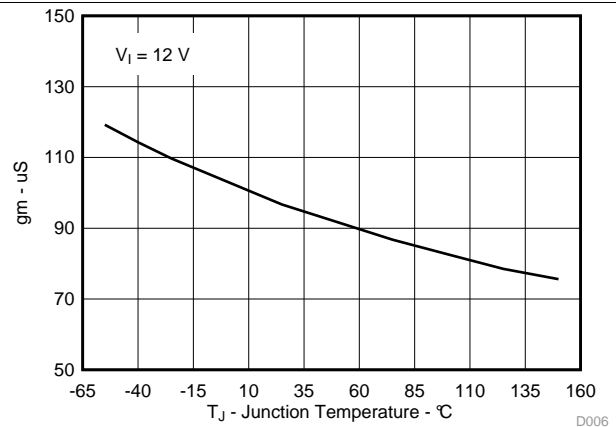
Figure 9. EA Transconductance vs Junction Temperature
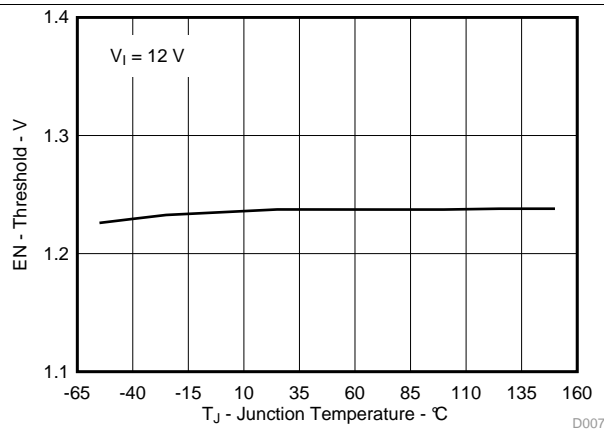
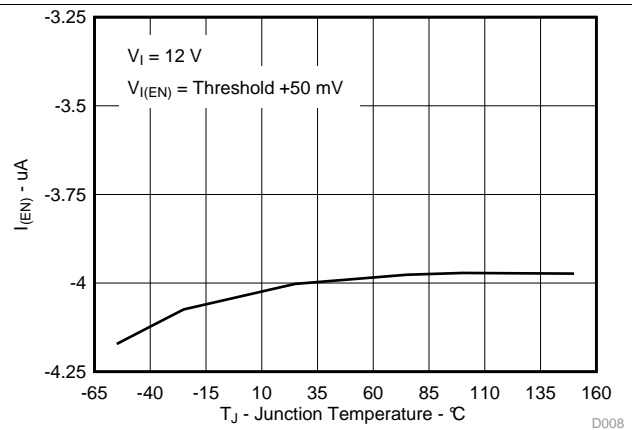**Figure 10. EN Pin Voltage vs Junction Temperature**
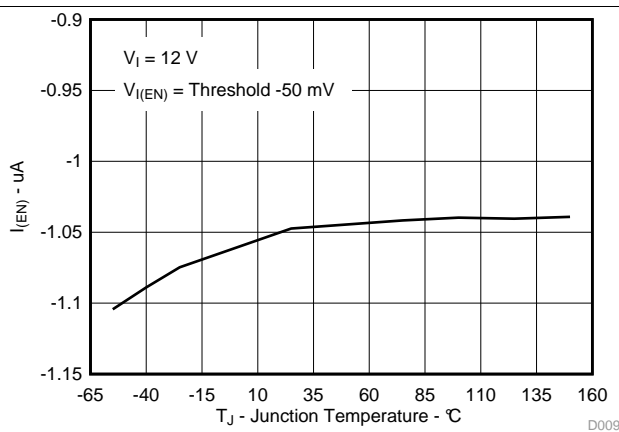
Figure 11. EN Pin Current vs Junction Temperature
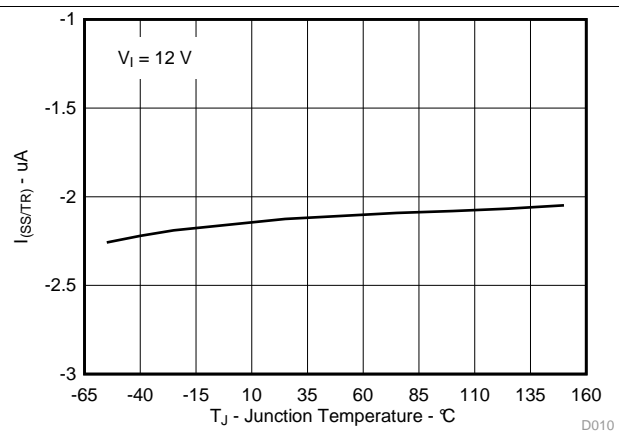
Figure 12. EN Pin Current vs Junction Temperature

Figure 13. SS/TR Charge Current vs Junction Temperature
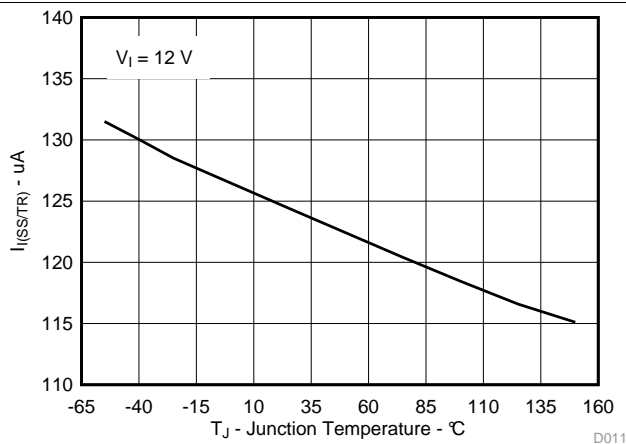
Typical Characteristics (continued)

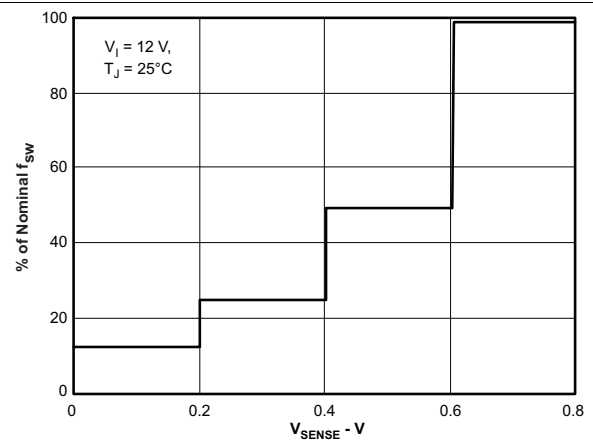**Figure 14. SS/TR Discharge Current vs Junction Temperature**
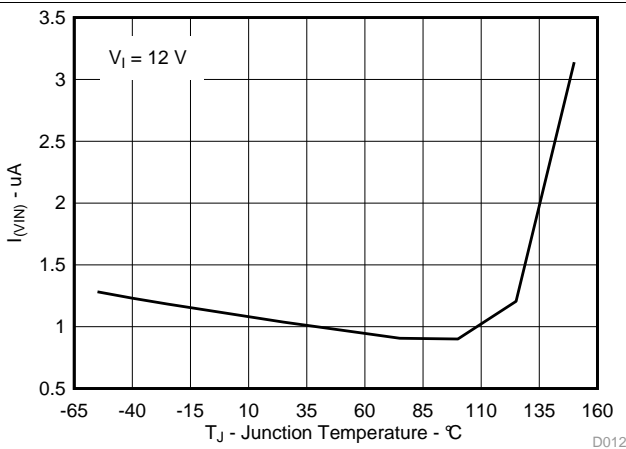
Figure 15. Switching Frequency vs V$_{SENSE}$

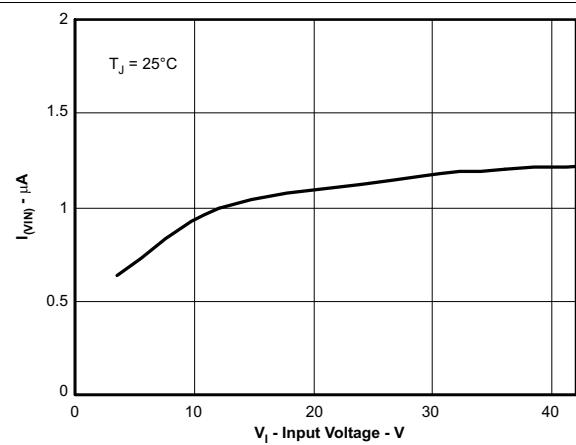**Figure 16. Shutdown Supply Current vs Junction Temperature**

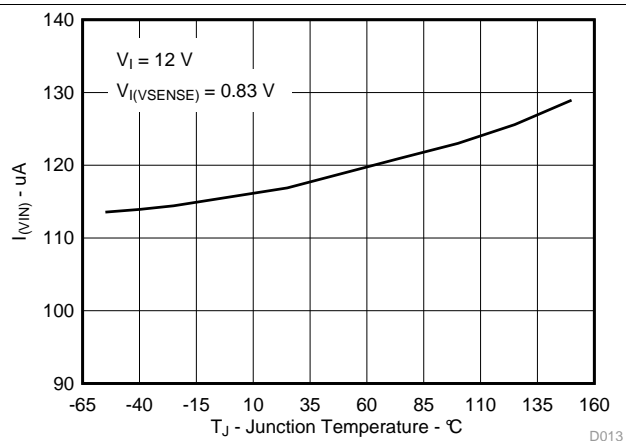**Figure 17. Shutdown Supply Current vs Input Voltage (V$_{IN}$)**

Figure 18. V$_{IN}$ Supply Current vs Junction Temperature
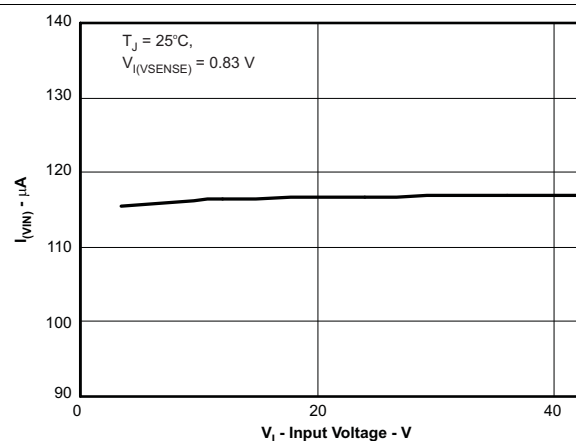
Figure 19. V$_{IN}$ Supply Current vs Input Voltage

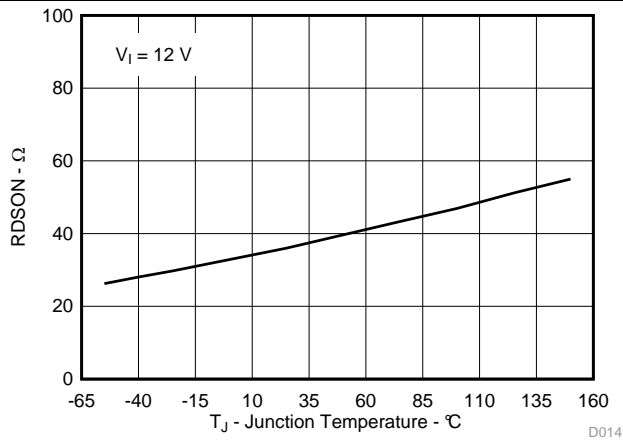## Typical Characteristics (continued)

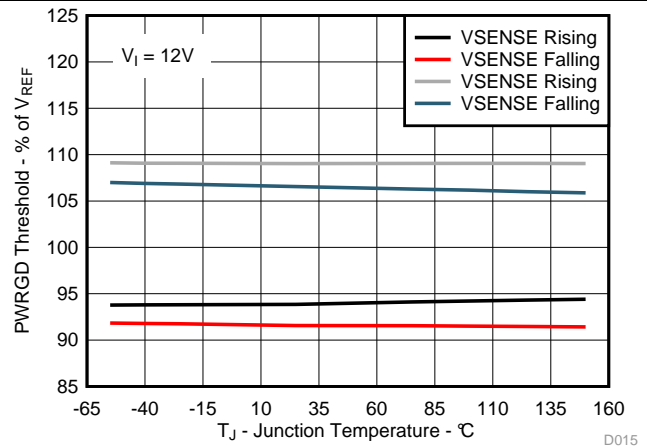**Figure 20. PWRGD On Resistance vs Junction Temperature**

Figure 21. PWRGD Threshold vs Junction Temperature
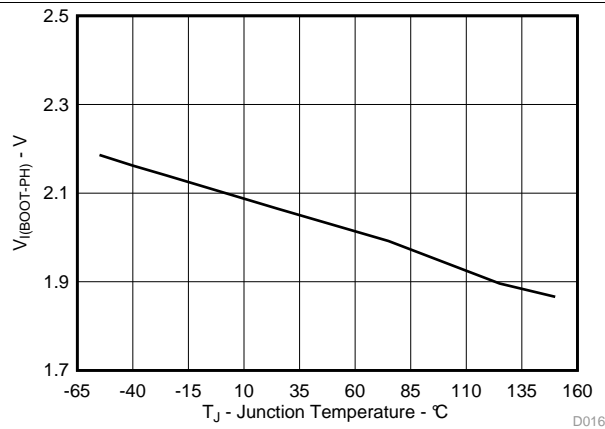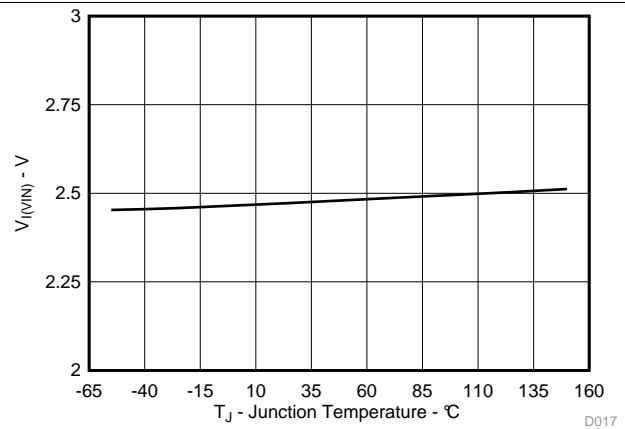

Figure 22. BOOT-PH UVLO vs Junction Temperature
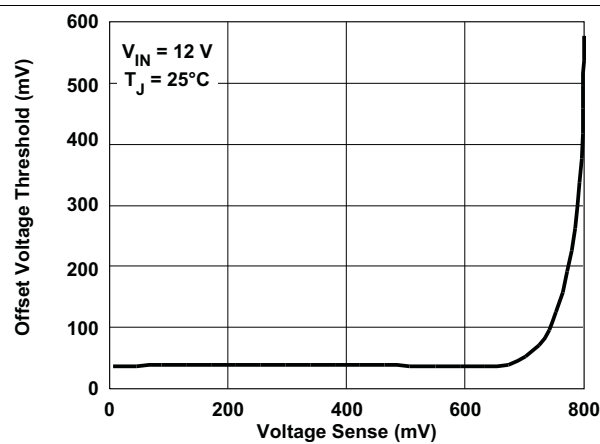

Figure 23. Input Voltage (UVLO) vs Junction Temperature

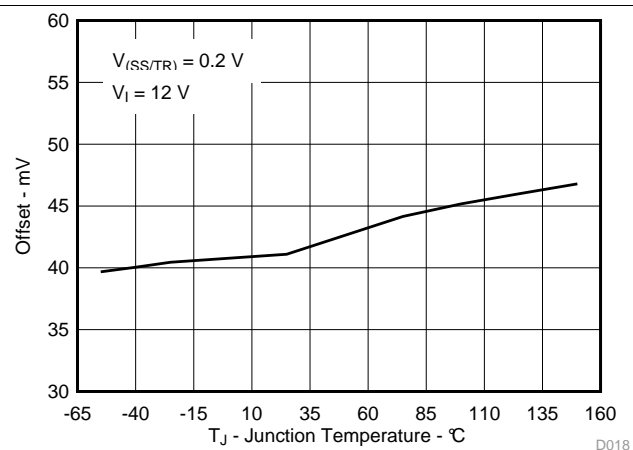**Figure 24. SS/TR to V$_{SENSE}$ Offset vs V$_{SENSE}$**

Figure 25. SS/TR to V$_{SENSE}$ Offset vs Temperature

7 Detailed Description

7.1 Overview

The TPS57140-EP device is a 42-V, 1.5-A, step-down (buck) regulator with an integrated high-side n-channel MOSFET. To improve performance during line and load transients, the device implements a constant-frequency, current-mode control which reduces output capacitance and simplifies external frequency-compensation design. The wide switching frequency of 100 to 2500 kHz allows for efficiency and size optimization when selecting the output-filter components. Using a resistor to ground on the RT/CLK pin adjusts the switching frequency. The device has an internal phase-locked loop (PLL) on the RT/CLK pin that synchronizes the power-switch turnon to a falling edge of an external system clock.

