TEXAS INSTRUMENTS

TPS61175

SLVS892C – DECEMBER 2008 – REVISED AUGUST 2014

TPS61175 3-A High-Voltage Boost Converter With Soft-Start and Programmable Switching Frequency

1 Features

- 2.9-V to 18-V Input Voltage Range
- 3-A, 40-V Internal Switch
- High-Efficiency Power Conversion: Up to 93%
- Frequency Set by External Resistor: 200 kHz to 2.2 MHz
- Synchronous External Switching Frequency
- User-Defined Soft-Start into Full Load
- Skip-Switching Cycle for Output Regulation at Light Load
- 14-Pin HTSSOP Package With PowerPad™

2 Applications

- 5-V to 12-V, 24-V Power Conversion
- Supports SEPIC and Flyback Topologies
- ADSL Modems
- TV Tuners

3 Description

The TPS61175 device is a monolithic switching regulator with integrated 3-A, 40-V power switch. The device can be configured in several standard switching-regulator topologies, including boost, SEPIC, and flyback. The device has a wide input voltage range to support application with input voltage from multicell batteries or regulated 5-V, 12-V power rails.

The TPS61175 regulates the output voltage with current mode PWM (pulse width modulation) control. The switching frequency of PWM is set by either an external resistor or an external clock signal. The user can program the switching frequency from 200 kHz to 2.2 MHz.
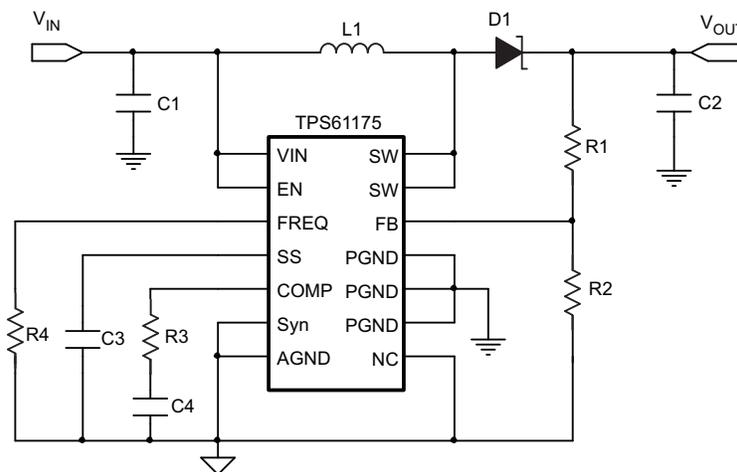
The device features a programmable soft-start function to limit inrush current during start-up, and has other built-in protection features, such as pulse-by-pulse overcurrent limit and thermal shutdown. The TPS61175 is available in 14-pin HTSSOP package with PowerPad.

Device Information[(1)]

PART NUMBER	PACKAGE	BODY SIZE (NOM)
TPS61175	HTSSOP (14)	5.00 mm x 4.40 mm

(1) For all available packages, see the orderable addendum at the end of the datasheet.

4 Simplified Schematic

Table of Contents

5 Revision History

Changes from Revision B (February 2012) to Revision C   **Page**

• Added *Handling Rating* table, *Feature Description* section, *Device Functional Modes*, *Application and Implementation* section, *Power Supply Recommendations* section, *Layout* section, *Device and Documentation Support* section, and *Mechanical, Packaging, and Orderable Information* section 1

Changes from Revision A (October 2010) to Revision B   **Page**

• Replaced the Dissipation Ratings Table with the Thermal Information Table 4
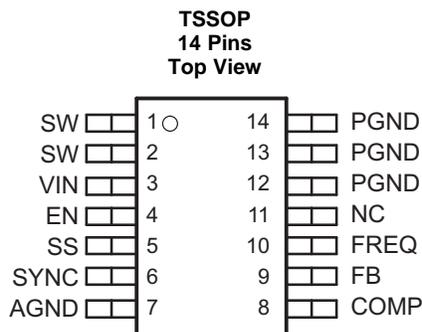
Changes from Original (December 2008) to Revision A   **Page**

• Changed the Ordering Information table - Part Number From: TPS61175 To: TPS61175PWP; Removed the Package Marking column 1

Product Folder Links: *TPS61175*

6 Pin Configuration and Functions

TSSOP
14 Pins
Top View

```
        SW  ┌────┐ 1 ○   14 ┌────┐  PGND
        SW  └────┘ 2      13 └────┘  PGND
       VIN  ┌────┐ 3      12 ┌────┐  PGND
        EN  └────┘ 4      11 └────┘  NC
        SS  ┌────┐ 5      10 ┌────┐  FREQ
      SYNC  └────┘ 6       9 └────┘  FB
      AGND  ┌────┐ 7       8 ┌────┐  COMP
```

Pin Functions

PIN		I/O	DESCRIPTION
NAME	NO.		
AGND	7	I	Signal ground of the IC
COMP	8	O	Output of the internal transconductance error amplifier. An external RC network is connected to this pin to compensate the regulator.
EN	4	I	Enable pin. When the voltage of this pin falls below the enable threshold for more than 10ms, the IC turns off.
FB	9	I	Feedback pin for positive voltage regulation. Connect to the center tap of a resistor divider to program the output voltage.
FREQ	10	O	Switch frequency program pin. An external resistor is connected to this pin to set switch frequency. See application section for information on how to size the FREQ resistor.
NC	11	I	Reserved pin. Must connect this pin to ground.
PGND	12,13,14	I	Power ground of the IC. It is connected to the source of the PWM switch.
SS	5	O	Soft start programming pin. A capacitor between the SS pin and GND pin programs soft start timing. See application section for information on how to size the SS capacitor.
SW	1,2	I	This is the switching node of the IC. Connect SW to the switched side of the inductor.
SYNC	6	I	Switch frequency synchronous pin. Customers can use an external signal to set the IC switch frequency between 200-kHz and 2.2-MHz. If not used, this pin should be tied to AGND as short as possible to avoid noise coupling.
Thermal Pad			The thermal pad should be soldered to the analog ground. If possible, use thermal via to connect to top and internal ground plane layers for ideal power dissipation.
VIN	3	I	The input supply pin for the IC. Connect VIN to a supply voltage between 2.9V and 18V. It is acceptable for the voltage on the pin to be different from the boost power stage input for applications requiring voltage beyond VIN range.

7 Specifications

7.1 Absolute Maximum Ratings[1]

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

	MIN	MAX	UNIT
Supply Voltages on pin VIN[2]	−0.3	20	V
Voltages on pins EN[2]	−0.3	20	V
Voltage on pin FB, FREQ and COMP[2]	−0.3	3	V
Voltage on pin SYNC, SS[2]	−0.3	7	V
Voltage on pin SW[2]	−0.3	40	V
Continuous Power Dissipation	See *Thermal Information*		
Operating Junction Temperature Range	−40	150	°C

(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) All voltage values are with respect to network ground terminal.

7.2 Handling Ratings

			MIN	MAX	UNIT
T_{stg}	Storage temperature range		−65	+150	°C
$V_{(ESD)}$	Electrostatic discharge	Human body model (HBM), per ANSI/ESDA/JEDEC JS-001, all pins[1]	-2000	2000	V
		Charged device model (CDM), per JEDEC specification JESD22-C101, all pins[2]	-500	500	

(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.

7.3 Recommended Operating Conditions

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

		MIN	NOM	MAX	UNIT
V_{IN}	Input voltage range	2.9		18	V
V_O	Output voltage range	V_{IN}		38	V
L	Inductor[1]	4.7		47	µH
f_{SW}	Switching frequency	200		2200	kHz
C_I	Input Capacitor	4.7			µF
C_O	Output Capacitor	4.7			µF
V_{SYN}	External Switching Frequency Logic			5	V
T_A	Operating ambient temperature	−40		85	°C
T_J	Operating junction temperature	−40		125	°C

(1) The inductance value depends on the switching frequency and end application. While larger values may be used, values between 4.7-µH and 47-µH have been successfully tested in various applications. Refer to *Selecting the Inductor* for detail.

