TPS65301-Q1 3-MHz Step-Down Regulator and Triple Linear Regulators and Protected Sensor Supply

1 Features

- Qualified for Automotive Applications
- AEC-Q100 Qualified with the Following Results
  - Device Temperature Grade 1: –40°C to 125°C Ambient Operating Temperature Range
  - Device HBM ESD Classification Level H2
  - Device CDM ESD Classification Level C4B
- Input VIN Range 5.6 V to 40 V, With Transients up to 45 V
- 5.45-V Switch-Mode Regulator With Integrated High-Side Switch
  - Recommended Switch-Mode Frequency Range 2 MHz to 3 MHz
  - Overcurrent Protection and 1.2-A Peak Switch Current
- One Linear Regulator 5 V ±2%
- Two Linear Regulator Controllers With 3.3-V and 1.2 V ±2%
- Status Indicator Output of IGN_EN Input
- Soft Start on IGN_EN and EN
- External Clock Input for Synchronization
- Programmable Power-On-Reset Delay, Reset-Function Filter Timer for Fast Negative Transients
- Voltage Supervisor for the Following Supplies
  - VREG, 3.3 V, 1.2 V
- Thermally Enhanced 24-Pin HTSSOP or 24-Pin VQFN Package
- Protected 5-V Sensor Supply Output, Which Tracks 3.3-V Supply

2 Applications

- Power Supply for TMS570 Microcontrollers
- Power Supply for C28XXX DSP
- General-Purpose Power Supply for Automotive Applications

3 Description

The TPS65301-Q1 power supply is a combination of a single switch-mode buck power supply and three linear regulators. This is a monolithic high-voltage switching regulator with an integrated 1.2-A peak current switch, 45-V power MOSFET, one low-voltage linear regulator, two voltage-regulator controllers and a protected sensor supply.

The device has a voltage supervisor which monitors the output of the switch-mode power supply, the 3.3-V linear regulator, and the 1.2-V linear regulator. An external timing capacitor is used to set the power-on delay and the release of the reset output nRST. This reset output is also used to indicate if the switch-mode supply, the 3.3-V linear regulator supply, or the 1.2-V linear regulator supply is outside the set limits. The protected sensor supply 5VS tracks the 3.3-V linear regulator within the specified limits.

The TPS65301-Q1 device has a switching frequency range from 2 MHz to 3 MHz, allowing the use of low-profile inductors and low-value input and output ceramic capacitors. External loop compensation gives the user the flexibility to optimize the converter response for the appropriate operating conditions.

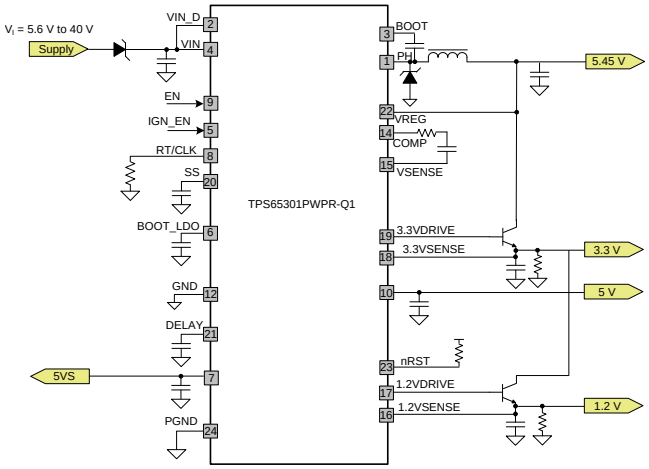
This device has built-in protection features such as soft start on IGN_EN ON or enables cycle, pulse-by-pulse current limit, thermal sensing, and shutdown due to excessive power dissipation.

Device Information[1]

PART NUMBER	PACKAGE	BODY SIZE (NOM)
TPS65301-Q1	HTSSOP (24)	7.80 mm × 4.40 mm
	VQFN (24)	5.00 mm × 4.00 mm

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

Typical Application Schematic



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4 Revision History

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

Changes from Revision B (December 2013) to Revision C — **Page**

- Changed the *Features* section ... 1
- Changed the minimum VIN value from 5.75 to 5.6 in the *Features* and *Power Supply Recommendations* sections ... 1
- Added *Pin Configuration and Functions* section, *ESD Ratings* 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
- Updated the pin types and descriptions in the *Pin Functions* table ... 4
- Changed the maximum PH buck regulator voltage in the *Absolute Maximum Ratings* table ... 4
- Moved the TA value from the *Recommended Operating Conditions* table to the *Absolute Maximum Ratings* table ... 4
- Changed *DC CHARACTERISTICS* condition statement from T_J = –40°C to 150°C to $T_{J\text{-}Max}$ = 150°C ... 5

Changes from Revision A (November 2013) to Revision B — **Page**

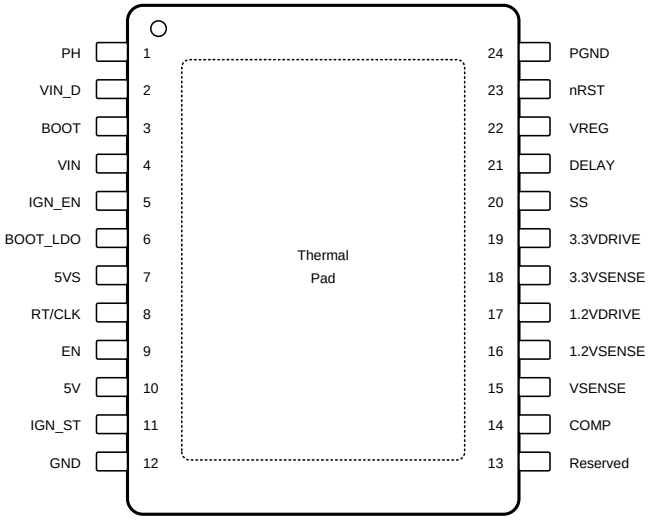
- Changed the Operating Junction Temperature Range from –40°C to 150°C to up to 150°C in the *FEATURES* list ... 1
- Changed *DC CHARACTERISTICS* condition statement from T_J = –40°C to 150°C to $T_{J\text{-}Max}$ = 150°C ... 5
- Added *Output Voltage vs Output Current* graph to *TYPICAL CHARACTERISTICS* section ... 8
- Changed Y-axis name from *Current (mA)* to *Efficiency* in the *EFFICIENCY vs OUTPUT CURRENT ON VREG* graph in the *TYPICAL CHARACTERISTICS* section ... 24

Changes from Original (October 2013) to Revision A — **Page**

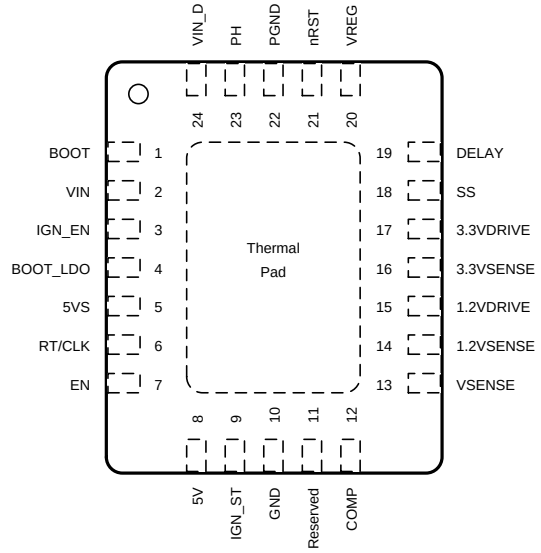
- Changed document status from *Product Preview* to *Production Data* ... 1
- Changed min value for V_{IL} in the *DC CHARACTERISTICS* table from 2 to 2.2 ... 5
- Deleted the 5VS oft-start time, T_{SS}, parameter from the *DC CHARACTERISTICS* table ... 7
- Changed the min, typ, and max values for the nRST parameter for VREG output from 0.87, 0.9, and 0.93 to 0.845, 0.875, and 0.905 respectively ... 7