The TPS57140-EP has a default start-up voltage of approximately 2.5 V. The EN pin has an internal pullup current source that the designer can use to adjust the input-voltage UVLO threshold with two external resistors. In addition, the pullup current provides a default condition. When the EN pin is floating, the device operates. The operating current is 116 μA when not switching and under no load. With the device disabled, the supply current is 1.5 μA.

The integrated 200-mΩ high-side MOSFET allows for high-efficiency power-supply designs capable of delivering 1.5 A of continuous current to a load. The TPS57140-EP reduces the external component count by integrating the boot recharge diode. A capacitor between the BOOT and PH pins supplies the bias voltage for the integrated high-side MOSFET. A UVLO circuit monitors the boot-capacitor voltage and turns the high-side MOSFET off when the boot voltage falls below a preset threshold. The TPS57140-EP can operate at high duty cycles because of the boot UVLO. Stepping down of the output voltage can extend as low as the 0.8-V reference.

The TPS57140-EP has a power-good comparator (PWRGD) which asserts when the regulated output voltage is <92% or >109% of the nominal output voltage. The PWRGD pin is an open-drain output which deasserts when the VSENSE pin voltage is between 94% and 107% of the nominal output voltage, allowing the pin to transition high when a pullup resistor is used.

The TPS57140-EP minimizes excessive output overvoltage (OV) transients by taking advantage of the OV power-good comparator. With the OV comparator activated, the high-side MOSFET turns off and remains masked from turning on until the output voltage is lower than 107%.

Use the SS/TR (slow-start/tracking) pin to minimize inrush currents or provide power-supply sequencing during power up. Couple a small-value capacitor to the pin to adjust the slow-start time. The designer can couple a resistor divider to the pin for critical power-supply sequencing requirements. Discharge of the SS/TR pin occurs before the output powers up. This discharging ensures a repeatable restart after an overtemperature fault, UVLO fault, or a disabled condition.

The TPS57140-EP also discharges the slow-start capacitor during overload conditions with an overload-recovery circuit. The overload-recovery circuit slow-starts the output from the fault voltage to the nominal regulation voltage on removal of a fault condition. A frequency-foldback circuit reduces the switching frequency during start-up and overcurrent fault conditions to help control the inductor current.

7.2 Functional Block Diagram
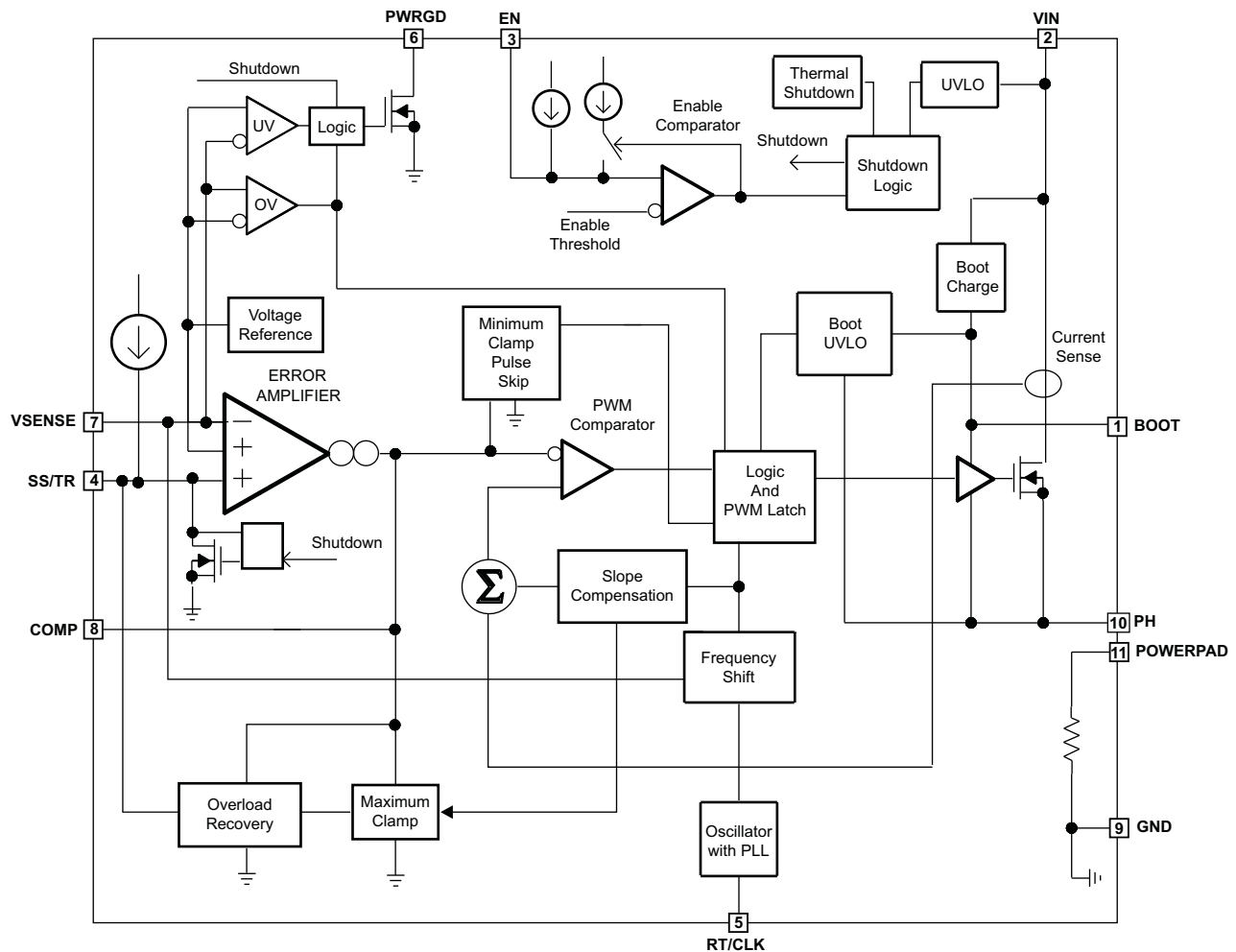
7.3 Feature Description

7.3.1 Fixed Frequency PWM Control

The TPS57140-EP uses an adjustable fixed-frequency, peak-current mode control. External resistors on the VSENSE pin compare the output voltage to an internal voltage reference by an error amplifier which drives the COMP pin. An internal oscillator initiates the turnon of the high-side power switch. The device compares the error-amplifier output to the high-side power-switch current. When the power-switch current reaches the COMP voltage level, the power switch turns off. The COMP pin voltage increases and decreases as the output current increases and decreases. The device implements a current limit by clamping the COMP pin voltage to a maximum level. A minimum clamp on the COMP pin implements the Eco-mode control scheme.

7.3.2 Slope-Compensation Output Current

The TPS57140-EP adds a compensating ramp to the switch-current signal. This slope compensation prevents subharmonic oscillations. The available peak inductor current remains constant over the full duty-cycle range.

Feature Description (continued)

7.3.3 Bootstrap Voltage (Boot)

The TPS57140-EP has an integrated boot regulator and requires a small ceramic capacitor between the BOOT and PH pins to provide the gate-drive voltage for the high-side MOSFET. The value of the ceramic capacitor should be 0.1 μF. TI recommends a ceramic capacitor with an X7R or X5R grade dielectric because of the stable characteristics over temperature and voltage. To improve dropout, the TPS57140-EP operates at 100% duty cycle as long as the BOOT-to-PH pin voltage is greater than 2.1 V. When the voltage from BOOT to PH drops below 2.1 V, the high-side MOSFET turns off using a UVLO circuit, allowing for the low-side diode to conduct, which allows refreshing of the BOOT capacitor. Because the supply current sourced from the BOOT capacitor is low, the high-side MOSFET can remain on for more switching cycles than it refreshes; thus, the effective duty-cycle limitation attributed to the boot regulator system is high.

7.3.4 Low-Dropout Operation

The voltage drops across the power MOSFET, inductor, low-side diode, and PCB resistance mainly determine the duty cycle during dropout of the regulator. During operating conditions in which the input voltage drops, the high-side MOSFET can remain on for 100% of the duty cycle to maintain output regulation until the BOOT-to-PH voltage falls below 2.1 V.

After the high side is off, the low-side diode conducts and the BOOT capacitor recharges. During this boot-capacitor recharge time, the inductor current ramps down until the high-side MOSFET turns on. The recharge time is longer than the typical high-side off-time of previous switching cycles, and thus the inductor-current ripple is larger, resulting in more ripple voltage on the output. The recharge time is a function of the input voltage, boot-capacitor value, and the impedance of the internal boot-recharge diode.

Pay attention in maximum-duty-cycle applications which experience extended time periods without a load current. When the voltage across the BOOT capacitors falls below the 2.1-V threshold in applications that have a difference in the input voltage and output voltage that is <3 V, the high-side MOSFET turns off, but there is not enough current in the inductor to pull the PH pin down to recharge the boot capacitor. The regulator does not switch because the boot capacitor is less than 2.1 V, and the output capacitor decays until the difference between the input voltage and output voltage is 2.1 V. At this time, the boot UVLO is exceeded and the device switches until reaching the desired output voltage.

Figure 26 and Figure 27 show the start and stop voltages for 3.3-V and 5-V applications. The graphs plot voltages versus the load current. The definition of start voltage is the input voltage needed to regulate within 1%. The definition of stop voltage is the input voltage at which the output drops by 5% or stops switching.
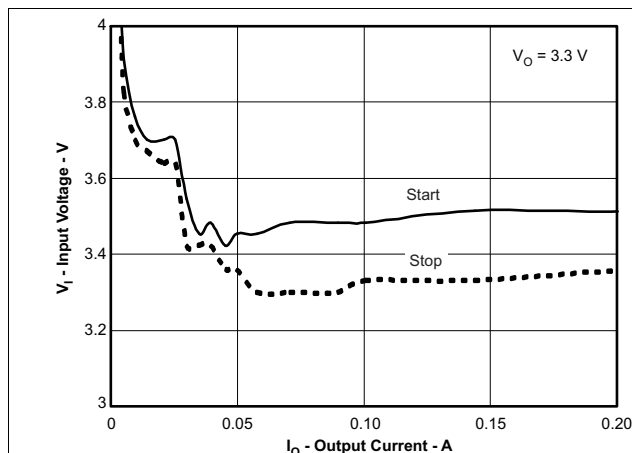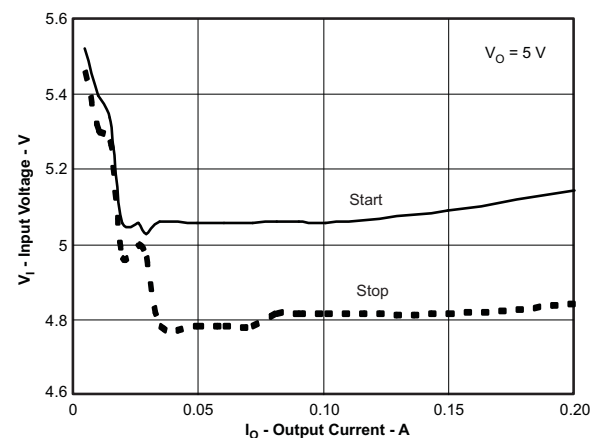

Figure 26. 3.3-V Start and Stop Voltage

Figure 27. 5-V Start and Stop Voltage

Feature Description (continued)

7.3.5 Error Amplifier

The TPS57140-EP has a transconductance amplifier for the error amplifier. The error amplifier compares the VSENSE voltage to the lower of the SS/TR pin voltage or the internal 0.8-V voltage reference. The transconductance (gm) of the error amplifier is 97 μS during normal operation. During the slow-start operation, the transconductance is a fraction of the normal operating gm. When the voltage of the VSENSE pin is below 0.8 V and the device is regulating using the SS/TR voltage, the gm is 25 μS.

The frequency-compensation components (capacitor, series resistor, and capacitor) are added from the COMP pin to ground.

7.3.6 Voltage Reference

The voltage-reference system produces a precise ±2% voltage reference over temperature by scaling the output of a temperature-stable band-gap circuit.

7.3.7 Adjusting the Output Voltage

A resistor divider from the output node to the VSENSE pin sets the output voltage. TI recommends to use 1% tolerance or better divider resistors. Start with 10 kΩ for the R2 resistor and use Equation 1 to calculate R1. To improve efficiency at very-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 VSENSE input current become noticeable.

$$R1 = R2 \times \left(\frac{Vout - 0.8V}{0.8\ V} \right)$$

(1)

7.3.8 Enable and Adjusting UVLO

The VIN pin voltage falling below 2.5 V disables the TPS57140-EP. If an application requires a higher UVLO, use the EN pin as shown in Figure 28 to adjust the input-voltage UVLO by using two external resistors. Though it is not necessary to use the UVLO adjust resistors, for operation, TI highly recommends providing consistent power-up behavior. The EN pin has an internal pullup current source, I1, of 0.9 μA that provides the default condition of the TPS57140-EP operating when the EN pin floats. When the EN pin voltage exceeds 1.25 V, an additional 2.9 μA of hysteresis, Ihys, is added. This additional current facilitates input-voltage hysteresis. Use Equation 2 to set the external hysteresis for the input voltage. Use Equation 3 to set the input start voltage.
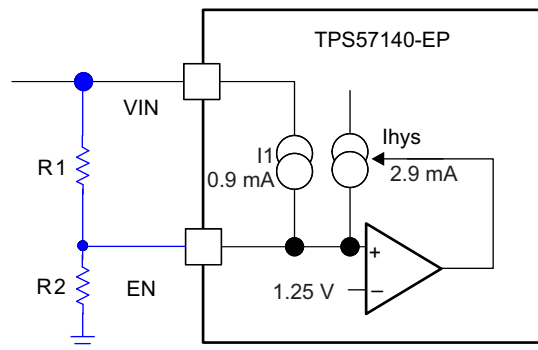

Figure 28. Adjustable UVLO

$$R1 = \frac{V_{START} - V_{STOP}}{I_{HYS}}$$

(2)

$$R2 = \frac{V_{ENA}}{\dfrac{V_{START} - V_{ENA}}{R1} + I_1}$$

(3)

Feature Description (continued)

Figure 29 shows another technique to add input-voltage hysteresis. The designer can use this method if the resistance values are high from the previous method and there is a need for a wider voltage hysteresis. Resistor R3 sources additional hysteresis current into the EN pin.
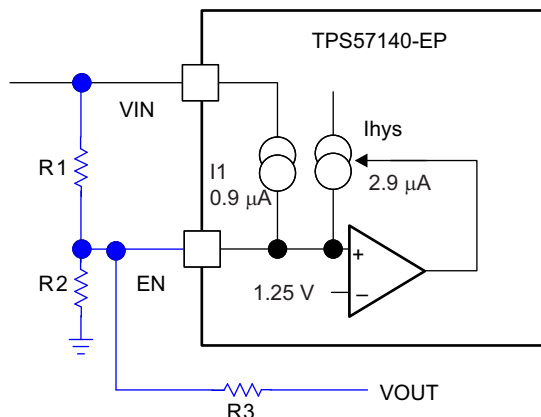
Figure 29. Adding Additional Hysteresis

$$R1 = \frac{V_{START} - V_{STOP}}{I_{HYS} + \dfrac{V_{OUT}}{R3}}$$

(4)

$$R2 = \frac{V_{ENA}}{\dfrac{V_{START} - V_{ENA}}{R1} + I_1 - \dfrac{V_{ENA}}{R3}}$$

(5)

Do not place a low-impedance voltage source with >5 V directly on the EN pin. Do not place a capacitor directly on the EN pin if V_{EN} > 5 V when using a voltage divider to adjust the start and stop voltage. The node voltage (see Figure 30) must remain ≤5.8 V. The Zener diode can sink up to 100 μA. The EN pin voltage can be >5 V if the V_{IN} voltage source has a high impedance and does not source more than 100 μA into the EN pin.