7.4 Thermal Information

THERMAL METRIC[1]	TPS61175 PWP 14 PINS	UNIT
θ_{JA} Junction-to-ambient thermal resistance	45.2	°C/W
θ_{JCtop} Junction-to-case (top) thermal resistance	34.9	
θ_{JB} Junction-to-board thermal resistance	30.1	
ψ_{JT} Junction-to-top characterization parameter	1.5	
ψ_{JB} Junction-to-board characterization parameter	29.9	
θ_{JCbot} Junction-to-case (bottom) thermal resistance	5.8	

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

7.5 Electrical Characteristics

FSW = 1.2 MHz (R_{freq} = 80 kΩ), V_{IN} = 3.6V, T_A = –40°C to 85°C, typical values are at T_A = 25°C (unless otherwise noted)

	PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
SUPPLY CURRENT						
V_{IN}	Input voltage range		2.9		18	V
I_Q	Operating quiescent current into Vin	Device PWM switching without load			3.5	mA
I_{SD}	Shutdown current	EN=GND			1.5	µA
V_{UVLO}	Under-voltage lockout threshold			2.5	2.7	V
V_{hys}	Under-voltage lockout hysteresis			130		mV
ENABLE AND REFERENCE CONTROL						
V_{enh}	EN logic high voltage	Vin = 2.9 V to 18 V	1.2			V
V_{enl}	EN logic low voltage	Vin = 2.9 V to 18 V			0.4	V
V_{SYNh}	SYN logic high voltage		1.2			V
V_{SYNl}	SYN logic low voltage				0.4	V
R_{en}	EN pull down resistor		400	800	1600	kΩ
T_{off}	Shutdown delay, SS discharge	EN high to low	10			ms
VOLTAGE AND CURRENT CONTROL						
V_{REF}	Voltage feedback regulation voltage		1.204	1.229	1.254	V
I_{FB}	Voltage feedback input bias current				200	nA
I_{sink}	Comp pin sink current	V_{FB} = V_{REF} + 200 mV, V_{COMP} = 1 V		50		µA
I_{source}	Comp pin source current	V_{FB} = V_{REF} –200 mV, V_{COMP} = 1 V		130		µA
V_{CCLP}	Comp pin Clamp Voltage	High Clamp, V_{FB} = 1 V Low Clamp, V_{FB} = 1.5 V		3 0.75		V
V_{CTH}	Comp pin threshold	Duty cycle = 0%		0.95		V
G_{ea}	Error amplifier transconductance		240	340	440	µmho
R_{ea}	Error amplifier output resistance			10		MΩ
f_{ea}	Error amplifier crossover frequency			500		KHz
FREQUENCY						
f_S	Oscillator frequency	Rfreq = 480 kΩ	0.16	0.21	0.26	MHz
		Rfreq = 80 kΩ	1.0	1.2	1.4	
		Rfreq = 40 kΩ	1.76	2.2	2.64	
D_{max}	Maximum duty cycle	V_{FB} = 1.0 V, Rfreq = 80 kΩ	89%	93%		
V_{FREQ}	FREQ pin voltage			1.229		V
T_{min_on}	Minimum on pulse width	Rfreq = 80 kΩ		60		ns

Electrical Characteristics (continued)

FSW = 1.2 MHz (R_{freq} = 80 kΩ), V_{IN} = 3.6V, T_A = –40°C to 85°C, typical values are at T_A = 25°C (unless otherwise noted)

	PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
POWER SWITCH						
$R_{DS(ON)}$	N-channel MOSFET on-resistance	V_{IN} = V_{GS} = 3.6 V V_{IN} = V_{GS} = 3.0 V		0.13 0.13	0.25 0.3	Ω
I_{LN_NFET}	N-channel leakage current	V_{DS} = 40 V, T_A = 25°C			1	µA
OC, OVP AND SS						
I_{LIM}	N-Channel MOSFET current limit	D = D_{max}	3	3.8	5	A
I_{SS}	Soft start bias current	Vss = 0 V		6		µA
THERMAL SHUTDOWN						
$T_{shutdown}$	Thermal shutdown threshold			160		°C
$T_{hysteresis}$	Thermal shutdown threshold hysteresis			15		°C

7.6 Typical Characteristics

Circuit of Figure 1; L1 = D104C2-10µH; D1 = SS3P6L-E3/86A, R4 = 80kΩ, R3 = 10kΩ, C4 = 22nF, C2 = 10µF; V_{IN} = 5V, V_{OUT} = 24V, I_{OUT} = 200mA (unless otherwise noted)

Table 1. Table Of Graphs

		FIGURE
Efficiency	VIN = 5V, Vout = 12V, 24V, 35V	Figure 1
Efficiency	VIN = 5V, 12V; Vout = 24V	Figure 2
Error amplifier transconductance	vs Temperature	Figure 3
Switch current limit	vs Temperature	Figure 4
Switch current limit	vs Duty cycle	Figure 5
FB accuracy	vs Temperature	Figure 6
Line transient response	VIN = 4.5 V to 5 V	Figure 12
Load transient response	IOUT = 100 mA to 300 mA; refer to 'compensating the control loop' for optimization	Figure 13
PWM Operation		Figure 14
Pulse skipping	No load	Figure 15
Start-up	C3 = 47 nF	Figure 16

Figure 1. Efficiency vs Output Current

Figure 2. Efficiency vs Output Current
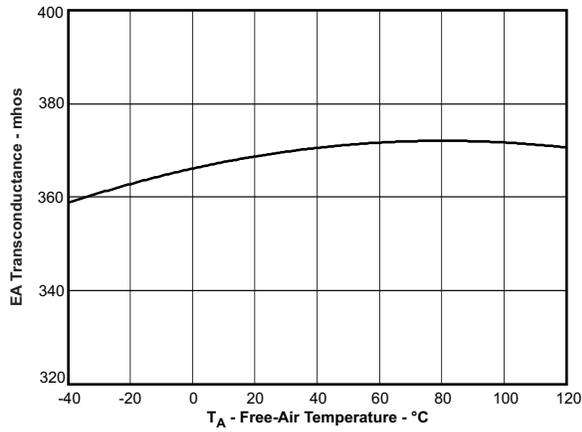
Product Folder Links: *TPS61175*
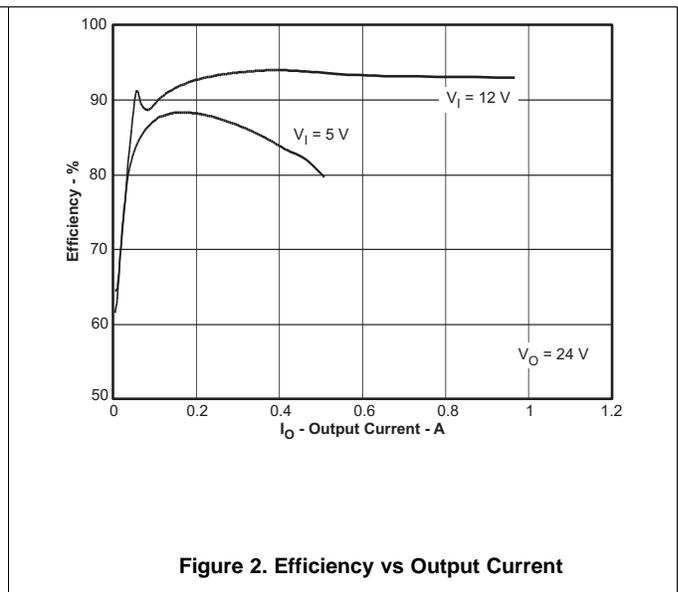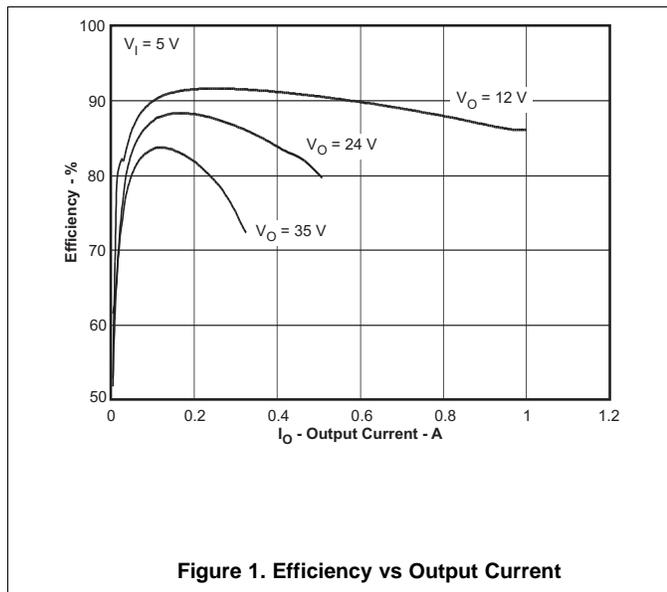
Figure 3. Error Amplifier Transconductance vs Free-Air Temperature
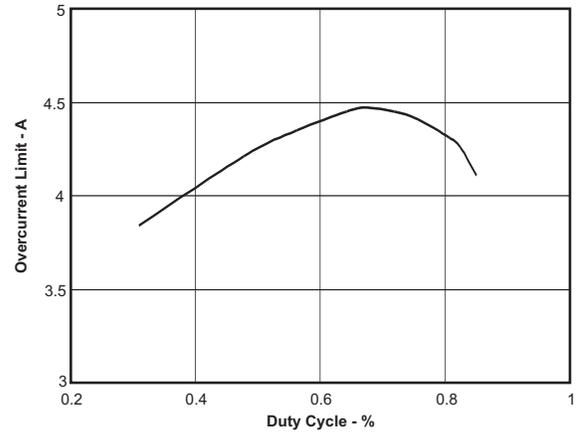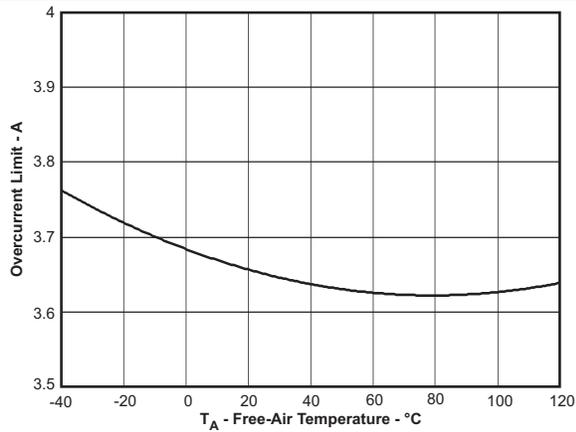
Figure 4. Overcurrent Limit vs Duty Cycle