5 Pin Configuration and Functions

PWP Package
24-Pin HTSSOP With Exposed Thermal Pad
Top View



RHF Package
24-Pin VQFN With Exposed Thermal Pad
Top View



Pin Functions

PIN			TYPE[1]	DESCRIPTION
NAME	NO.			
	HTSSOP	VQFN		
1.2VDRIVE	17	15	PWR	Output current source to drive the base of an external bipolar transistor to regulate the 1.2-V supply
1.2VSENSE	16	14	I	Feedback node of 1.2-V supply
3.3VDRIVE	19	17	PWR	Output current source to drive the base of an external bipolar transistor to regulate the 3.3-V supply
3.3VSENSE	18	16	I	Feedback node of 3.3-V supply
5V	10	8	O	Regulated output, external capacitor to ground for stability of regulated output
5VS	7	5	PWR	Regulated output, external capacitor to ground for stability of regulated output
BOOT	3	1	O	External bootstrap capacitor connected to PH (pin 1) to drive gate of internal switching FET
BOOT_LDO	6	4	O	External capacitor connected to ground for stability of internal regulator
COMP	14	12	O	Error amplifier output to connect external compensation components
DELAY	21	19	O	External capacitor to ground to program the power-on-reset delay
EN	9	7	I	A high logic-level input signal to enable and low signal to disable device. Internally pulled down to ground
GND	12	10	GND	GND pin, must be electrically connected to the exposed copper pad on PCB
IGN_EN	5	3	I	Ignition input, (high-voltage tolerant) internally pulls to ground. Must be externally pulled up to enable
IGN_ST	11	9	O	Active-low, open-drain ignition input indicator, output connected to external bias voltage through a resistor. Asserted high after ignition input is high
Reserved	13	11	—	Should be grounded in the application
nRST	23	21	O	Active-low, open-drain reset output connected to external bias voltage through a resistor. This output is floating and pulled high by an external resistor after the preregulator, 3.3-V, and 1.2-V regulator outputs are regulating and the delay timer has expired. Also, output is asserted low if any one of these three supplies is out of the set regulation, this threshold is internally set.
PGND	24	22	GND	Power ground pin, must be connected to the exposed copper pad on PCB for proper electrical and thermal performance
PH	1	23	PWR	Source of internal switching FET
RT/CLK	8	6	I/O	External resistor connected ground to program the internal oscillator. Alternative option is to feed an external clock to provide reference for switching frequency.

(1) PWR = power, I = input, O = output, I/O = input-output, GND = ground, — = not applicable

Pin Functions (continued)

PIN			TYPE(1)	DESCRIPTION
NAME	NO.			
	HTSSOP	VQFN		
SS	20	18	O	External capacitor to ground to program soft-start time
VIN	4	2	PWR	Unregulated input voltage supply. Pin 2 and pin 4 must be connected together externally.
VIN_D	2	24	PWR	Drain input for internal high side MOSFET. Pin 2 and pin 4 must be connected together externally.
VREG	22	20	I	Connect this pin to the buck converter output. Integrated internal low-side FET to load output during start-up or limit voltage overshoot
VSENSE	15	13	I	Inverting node of error amplifier for voltage-mode control of preregulated supply
Thermal pad			—	Electrically connect to ground and solder to ground plane of PCB for thermal efficiency

6 Specifications

6.1 Absolute Maximum Ratings

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

		MIN	MAX	UNIT
Buck regulator	VIN, VIN_D	–0.3	45	V
	BOOT	–0.3	50	V
	PH	–1 –2 for 30 ns	45	V
	VSENSE	–0.3	5.5	V
Control	IGN_EN	–0.3	45	V
	EN	–0.3	5.5	V
	3.3VSENSE	–0.3	5.5	V
	1.2VSENSE	–0.3	5.5	V
	RT/CLK	–0.3	5.5	V
	VREG	–0.3	8	V
Output	3.3VDRIVE	–0.3	8	V
	1.2VDRIVE	–0.3	8	V
	nRST	–0.3	5.5	V
	IGN_ST	–0.3	5.5	V
	SS	–0.3	7	V
	DELAY	–0.3	7	V
	COMP	–0.3	7	V
	BOOT_LDO	–0.3	9	V
	5V	–0.3	7	V
	5VS	–1	45	V
Temperature	Operating junction, T_J	–40	150	°C
	Operating ambient temperature, T_A	–40	125	°C
	Storage, T_S	–55	165	°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.

6.2 ESD Ratings

				VALUE	UNIT
$V_{(ESD)}$	Electrostatic discharge	Human-body model (HBM), per AEC Q100-002(1)		±2000	V
		Charged-device model (CDM), per AEC Q100-011	All pins	±500	
			Corner pins (1, 12, 13, and 24)	±750	

(1) AEC Q100-002 indicates that HBM stressing shall be in accordance with the ANSI/ESDA/JEDEC JS-001 specification.

6.3 Recommended Operating Conditions

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

	MIN	NOM	MAX	UNIT
VIN, VIN_D	5.6		40	V
BOOT	5.6		48	V
PH	–1		40	V
IGN_EN	0		40	V
EN, VSENSE, 3.3VSENSE, 1.2VSENSE, RT/CLK, nRST, IGN_ST	0		5.25	V
VREG, 3.3VDRIVE, 1.2VDRIVE	0		7.5	V
SS, DELAY, COMP	0		6.5	V
BOOT_LDO	0		8.1	V

6.4 Thermal Information

THERMAL METRIC[1]		TPS65301-Q1		UNIT
		PWP (HTSSOP)	RHF (VQFN)	
		24 PINS	24 PINS	
$R_{\theta JA}$	Junction-to-ambient thermal resistance	33.6	30.3	°C/W
$R_{\theta JC(top)}$	Junction-to-case (top) thermal resistance	16.6	30.5	°C/W
$R_{\theta JB}$	Junction-to-board thermal resistance	14.5	8.7	°C/W
ψ_{JT}	Junction-to-top characterization parameter	0.4	0.3	°C/W
ψ_{JB}	Junction-to-board characterization parameter	14.3	8.8	°C/W
$R_{\theta JC(bot)}$	Junction-to-case (bottom) thermal resistance	1.3	1.6	°C/W

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

6.5 DC Characteristics

VIN = VIN_D = 6 V to 27 V, IGN_EN = VIN, T_J = –40°C to 150°C, unless otherwise noted

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
VIN, VIN_D (Input Power Supply)						
VIN, VIN_D	Supply voltage on VIN, line	Normal mode, after initial startup	5.6	14	40	V
$I_{q\text{-Normal}}$	Current normal mode	Open-loop test		5.57		mA
$I_{SD\ VIN}$	Shut down	IGN = 0 V, VIN = 12 V, T_A = –40°C to 125°C		2.2	15	µA
$I_{SD\ VIND}$		IGN = 0 V, VIN = 12 V, T_A = –40°C to 125°C		2.2	15	
IGN_EN (Ignition Input)						
V_{IGN_EN}	Input voltage range	Input into IGN_EN pin		14	40	V
V_{IH}	Input high	Enable device to be ON (rising signal)		3.16	3.6	V
V_{IL}	Input low	Enable device to be OFF (falling signal)	2.2	3.03		V
I_{IH}	Input high	Enable device to be ON, V_{IGN_EN} = 18 V		23.7	50	µA
		Enable device to be ON, V_{IGN_EN} = 3.7 V		4	7	
EN (Logic Level Enable)						
V_{IH}	Input high	Enable device to be ON (rising signal)		1.7	2.3	V
V_{IL}	Input low	Enable device to be OFF (falling signal)	0.7	1.53		V