Figure 30. Node Voltage

Feature Description (continued)

7.3.9 Slow-Start or Tracking Pin (SS/TR)

The TPS57140-EP effectively uses the lower voltage of the internal voltage reference or the SS/TR pin voltage as the power-supply reference voltage and regulates the output accordingly. A capacitor on the SS/TR pin to ground impleme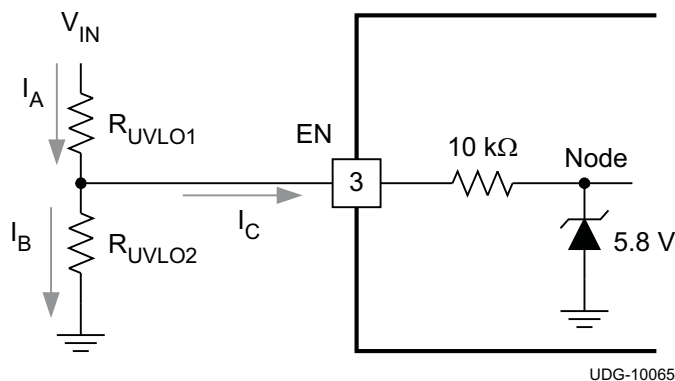nts a slow-start time. The TPS57140-EP has an internal pullup current source of 2 μA that charges the external slow-start capacitor. The calculations for the slow-start (Equation 6) show the time (10% to 90%). The voltage reference (V_{REF}) is 0.8 V and the slow-start current (I_{SS}) is 2 μA. The slow-start capacitor should remain lower than 0.47 μF and greater than 0.47 nF.

$$Css(nF) = \frac{Tss(ms) \times Iss(\mu A)}{Vref\ (V) \times 0.8}$$

(6)

At power up, the TPS57140-EP does not start switching until the slow-start pin discharges to <40 mV; to ensure proper power up, see Figure 31.

During normal operation, the TPS57140-EP stops switching and the SS/TR must discharge to 40 mV when the VIN UVLO is exceeded, EN pin is pulled below 1.2 V, or a thermal shutdown event occurs.

The VSENSE voltage follows the SS/TR pin voltage with a 45-mV offset up to 85% of the internal voltage reference. When the SS/TR voltage is >85% of the internal reference voltage, the offset increases as the effective system reference transitions from the SS/TR voltage to the internal voltage reference (see Figure 24). The SS/TR voltage ramps linearly until clamped at 1.7 V.
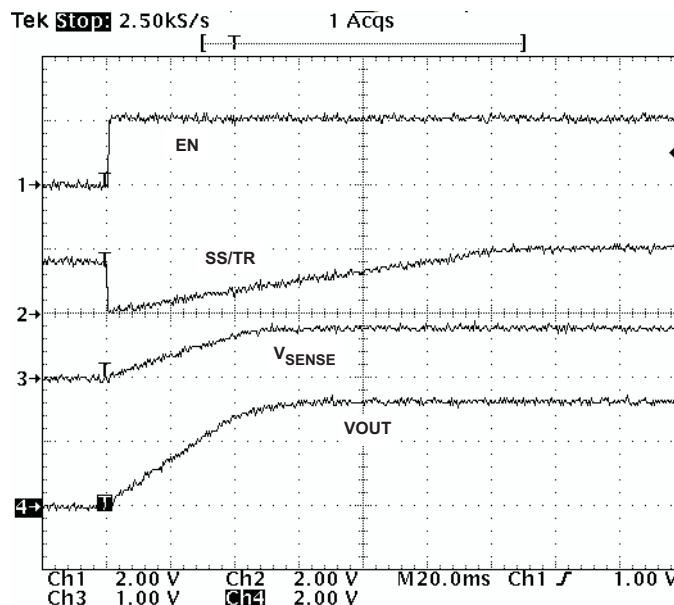
Figure 31. Operation of SS/TR Pin When Starting

7.3.10 Overload Recovery Circuit

The TPS57140-EP has an overload recovery (OLR) circuit. The OLR circuit slow-starts the output from the overload voltage to the nominal regulation voltage on removal of the fault condition. The OLR circuit discharges the SS/TR pin to a voltage slightly greater than the VSENSE pin voltage using an internal pulldown of 100 μA when the error amplifier is changed to a high voltage from a fault condition. On removal of the fault condition, the output slow-starts from the fault voltage to the nominal output voltage.

Feature Description (continued)

7.3.11 Constant Switching Frequency and Timing Resistor (RT/CLK Pin)

The switching frequency of the TPS57140-EP is adjustable over a wide range from approximately 100-kHz to 2500-kHz by placing a resistor on the RT/CLK pin. The RT/CLK pin voltage is typically 0.5 V and must have a resistor to ground to set the switching frequency. To determine the timing resistance for a given switching frequency, use Equation 7 or the curves in Figure 32 or Figure 33. To reduce the solution size, the designer would typically set the switching frequency as high as possible, but consider tradeoffs of the supply efficiency, maximum input voltage, and minimum controllable on-time.

The minimum controllable on time is typically 130 ns, which limits the maximum operating input voltage.

The frequency-shift circuit also limits the maximum switching frequency. The following contains more discussion on the details of the maximum switching frequency.

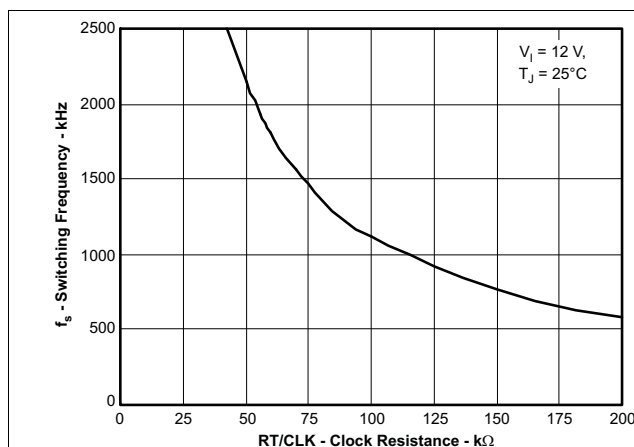$$RT \text{ (k}\Omega) = \frac{206033}{f_{SW} \text{ (kHz)}^{1.0888}}$$
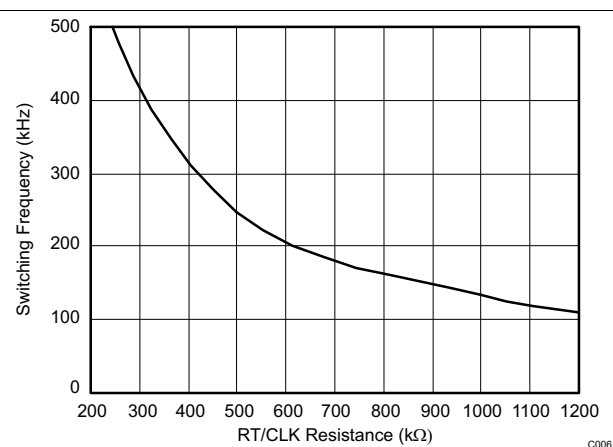
(7)

Figure 32. High-Range RT

Figure 33. Low-Range RT

7.3.12 Overcurrent Protection and Frequency Shift

The TPS57140-EP implements current-mode control, which uses the COMP pin voltage to turn off the high-side MOSFET on a cycle-by-cycle basis. During each cycle, the device compares the switch current and COMP pin voltage. When the peak switch current intersects the COMP voltage, the high-side switch turns off. During overcurrent conditions that pull the output voltage low, the error amplifier responds by driving the COMP pin high, increasing the switch current. The internal clamping of the error amplifier output functions as a switch current limit.

To increase the maximum operating switching frequency at high input voltages, the TPS57140-EP implements a frequency shift. The divisor of the switching frequency goes to 8, 4, 2, and 1 as the voltage ramps from 0 to 0.8 V on the VSENSE pin.

The device implements a digital frequency shift to enable synchronizing to an external clock during normal start-up and fault conditions. Because the device can only divide the switching frequency by 8, there is a maximum input voltage limit in which the device operates and still has frequency-shift protection.

During short-circuit events (particularly with high-input-voltage applications), the control loop has a finite minimum controllable on-time and the output has a very-low voltage. During the switch on-time, the inductor current ramps to the peak current limit because of the high input voltage and minimum on-time. During the switch off time, the inductor would normally not have enough off-time and output voltage for the inductor to ramp down by the ramp-up amount. The frequency shift effectively increases the off-time, allowing the current to ramp down.

Feature Description (continued)

7.3.13 Selecting the Switching Frequency

The selected switching frequency should be the lower value of the two equations, Equation 8 and Equation 9. Equation 8 is the maximum switching frequency limitation set by the minimum controllable on-time. Setting the switching frequency above this value causes the regulator to skip switching pulses.

Equation 9 is the maximum switching-frequency limit set by the frequency-shift protection. To have adequate output short-circuit protection at high input voltages, set the switching frequency to be less than the $f_{SW(maxshift)}$ frequency. In Equation 9, to calculate the maximum switching frequency, take into account that the output voltage decreases from the nominal voltage to 0 V and that the f_{DIV} integer increases from 1 to 8, corresponding to the frequency shift.

In Figure 34, the solid line illustrates a typical safe operating area regarding frequency shift and assumes the output voltage is 0 V, the resistance of the inductor is 0.1 Ω, the FET on-resistance is 0.2 Ω, and the voltage drop of the diode is 0.5 V. The dashed line is the maximum switching frequency to avoid pulse skipping. Enter these equations in a spreadsheet or other software, or use the SwitcherPro design software to determine the switching frequency.

$$f_{SW(max\,skip)} = \frac{1}{t_{ON}} \times \left(\frac{I_L \times R_{dc} + V_{OUT} + V_d}{V_{IN} - I_L \times R_{DS(on)} + V_d} \right)$$

(8)

$$f_{SWshift} = \frac{f_{DIV}}{t_{ON}} \times \left(\frac{I_L \times R_{dc} + V_{OUT(sc)} + V_d}{V_{IN} - I_L \times R_{DS(on)} + V_d} \right)$$

where

- I_L = Inductor current
- Rdc = Inductor resistance
- V_{IN} = Maximum input voltage
- V_{OUT} = Output voltage
- V_{OUTSC} = Output voltage during short
- Vd = Diode voltage drop
- $r_{DS(on)}$ = Switch on-resistance
- t_{ON} = Controllable on-time
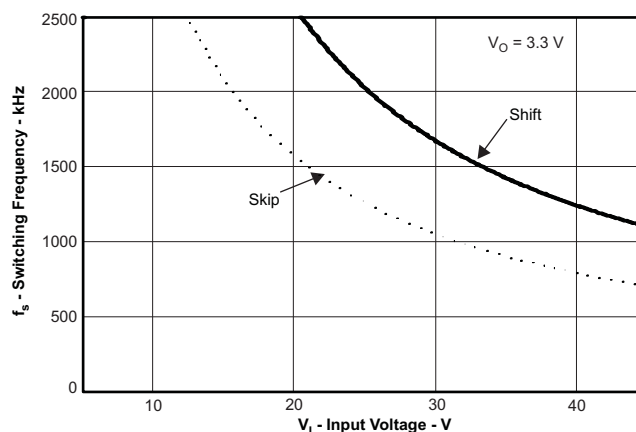- f_{DIV} = Frequency divide (equals 1, 2, 4, or 8)

(9)

Figure 34. Maximum Switching Frequency vs Input Voltage

Feature Description (continued)

7.3.14 How to Interface to RT/CLK Pin

The designer can use the RT/CLK pin to synchronize the regulator to an external system clock. To implement the synchronization feature, connect a square wave to the RT/CLK pin through the circuit network shown in Figure 35. The square wave amplitude must transition lower than 0.5 V and higher than 2.2 V on the RT/CLK pin and have an on-time greater than 40 ns and an off-time greater than 40 ns. The synchronization frequency range is 300 to 2200 kHz. The rising edge of PH synchronizes to the falling edge of the signal on the RT/CLK pin. Design the external synchronization circuit in such a way that the device has the default frequency-set resistor connected from the RT/CLK pin to ground should the synchronization signal turn off. TI recommends using a frequency-set resistor connected as shown in Figure 35 through a 50-Ω resistor to ground. The resistor should set the switching frequency close to the external CLK frequency. TI recommends ac-coupling the synchronization signal through a 10-pF ceramic capacitor and a 4-kΩ series resistor to the RT/CLK pin. The series resistor reduces PH jitter in heavy-load applications when synchronizing to an external clock, and in applications which transition from synchronizing to RT mode. The first time CLK rises above the CLK threshold, the device switches from the RT resistor frequency to PLL mode. The internal 0.5-V voltage source opens and the CLK pin becomes high-impedance as the PLL starts to lock onto the external signal. Because there is a PLL on the regulator, the switching frequency can be higher or lower than the frequency set with the external resistor. The device transitions from the resistor mode to the PLL mode and then increases or decreases the switching frequency until the PLL locks onto the CLK frequency within 100 µs.

When the device transitions from the PLL to resistor mode, the switching frequency slows down from the CLK frequency to 150 kHz; then reapply the 0.5-V voltage, and the resistor then sets the switching frequency. The divisor of the switching frequency goes to 8, 4, 2, and 1 as the voltage ramps from 0 to 0.8 V on the VSENSE pin. The device implements a digital frequency shift to enable synchronizing to an external clock during normal start-up and fault conditions. Figure 36, Figure 37, and Figure 38 show the device synchronized to an external system clock in continuous-conduction mode (CCM), discontinuous-conduction mode (DCM), and pulse-skip mode (PSM).
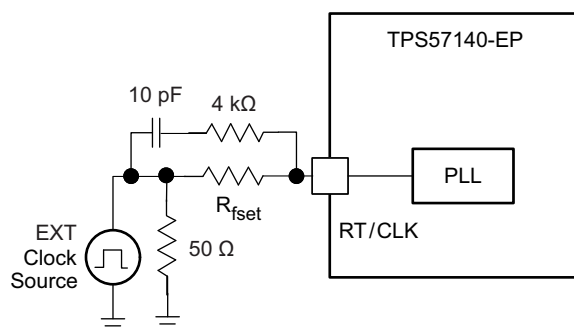

Figure 35. Synchronizing to a System Clock

Feature Description (continued)
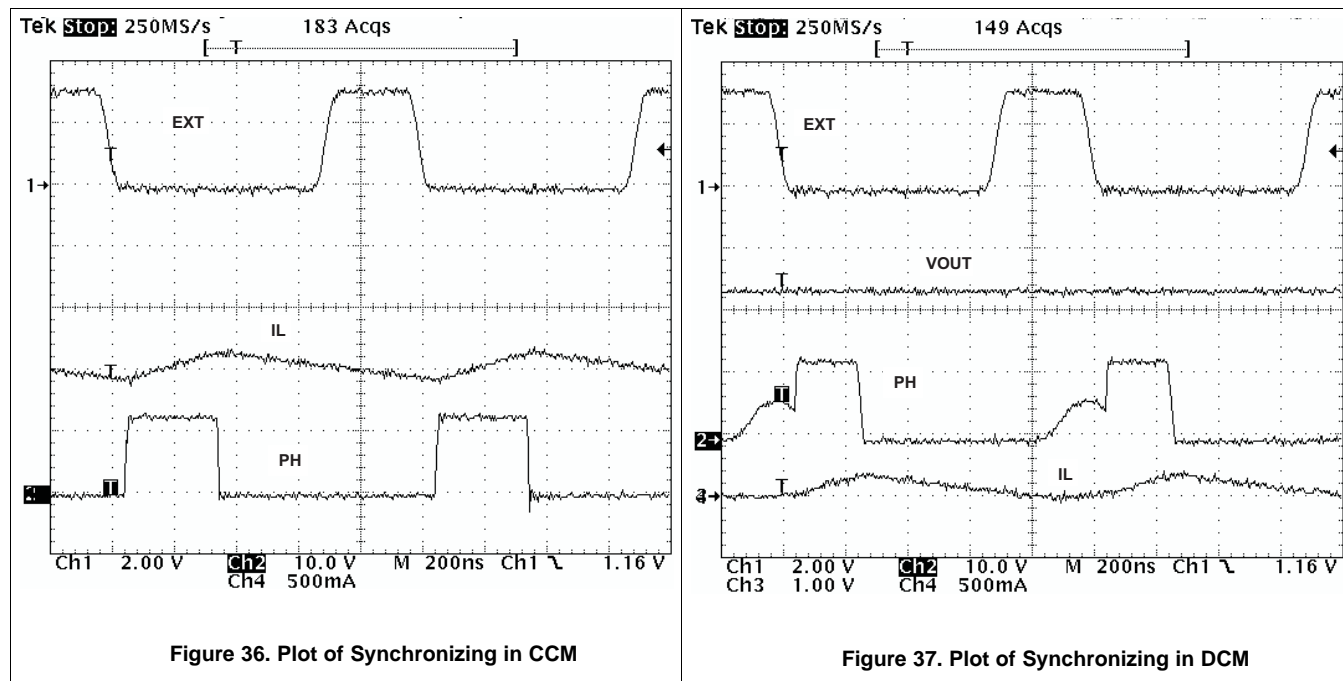
Figure 36. Plot of Synchronizing in CCM