Figure 5. Overcurrent Limit vs Free-Air Temperature
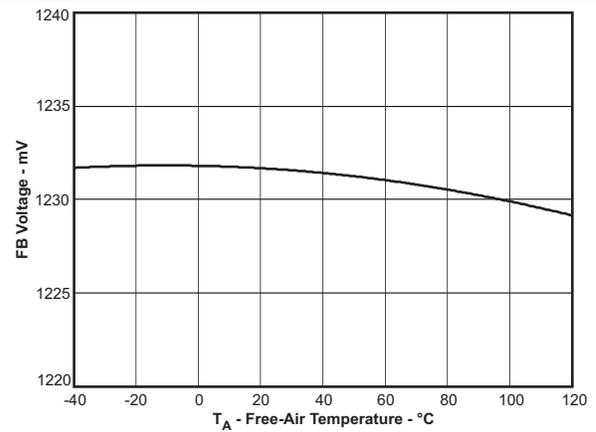
Figure 6. FB Voltage vs Free-Air Temperature

8 Detailed Description

8.1 Overview

The TPS61175 integrates a 40-V low side switch FET for up to 38-V output. The device regulates the output with current mode PWM (pulse width modulation) control. The PWM control circuitry turns on the switch at the beginning of each switching cycle. The input voltage is applied across the inductor and stores the energy as inductor current ramps up. During this portion of the switching cycle, the load current is provided by the output capacitor. When the inductor current rises to the threshold set by the error amplifier output, the power switch turns off and the external Schottky diode is forward biased. The inductor transfers stored energy to replenish the output capacitor and supply the load current. This operation repeats each every switching cycle. As shown in the block diagram, the duty cycle of the converter is determined by the PWM control comparator which compares the error amplifier output and the current signal. The switching frequency is programmed by the external resistor or synchronized to an external clock signal.

A ramp signal from the oscillator is added to the current ramp to provide slope compensation. Slope compensation is necessary to avoid subharmonic oscillation that is intrinsic to the current mode control at duty cycle higher than 50%. If the inductor value is lower than 4.7µH, the slope compensation may not be adequate.

The feedback loop regulates the FB pin to a reference voltage through a transconductance error amplifier. The output of the error amplifier is connected to the COMP pin. An external RC compensation network is connected to the COMP pin to optimize the feedback loop for stability and transient response.

8.2 Functional Block Diagram
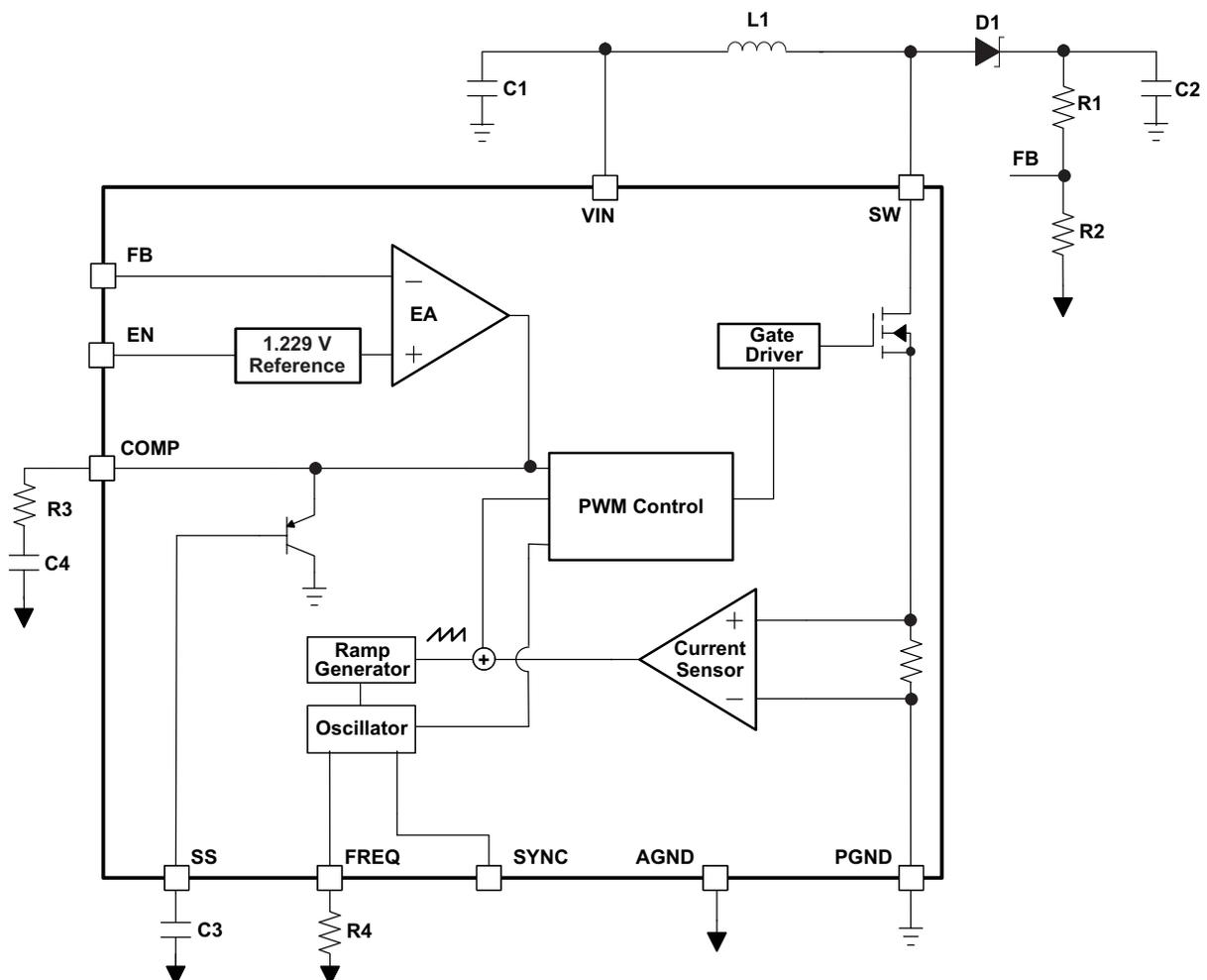
8.3 Feature Description

8.3.1 Switching Frequency

The switch frequency is set by a resistor (R4) connected to the FREQ pin of the TPS61175. Do not leave this pin open. A resistor must always be connected for proper operation. See Table 2 and Figure 7 for resistor values and corresponding frequencies.
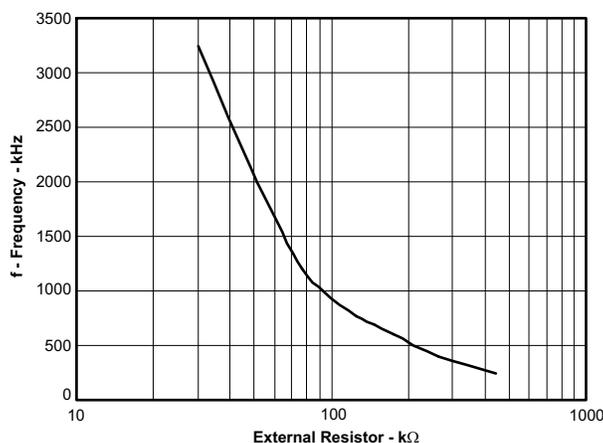
Table 2. Switching Frequency vs External Resistor

R4 (kΩ)	f_{SW} (kHz)
443	240
256	400
176	600
80	1200
51	2000

Figure 7. Switching Frequency vs External Resistor

Alternatively, the TPS61175 switching frequency will synchronize to an external clock signal that is applied to the SYNC pin. The logic level of the external clock is shown in the specification table. The duty cycle of the clock is recommended in the range of 10% to 90%. The resistor also must be connected to the FREQ pin when IC is switching by the external clock. The external clock frequency must be within ±20% of the corresponding frequency set by the resistor. For example, if the corresponding frequency as set by a resistor on the FREQ pin is 1.2-MHz, the external clock signal should be in the range of 0.96-MHz to 1.44-MHz.