DC Characteristics (continued)

VIN = VIN_D = 6 V to 27 V, IGN_EN = VIN, T_J = –40°C to 150°C, unless otherwise noted

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
Switch-Mode Output 5.45 V						
VREG	Regulator output internal resistor network	Fixed output based on internal resistor network	5.30	5.45	5.70	V
C_O	Output capacitor for 5.45 V	ESR = 0.001 Ω to 100 mΩ; large output capacitance may be required for load transients	10			µF
$r_{ds(on)}$	Internal switch resistance	Measured across VIN_D and PH pins, I_{VREG} = 1 A		0.3		Ω
I_{O-CL}	Switch current limit	VIN = 12 V	1.2	2	3	A
t_{ON-min}	Minimum ON time			40		ns
D_{max}	Maximum duty cycle			97%		
VSENSE (Internal Reference Voltage)						
VREG ref	Internal reference voltage		1.954	2	2.046	V
SS (Soft-Start Timer for Switch-Mode Converter)						
I_{SS}	Soft-start source current	Css = 0.001 µF to 0.01 µF	40	50	60	µA
IGN_ST (Ignition Input Status)						
V_{OL}	Output low	Output asserted low when IGN_EN < 2.2 V, I_{OL} = 1 mA		0.056	0.4	V
I_{IH}	Leakage test	IGN_ST = 5 V		0.05	2	µA
5V (5-V Linear Regulator)						
$5V_O$	Output voltage	I_O = 1 mA, VREG = 5.45 V	4.9	5	5.1	V
ΔV_{O-Line}	Line regulation	5.15 V < VREG < 5.45 V, I_O = 1 mA, VIN = 12 V		10	20	mV
ΔV_{O-Load}	Load regulation	1 mA < I_O < 200 mA, VREG = 5.45 V, VIN = 12 V		10	30	mV
V_{DO}	Dropout voltage	I_O = 150 mA, measure VREG when V_O(nom) – 0.1 V, then V_{DO} = VREG – ($5V_O$ – 0.1) V, VREG > 5 V		0.15	0.26	V
I_{5V-CL}	Current limit	$5V_O$ = 0.8 x $5V_O$ (nom)	350	1080		mA
C_O	Output capacitor	ESR = 0.001 Ω to 2 Ω. Larger output capacitance may be required for load transients.	1	2.2	10	µF
PSRR	Power-supply rejection ratio	f = 100 Hz, VREG = 5.45 V, I_O = 100 mA, VIN = 12 V	45	60	75	dB
$V_{soft-start}$	Soft start on enable cycle	$5V_O$ = 0 V (initially) with f_{sw} = 2.5 MHz		13		ms
3.3-V Linear Regulator Controller (3.3VSENSE)						
$3.3V_O$	Output voltage	I_O = 5 mA, Vnpn_power input = 5.3 V	3.234	3.3	3.366	V
$\Delta 3.3V_{O-Line}$	Line regulation	3.8 V < Vnpn_power input < 7 V (with nRST not triggered)		1	10	mV
$\Delta 3.3V_{O-Load}$	Load regulation	5 mA < I_O < 550 mA		7.5	30	mV
C_O	Output capacitor for 3.3 V	ESR = 0.001 Ω to 2 Ω. Large output capacitance may be required for load transients.	1	4.7	10	µF
PSRR	Power-supply rejection ratio	f = 100 Hz, VREG = 5.45 V, I_O = 200 mA, VIN = 12 V	45	60	75	dB
t_{ss}	Soft-start time	$3.3V_O$ = 0 V (initially) with f_{sw} = 2.5 MHz		12.3		ms
3.3VDRIVE (Example: Switch Control Output)						
I_{OH}	Base drive current. NPN turn ON	3.3VDRIVE – 3.3VSENSE = 1 V	10	28	50	mA
I_{OL}	NPN turn off	3.3VDRIVE – 3.3VSENSE at 0.2 V	0.1	0.412		mA
1.2-V Linear Regulator Controller (1.2VSENSE)						
$1.2V_O$	Output voltage	I_O = 5 mA, Vnpn_power input = 5.3 V	1.176	1.2	1.224	V
$\Delta 1.2V_{O-Line}$	Line regulation	3.25 V < Vnpn_power input < 7 V (with nRST not triggered)		1	10	mV
$\Delta 1.2V_{O-Load}$	Load regulation	5 mA < I_O < 350 mA		5	15	mV

DC Characteristics (continued)

VIN = VIN_D = 6 V to 27 V, IGN_EN = VIN, T_J = –40°C to 150°C, unless otherwise noted

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
C_O	Output capacitor for 1.2 V	ESR = 0.001 Ω to 100 mΩ. Large output capacitance may be required for load transients.	8	10	12	µF
PSRR	Power-supply rejection ratio	f = 100 Hz, VREG = 6 V, I_O = 200 mA, VIN = 12 V	45	60	75	dB
t_{ss}	Soft-start time	1.2V_O = 0 V (initially) with f_{sw} = 2.5 MHz		8.5		ms
1.2VDRIVE (Example: Switch Control Output)						
I_{OH}	Base drive current. NPN turn ON	1.2VDRIVE – 1.2VSENSE = 1 V	10	27	50	mA
I_{OL}	NPN turn off	1.2VDRIVE – 1.2VSENSE at 0.2 V	0.1	0.47		mA
5VS (Protected Sensor Supply Linear Regulator)						
V_{SENSOR}	Output tolerant range	V_{SENSOR} output shorted fault conditions	–1		VIN	V
VSENSOR	Output voltage	I_O = 1 mA to 100 mA, VREG = 5.45 V	4.9	5	5.1	V
I_{5VS_SC}	Short circuit current	5VS = 45 V		2.25		mA
I_{5VS}	Output current	VREG = 5.45 V			150	mA
$\Delta 5VS_{LOAD}$	Load regulation	1 mA < I_{5VS} < 75 mA, VREG = 5.45 V, VIN = 12 V		15		mV
V_{DO}	Drop out voltage	IO = 150 mA, Measure VREG when 5VS (nom) – 0.1 V Then V_{DO} = VREG – (5VS – 0.1) V, VREG > 5.1 V			0.4	V
C_O	Output capacitor for protected 5-V supply	ESR = 0.001 Ω to 2 Ω, Larger output capacitance may be required for load transients	1		10	µF
I_{5VS-CL}	Current limit	5VS = 0.8 x 5VS (nom)	180	320	650	mA
I_{Lkg}	Leakage current	EN_LIN_REG = 0 V with VIN = 14 V			5	µA
PSRR	Power-supply rejection ratio	f = 100 Hz, VREG = 6 V, I_{5VS} = 75 mA, VIN = 12 V		60		dB
DELAY (Power-On-Reset Delay)						
$V_{Threshold}$	Threshold voltage	Threshold to release nRST high	1.3	2.05	2.6	V
I_{Charge}	Capacitor charging current		1.4	2	2.6	µA
nRST (Reset Indicator)						
V_{OL}	Output low	Reset asserted due to falling VREG or 3.3 V_O or 1.2 V_O output voltages, I_{OL} = 1 mA	0	0.16	0.4	V
$t_{nRSTdly}$	Filter time	Delay before nRST is asserted low		11		µs
V_{TH_VREG}	Trigger nRST for VREG output	VREG ramp down	0.845	0.875	0.905	VREG
	Trigger nRST for 3.3 V_O	VREG ramp down	0.9	0.93	0.96	3.3 V_O
	Trigger nRST for 1.2 V_O	VREG ramp down	0.9	0.93	0.96	1.2 V_O
I_{IH}	Leakage test	Reset = 5 V		0.07	2	µA
RT/CLK (Oscillator Setting of External Clock Input)						
f_{sw}	Switching freq using RT mode		2		3	MHz
	Switching freq using CLK mode		2		3	
	Minimum clock input pulse duration			40		ns
	Internal oscillator frequency	Switching frequency tolerance for clock	–14%		14%	
	External clock input		–20%		10%	
V_{IH}	Input high				2.3	V
V_{IL}	Input low		0.6			V