Figure 37. Plot of Synchronizing in DCM
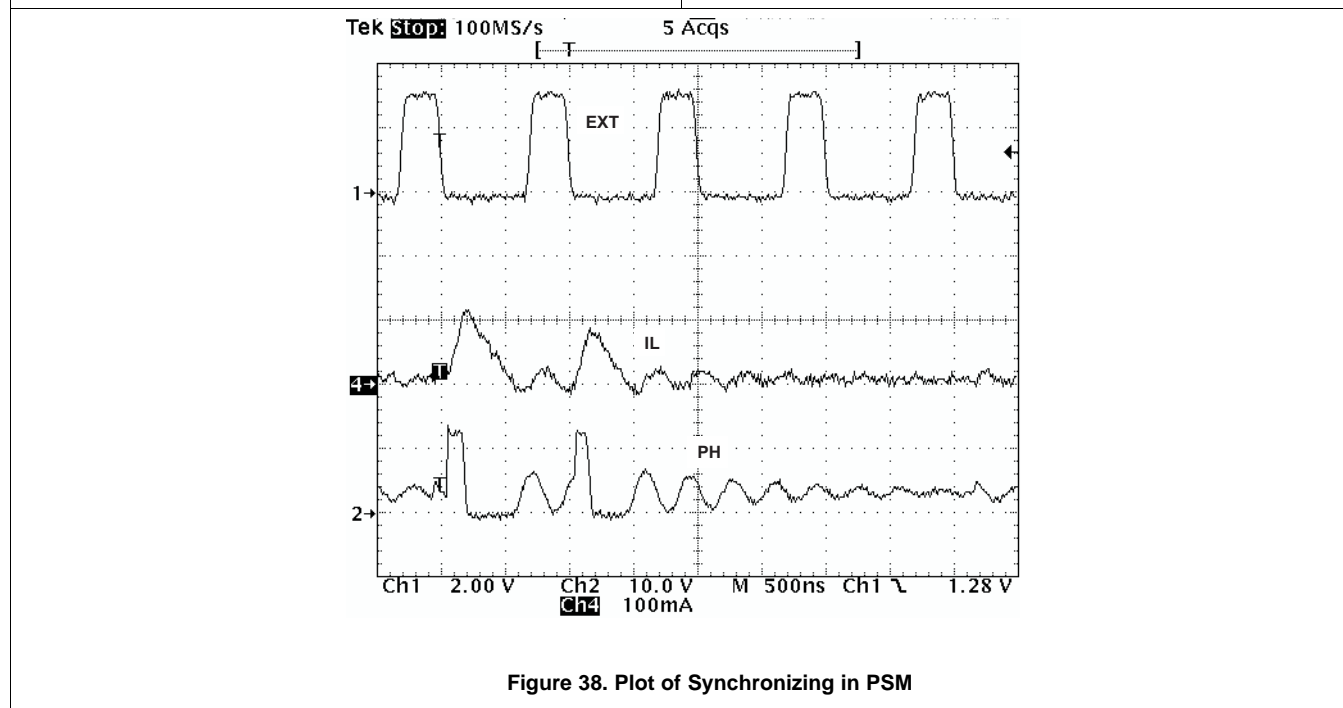
Figure 38. Plot of Synchronizing in PSM

7.3.15 Power Good (PWRGD Pin)

The PWRGD pin is an open-drain output. When the VSENSE pin is between 94% and 107% of the internal voltage reference, the PWRGD pin deasserts and the pin floats. TI recommends using a pullup resistor between the values of 1 and 100 kΩ to a voltage source that is ≤5.5 V. PWRGD is in a defined state when the VIN input voltage is greater than 1.5 V, but with reduced current-sinking capability. PWRGD achieves full current-sinking capability as the VIN input voltage approaches 3 V.

The PWRGD pin goes low when VSENSE is lower than 92% or greater than 109% of the nominal internal reference voltage. Also, PWRGD goes low if UVLO or thermal shutdown asserts or the EN pin goes low.

Feature Description (continued)

7.3.16 Overvoltage Transient Protection (OVTP)

The TPS57140-EP incorporates an OVTP circuit to minimize voltage overshoot when recovering from output fault conditions or strong unload transients on power-supply designs with low-value output capacitance. For example, with the power-supply output overloaded, the error amplifier compares the actual output voltage to the internal reference voltage. If the VSENSE pin voltage is lower than the internal reference voltage for a considerable time, the output of the error amplifier responds by clamping the error-amplifier output to a high voltage, thus requesting the maximum output current. On removal of the condition, the regulator output rises and the error-amplifier output transitions to the steady-state duty cycle. In some applications, the power-supply output voltage can respond faster than the error-amplifier output can respond; this actuality leads to the possibility of an output overshoot. The OVTP feature minimizes the output overshoot when using a low-value output capacitor by implementing a circuit to compare the VSENSE pin voltage to OVTP threshold, which is 109% of the internal voltage reference. A VSENSE pin voltage greater than the OVTP threshold disables the high-side MOSFET, preventing current from flowing to the output and minimizing output overshoot. The VSENSE voltage dropping lower than the OVTP threshold allows the high-side MOSFET to turn on at the next clock cycle.

7.3.17 Thermal Shutdown

The device implements an internal thermal shutdown to protect itself if the junction temperature exceeds 182°C. The thermal shutdown forces the device to stop switching when the junction temperature exceeds the thermal trip threshold. When the die temperature decreases below 182°C, the device reinitiates the power-up sequence by discharging the SS/TR pin.

7.3.18 Small-Signal Model for Loop Response

Figure 39 shows an equivalent model for the TPS57140-EP control loop which the designer can model in a circuit-simulation program to check frequency response and dynamic load response. The error amplifier is a transconductance amplifier with a gm_{EA} of 97 μS. The designer can model the error amplifier using an ideal voltage-controlled current source. Resistor R_o and capacitor C_o model the open-loop gain and frequency response of the amplifier. The 1-mV ac voltage source between nodes a and b effectively breaks the control loop for the frequency-response measurements. Plotting c / a shows the small-signal response of the frequency compensation. Plotting a / b shows the small-signal response of the overall loop. The designer can check the dynamic loop response in a time-domain analysis by replacing R_L with a current source having the appropriate load-step amplitude and step rate. This equivalent model is only valid for CCM designs.
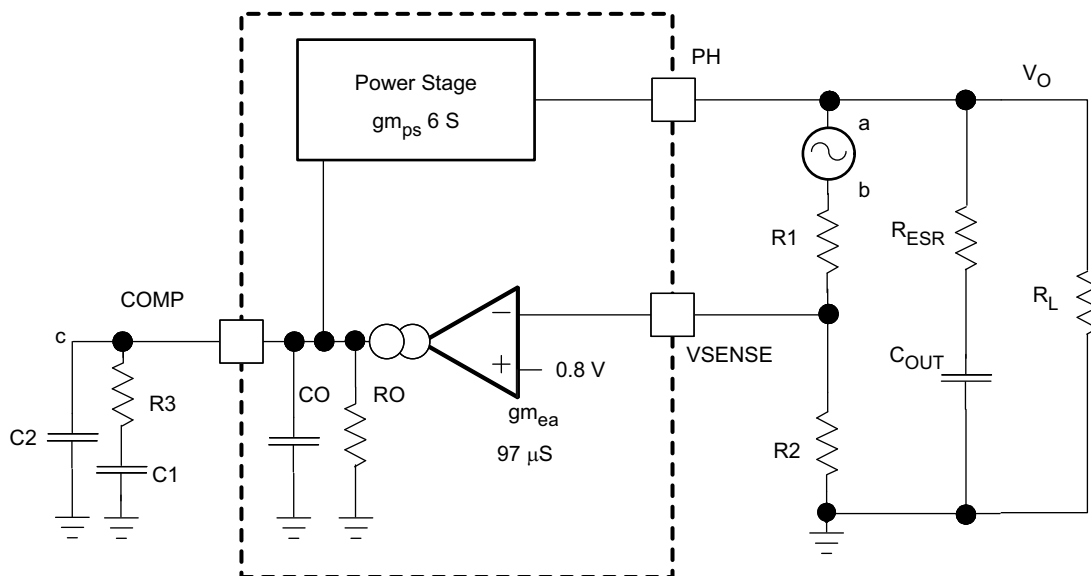

Figure 39. Small-Signal Model for Loop Response

Feature Description (continued)

7.3.19 Simple Small-Signal Model for Peak-Current-Mode Control

Figure 40 describes a simple small-signal model that the designer can use to understand how to design the frequency compensation. The designer can approximate the TPS57140-EP power stage by a voltage-controlled current source (duty-cycle modulator) supplying current to the output capacitor and load resistor. Equation 10 shows the control-to-output transfer function, which consists of a dc gain, one dominant pole, and one ESR zero. The quotient of the change in switch current and the change in COMP pin voltage (node c in Figure 39) is the power-stage transconductance. The gm_{PS} for the TPS57140-EP is 6 S. The low-frequency gain of the power-stage frequency response is the product of the transconductance and the load resistance as shown in Equation 11.

As the load current increases and decreases, the low-frequency gain decreases and increases, respectively. This variation with the load may seem problematic at first glance, but fortunately the dominant pole moves with the load current (see Equation 12). The combined effect is highlighted by the dashed line in the right half of Figure 40. As the load current decreases, the gain increases and the pole frequency lowers, keeping the 0-dB crossover frequency the same for the varying load conditions, which makes it easier to design the frequency compensation. The type of output capacitor chosen determines whether the ESR zero has a profound effect on the frequency-compensation design. Using high-ESR aluminum electrolytic capacitors may reduce the number of frequency-compensation components needed to stabilize the overall loop because the phase margin increases from the ESR zero at the lower frequencies (see Equation 13).
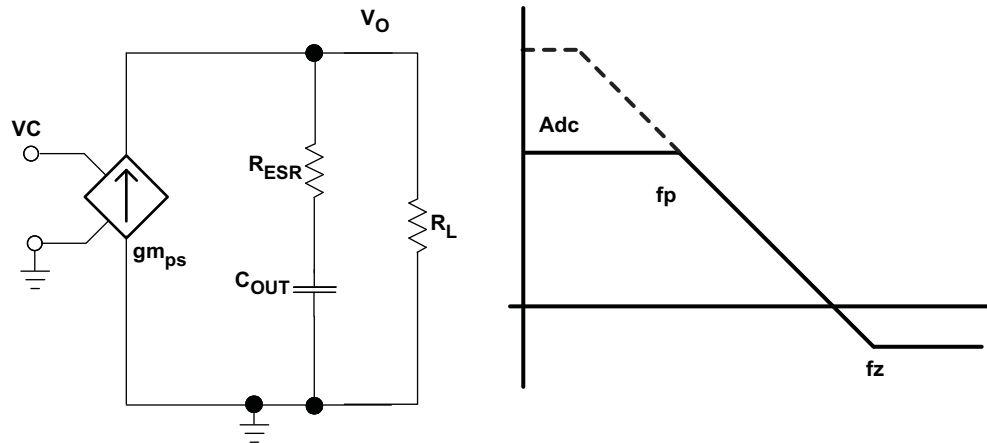
Figure 40. Simple Small-Signal Model and Frequency Response for Peak-Current-Mode Control

$$\frac{V_{OUT}}{V_C} = Adc \times \frac{\left(1 + \dfrac{s}{2\pi \times f_Z}\right)}{\left(1 + \dfrac{s}{2\pi \times f_P}\right)}$$

(10)

$$Adc = gm_{ps} \times R_L$$

(11)

$$f_P = \frac{1}{C_{OUT} \times R_L \times 2\pi}$$

(12)

$$f_Z = \frac{1}{C_{OUT} \times R_{ESR} \times 2\pi}$$

(13)

Feature Description (continued)

7.3.20 Small-Signal Model for Frequency Compensation

The TPS57140-EP uses a transconductance amplifier for the error amplifier and readily supports three of the commonly-used frequency compensation circuits. Figure 41 shows compensation circuits Type 2A, Type 2B, and Type 1. Type 2 circuits are most likely used in high-bandwidth power-supply designs using low-ESR output capacitors. Power-supply designs with high-ESR aluminum electrolytic or tantalum capacitors likely use the Type 1 circuit. Equation 14 and Equation 15 show how to relate the frequency response of the amplifier to the small-signal model in Figure 41. Modeling of the open-loop gain and bandwidth uses the R_O and C_O shown in Figure 41. See *Application and Implementation* for a design example using a Type 2A network with a low-ESR output capacitor.

Equation 14 through Equation 23 are a reference for those who prefer to compensate using the preferred methods. Those who prefer to use a prescribed method must use the method outlined in *Application and Implementation* or use switched information.
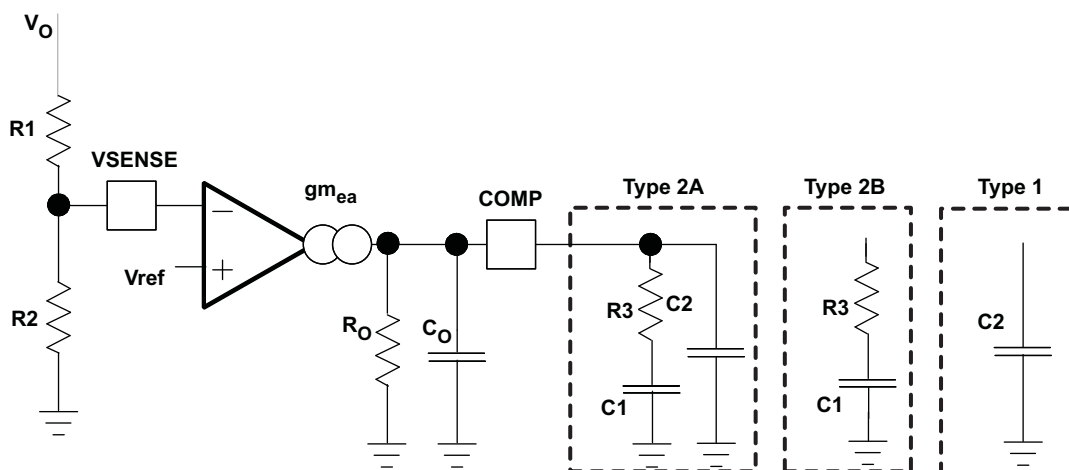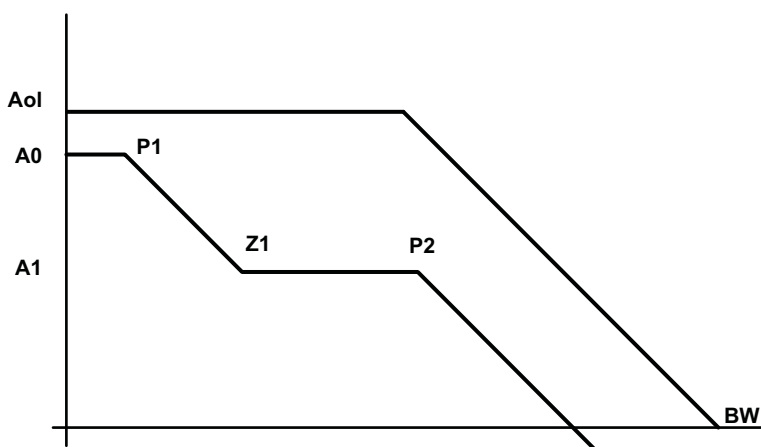

Figure 41. Types of Frequency Compensation

Figure 42. Frequency Response of the Type 2A and Type 2B Frequency Compensation

$$Ro = \frac{Aol(V/V)}{gm_{ea}}$$

(14)

$$C_{OUT} = \frac{gm_{ea}}{2\pi \times BW \,(Hz)}$$

(15)

Feature Description (continued)

$$EA = A0 \times \frac{\left(1 + \dfrac{s}{2\pi \times f_{Z1}}\right)}{\left(1 + \dfrac{s}{2\pi \times f_{P1}}\right) \times \left(1 + \dfrac{s}{2\pi \times f_{P2}}\right)} \tag{16}$$

$$A0 = gm_{ea} \times Ro \times \frac{R2}{R1 + R2} \tag{17}$$

$$A1 = gm_{ea} \times Ro||R3 \times \frac{R2}{R1 + R2} \tag{18}$$

$$P1 = \frac{1}{2\pi \times Ro \times C1} \tag{19}$$

$$Z1 = \frac{1}{2\pi \times R3 \times C1} \tag{20}$$

$$P2 = \frac{1}{2\pi \times R3||R \times (C2 + C_{OUT})} \text{ type 2a} \tag{21}$$

$$P2 = \frac{1}{2\pi \times R3||R \times C_{OUT}} \text{ type 2b} \tag{22}$$

$$P2 = \frac{1}{2\pi \times R \times (C2 + C_{OUT})} \text{ type 1} \tag{23}$$

7.4 Device Functional Modes

7.4.1 Sequencing

The designer can implement many of the common power-supply sequencing methods using the SS/TR, EN, and PWRGD pins. Implement the sequential method using an open-drain output of a power-on-reset pin of another device. Figure 43 shows the sequential method using two TPS57140-EP devices. The power-good pin connects to the EN pin on the TPS57140-EP, which enables the second power supply when the primary supply reaches regulation. If needed, a 1-nF ceramic capacitor on the EN pin of the second power supply provides a 1-ms start-up delay. Figure 44 shows the results of Figure 43.