If the external clock signal is higher than the frequency per the resistor on the FREQ pin, the maximum duty cycle specification (D_{MAX}) should be lowered by 2%. For instance, if the resistor set value is 2.5MHz, and the external clock is 3MHz, D_{MAX} is 87% instead of 89%.

8.3.2 Soft Start

The TPS61175 has a built-in soft start circuit which significantly reduces the start-up current spike and output voltage overshoot. When the IC is enabled, an internal bias current (6-µA typically) charges a capacitor (C3) on the SS pin. The voltage at the capacitor clamps the output of the internal error amplifier that determines the duty cycle of PWM control, thereby the input inrush current is eliminated. Once the capacitor reaches 1.8-V, the soft start cycle is completed and the soft start voltage no longer clamps the error amplifier output. Refer to Figure 7 for the soft start waveform. See Table 3 for C3 and corresponding soft start time. A 47-nF capacitor eliminates the output overshoot and reduces the peak inductor current for most applications.

Table 3. Soft Start Time vs C3

Vin (V)	Vout (V)	Load (A)	Cout (µF)	f_{SW} (MHz)	C3 (nF)	t_{SS}(ms)	Overshot (mV)
5	24	0.4	10	1.2	47	4	none
					10	0.8	210
12	35	0.6	10	2	100	6.5	none
					10	0.4	300

When the EN is pulled low for 10-ms, the IC enters shutdown and the SS capacitor discharges through a 5kΩ resistor for the next soft start.

8.3.3 Overcurrent Protection

The TPS61175 has a cycle-by-cycle overcurrent limit protection that turns off the power switch once the inductor current reaches the overcurrent limit threshold. The PWM circuitry resets itself at the beginning of the next switch cycle. During an overcurrent event, the output voltage begins to droop as a function of the load on the output. When the FB voltage drops lower than 0.9-V, the switching frequency is automatically reduced to 1/4 of the set value. The switching frequency does not reset until the overcurrent condition is removed. This feature is disabled during soft start.

8.3.4 Enable and Thermal Shutdown

The TPS61175 enters shutdown when the EN voltage is less than 0.4-V for more than 10-ms. In shutdown, the input supply current for the device is less than 1.5-µA (max). The EN pin has an internal 800-kΩ pull down resistor to disable the device when it is floating.

An internal thermal shutdown turns off the device when the typical junction temperature of 160°C is exceeded. The IC restarts when the junction temperature drops by 15°C.

8.3.5 Under Voltage Lockout (UVLO)

An under voltage lockout circuit prevents mis-operation of the device at input voltages below 2.5-V (typical). When the input voltage is below the under voltage threshold, the device remains off and the internal switch FET is turned off. The under voltage lockout threshold is set below minimum operating voltage of 2.9V to avoid any transient VIN dip triggering the UVLO and causing the device to reset. For the input voltages between UVLO threshold and 2.9V, the device attempts to operate, but the specifications are not ensured.

8.4 Device Functional Modes

8.4.1 Minimum ON Time and Pulse Skipping

Once the PWM switch is turned on, the TPS61175 has minimum ON pulse width of 60-ns. This sets the limit of the minimum duty cycle of the PWM switch, and it is independent of the set switching frequency. When operating conditions result in the TPS61175 having a minimum ON pulse width less than 60-ns, the IC enters pulse-skipping mode. In this mode, the device keeps the power switch off for several switching cycles to keep the output voltage from rising above the regulated voltage. This operation typically occurs in light load condition when the PWM operates in discontinuous mode. Pulse skipping increases the output voltage ripple, see Figure 15.

If the switching frequency is above 1.2 MHz, the pulse-skipping operation may not function. The TPS61175 will always run in PWM mode with minimum ON pulse width. To keep the output voltage in regulation, a minimum load is required. The minimum load is related to the input voltage, output voltage, switching frequency, external inductor value and the maximum value of the minimum ON pulse width. Use Equation 1 and Equation 2 to calculate the required minimum load at the worst case. The maximum t_{min_ON} could be estimated to 80 ns. CSW is the total parasite capacitance at the switching node SW pin. It could be estimated to 100 pF.

$$I_{(min\_load)} = \frac{1}{2} \times \frac{\left(V_{IN} \times t_{min\_ON} + (V_{OUT} + V_D - V_{IN}) \times \sqrt{L \times C_{SW}}\right)^2 \times f_{SW}}{L \times (V_{OUT} + V_D - V_{IN})} \quad \text{When } V_{OUT} + V_D - V_{IN} < V_{IN}$$

(1)

$$I_{(min\_load)} = \frac{1}{2} \times \frac{\left(V_{IN} \times t_{min\_ON} + V_{IN} \times \sqrt{L \times C_{SW}}\right)^2 \times f_{SW}}{L \times (V_{OUT} + V_D - V_{IN})} \quad \text{When } V_{OUT} + V_D - V_{IN} > V_{IN}$$

(2)

Product Folder Links: *TPS61175*

9 Application and Implementation

9.1 Application Information

The following section provides a step-by-step design approach for configuring the TPS61175 as a voltage regulating boost converter, as shown in Figure 8. When configured as SEPIC or flyback converter, a different design approach is required.

9.2 Typical Application
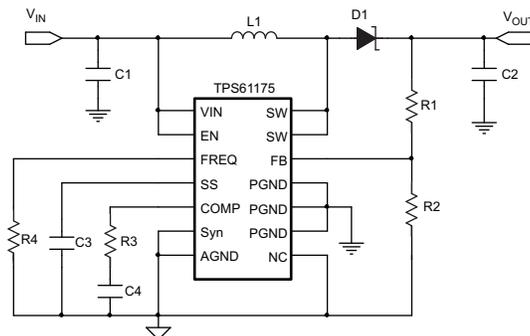
Figure 8. Boost Converter Configuration

9.2.1 Design Requirements

Table 4. Design Parameters

PARAMETERS	VALUES
Input voltage	5 V
Output voltage	24 V
Operating frequency	1.2 MHz

9.2.2 Detailed Design Procedure

9.2.2.1 Determining the Duty Cycle

The TPS61175 has a maximum worst case duty cycle of 89% and a minimum on time of 60 ns. These two constraints place limitations on the operating frequency that can be used for a given input to output conversion ratio. The duty cycle at which the converter operates is dependent on the mode in which the converter is running. If the converter is running in discontinuous conduction mode (DCM), where the inductor current ramps to zero at the end of each cycle, the duty cycle varies with changes to the load much more than it does when running in continuous conduction mode (CCM). In continuous conduction mode, where the inductor maintains a dc current, the duty cycle is related primarily to the input and output voltages as computed in Equation 3:

$$D = \frac{V_{OUT} + V_D - V_{IN}}{V_{OUT} + V_D}$$

(3)

In discontinuous mode the duty cycle is a function of the load, input and output voltages, inductance and switching frequency as computed in Equation 4:

$$D = \frac{2 \times (V_{OUT} + V_D) \times I_{OUT} \times L \times f_{SW}}{V_{IN}^2}$$

(4)

All converters using a diode as the freewheeling or catch component have a load current level at which they transition from discontinuous conduction to continuous conduction. This is the point where the inductor current just falls to zero. At higher load currents, the inductor current does not fall to zero but remains flowing in a positive direction and assumes a trapezoidal wave shape as opposed to a triangular wave shape. This load boundary between discontinuous conduction and continuous conduction can be found for a set of converter parameters in Equation 5:

$$I_{OUT(crit)} = \frac{(V_{OUT} + V_D - V_{IN}) \times V_{IN}^2}{2 \times (V_{OUT} + V_D)^2 \times f_{SW} \times L}$$

(5)

For loads higher than the result of the equation above, the duty cycle is given by Equation 3 and for loads less that the results of Equation 4, the duty cycle is given Equation 5. For Equation 3 through Equation 5, the variable definitions are as follows:

- V_{OUT} is the output voltage of the converter in V
- V_D is the forward conduction voltage drop across the rectifier or catch diode in V
- V_{IN} is the input voltage to the converter in V
- I_{OUT} is the output current of the converter in A
- L is the inductor value in H
- f_{SW} is the switching frequency in Hz

Unless otherwise stated, the design equations that follow assume that the converter is running in continuous mode.