6.6 Typical Characteristics

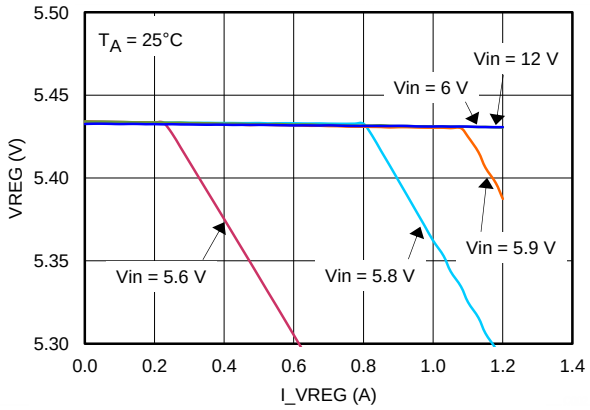


Figure 1. Output Voltage vs Output Current

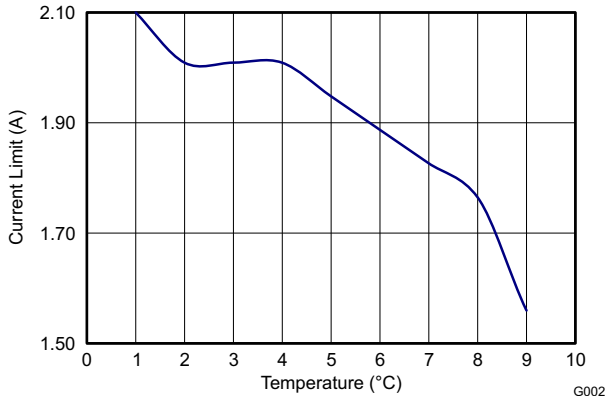


Figure 2. Buck Current Limit vs Temperature

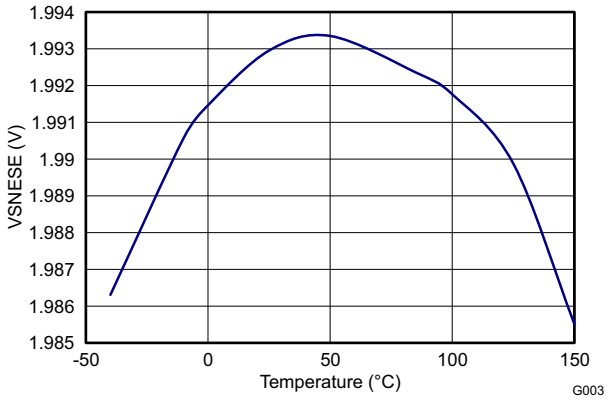


Figure 3. VSENSE Reference Voltage vs Temperature

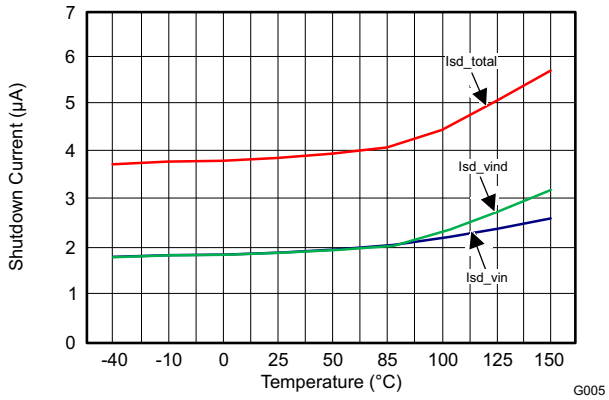


Figure 4. Current Consumption IIN vs Temperature

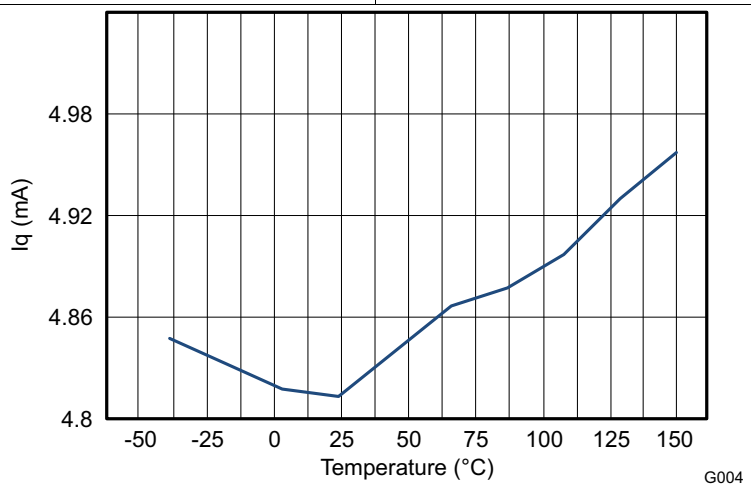


Figure 5. Quiescent Current vs Temperature

6.6.1 5-V Linear Regulator (5 V_O)

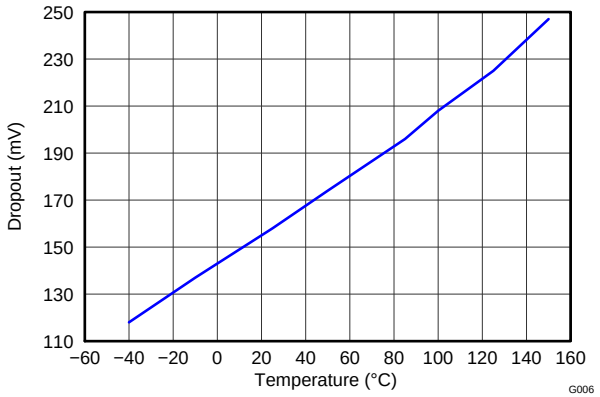


Figure 6. Dropout Voltage vs Temperature

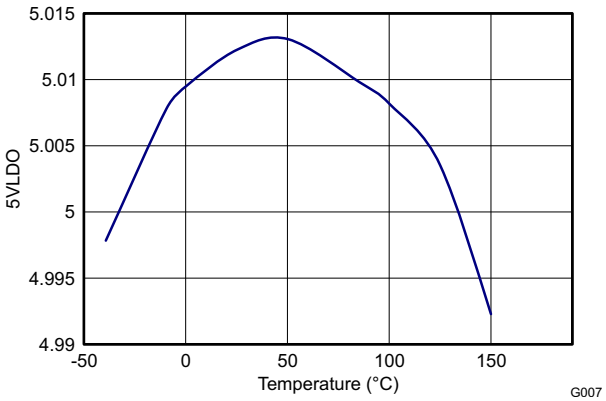


Figure 7. Output Voltage 5VO vs Temperature

6.6.2 3.3-V Linear Regulator Controller (3.3 V_O)