Figure 45 shows a method for a ratiometric start-up sequence by connecting the SS/TR pins together. The regulator outputs ramp up and reach regulation at the same time. When calculating the slow-start time, the pullup current source must be doubled in Equation 6. Figure 46 shows the results of Figure 45.
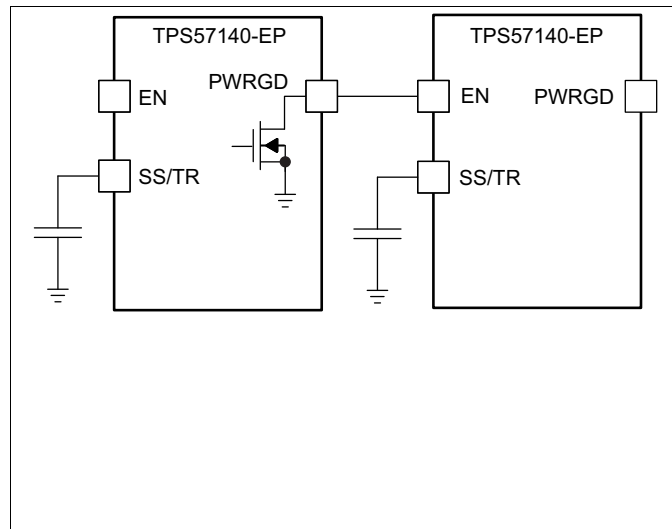
Device Functional Modes (continued)

Figure 43. Schematic for Sequential Start-Up Sequence
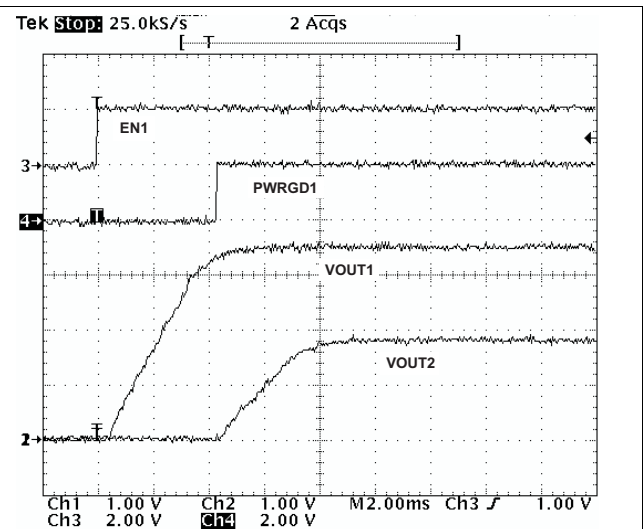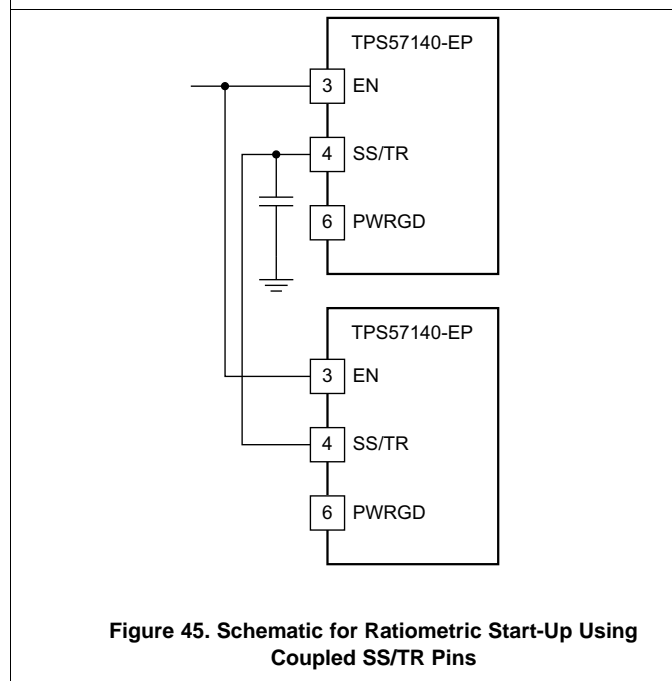

Figure 44. Sequential Start-Up Using EN and PWRGD

Figure 45. Schematic for Ratiometric Start-Up Using Coupled SS/TR Pins
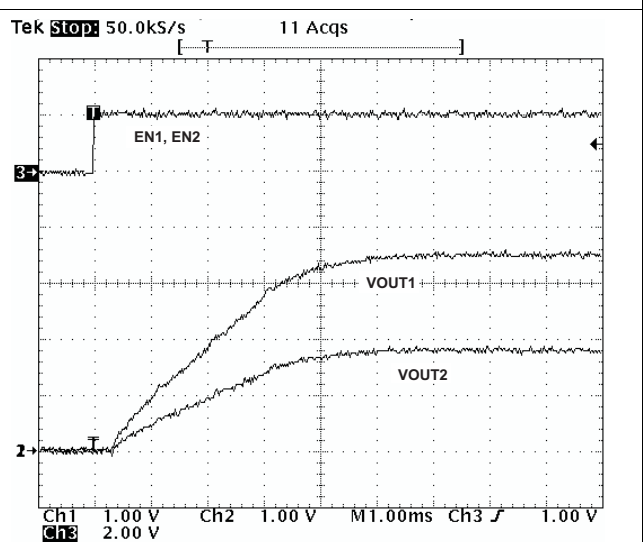

Figure 46. Ratiometric Start-Up Using Coupled SS/TR Pins
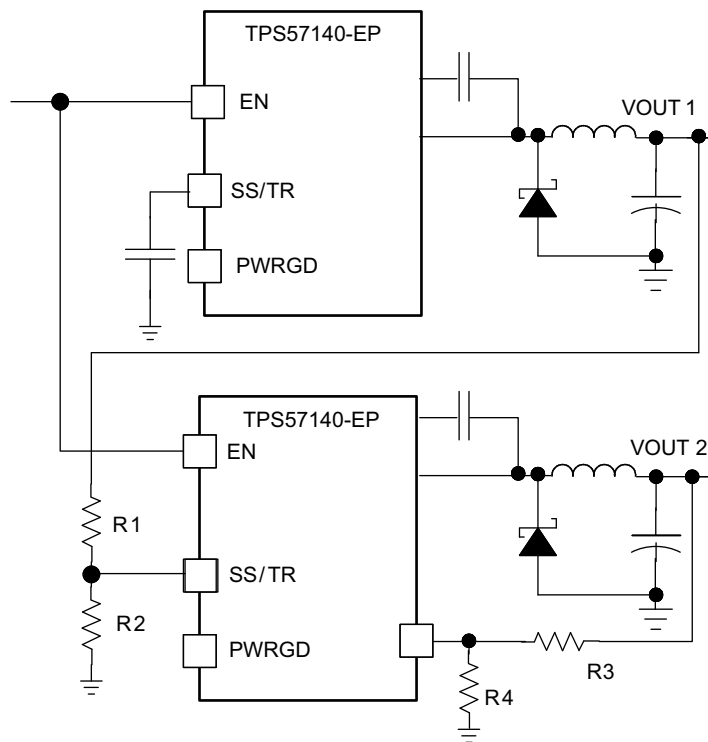
Device Functional Modes (continued)

Figure 47. Schematic for Ratiometric and Simultaneous Start-Up Sequence

The designer can implement ratiometric and simultaneous power-supply sequencing by connecting the resistor network of R1 and R2 shown in Figure 47 to the output of the power supply that requires tracking, or to another voltage reference source. Using Equation 24 and Equation 25, calculate values for the tracking resistors to initiate the Vout2 slightly before, after, or at the same time as Vout1. Equation 26 is the voltage difference between Vout1 and Vout2 at 95% of nominal output regulation.

The deltaV variable is 0 V for simultaneous sequencing. To minimize the effect of the inherent SS/TR-to-VSENSE offset (Vssoffset) in the slow-start circuit and the offset created by the pullup current source (Iss) and tracking resistors, the equations include Vssoffset and Iss as variables.

To design a ratiometric start-up in which the Vout2 voltage is slightly greater than the Vout1 voltage when Vout2 reaches regulation, use a negative number in Equation 24 through Equation 26 for deltaV. Equation 26 results in a positive number for applications in which Vout2 is slightly lower than Vout1 when Vout2 regulation is achieved.

Because the SS/TR pin must be below 40 mV before starting after an EN, UVLO, or thermal shutdown fault, a design requires careful selection of the tracking resistors to ensure the device restarts after a fault. Make sure the calculated R1 value from Equation 24 is greater than the value calculated in Equation 27 to ensure the device can recover from a fault.

As the SS/TR voltage becomes more than 85% of the nominal reference voltage, Vssoffset becomes larger as the slow-start circuits gradually hand off the regulation reference to the internal voltage reference. The SS/TR pin voltage must be greater than 1.3 V for a complete handoff to the internal voltage reference as shown in Figure 24.

Device Functional Modes (continued)

$$R1 = \frac{Vout2 + deltaV}{VREF} \times \frac{Vssoffset}{Iss} \tag{24}$$

$$R2 = \frac{VREF \times R1}{Vout2 + deltaV - VREF} \tag{25}$$

$$deltaV = Vout1 - Vout2 \tag{26}$$

$$R1 > 2800 \times Vout1 - 180 \times deltaV \tag{27}$$

Figure 48. Ratiometric Start-Up With V$_{OUT2}$ Leading V$_{OUT1}$

Figure 49. Ratiometric Start-Up With V$_{OUT1}$ Leading V$_{OUT2}$

Figure 50. Simultaneous Start-Up With Tracking Resistor
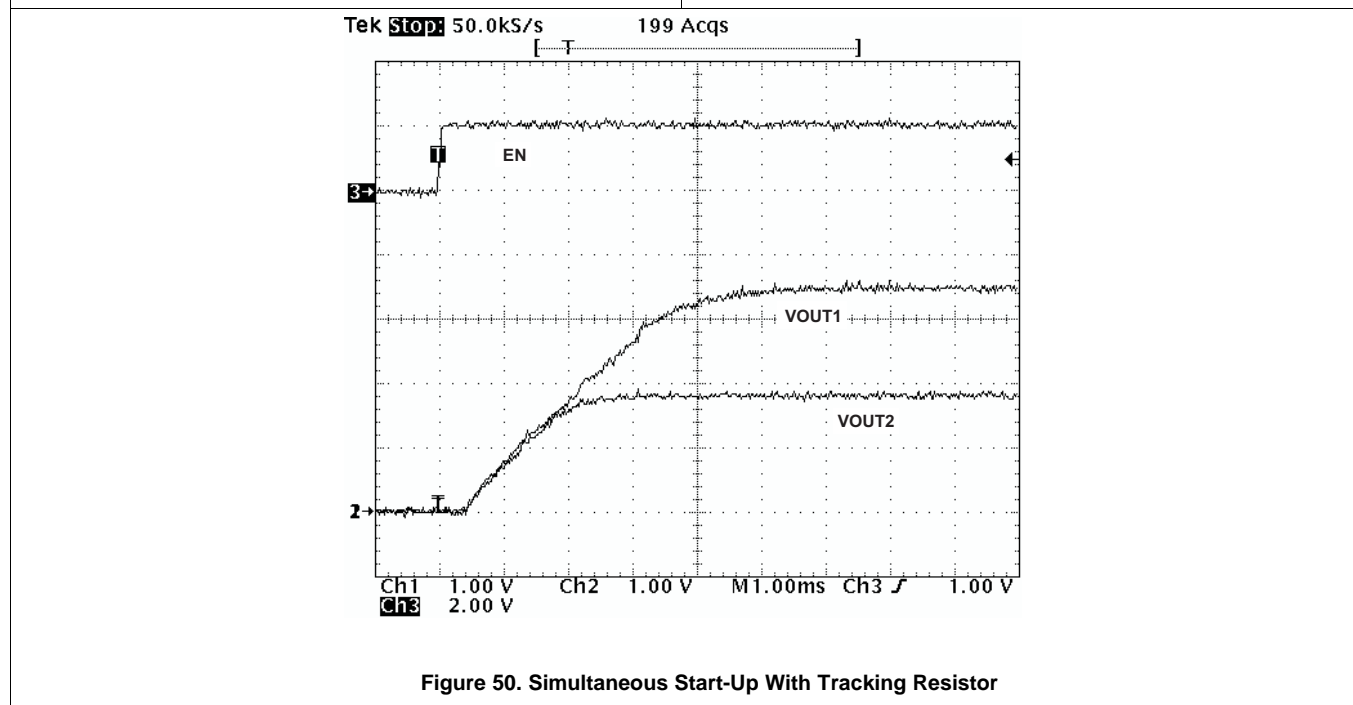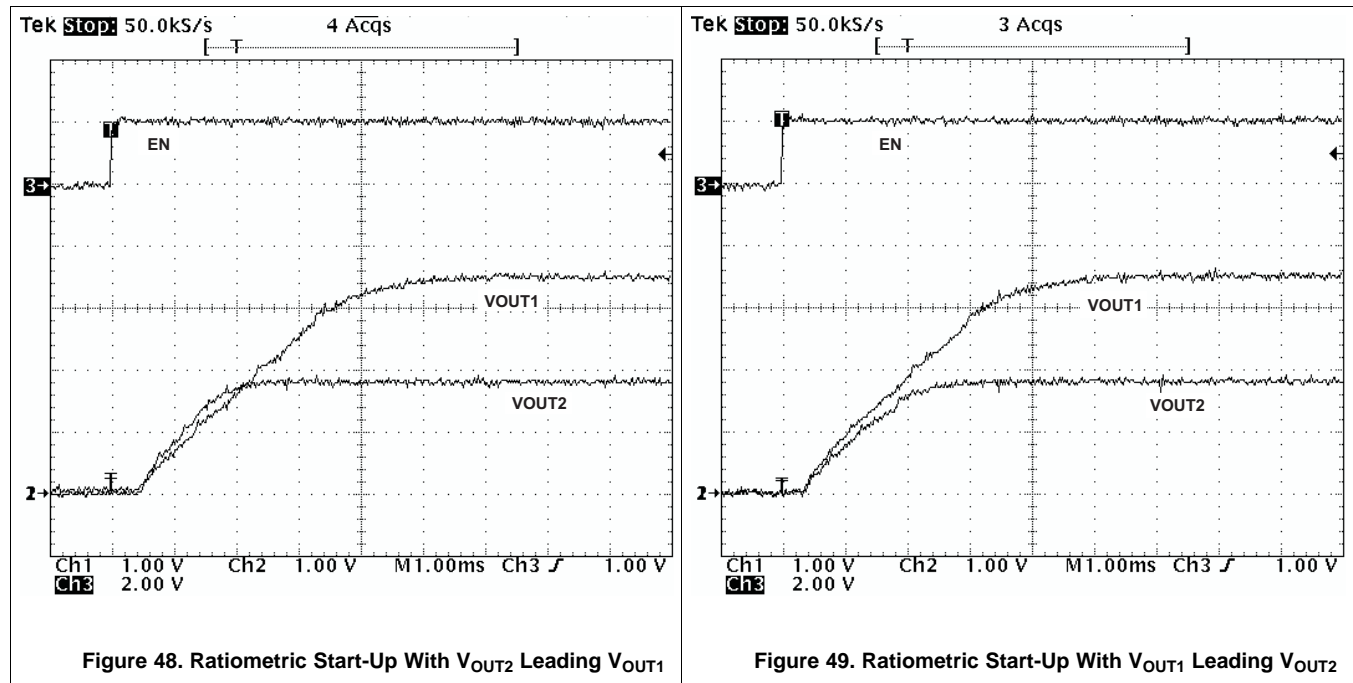
Device Functional Modes (continued)

7.4.2 Pulse-Skip Eco-mode Control Scheme

The TPS57140-EP enters the pulse-skip mode when the voltage on the COMP pin is the minimum clamp value. The TPS57140-EP operates in a pulse-skip mode at light load currents to improve efficiency. The peak switch current during the pulse-skip mode is the greater value of either 50 mA or the peak inductor current that is a function of the minimum on-time, input voltage, output voltage, and inductance value. When the load current is low and the output voltage is within regulation, the device enters a sleep mode and draws only 116 μA of input quiescent current. While the device is in sleep mode, the output capacitor delivers the output power. As the load current decreases, the time the output capacitor supplies the load current increases and the switching frequency decreases, reducing gate-drive and switching losses. As the output voltage drops, the TPS57140-EP wakes up from sleep mode and the power switch turns on to recharge the output capacitor, see Figure 51. The internal PLL remains operating when in sleep mode. When operating at light load currents in pulse-skip mode, the switching transitions occur synchronously with the external clock signal.