9.2.2.2 Selecting the Inductor

The selection of the inductor affects steady state operation as well as transient behavior and loop stability. These factors make it the most important component in power regulator design. There are three important inductor specifications, inductor value, DC resistance and saturation current. Considering inductor value alone is not enough.

Inductor values can have ±20% tolerance with no current bias. When the inductor current approaches saturation level, its inductance can fall to some percentage of its 0-A value depending on how the inductor vendor defines saturation current. For CCM operation, the rule of thumb is to choose the inductor so that its inductor ripple current (ΔI_L) is no more than a certain percentage (RPL% = 20–40%) of its average DC value ($I_{IN(AVG)} = I_{L(AVG)}$).

$$\Delta I_L = \frac{V_{IN} \times D}{L \times f_{SW}} = \frac{(V_{OUT} + V_D - V_{IN}) \times (1 - D)}{L \times f_{SW}} = \frac{1}{\left[L \times f_{SW} \times \left(\frac{1}{V_{OUT} + V_D - V_{IN}} + \frac{1}{V_{IN}} \right) \right]}$$

$$\leq RPL\% \times \frac{P_{OUT}}{V_{IN} \times \eta_{est}}$$

(6)

Rearranging and solving for L gives:

$$L \leq \frac{\eta_{est} \times V_{IN}}{\left[f_{SW} \left(\frac{1}{V_{OUT} + V_D - V_{IN}} + \frac{1}{V_{IN}} \right) \right] \times RPL\% \ P_{OUT}}$$

(7)

Choosing the inductor ripple current to closer to 20% of the average inductor current results in a larger inductance value, maximizes the converter's potential output current and minimizes EMI. Choosing the inductor ripple current closer to 40% of $I_{L(AVG)}$ results in a smaller inductance value, and a physically smaller inductor, improves transient response but results in potentially higher EMI and lower efficiency if the DCR of the smaller packaged inductor is significantly higher. Using an inductor with a smaller inductance value than computed above may result in the converter operating in DCM. This reduces the boost converter's maximum output current, causes larger input voltage and output ripple and typically reduces efficiency. Table 5 lists the recommended inductor for the TPS61175.

Table 5. Recommended Inductors for TPS61175

PART NUMBER	L (µH)	DCR MAX (mΩ)	SATURATION CURRENT (A)	SIZE (L × W × H mm)	VENDOR
D104C2	10	44	3.6	10.4x10.4x4.8	TOKO
VLF10040	15	42	3.1	10.0x9.7x4.0	TDK
CDRH105RNP	22	61	2.9	10.5x10.3x5.1	Sumida
MSS1038	15	50	3.8	10.0x10.2x3.8	Coilcraft

Product Folder Links: *TPS61175*

The device has built-in slope compensation to avoid subharmonic oscillation associated with current mode control. If the inductor value is lower than 4.7μH, the slope compensation may not be adequate, and the loop can be unstable. Applications requiring inductors above 47μH have not been evaluated. Therefore, the user is responsible for verifying operation if they select an inductor that is outside the 4.7μH–47μH recommended range.

9.2.2.3 Computing the Maximum Output Current

The over-current limit for the integrated power FET limits the maximum input current and thus the maximum input power for a given input voltage. Maximum output power is less than maximum input power due to power conversion losses. Therefore, the current limit setting, input voltage, output voltage and efficiency can all change the maximum current output ($I_{OUT(MAX)}$). The current limit clamps the peak inductor current, therefore the ripple has to be subtracted to derive maximum DC current.

$$I_{OUT(max)} = \frac{V_{IN(min)} \times \eta_{est}}{V_{OUT}} = \frac{V_{IN(min)} \times I_{LIM} \times (1 - \%RPL/2) \times \eta_{est}}{V_{OUT}}$$

(8)

where

    I_{LIM} = over current limit

    η_{est}= efficiency estimate based on similar applications or computed above

For instance, when VIN = 12-V is boosted to VOUT = 24-V, the inductor is 10-uH, the Schottky forward voltage is 0.4-V and the switching frequency is 1.2-MHz; then the maximum output current is 1.2-A in typical condition, assuming 90% efficiency and a %RPL = 20%.

9.2.2.4 Setting Output Voltage

To set the output voltage in either DCM or CCM, select the values of R1 and R2 according to Equation 9:

$$\text{Vout} = 1.229 \text{ V} \times \left(\frac{R1}{R2} + 1 \right)$$

$$R1 = R2 \times \left(\frac{\text{Vout}}{1.229V} - 1 \right)$$

(9)

Considering the leakage current through the resistor divider and noise decoupling into FB pin, an optimum value for R2 is around 10k. The output voltage tolerance depends on the V_{FB} accuracy and the tolerance of R1 and R2.

9.2.2.5 Setting the Switching Frequency

Choose the appropriate resistor from the resistance versus frequency table Table 2 or graph Figure 7. A resistor must be placed from the FREQ pin to ground, even if an external oscillation is applied for synchronization.

Increasing switching frequency reduces the value of external capacitors and inductors, but also reduces the power conversion efficiency. The user should set the frequency for the minimum tolerable efficiency.

9.2.2.6 Setting the Soft Start Time

Choose the appropriate capacitor from the soft start table, Table 3. Increasing the soft start time reduces the overshoot during start-up.

9.2.2.7 Selecting the Schottky Diode

The high switching frequency of the TPS61175 demands a high-speed rectification for optimum efficiency. Ensure that the diode's average and peak current rating exceed the average output current and peak inductor current. In addition, the diode's reverse breakdown voltage must exceed the switch FET rating voltage of 40V. So, the VISHAY SS3P6L-E3/86A is recommended for TPS61175. The power dissipation of the diode's package must be larger than $I_{OUT(max)}$ x V_D.

9.2.2.8 Selecting the Input and Output Capacitors

The output capacitor is mainly selected to meet the requirements for the output ripple and load transient. Then the loop is compensated for the output capacitor selected. The output ripple voltage is related to the capacitor's capacitance and its equivalent series resistance (ESR). Assuming a capacitor with zero ESR, the minimum capacitance needed for a given ripple can be calculated using Equation 10:

$$C_{out} = \frac{(V_{OUT} - V_{IN})I_{out}}{V_{OUT} \times Fs \times V_{ripple}}$$

(10)

where V_{ripple} = peak to peak output ripple. The additional output ripple component caused by ESR is calculated using:

$$V_{ripple\_ESR} = I \times R_{ESR}$$

Due to its low ESR, V_{ripple_ESR} can be neglected for ceramic capacitors, but must be considered if tantalum or electrolytic capacitors are used.

The minimum ceramic output capacitance needed to meet a load transient requirement can be estimated using Equation 11:

$$C_{OUT} = \frac{\Delta I_{TRAN}}{2 \times \pi \times f_{LOOP\text{-}BW} \times \Delta V_{TRAN}}$$

(11)

Where

ΔI_{TRAN} is the transient load current step

ΔV_{TRAN} is the allowed voltage dip for the load current step

$f_{LOOP\text{-}BW}$ is the control loop bandwidth (i.e., the frequency where the control loop gain crosses zero).

Care must be taken when evaluating a ceramic capacitor's derating under dc bias, aging and AC signal. For example, larger form factor capacitors (in 1206 size) have their self resonant frequencies in the range of the switching frequency. So the effective capacitance is significantly lower. The DC bias can also significantly reduce capacitance. Ceramic capacitors can loss as much as 50% of its capacitance at its rated voltage. Therefore, one must add margin on the voltage rating to ensure adequate capacitance at the required output voltage.

For a typical boost converter implementation, at least 4.7µF of ceramic input and output capacitance is recommended. Additional input and output capacitance may be required to meet ripple and/or transient requirements.