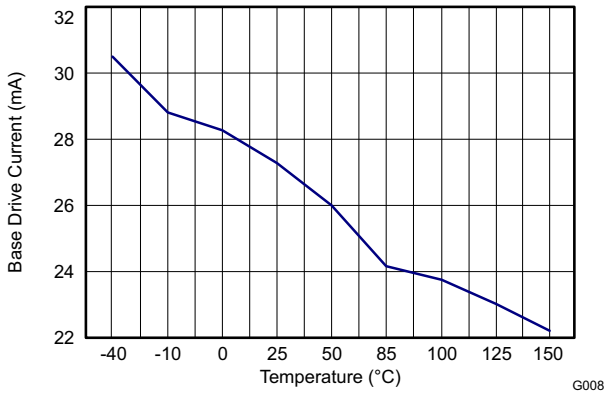


Figure 8. Output Base Drive vs Temperature

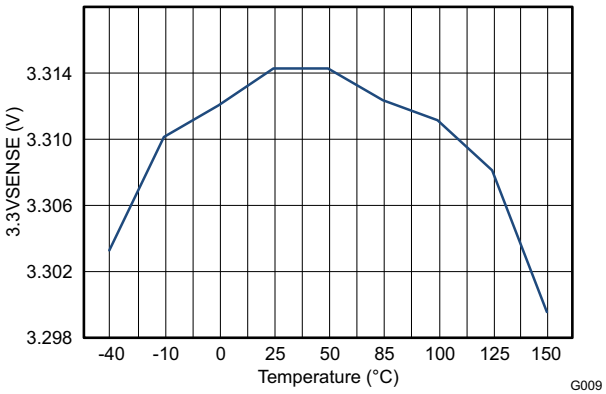


Figure 9. 3.3VSENSE vs Temperature

6.6.3 1.2-V Linear Regulator Controller (1.2 V_O)

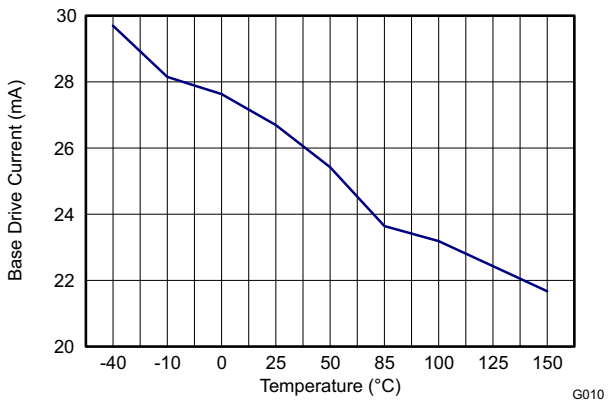


Figure 10. Output Base Drive vs Temperature

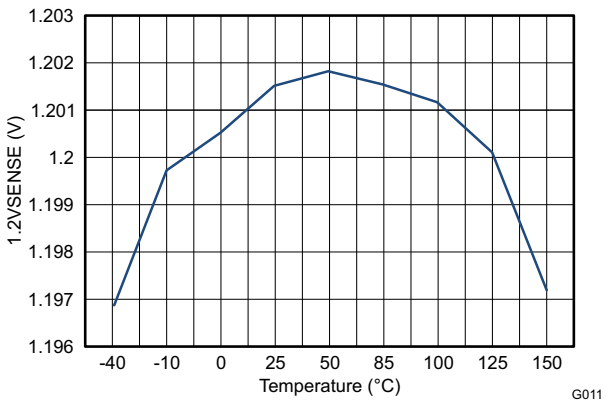


Figure 11. 1.2VSENSE vs Temperature

7 Detailed Description

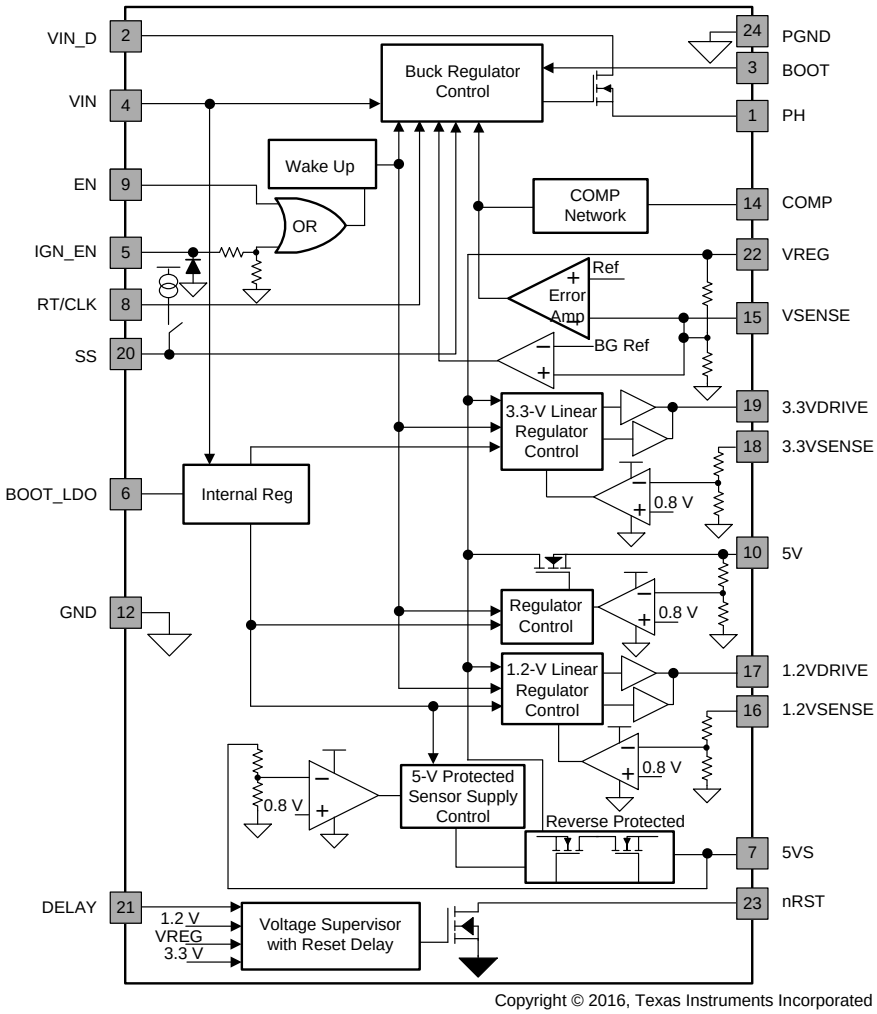
7.1 Overview

The device integrates an asynchronous switch-mode power-supply converter with a internal FET that converts the input battery voltage to a 5.45-V preregulator output. This 5.45-V output supplies the other regulators. The switching frequency range is from 2 MHz to 3 MHz, allowing the use of low-profile inductors and low value input and output capacitors. External loop compensation provides flexibility which optimizes the converter response for the appropriate operating condition.

A fixed 5-V linear regulator with an internal FET is integrated as an external peripheral supply. A fixed 3.3-V linear regulator controller with external bi-polar transistor is used for an IO supply, for example. A fixed 1.2-V linear regulator controller with external bi-polar transistor is used for a CPU Core supply, for example. The device has a voltage supervisor which monitors the output of the switch-mode power supply, the 3.3-V linear regulator, and the 1.2-V linear regulator.

An external timing capacitor sets the power-on delay and the release of the reset output nRST. This reset output is also used to indicate if the switch-mode supply, the 3.3-V linear regulator supply, or the 1.2-V linear regulator supply is outside the set limits. The 5-V regulator tracks the 3.3-V linear regulator within the specified limits.

7.2 Functional Block Diagram



Pin numbers apply to the PWP package.

7.3 Feature Description

7.3.1 Buck Converter

7.3.1.1 PWM Operation

The switch-mode power supply (SMPS) operates in a fixed-frequency pulse-width modulation (PWM) mode. The switching frequency is set by an external resistor or synchronized with an external clock input. The internal N-channel MOSFET is turned on at the beginning of each cycle. This MOSFET is turned off when the PWM comparator resets the latch. Once the high external FET is turned off, the external Schottky diode recirculates the energy stored in the inductor for the remainder of the switching period.

The external bootstrap capacitor acts as a voltage supply for the internal high-side MOSFET. This capacitor is recharged on every recirculation cycle (when the internal high-side MOSFET is turned OFF). In case of a VIN close to the desired output voltage, requiring a nearly 100% duty cycle for the internal high side MOSFET, the device automatically revert to 87% to allow the bootstrap capacitor to recharge.