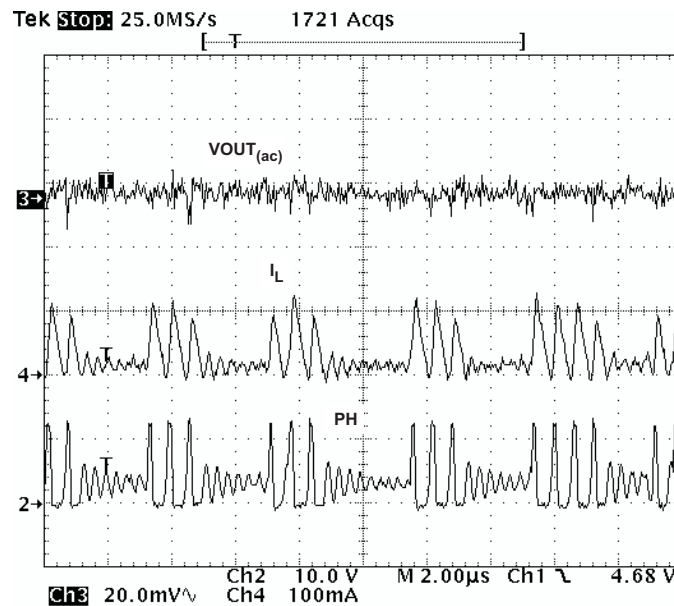**Figure 51. Operation in Pulse-Skip Mode**

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 TPS57140-EP DC-DC converter is designed to provide up to a 1.5-A output from an input voltage source of 3.5 V to 42 V. The high-side MOSFET is incorporated inside the TPS57140-EP package along with the gate-drive circuitry. The low drain-to-source on-resistance of the MOSFET allows the TPS57140-EP device to achieve high efficiencies and helps keep the junction temperature low at high output currents. The compensation components are external to the integrated circuit (IC), and an external divider allows for an adjustable output voltage. Additionally, the TPS57140-EP device provides adjustable slow start and undervoltage-lockout inputs.

8.2 Typical Application
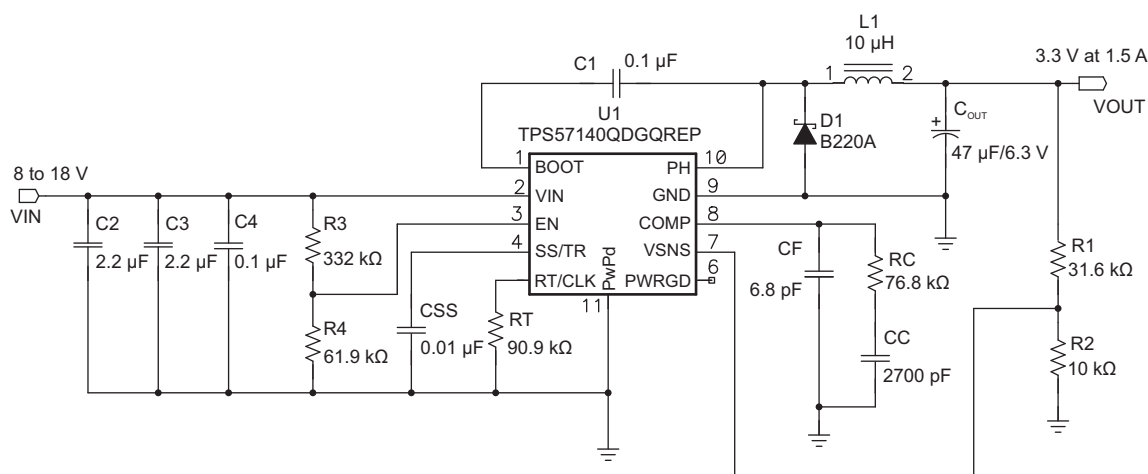
Figure 52. High-Frequency, 3.3-V Output Power-Supply Design With Adjusted UVLO

8.2.1 Design Requirements

This example details the design of a high-frequency switching regulator design using ceramic output capacitors. The designer must know a few parameters in order to start the design process. The determination of these parameters is typically at the system level. This example starts with the following known parameters in Table 1.

Table 1. Design Parameters

DESIGN PARAMETER	VALUE
Output voltage	3.3 V
Transient response, 0- to 1.5-A load step	ΔVout = 4%
Maximum output current	1.5 A
Input voltage	12-V nominal, 8 to 18 V
Output voltage ripple	<33 mV$_{pp}$
Start input voltage (rising VIN)	7.25 V
Stop input voltage (falling VIN)	6.25 V

8.2.2 Detailed Design Procedure

8.2.2.1 Selecting the Switching Frequency

The first step is to decide on a switching frequency for the regulator. Typically, the user wants to choose the highest switching frequency possible, because this produces the smallest solution size. The high switching frequency allows for lower-valued inductors and smaller output capacitors compared to a power supply that switches at a lower frequency. The switching frequency that the designer can select has limits imposed by the minimum on-time of the internal power switch, the input voltage, the output voltage, and the frequency-shift limitation.

Use Equation 8 and Equation 9 to find the maximum switching frequency for the regulator; choose the lower value of the two equations. Switching frequencies higher than this value result in pulse skipping or a lack of overcurrent protection during a short circuit.

The typical minimum on-time, t_{onmin}, is 130 ns for the TPS57140-EP. For this example, the output voltage is 3.3 V and the maximum input voltage is 18 V, which allows for a maximum switch frequency up to 1600 kHz when including the inductor resistance, on-resistance, and diode voltage in Equation 8. To ensure overcurrent runaway is not a concern during short circuits in the design, use Equation 9 or the solid curve in Figure 34 to determine the maximum switching frequency. With a maximum input voltage of 20 V, assuming a diode voltage of 0.5 V, inductor resistance of 100 mΩ, switch resistance of 200 mΩ, and an output current of 2.8 A, the maximum switching frequency is approximately 1600 kHz.

Choosing the lower of the two values and adding some margin, this example uses a switching frequency of 1200 kHz. To determine the timing resistance for a given switching frequency, use Equation 7 or the curve in Figure 32.

Resistor R_t sets the switching frequency as shown in Figure 52.

8.2.2.2 Output Inductor Selection (L_O)

To calculate the minimum value of the output inductor, use Equation 28.

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

The output capacitor filters the inductor ripple current. Therefore, choosing high inductor ripple currents impacts the selection of the output capacitor, because the output capacitor must have a ripple-current rating equal to or greater than the inductor ripple current. In general, the inductor ripple value is at the discretion of the designer; however, the designer can use the following guidelines.

For designs using low-ESR output capacitors such as ceramics, use a value as high as K_{IND} = 0.3. When using higher-ESR output capacitors, K_{IND} = 0.2 yields better results. Because the inductor ripple current is part of the PWM control system, the inductor ripple current should always be >100 mA for dependable operation. In a wide-input voltage regulator, it is best to choose an inductor ripple current on the larger side. This allows the inductor to still have a measurable ripple current with the input voltage at its minimum.

For this design example, use K_{IND} = 0.2 and the calculated minimum inductor value is 7.6 µH. For this design, the choice was a nearest standard value of 10 µH. For the output filter inductor, it is important not to exceed the RMS-current and saturation-current ratings. Find the RMS and peak inductor current from Equation 30 and Equation 31.

For this design, the RMS inductor current is 1.506 A and the peak inductor current is 1.62 A. The chosen inductor is a MSS6132-103. It has a saturation current rating of 1.64 A and an RMS current rating of 1.9 A.

As the equation set demonstrates, lower ripple currents reduce the output voltage ripple of the regulator but require a larger value of inductance. Selecting higher ripple currents increases the output-voltage ripple of the regulator but allows for a lower inductance value.

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

$$Lo\ min = \frac{Vinmax - Vout}{Io \times K_{IND}} \times \frac{Vout}{Vinmax \times f sw}$$

(28)

$$I_{RIPPLE} \leq I_O \times K_{IND} \tag{29}$$

$$I_{L(rms)} = \sqrt{(I_O)^2 + \frac{1}{12} \times \left(\frac{V_{OUT} \times (Vinmax - V_{OUT})}{Vinmax \times L_O \times f_{SW}} \right)^2} \tag{30}$$

$$ILpeak = Iout + \frac{Iripple}{2} \tag{31}$$

8.2.2.3 *Output Capacitor*

The three primary considerations for selecting the value of the output capacitor are: the output capacitor determines the modulator pole, the output-voltage ripple, and how the regulator responds to a large change in load current. Select the output capacitance based on the most 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 would occur if there are desired hold-up times for the regulator where the output capacitor must hold the output voltage above a certain level for a specified amount of time after the input power is removed. The regulator also temporarily is not able to supply sufficient output current if there is a large, fast increase in the current needs of the load, such as transitioning from no load to a full load. The regulator usually needs two or more clock cycles for the control loop to see the change in load current and output voltage and adjust the duty cycle to react to the change. The output-capacitor size must be adequate to supply the extra current to the load until the control loop responds to the load change. The output capacitance must be large enough to supply the difference in current for two clock cycles while only allowing a tolerable amount of droop in the output voltage. Calculate the minimum output capacitance necessary to accomplish this using Equation 32.

For this example, the transient load response is specified as a 4% change in Vout for a load step from 0 A (no load) to 1.5 A (full load). For this example, ΔIout = 1.5 − 0 = 1.5 A and ΔVout = 0.04 × 3.3 V = 0.132 V. Using these numbers gives a minimum capacitance of 18.9 μF. This value does not take the ESR of the output capacitor into account in the output-voltage change. For ceramic capacitors, the ESR is usually small enough to ignore in this calculation. Aluminum electrolytic and tantalum capacitors have higher ESR that the designer must take into account.

The catch diode of the regulator cannot sink current, so any stored energy in the inductor produces an output voltage overshoot when the load current rapidly decreases, see Figure 53. The output capacitor must also be sized to absorb energy stored in the inductor when transitioning from a high load current to a lower load current. The excess energy that is stored in the output capacitor increases the voltage on the capacitor. The capacitor must be sized to maintain the desired output voltage during these transient periods. UseEquation 33 to calculate the minimum capacitance to keep the output voltage overshoot to a desired value. For this example, the worst-case load step is from 1.5 to 0 A. The output voltage increases during this load transition, and the stated maximum in our specification is 4% of the output voltage. This makes Vf = 1.04 × 3.3 = 3.432. Vi is the initial capacitor voltage, which is the nominal output voltage of 3.3 V. Using these numbers in Equation 33 yields a minimum capacitance of 25.3 μF.

Use Equation 34 to calculate the minimum output capacitance needed to meet the ripple specification for output voltage. Equation 34 yields 0.7 μF.

Use Equation 35 to calculate the maximum ESR an output capacitor can have to meet the ripple specification for the output voltage. Equation 35 indicates the ESR should be less than 147 mΩ.

The most stringent criterion for the output capacitor is 25.3 μF of capacitance to keep the output voltage in regulation during an unload transient.

Factor in additional capacitance deratings for aging, temperature, and dc bias, increasing this minimum value. For this example, select a 47-μF 6.3-V X7R ceramic capacitor with 5 mΩ of ESR.

Capacitors generally have limits to the amount of ripple current they can handle without failing or producing excess heat. Specify an output capacitor that can support the inductor ripple current. Some capacitor data sheets specify the root-mean-square (RMS) value of the maximum ripple current. Use Equation 36 to calculate the RMS ripple current that the output capacitor must support. For this application, Equation 36 yields 64.8 mA.

$$C_{OUT} > \frac{2 \times \Delta I_{OUT}}{f_{SW} \times \Delta V_{OUT}}$$

where

- ΔIout is the change in output current.
- f_{SW} is the switching frequency of the regulator.
- ΔVout is the allowable change in the output voltage.

(32)

$$C_{OUT} > L_O \times \frac{\left((I_{OH})^2 - (I_{OL})^2\right)}{\left((V_f)^2 - (V_i)^2\right)}$$

where

- L is the value of the inductor.
- I_{OH} is the output current under heavy load.
- I_{OL} is the output under light load.
- VF is the final peak output voltage.
- Vi is the initial capacitor voltage

(33)

$$C_{OUT} > \frac{1}{8 \times f_{SW}} \times \frac{1}{\left(\dfrac{V_{OUT(ripple)}}{I_{RIPPLE}}\right)}$$

where

- f_{SW} is the switching frequency.
- $V_{oripple}$ is the maximum allowable output-voltage ripple.
- I_{ripple} is the ripple current of the inductor.

(34)

$$R_{ESR} = \frac{V_{OUT(ripple)}}{I_{RIPPLE}}$$

(35)

$$I_{COUT(rms)} = \frac{V_{OUT} \times \left(V_{IN(max)} - V_{OUT}\right)}{\sqrt{12} \times V_{IN(max)} \times L_O \times f_{SW}}$$

(36)

8.2.2.4 Catch Diode

The TPS57140-EP requires an external catch diode between the PH pin and GND. The selected diode must have a reverse voltage rating equal to or greater than Vinmax. The peak current rating of the diode must be greater than the maximum inductor current. The diode should also have a low forward voltage. Schottky diodes are typically a good choice for the catch diode due to their low forward voltage. The lower the forward voltage of the diode, the higher the efficiency of the regulator is.

Typically, the higher the voltage and current ratings the diode has, the higher the forward voltage is. Because the design example has an input voltage up to 18 V, select a diode with a minimum of 20-V reverse voltage.

For the example design, Schottky diode selection is the B220A for its lower forward voltage, and it comes in a larger package size, which has good thermal characteristics over small devices. The typical forward voltage of the B220A is 0.5 V.

The diode selection must also have an appropriate power rating. The diode conducts the output current during the off-time of the internal power switch. The off-time of the internal switch is a function of the maximum input voltage, the output voltage, and the switching frequency. The output current during the off-time multiplied by the forward voltage of the diode equals the conduction losses of the diode. At higher switch frequencies, take the ac losses of the diode into account. The ac losses of the diode are due to the charging and discharging of the junction capacitance and reverse recovery. Use Equation 37 to calculate the total power dissipation, conduction losses plus ac losses, of the diode.

The B220A has a junction capacitance of 120 pF. Using Equation 37, the selected diode dissipates 0.632 W. This power dissipation, depending on mounting techniques, should produce a 16°C temperature rise in the diode when the input voltage is 18 V and the load current is 1.5 A.

If the power supply spends a significant amount of time at light load currents or in sleep mode, consider using a diode which has a low leakage current and slightly higher forward voltage drop.

$$P_D = \frac{\left(V_{IN(max)} - V_{OUT}\right) \times I_{OUT} \times V\!f\mathrm{d}}{V_{IN(max)}} + \frac{C_j \times f_{SW} \times \left(V_{IN} + V\!f\mathrm{d}\right)^2}{2}$$

(37)

8.2.2.5 Input Capacitor

The TPS57140-EP requires a high-quality ceramic, type X5R or X7R, input decoupling capacitor of at least 3 μF of effective capacitance, and in some applications a bulk capacitance. The effective capacitance includes any dc-bias effects. The voltage rating of the input capacitor must be greater than the maximum input voltage. The capacitor must also have a ripple-current rating greater than the maximum input-current ripple of the TPS57140-EP. Calculate the input-ripple current using Equation 38.

The value of a ceramic capacitor varies significantly over temperature and the amount of dc bias applied to the capacitor. The designer can minimize the capacitance variations due to temperature by selecting a dielectric material that is stable over temperature. Designers usually select X5R and X7R ceramic dielectrics for power regulator capacitors because they have a high capacitance-to-volume ratio and are fairly stable over temperature. The designer must also take the dc bias into account for output capacitor selection. The capacitance value of a capacitor decreases as the dc bias across a capacitor increases.