The popular vendors for high value ceramic capacitors are:

TDK (http://www.component.tdk.com/components.php)

Murata (http://www.murata.com/cap/index.html)

9.2.2.9 Compensating the Small Signal Control Loop

All continuous mode boost converters have a right half plane zero (f_{RHPZ}) due to the inductor being removed from the output during charging. In a traditional voltage mode controlled boost converter, the inductor and output capacitor form a small signal double pole. For a negative feedback system to be stable, the fed back signal must have a gain less than 1 before having 180 degrees of phase shift. With its double pole and RHPZ all providing phase shift, voltage mode boost converters are a challenge to compensate. In a converter with current mode control, there are essentially two loops, an inner current feedback loop created by the inductor current information sensed across R_{SENSE} (40mΩ) and the output voltage feedback loop. The inner current loop allows the switch, inductor and modulator to be lumped together into a small signal variable current source controlled by the error amplifier, as shown in Figure 9.

Product Folder Links: *TPS61175*
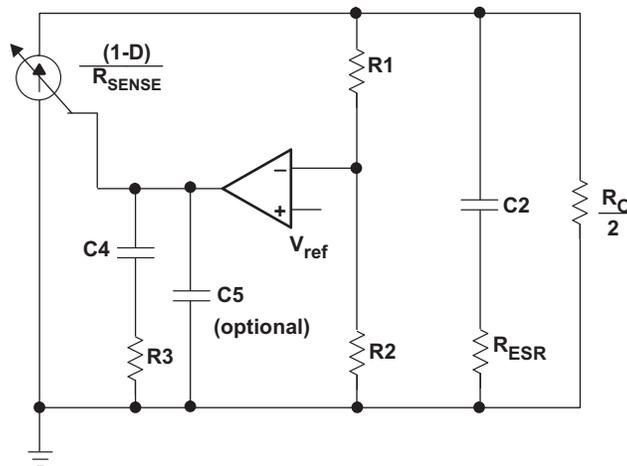
Figure 9. Small Signal Model of a Current Mode Boost in CCM

The new power stage, including the slope compensation, small signal model becomes:

$$G_{PS}(s) = \frac{R_{OUT} \times (1-D)}{2 \times R_{SENSE}} \times \frac{\left(1 + \frac{s}{2 \times \pi \times f_{ESR}}\right)\left(1 - \frac{s}{2 \times \pi \times f_{RHPZ}}\right)}{1 + \frac{s}{2 \times \pi \times f_{P}}} \times He(s)$$

(12)

Where

$$f_P = \frac{2}{2\pi \times R_O \times C2}$$

(13)

$$f_{ESR} \approx \frac{1}{2\pi \times R_{ESR} \times C2}$$

(14)

$$f_{RHPZ} = \frac{R_O}{2\pi \times L} \times \left(\frac{V_{IN}}{V_{OUT}}\right)^2$$

(15)

And

$$He(s) = \frac{1}{1 + \frac{s \times \left[\left(1 + \frac{Se}{Sn}\right) \times (1-D) - 0.5\right]}{f_{SW}} + \frac{s^2}{(\pi \times f_{SW})^2}}$$

(16)

He(s) models the inductor current sampling effect as well as the slope compensation effect on the small signal response. Note that if Sn > Se, e.g., when L is smaller than recommended, the converter operates as a voltage mode converter and the above model no longer holds.

The slope compensation in TPS61175 is shown as follows:

$$Sn = \frac{V_{OUT} + V_D - V_{IN}}{L} \times R_{SENSE}$$

(17)

$$Se = \frac{0.32 \text{ V} / R4}{16 \times (1-D) \times 6pF} + \frac{0.5 \text{ } \mu A}{6 \text{ pF}}$$

Where R4 is the frequency setting resistor

(18)

Figure 10 shows a bode plot of a typical CCM boost converter power stage.
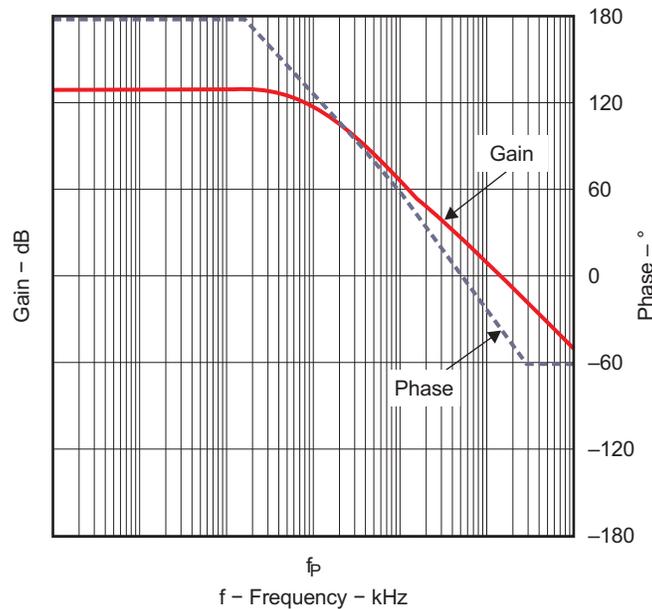
Figure 10. Bode Plot of Power Stage Gain and Phase

The TPS61175 COMP pin is the output of the internal trans-conductance amplifier. Equation 19 shows the equation for feedback resistor network and the error amplifier.

$$H_{EA} = G_{EA} \times R_{EA} \times \frac{R2}{R2 + R1} \times \frac{1 + \dfrac{s}{2 \times \pi \times f_Z}}{\left(1 + \dfrac{s}{2 \times \pi \times f_{P1}}\right) \times \left(1 + \dfrac{s}{2 \times \pi \times f_{P2}}\right)}$$

(19)

where G_{EA} and R_{EA} are the amplifier's trans-conductance and output resistance located in the *Electrical Characteristics* table.

$$f_{P1} = \frac{1}{2\pi \times R_{EA} \times C4}$$

(20)

$$f_{P2} = \frac{1}{2\pi \times R3 \times C5} \text{ (optional)}$$

C5 is optional and can be modeled as 10 pF stray capacitance.

(21)

and

$$f_Z = \frac{1}{2\pi \times R3 \times C4}$$

(22)

Figure 11 shows a typical bode plot for transfer function H(s).
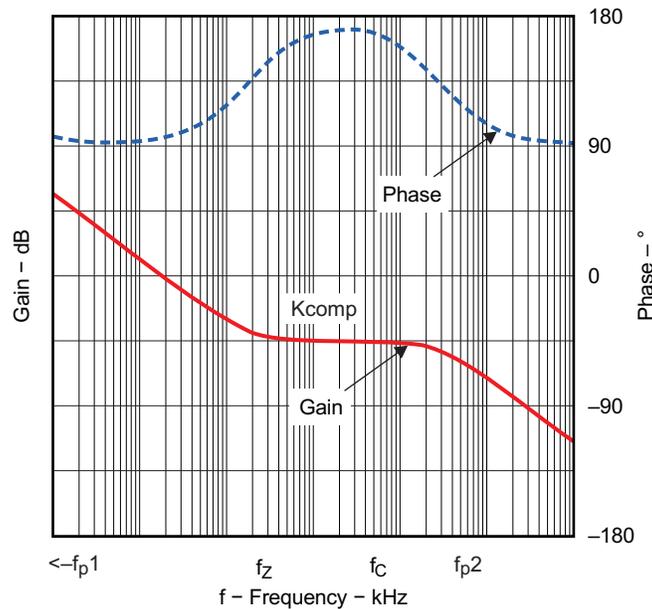
Figure 11. Bode Plot of Feedback Resistors and Compensated Amplifier Gain and Phase

The next step is to choose the loop crossover frequency, f_C. The higher in frequency that the loop gain stays above zero before crossing over, the faster the loop response will be and therefore the lower the output voltage will droop during a step load. It is generally accepted that the loop gain cross over no higher than the lower of either 1/5 of the switching frequency, f_{SW}, or 1/3 of the RHPZ frequency, f_{RHPZ}. To approximate a single pole roll-off up to f_{P2}, select R3 so that the compensation gain, K_{COMP}, at f_C on Figure 11 is the reciprocal of the gain, K_{PW}, read at frequency f_C from the Figure 10 bode plot, or more simply:

$$K_{COMP}(f_C) = 20 \times \log(G_{EA} \times R3 \times R2/(R2+R1)) = 1/K_{PW}(f_C)$$

This makes the total loop gain, $T(s) = G_{PS}(s) \times H_{EA}(s)$, zero at the f_C. Then, select C4 so that $f_Z \cong f_C/10$ and optional $f_{P2} > f_C *10$. Following this method should lead to a loop with a phase margin near 45 degrees. Lowering R3 while keeping $f_Z \cong f_C/10$ increases the phase margin and therefore increases the time it takes for the output voltage to settle following a step load.