7.3.1.2 Voltage-Mode Control Loop

The voltage-mode control monitors the set output voltage and processes the signal to control the internal MOSFET. A voltage feedback signal is compared to a constant ramp waveform, resulting in a PWM modulation pulse. An input line-voltage feedforward technique is incorporated to compensate for changes in the input voltage and ensures the output voltage is stable by adjusting the ramp waveform for the correct duty cycle. The internal MOSFET is protected from excess power dissipation with a current limit and frequency foldback circuitry during an output-to-ground short-circuit event.

A combination of internal and external components forms a compensation network to ensure error-amplifier gain does not cause instability due to input voltage changes or load perturbations.

7.3.1.3 Output Voltage 5.45 V (VREG)

Output voltage VREG is generated by the converter supplied from the battery voltage VIN and the external components (L, C). The output is sensed through an internal resistor divider and compared with an internal reference voltage.

This output requires larger output capacitors (4.7-µF to 10-µF range) to ensure that during load transients the output does not drop below the reset threshold for a period longer than the reset deglitch filter time.

An internal load is enabled for a short period whenever

- a start-up condition occurs, that is, during power up or when IGN_EN or EN is toggled.
- an overvoltage condition exists on this output.

7.3.1.4 Switching Frequency (RT/CLK)

The oscillator frequency of the buck regulator is selectable by means of a resistor placed at the RT/CLK pin to ground. The switching frequency (f_{SW}) can be set in the range 2 MHz to 3 MHz in this resistor mode. Alternatively, if there is an external clock input signal, the internal oscillator synchronizes to this signal within 10 µs.

The following equation determines the value of resistor (RT) for the required switching frequency f_{SW}.

$$RT = \frac{98.4 \times 10^9}{f_{SW}} \quad \text{(Ohms)} \tag{1}$$

7.3.1.5 Boost Capacitor (BOOT)

This capacitor provides the gate-drive voltage for the internal MOSFET switch. X7R and X5R grade dielectrics are recommended due to their stable values over temperature. Usually, a 0.1-µF capacitor is used for the boot capacitor.

Feature Description (continued)

7.3.1.6 Soft Start (SS)

To limit the start-up inrush current for the switch-mode supply, an internal soft-start circuit is used to ramp up the reference voltage from 0 V to the final value of 0.8 V. The regulator uses the internal reference or the SS-pin voltage as the power-supply reference voltage to regulate the output accordingly. The following equation determines the soft-start timing.

$$\text{Time } (t_{SS}) = \frac{C \times 0.8\ V}{50\ \mu A} \quad (2)$$

where

- C = Capacitor on SS pin, usually 0.1 µF or lower

7.3.1.7 Power-On Delay (DELAY)

The power-on delay function delays the release of the nRST line. The method of operation is to detect when all VREG (5.45 V), 3.3-V and 1.2-V power-supply outputs are above 90% (typical) of the set value. This then triggers a current source to charge the external capacitor on the DELAY pin. Once this capacitor is charged to approximately 2 V, the nRST line is asserted high. The delay time is calculated using the following equation:

$$t_{DELAY} = \frac{2\ V \times C}{2\ \mu A} \quad (3)$$

where

- C = capacitor on DELAY pin

Example: For a 20-ms delay, C = 20 nf.

7.3.1.8 Reset (nRST)

The nRST pin is an open-drain output. The power-on reset signal is a voltage supervisor output to indicate the output voltages on VREG (5.45 V), 3.3 V, and 1.2 V are within the specified tolerance of their set regulated voltages. Additionally, whenever both the IGN_EN and EN pins are low or open, nRST is immediately asserted low regardless of the output voltage. If a thermal shutdown occurs due to excessive thermal conditions, this pin is asserted low.

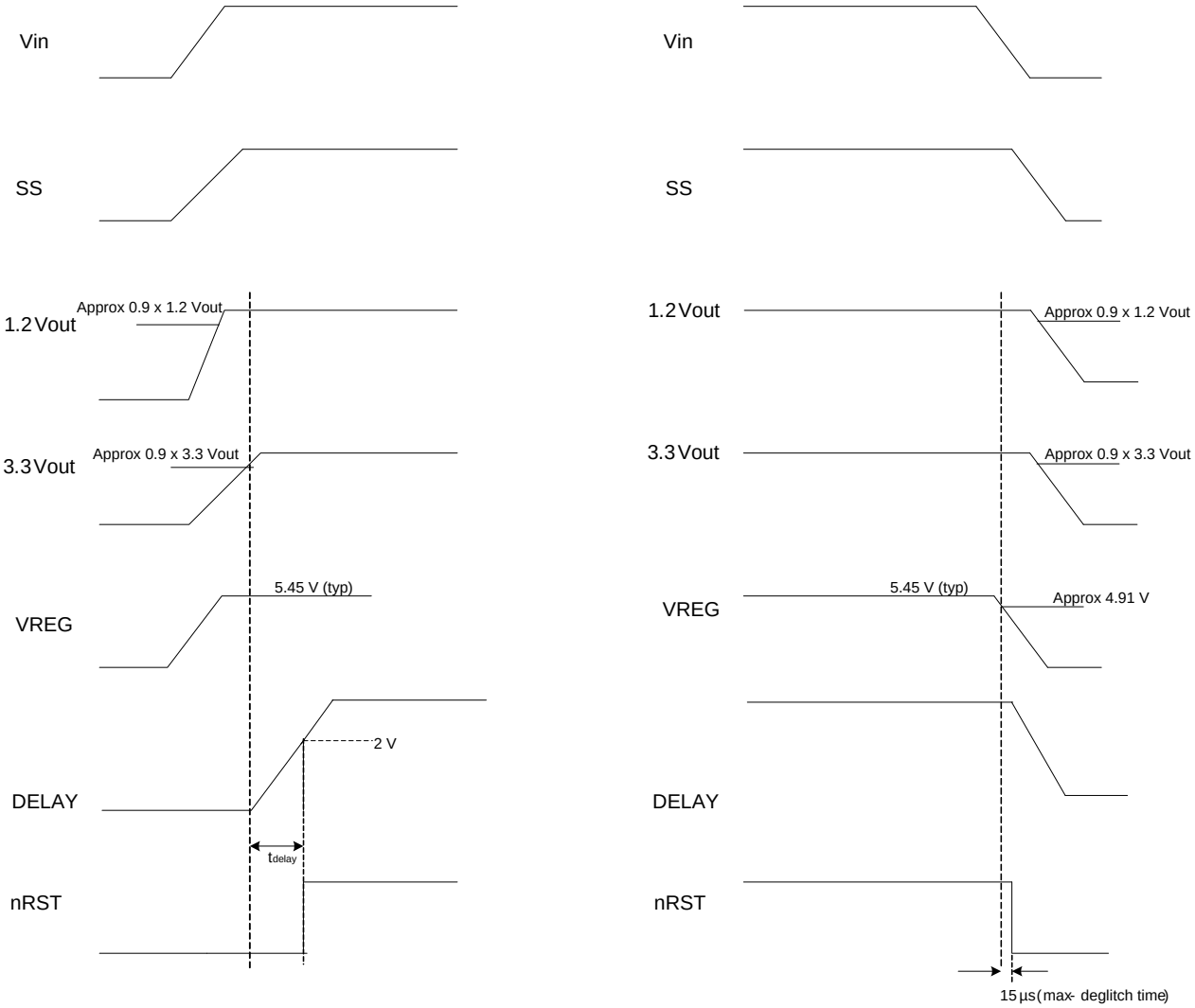
Conversely on power down, once the VREG or 3.3V or 1.2V output voltage falls below 90% of its respective set threshold, nRST is pulled low after a de-glitch filter delay of approximately 15 µs (max). This is implemented to prevent nRST from being invoked due to noise on the output supplies.