This example design requires a ceramic capacitor with at least a 20-V voltage rating to support the maximum input voltage. Common standard ceramic capacitor voltage ratings include 4 V, 6.3 V, 10 V, 16 V, 25 V, 50 V, and 100 V, so select a 25-V capacitor. For this example, the selection is two 2.2-μF, 25-V capacitors in parallel. Table 2 shows a selection of high-voltage capacitors. The input capacitance value determines the input-voltage ripple of the regulator. Calculate the input-voltage ripple using Equation 39. Using the design example values, Ioutmax = 1.5 A, Cin = 4.4 μF, f_{SW} = 1200 kHz, yields an input-voltage ripple of 71 mV and an RMS input-ripple current of 0.701 A.

$$\mathrm{Icirms} = \mathrm{Iout} \times \sqrt{\frac{\mathrm{Vout}}{\mathrm{Vin\ min}} \times \frac{\left(\mathrm{Vin\ min} - \mathrm{Vout}\right)}{\mathrm{Vin\ min}}}$$

(38)

$$\Delta\mathrm{Vin} = \frac{\mathrm{Iout\ max} \times 0.25}{\mathrm{Cin} \times f\mathrm{sw}}$$

(39)

Table 2. Capacitor Types

VENDOR	VALUE (µF)	EIA Size	VOLTAGE	DIELECTRIC	COMMENTS
Murata	1 to 2.2	1210	100 V	X7R	GRM32 series
	1 to 4.7		50 V		
	1	1206	100 V		GRM31 series
	1 to 2.2		50 V		
Vishay	1 10 1.8	2220	50 V		VJ X7R series
	1 to 1.2		100 V		
	1 to 3.9	2225	50 V		
	1 to 1.8		100 V		
TDK	1 to 2.2	1812	100 V		C series C4532
	1.5 to 6.8		50 V		
	1 to 2.2	1210	100 V		C series C3225
	1 to 3.3		50 V		
AVX	1 to 4.7	1210	50 V		X7R dielectric series
	1		100 V		
	1 to 4.7	1812	50 V		
	1 to 2.2		100 V		

8.2.2.6 Slow-Start Capacitor

The slow-start capacitor determines the minimum amount of time it takes for the output voltage to reach its nominal programmed value during power up. This is useful if a load requires a controlled voltage-slew rate. This is also used if the output capacitance is very large and would require large amounts of current to charge the capacitor quickly to the output-voltage level. The large currents necessary to charge the capacitor may make the TPS57140-EP reach the current limit, or excessive current draw from the input power supply may cause the input voltage rail to sag. Limiting the output-voltage slew rate solves both of these problems.

The slow-start time must be long enough to allow the regulator to charge the output capacitor up to the output voltage without drawing excessive current. Use Equation 40 to find the minimum slow start time, t_{SS}, necessary to charge the output capacitor, Cout, from 10% to 90% of the output voltage, Vout, with an average slow-start current of Issavg. In the example, to charge the 47-µF output capacitor up to 3.3 V while only allowing the average input current to be 0.125 A would require a 1-ms slow-start time.

After the slow-start time is known, calculate the slow-start capacitor value using Equation 6. For the example circuit, the slow-start time is not too critical, because the output capacitor value is 47 µF, which does not require much current to charge to 3.3 V. The example circuit has the slow-start time set to an arbitrary value of 1 ms, which requires a 3.3-nF capacitor.

$$\text{Tss} > \frac{\text{Cout} \times \text{Vout} \times 0.8}{\text{Issavg}}$$

(40)

8.2.2.7 Bootstrap Capacitor Selection

Connect a 0.1-µF ceramic capacitor between the BOOT and PH pins for proper operation. TI recommends using a ceramic capacitor with X5R or better grade dielectric. The capacitor should have a 10-V or higher voltage rating.

8.2.2.8 UVLO Set Point

The designer can adjust the UVLO using an external voltage divider on the EN pin of the TPS57140-EP. The UVLO has two thresholds, one for power up when the input voltage is rising and one for power down or brownouts when the input voltage is falling. For the example design, the supply should turn on and start switching when the input voltage increases above 7.25 V (enabled). After the regulator starts switching, it should continue to do so until the input voltage falls below 6.25 V (UVLO stop).

Set the programmable UVLO and enable voltages by using a resistor divider between Vin and ground to the EN pin. Use Equation 2 through Equation 3 to calculate the resistance values necessary. For the example application, 332 kΩ between Vin and EN and 61.9 kΩ between EN and ground are required to produce the 7.25- and 6.25-V start and stop voltages.

8.2.2.9 Output Voltage and Feedback Resistors Selection

For the example design, the selected resistance for R2 is 10 kΩ. Using Equation 1, the calculated value of R1 is 31.25 kΩ. The nearest standard 1% resistor is 31.6 kΩ. Due to current leakage of the VSENSE pin, the current flowing through the feedback network should be greater than 1 μA to maintain the output voltage accuracy. This requirement makes the maximum value of R2 equal to 800 kΩ. Choosing higher resistor values decreases quiescent current and improves efficiency at low output currents, but may introduce noise-immunity problems.

8.2.2.10 Compensation

Several industry techniques are used to compensate dc-dc regulators. The method presented here yields high phase margins. For most conditions, the regulator has a phase margin between 60° and 90°. The method presented here ignores the effects of the slope compensation that is internal to the TPS57140-EP. Ignoring the slope compensation usually causes the actual crossover frequency to be lower than the crossover frequency used in the calculations.

Use SwitcherPro software for a more accurate design.

The uncompensated regulator has a dominant pole, typically located between 300 Hz and 3 kHz, due to the output capacitor and load resistance, and a pole due to the error amplifier. One zero exists due to the output capacitor and the ESR. The zero frequency is higher than either of the two poles.

If left uncompensated, the double pole created by the error amplifier and the modulator would lead to an unstable regulator. Stabilizing the regulator requires one pole to be canceled out. One design approach is to locate a compensating zero at the modulator pole. Then, select a crossover frequency that is higher than the modulator pole. Calculate the gain of the error amplifier to achieve the desired crossover frequency. The capacitor used to create the compensation zero, along with the output impedance of the error amplifier, forms a low-frequency pole to provide a minus-one slope through the crossover frequency. Then, adding a compensating pole cancels the zero due to the output-capacitor ESR. If the ESR zero resides at a frequency higher than the switching frequency, then it can be ignored.

To compensate the TPS57140-EP using this method, first calculate the modulator pole and zero using the following equations:

$$f\text{p mod} = \frac{\text{Ioutmax}}{2 \times \pi \times \text{Vout} \times \text{Cout}}$$

where

- Ioutmax is the maximum output current.
- Cout is the output capacitance.
- Vout is the nominal output voltage. (41)

$$f\text{z mod} = \frac{1}{2 \times \pi \times \text{Resr} \times \text{Cout}}$$

(42)

For the example design, the location of the modulator pole is at 1.5 kHz and the ESR zero is at 338 kHz.

Next, the designer must select a crossover frequency which determines the bandwidth of the control loop. The crossover-frequency location must be at a frequency at least 5× higher than the modulator pole. The crossover-frequency selection must also be such that the available gain of the error amplifier at the crossover frequency is high enough to allow for proper compensation.

Use Equation 47 to calculate the maximum crossover frequency when the ESR-zero location is at a frequency that is higher than the desired crossover frequency. This is usually the case for ceramic or low-ESR tantalum capacitors. Aluminum electrolytic and tantalum capacitors typically produce a modulator zero at a low frequency due to their high ESR.

The example application uses a low-ESR ceramic capacitor with 10 mΩ of ESR, making the zero at 338 kHz.

This value is much higher than typical crossover frequencies, so calculate the maximum crossover frequency using both Equation 43 and Equation 46.

Using Equation 46 gives a minimum crossover frequency of 7.6 kHz and Equation 43 gives a maximum crossover frequency of 45.3 kHz.

Arbitrarily select a crossover frequency of 45 kHz from this range.

$$F_{c\,max} \leq 2100 \sqrt{\frac{F_{pmod}}{V_{out}}} \text{ for ceramic capacitors.}$$
(43)

$$F_{c\,max} \leq \frac{51442}{\sqrt{V_{out}}} \text{ for Tantalum or Aluminum capacitors.}$$
(44)

$$F_{c\,max} \leq \frac{F_{sw}}{5} \text{ for all cases.}$$
(45)

$$F_{c\,min} \geq 5 \times F_{pmod} \text{ for all cases.}$$
(46)

After selection of a crossover frequency, F_c, calculate the gain of the modulator at the crossover frequency using Equation 47.

$$G_{MOD(fc)} = \frac{gm_{(PS)} \times R_{LOAD} \times (2\pi \times f_C \times C_{OUT} \times R_{ESR} + 1)}{2\pi \times f_C \times C_{OUT} \times (R_{LOAD} + R_{ESR}) + 1}$$
(47)

For the example problem, the gain of the modulator at the crossover frequency is 0.542. Next, calculate the compensation components. Use a resistor in series with a capacitor to create a compensating zero. A capacitor in parallel to these two components forms the compensating pole. However, calculating the values of these components varies, depending on whether the ESR-zero location is above or below the crossover frequency. For ceramic or low-ESR tantalum output capacitors, the zero location is usually above the crossover frequency. For aluminum electrolytic and tantalum capacitors, the modulator-zero location is usually lower in frequency than the crossover frequency. For cases where the modulator zero is higher than the crossover frequency (ceramic capacitors):

$$R_C = \frac{V_{OUT}}{G_{MOD(fc)} \times gm_{(EA)} \times V_{REF}}$$
(48)

$$Cc = \frac{1}{\pi \times Rc \times f_{p\,mod}}$$
(49)

$$Cf = \frac{Co \times Resr}{Rc}$$
(50)

For cases where the modulator zero is less than the crossover frequency (aluminum or tantalum capacitors), the equations are:

$$R_C = \frac{V_{OUT}}{G_{MOD(fc)} \times f_{Z(mod)} \times gm_{(EA)} \times V_{REF}}$$
(51)

$$Cc = \frac{1}{\pi \times Rc \times f_{p\,mod}}$$
(52)

$$Cf = \frac{1}{2 \times \pi \times Rc \times f_{z\,mod}}$$
(53)

For the example problem, the ESR-zero location is at a higher frequency compared to the crossover frequency, so use Equation 50 through Equation 53 to calculate the compensation components. For the example problem, the calculated components are: Rc = 76.2 kΩ, Cc = 2710 pF, and Cf = 6.17 pF.

The calculated value of the Cf capacitor is not a standard value, so use a value of 2700 pF. Use 6.8 pF for Cc. The Rc resistor sets the gain of the error amplifier, which determines the crossover frequency. The calculated Rc resistor is not a standard value, so use 76.8 kΩ.
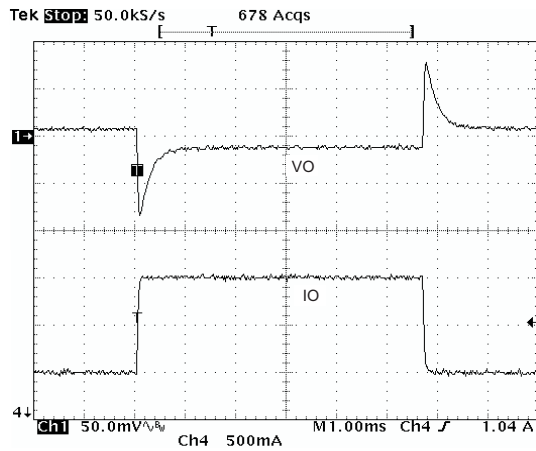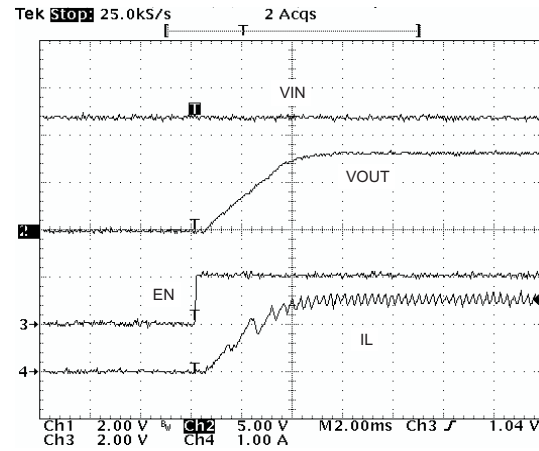
8.2.3 Application Curves

Figure 53. Load Transmit
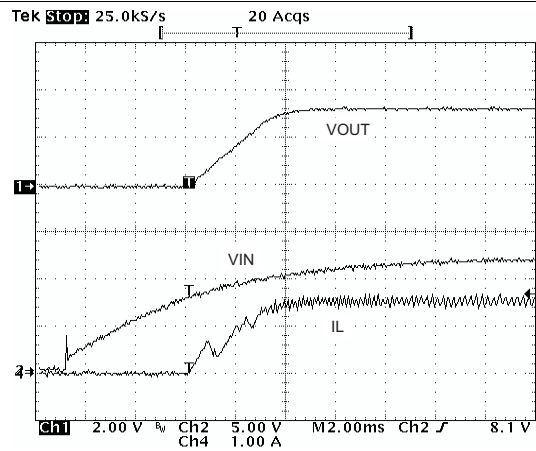

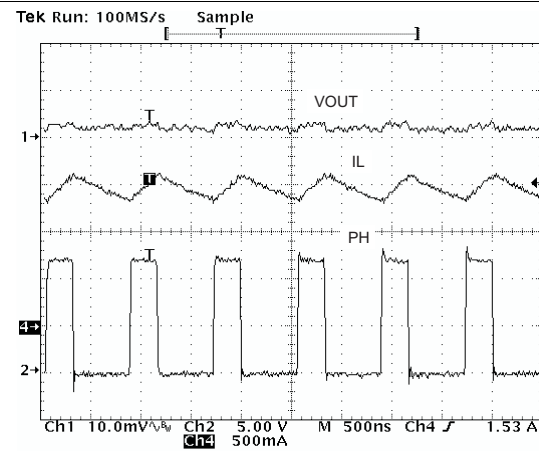**Figure 54. Startup With EN**

Figure 55. VIN Power Up

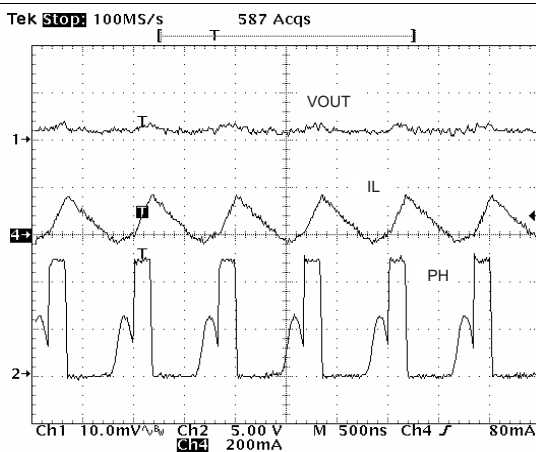**Figure 56. Output Ripple, CCM**

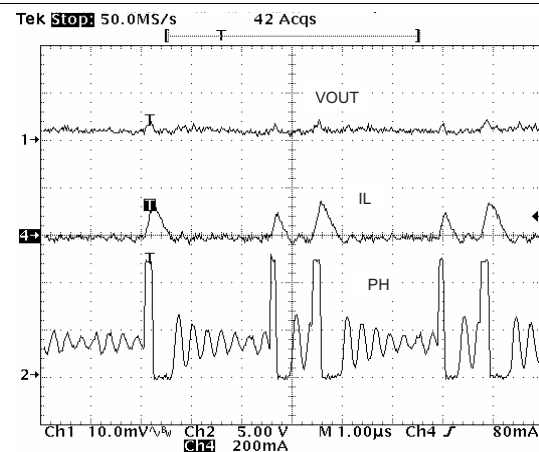**Figure 57. Output Ripple, DCM**
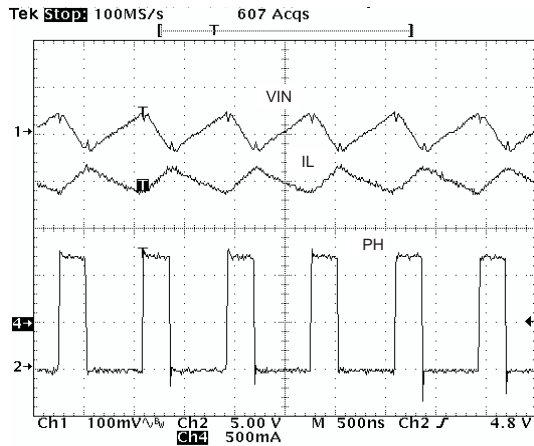