In the TPS61175, if the FB pin voltage changes suddenly due to a load step on the output voltage, the error amplifier increases its transconductance for 8-ms in an effort to speed up the IC's transient response and reduce output voltage droop due to the load step. For example, if the FB voltage decreases 10-mV due to load change, the error amplifier increases its source current through COMP by 5 times; if FB voltage increases 11-mV, the sink current through COMP is increased to 3.5 times normal value. This feature often results in saw tooth ringing on the output voltage, shown as Figure 13. Designing the loop for greater than 45 degrees of phase margin and greater than 10db gain margin minimizes the amplitude of this ringing. This feature is disabled during soft start.

9.2.3 Application Curves

Figure 12. Line Transient Response

Figure 13. Load Transient Response

Figure 14. PWM Operation

Figure 15. Pulse Skipping

Figure 16. Soft Startup
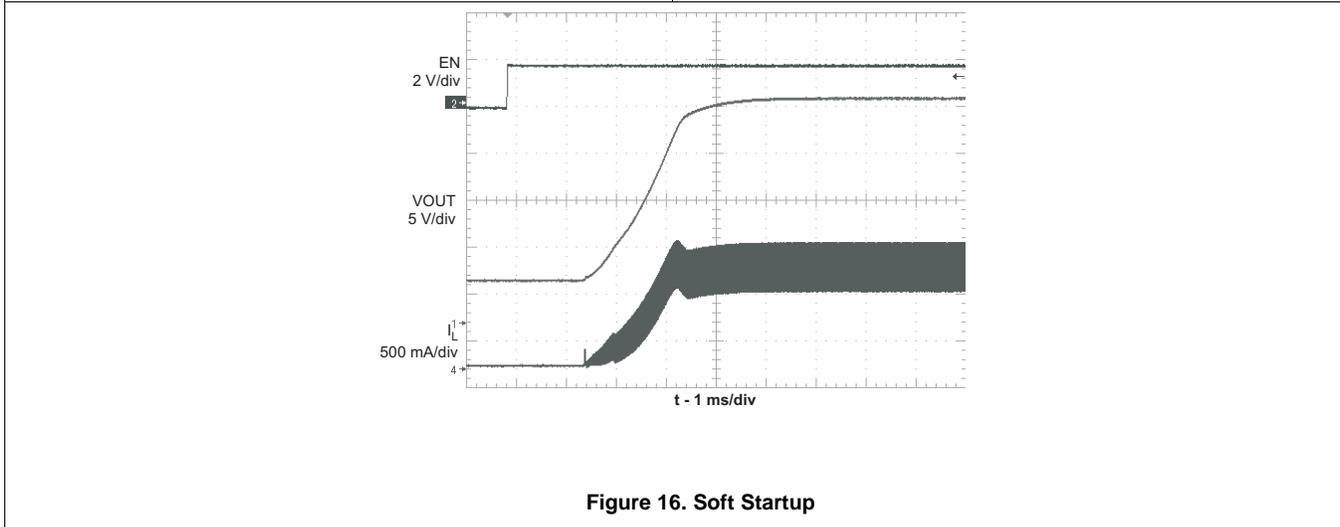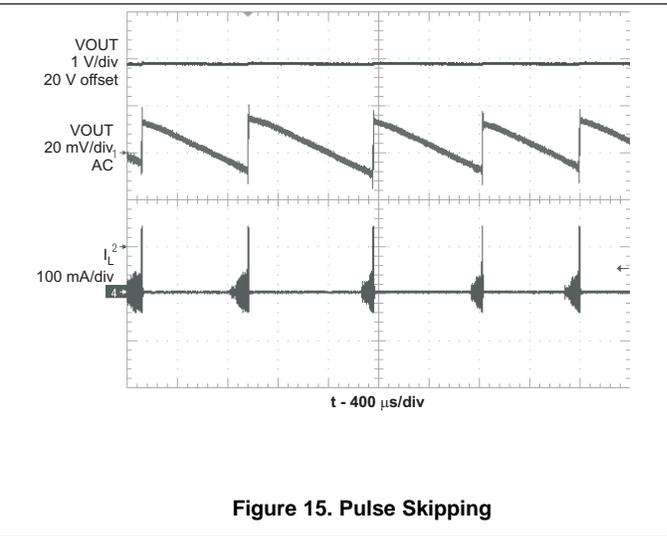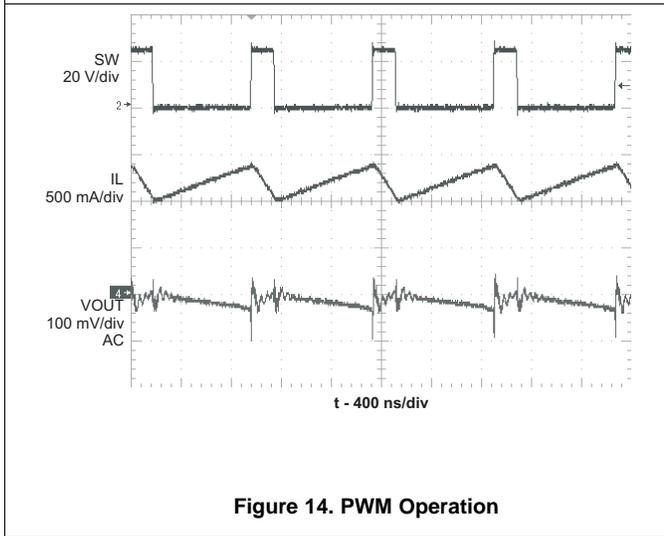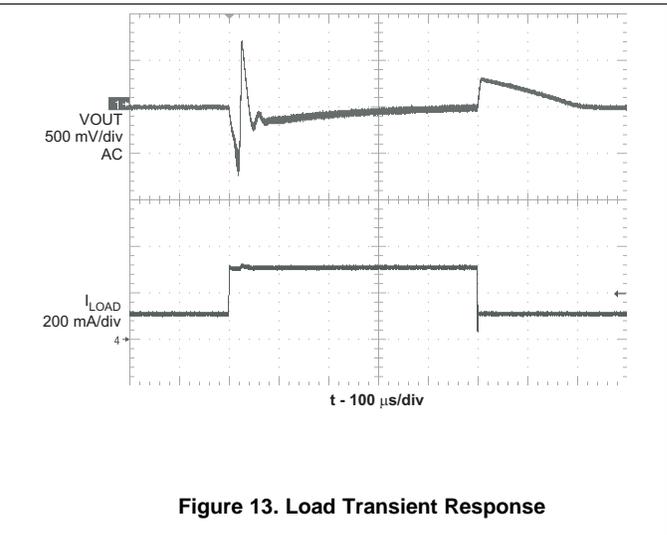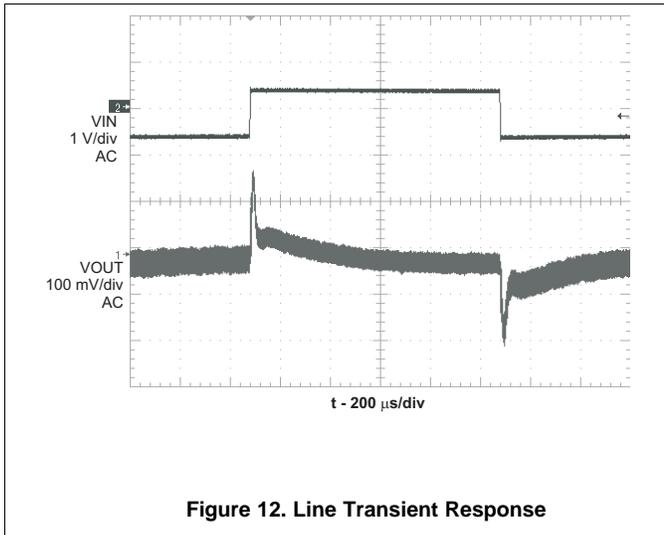
10 Power Supply Recommendations

The device is designed to operate from an input voltage supply range between 2.9 V and 18 V. The input power supply's output current needs to be rated according to the supply voltage, output voltage and output current of the TPS61175.

11 Layout

11.1 Layout Guidelines

- As for all switching power supplies, especially those running at high switching frequency and high currents, layout is an important design step. If layout is not carefully done, the regulator could suffer from instability as well as noise problems. To maximize efficiency, switch rise and fall times are very fast. To prevent radiation of high frequency noise (eg. EMI), proper layout of the high frequency switching path is essential.
- Minimize the length and area of all traces connected to the SW pin and always use a ground plane under the switching regulator to minimize interplane coupling.
- The high current path including the switch, Schottky diode, and output capacitor, contains nanosecond rise and fall times and should be kept as short as possible.
- The input capacitor needs not only to be close to the VIN pin, but also to the GND pin in order to reduce the Iinput supply ripple.