7.3.1.9 Thermal Shutdown

This device has two independent thermal-sensing circuits for the VREG (5.45 V), 5-V regulators; if either one of these circuits detects the power FET junction temperature to be greater than the set threshold, that particular output-power switch is turned OFF. The appropriate FET turns back on once it is allowed to cool sufficiently.

Feature Description (continued)

7.3.1.10 Reset Function



On power up, ALL three regulated supplies, VREG, 3.3 V and 1.2 V have to be more than 90% of their respective value before the delay timer capacitor on delay pin can start charging

On power down, if any one of the three regulated supplies, VREG, 3.3 V and 1.2 V drops below the 90% of it's value nRST is asserted low after a small deglitch filter time. Once nRST is asserted low, it can only go high again after ALL three supplies are above the 90% value and delay pin voltage higher than 2 V.

Figure 12. Reset Function

Feature Description (continued)

7.3.2 Linear Regulators

7.3.2.1 Fixed Linear Regulator Output (5 V)

This is a fixed, regulated output of 5 V ±2% over temperature and input supply using a precision voltage-sense resistor network. A low-ESR ceramic capacitor is required for loop stabilization; this capacitor must be placed close to the pin of the IC. This output is protected against shorts to ground by a foldback current limit for safe operating conditions, and a current limit for limiting inrush current due to depleted charge on the output capacitor. Initial IGN_EN or EN initiates power cycle of the soft-start circuit on this regulator. The soft-start takes typically 13 ms. This output may require a larger output capacitor to ensure that during load transients the output does not drop below the required regulated specifications.

7.3.2.2 Fixed Linear Regulator Controller (3.3 V)

The linear regulator controller requires an external NPN bipolar pass transistor of sufficient gain stage to support the maximum load current required. The base-drive output current is protected by current limiting both the source and sink drive circuitry. The 3.3VSENSE pin is the remote sense input of the output of the REG3 supply and controls the 3.3VDRIVE output accordingly. This regulator is fixed 3.3 V with ±2% tolerance using a precision voltage-sense resistor network. A low-ESR ceramic output capacitor is used for loop compensation of the regulator. A voltage on this pin of less than approximately 50% of the regulated value initiates a current limit on the 3.3VDRIVE output.

This output may require larger output capacitors to support load transients, so the output does not drop below 90% of 3.3 V.

7.3.2.3 Fixed Linear Regulator Controller (1.2 V)

The linear regulator controller requires an external NPN bipolar pass transistor of sufficient gain stage to support the maximum load current required. The 1.2VSENSE pin is the remote sense input of the output of 1.2-V supply and controls the 1.2VDRIVE output accordingly. This regulator output is 1.2 V with ±2% tolerance using a precision voltage-sense resistor network. A low-ESR ceramic output capacitor is used for loop compensation of the regulator. A voltage on this pin of less than approximately 50% of the regulated value initiates a current limit on the 1.2VDRIVE output.

This output may require larger output capacitors to support load transients, so the output does not drop below 90% of 1.2 V.

7.3.2.4 Protected Sensor Supply Output (5VS)

This is a fixed regulated output of 5 V ±2% over temperature and input supply using precision voltage sense resistor network. A low ESR ceramic capacitor is required for loop stabilization; this capacitor must be placed close to the pin of the IC. This output is protected against shorts to ground by a fold back current limit for safe operation conditions, and a current limit for limiting in-rush current due to depleted charge on the output capacitance. This output is also protected against shorts to battery voltage by limiting the reverse current. This supply can thus be used to power a sensor outside the electrical control unit ECU. On initial IGN_EN or EN power cycle the soft start circuit on this regulator is initiated. The soft-start takes typically 10 ms. This output may require larger output capacitor to ensure that during load transients the output does NOT drop below the required regulated specifications.

7.4 Device Functional Modes

7.4.1 Operational Mode

The purpose of the EN input is to keep the regulated supplies ON for a period for the microprocessor to log information into the memory locations once the ignition input is disabled. The microprocessor disables the power supplies by pulling EN low after this activity is complete (see Table 1).

Table 1. Enable Logic Table

IGN_EN	EN	nRST	OUTPUTS
H	H	H	ON
H	L	H	ON
L	H	H(1)	ON(1)
L	L	L	OFF

(1) If IGN_EN was high before.

7.4.2 Buck Converter Modes of Operation

7.4.2.1 Modes of Operation

The converter operates in different modes based on load current, input voltage, and component selection.

7.4.2.1.1 Continuous-Conduction Mode (CCM)

This mode of operation is typically when the inductor current is non-zero and the load current is greater than $I_{L\ CCM}$.

$$I_{IND\_CCM} \geq \frac{(1-D)\times VREG}{2\times f_{SW}\times L} \quad (4)$$

where

- I_{IND_CCM} = Inductor current in continuous-conduction mode
- D = duty cycle
- VREG = output voltage
- L = Inductor
- f_{SW} = switching frequency

In this mode, the duty cycle should always be greater than the minimum t_{ON} or the converter may go into burst mode.

7.4.2.1.2 Discontinuous Mode (DCM)

$$I_{IND\_DCM} \geq \frac{(1-D)\times VREG}{2\times f_{SW}\times L} \quad (5)$$

This mode of operation is typically when the inductor current goes to zero and the load current is less than $I_{IND\ DCM}$.

7.4.2.2 Tracking Mode

When the input voltage is low and the converter approaches approximately 100% duty cycle, the following equation determines the output voltage.

$$VREG = \left(1-\frac{t_{OFF\_MIN}}{T}\right)\times(VIN - I_{Load}\times R_{DS}) \quad (6)$$