Figure 58. Output Ripple, PSM
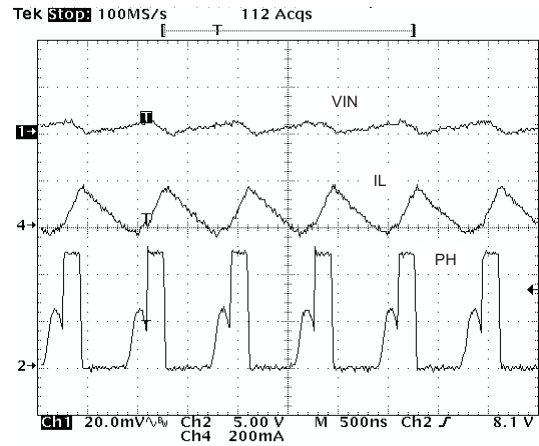
Figure 59. Input Ripple, CCM
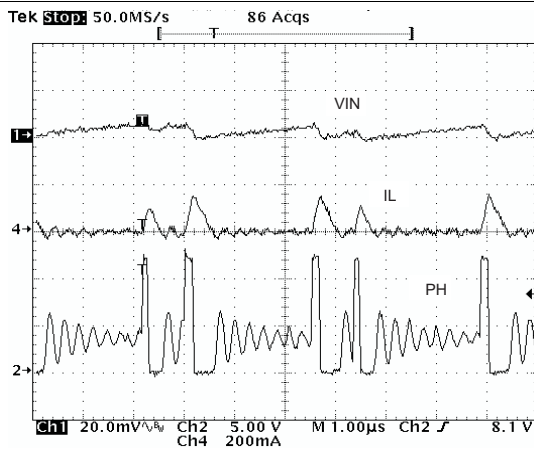

Figure 60. Input Ripple, DCM
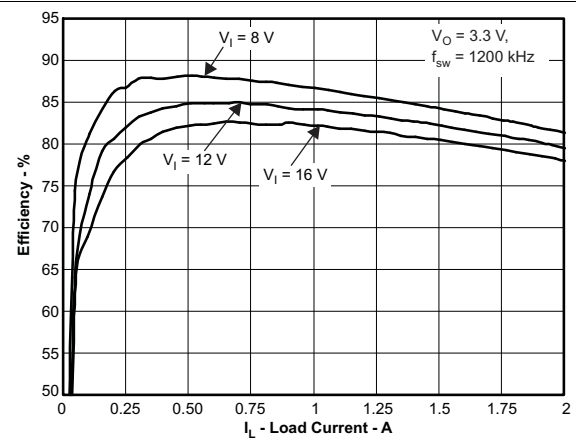

Figure 61. Input Ripple, PSM
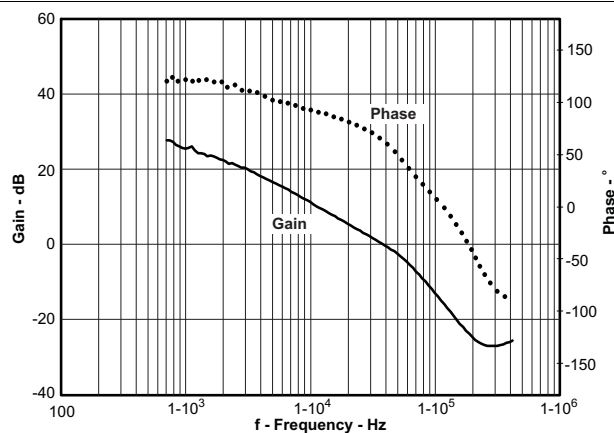

Figure 62. Efficiency vs Load Current

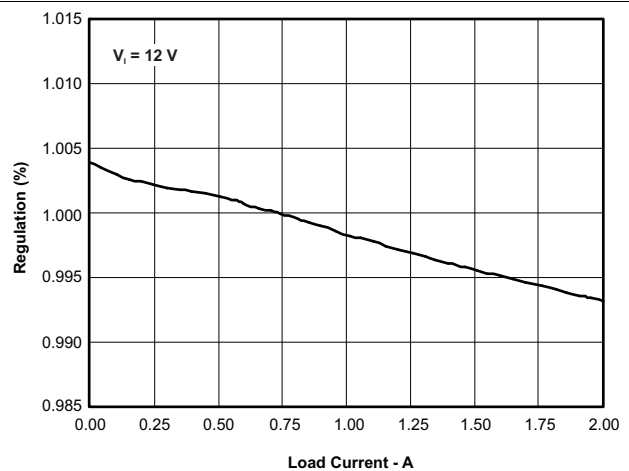Figure 63. Overall Loop Frequency Response

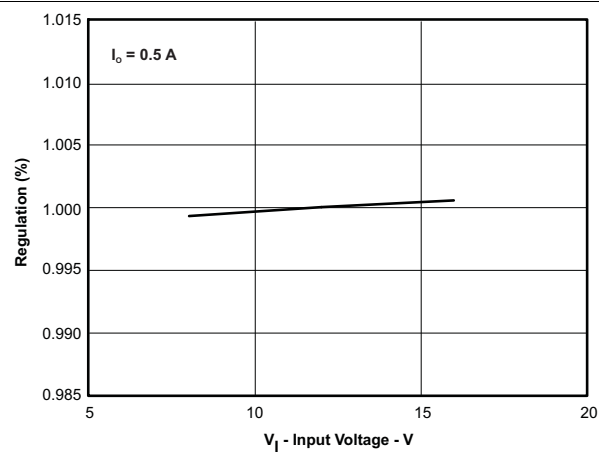Figure 64. Regulation vs Load Current

Figure 65. Regulation vs Input Voltage

9 Power Supply Recommendations

The input decoupling capacitors and bootstrap capacitor must be located as close as possible to the TPS57140-EP. In addition, the voltage set-point resistor divider components must also be kept close to the IC. The voltage divider network ties the output voltage to the point of regulation, the copper VOUT trace past the output capacitors. Ensure that input power supply is clean. TI recommends adding an additional input bulk capacitor depending on the board connection to the input supply.

10 Layout

10.1 Layout Guidelines

Layout is a critical portion of good power-supply design. There are several signals paths that conduct quickly changing currents or voltages that can interact with stray inductance or parasitic capacitance to generate noise or degrade the power-supply performance. To reduce these problems, bypass the VIN pin to ground with a low-ESR ceramic bypass capacitor with X5R or X7R dielectric. Take care to minimize the loop area formed by the bypass capacitor connections, the VIN pin, and the anode of the catch diode. See Figure 66 for a PCB layout example. Tie the GND pin directly to the thermal pad under the IC and the exposed thermal pad.

Connect the thermal pad to any internal PCB ground planes using multiple vias directly under the IC. Route the PH pin to the cathode of the catch diode and to the output inductor. Because the PH connection is the switching node, locate the catch diode and output inductor very close to the PH pins, and minimize the area of the PCB conductor to prevent excessive capacitive coupling. For operation at full-rated load, the top-side ground area must provide adequate heat dissipating area. The RT/CLK pin is sensitive to noise, so locate the RT resistor as close as possible to the IC and route the traces to minimize their lengths. Place the additional external components approximately as shown. It may be possible to obtain acceptable performance with alternate PCB layouts; however, this layout produces good results and can serve as a guideline.
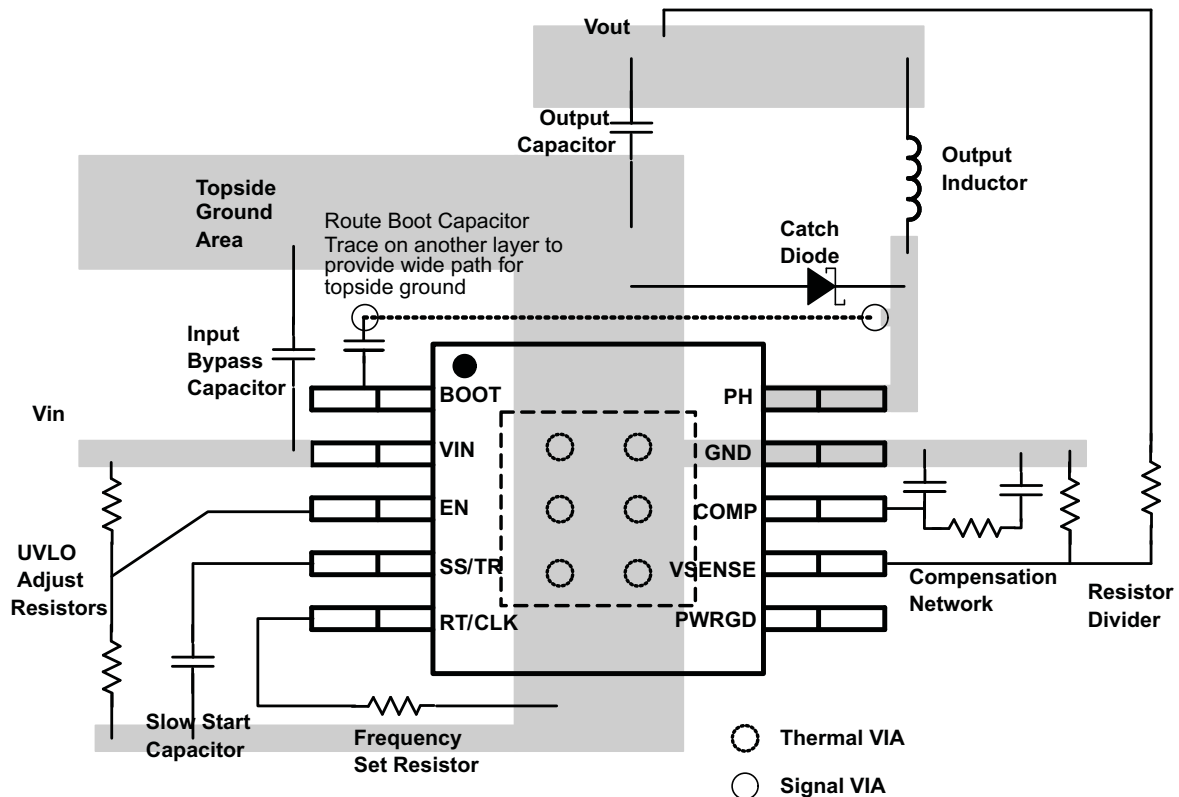
10.2 Layout Example

Figure 66. PCB Layout Example

10.3 Power-Dissipation Estimate

The following formulas show how to estimate the IC power dissipation under CCM operation. Do not use these equations used if the device is working in DCM.

The power dissipation of the IC includes conduction loss (Pcon), switching loss (Psw), gate-drive loss (Pgd), and supply current (Pq).

$$P_{con} = I_o^2 \times R_{DS(on)} \times \frac{V_{out}}{V_{in}} \tag{54}$$

$$P_{SW} = V_{IN}^2 \times f_{SW} \times I_O \times 0.25 \times 10^{-9} \, sec/V \tag{55}$$

$$P_{gd} = V_{IN} \times 3 \times 10^{-9} \, Asec \times f_{SW} \tag{56}$$

$$P_q = 116 \mu A \times V_{IN}$$

where

- I_{OUT} is the output current (A).
- R_{dson} is the on-resistance of the high-side MOSFET (Ω).
- V_{OUT} is the output voltage (V).
- V_{IN} is the input voltage (V).
- f_{SW} is the switching frequency (Hz). (57)

So,

$$P_{TOT} = P_{con} + P_{SW} + P_{gd} + P_q \tag{58}$$

For a given T_A,

$$T_J = T_A + R_{th} \times P_{TOT} \tag{59}$$

For a given $T_{JMAX} = 150°C$

$$T_{A(max)} = T_{J(max)} - R_{th} \times P_{TOT}$$

where

- P_{TOT} is the total device power dissipation (W).
- T_A is the ambient temperature (°C).
- T_J is the junction temperature (°C).
- R_{th} is the thermal resistance of the package (°C/W).
- $T_{J(max)}$ is maximum junction temperature (°C).
- $T_{A(max)}$ is maximum ambient temperature (°C). (60)

There are additional power losses in the regulator circuit, due to the inductor ac and dc losses, the catch diode, and trace resistance, that impact the overall efficiency of the regulator.

11 器件和文档支持

11.1 接收文档更新通知

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

11.2 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.

11.3 商标

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

11.4 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。

11.5 Glossary

SLYZ022 — *TI Glossary*.

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

12 机械、封装和可订购信息

以下页面包含机械、封装和可订购信息。这些信息是指定器件的最新可用数据。数据如有变更，恕不另行通知，且不会对此文档进行修订。如需获取此数据表的浏览器版本，请查阅左侧的导航栏。

PACKAGING INFORMATION

Orderable part number	Status (1)	Material type (2)	Package | Pins	Package qty | Carrier	RoHS (3)	Lead finish/ Ball material (4)	MSL rating/ Peak reflow (5)	Op temp (°C)	Part marking (6)
TPS57140MDRCREP	Active	Production	VSON (DRC) | 10	3000 | LARGE T&R	Yes	NIPDAU	Level-3-260C-168 HR	-55 to 125	5714M
TPS57140MDRCREP.A	Active	Production	VSON (DRC) | 10	3000 | LARGE T&R	Yes	NIPDAU	Level-3-260C-168 HR	-55 to 125	5714M
V62/15604-01YE	Active	Production	VSON (DRC) | 10	3000 | LARGE T&R	Yes	NIPDAU	Level-3-260C-168 HR	-55 to 125	5714M

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

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

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

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

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

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

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

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

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

OTHER QUALIFIED VERSIONS OF TPS57140-EP :

● Automotive : TPS57140-Q1

NOTE: Qualified Version Definitions:

● Automotive - Q100 devices qualified for high-reliability automotive applications targeting zero defects

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
TPS57140MDRCREP	VSON	DRC	10	3000	330.0	12.4	3.3	3.3	1.0	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)
TPS57140MDRCREP	VSON	DRC	10	3000	367.0	367.0	38.0

DRC 10

3 x 3, 0.5 mm pitch

PLASTIC SMALL OUTLINE - NO LEAD

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

4226193/A

Texas
INSTRUMENTS
www.ti.com

A

3.1
2.9

B

PIN 1 INDEX AREA

3.1
2.9

1.0
0.8

C

SEATING PLANE

0.05
0.00

0.08 | C

1.65±0.1

2X (0.5)

4X (0.25)

EXPOSED
THERMAL PAD

(0.2) TYP

5

6

2X
2

11

SYMM
℄

2.4±0.1

1

10

8X | 0.5

PIN 1 ID
(OPTIONAL)

SYMM
℄

10X | 0.30
0.18

⊕ | 0.1 (M) | C | A | B
  | 0.05 (M) | C

10X | 0.5
0.3

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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 optimal thermal and mechanical performance.

DRC0010J

(1.65)
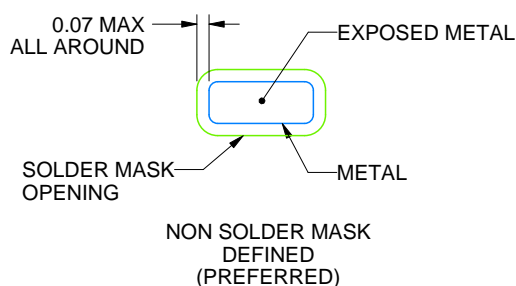
(0.5)

10X (0.6)

1

10X (0.24)

10

SYMM

11

(2.4)

(3.4)

(0.95)

8X (0.5)

5

6

(R0.05) TYP

(∅ 0.2) VIA
TYP

(0.25)

(0.575)

SYMM

(2.8)

LAND PATTERN EXAMPLE
EXPOSED METAL SHOWN
SCALE:20X

0.07 MAX
ALL AROUND

EXPOSED METAL

0.07 MIN
ALL AROUND

EXPOSED METAL

SOLDER MASK
OPENING

METAL

METAL UNDER
SOLDER MASK

SOLDER MASK
OPENING

NON SOLDER MASK
DEFINED
(PREFERRED)

SOLDER MASK
DEFINED
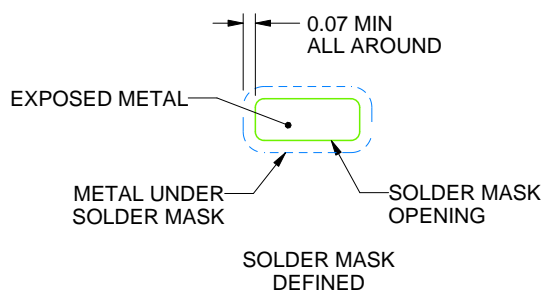
SOLDER MASK DETAILS

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NOTES: (continued)

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

SOLDER PASTE EXAMPLE
BASED ON 0.125 mm THICK STENCIL

EXPOSED PAD 11:
80% PRINTED SOLDER COVERAGE BY AREA
SCALE:25X

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NOTES: (continued)

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