11.2 Layout Example
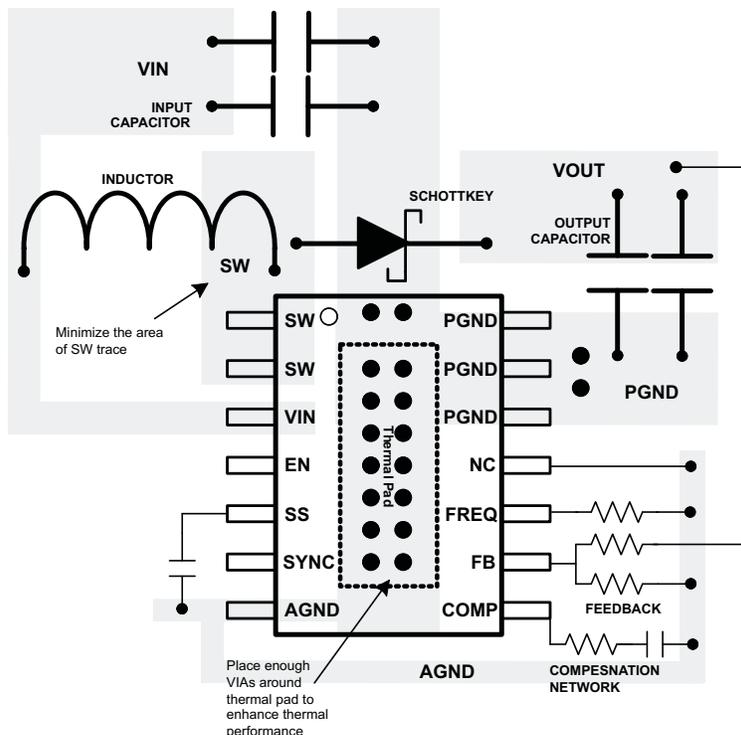
Figure 17. TPS61175 Layout

11.3 Thermal Considerations

The maximum IC junction temperature should be restricted to 125°C under normal operating conditions. This restriction limits the power dissipation of the TPS61175. Calculate the maximum allowable dissipation, $P_D(max)$, and keep the actual dissipation less than or equal to $P_D(max)$. The maximum-power-dissipation limit is determined using Equation 23:

$$P_{D(max)} = \frac{125°C - T_A}{R_{\theta JA}}$$

(23)

where, T_A is the maximum ambient temperature for the application. $R_{\theta JA}$ is the thermal resistance junction-to-ambient given in Power Dissipation Table.

The TPS61175 comes in a thermally enhanced TSSOP package. This package includes a thermal pad that improves the thermal capabilities of the package. The $R_{\theta JA}$ of the TSSOP package greatly depends on the PCB layout and thermal pad connection. The thermal pad must be soldered to the analog ground on the PCB. Using thermal vias underneath the thermal pad.

Product Folder Links: *TPS61175*

12 Device and Documentation Support

12.1 Device Support

12.1.1 Third-Party Products Disclaimer

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12.2 Trademarks

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12.3 Electrostatic Discharge Caution

These devices have limited built-in ESD protection. The leads should be shorted together or the device placed in conductive foam during storage or handling to prevent electrostatic damage to the MOS gates.

12.4 Glossary

SLYZ022 — *TI Glossary*.
This glossary lists and explains terms, acronyms, and definitions.

13 Mechanical, Packaging, and Orderable Information

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

PACKAGE 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
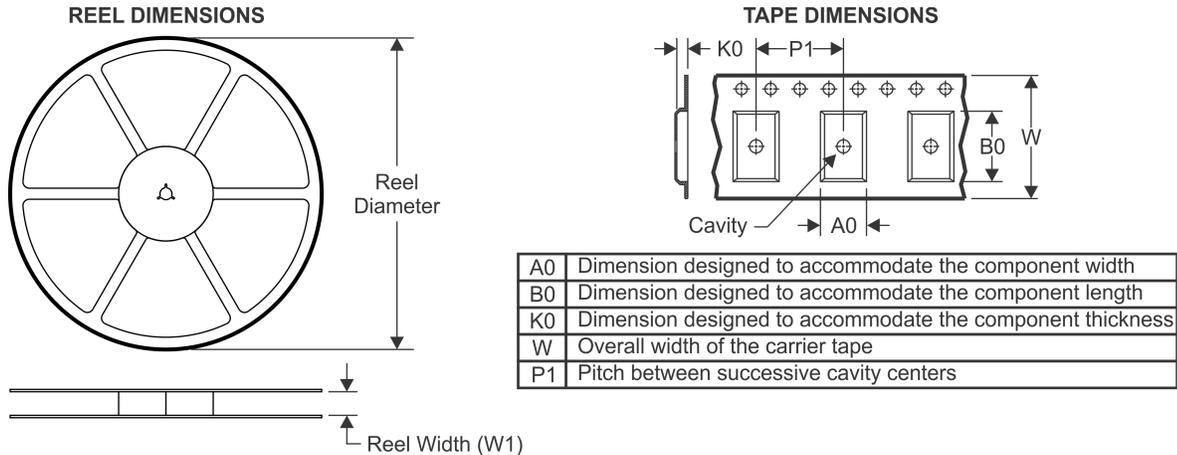
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE
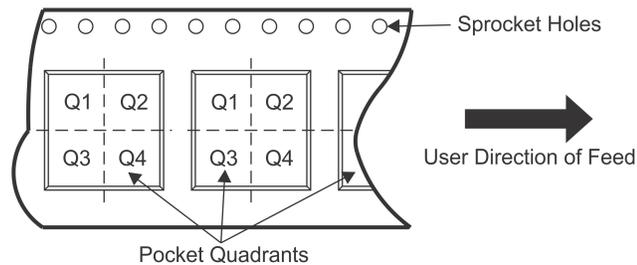
*All dimensions are nominal

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
TPS61175PWPR	HTSSOP	PWP	14	2000	330.0	12.4	6.9	5.6	1.6	8.0	12.0	Q1

TAPE AND REEL BOX DIMENSIONS
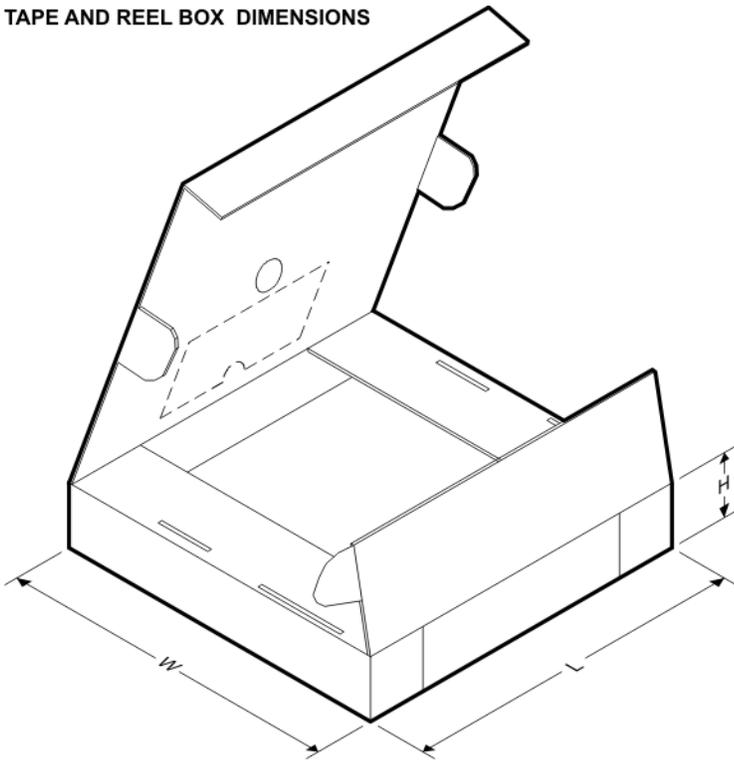
*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS61175PWPR	HTSSOP	PWP	14	2000	367.0	367.0	35.0

IMPORTANT NOTICE

Products		Applications	
Audio	www.ti.com/audio	Automotive and Transportation	www.ti.com/automotive
Amplifiers	amplifier.ti.com	Communications and Telecom	www.ti.com/communications
Data Converters	dataconverter.ti.com	Computers and Peripherals	www.ti.com/computers
DLP® Products	www.dlp.com	Consumer Electronics	www.ti.com/consumer-apps
DSP	dsp.ti.com	Energy and Lighting	www.ti.com/energy
Clocks and Timers	www.ti.com/clocks	Industrial	www.ti.com/industrial
Interface	interface.ti.com	Medical	www.ti.com/medical
Logic	logic.ti.com	Security	www.ti.com/security
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