where

- T = Period
- R_{DS} = Internal FET resistance
- I_{LOAD} = Output load current

8 Application and Implementation

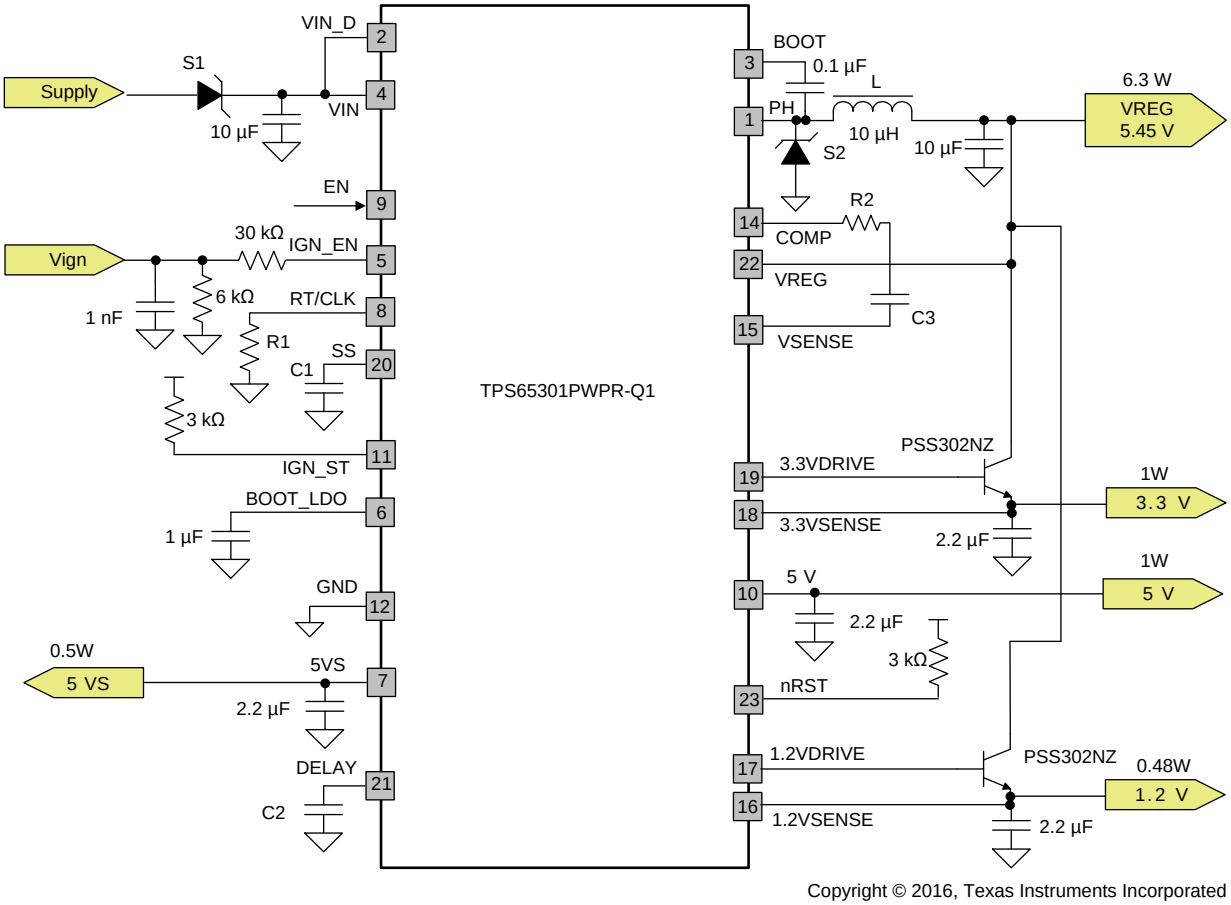
NOTE

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

This section is a starting point and theoretical representation of the values to be used for the application, further optimization of the components derived may be required to improve the performance of the device.

8.2 Typical Application



L: B82462G4103MOOO (EPCOS) or XFL4020 472MEB (Coilcraft)
S1: MBRS310T3 (ON Semiconductors) or SS3H10 (Vishay)
S2: B240A, SS16 (Vishay)
External BJT: PBSS302NZ (NXP)

Figure 13. Application Schematic

Typical Application (continued)

8.2.1 Design Requirements

For this design example, use the parameters listed in Table 2.

Table 2. Switching Regulator Requirements

PARAMETER	REQUIREMENT
Input voltage, V_I	6.5 V to 27 V, typical 14 V
Output voltage, 5.45 V	5.45 V_O ±2% at 6.3 W
Maximum output current $I_{5.45V\_max}$	1 A
Minimum output current $I_{5.45V\_min}$	0.01 A
Transient response 0.01 A to 0.8 A	5%
Reset threshold	90% of output voltage
5V	5 V_O at 1 W
3.3V	3.3 V_O at 1 W
1.2V	1.2 V_O at 0.5 W
5VS	5 V_O at 0.5 W
Switching frequency f_{SW}	2.5 MHz
Overvoltage threshold	106% of output voltage
Undervoltage threshold	95% of output voltage

8.2.2 Detailed Design Procedure

The following design procedure provides typical application procedures as well as the details of a switching regulator design using the requirements listed in Table 2.

8.2.2.1 Duty Cycle

Use Equation 7 to calculate the duty cycle.

$$D = \frac{V_O}{V_I} = \frac{5.45}{14} = 0.389 \quad (7)$$

where

- V_O = Output voltage
- V_I = Input voltage

8.2.2.2 Output Inductor Selection (L)

The minimum inductor value is calculated using the coefficient K_{IND} that represents the amount of inductor ripple current relative to the maximum output current. The inductor ripple current is filtered by the output capacitor, and so the typical range of this ripple current is in the range of K_{IND} = 0.2 to 0.3, depending on the ESR and the ripple-current rating of the output capacitor.

For this design example, use Equation 8 to calculate the inductor ripple current.

$$I_{Ripple} = K_{IND} \times I_O = 0.25 \times 1\ A = 0.25\ A \quad (8)$$

where

- I_O = Output current

The benefits of a low inductor value include the following:

- Low inductor value gives high di/dt, which allows for fewer output capacitors for good load transient response.
- Gives higher saturation current for the core due to fewer turns
- Fewer turns yields low DCR and therefore less dc inductor losses in the windings.
- High di/dt provides faster response to load steps.

The benefits of a high inductor value include the following:

- Low ripple current leads to lower conduction losses in MOSFETs
- Low ripple; means lower RMS ripple current for capacitors
- Low ripple; yields low ac inductor losses in the core (flux) and windings (skin effect)
- Low ripple; gives continuous inductor current flow over a wide load range

For this design example a value of 10 µH was selected because of variations in temperature and inductor tolerance. Use Equation 9 to find the value of L_{Min}.

$$L_{Min} = \frac{(V_{I-Max} - V_O) \times V_O}{f_{SW} \times I_{Ripple} \times V_{I-Max}} = \frac{(27\ V - 5.45\ V) \times 5.45\ V}{2.5\ MHz \times 0.25\ A \times 27\ V} = 7\ \mu H \quad (9)$$

where

- f_{SW} = the regulator switching frequency
- I_{Ripple} = Allowable ripple current in the inductor, typically ±20% of maximum output load I_O

For this design, use Equation 10 to calculate the inductor peak current.

$$I_{L\text{-}Peak} = I_O + \frac{I_{Ripple}}{2} = 1\ A + \frac{0.25\ A}{2} = 1.125\ A \quad (10)$$

8.2.2.3 Output Capacitor Selection (C_O)

The selection of the output capacitor determines several parameters in the operation of the converter, the modulator pole, the voltage droop on the output capacitor, and the output ripple.

During a load step from no load to full load or changes in the input voltage, the output capacitor must hold up the output voltage above a certain level for a specified time and not issue a reset until the main regulator control loop responds to the change. The capacitance value determines the modulator pole and the rolloff frequency due to the LC output-filter double pole—the output ripple voltage is a product of the output capacitor ESR and ripple current.

Use Equation 11 to calculate the minimum capacitance required to maintain desired output voltage during a high-to-low load transition and prevent overshoot.

$$C_O = \frac{L\left(\left(I_{O-max}\right)^2 - \left(I_{O-min}\right)^2\right)}{\left(V_{O-max}\right)^2 - \left(V_{O-min}\right)^2} = \frac{10\ \mu H\left(\left(1\ A\right)^2 - \left(0.01\ A\right)^2\right)}{\left(5.6\ V\right)^2 - \left(5.3\ V\right)^2} = 3.06\ \mu F \quad (11)$$

where

- $I_{o\text{-}max}$ is the maximum output current
- $I_{o\text{-}min}$ is the minimum output current

The difference between the output current, maximum to minimum, is the worst-case load step in the system.

- $V_{o\text{-}max}$ is maximum tolerance of regulated output voltage
- $V_{o\text{-}min}$ is the minimum tolerance of regulated output voltage

Use Equation 12 to calculate the output capacitor root-mean-square (RMS) ripple current I_{O_RMS}. This is to prevent excess heating or failure because of high ripple currents.

This parameter is sometimes specified by the manufacturer. Therefore, because of variations in temperature and manufacture, use a 10-µF capacitor with a voltage rating greater than the maximum 10-V output.

$$I_{O\_RMS} = \frac{V_O \times \left(V_{I\text{-}max} - V_O\right)}{\sqrt{12} \times V_{I\text{-}max} \times L \times f_{SW}} = \frac{5.45\ V \times \left(27\ V - 5.45\ V\right)}{\sqrt{12} \times 27\ V \times 10\ \mu H \times 2.5\ MHz} = 0.050\ A \quad (12)$$

(13)

8.2.2.4 External Schottky Diode (D) Power Dissipation

The TPS65301-Q1 device requires an external ultrafast Schottky diode with fast reverse-recovery time connected between the PH and power ground pins. The diode conducts the output current during the off-state of the internal power switch. This diode must have a reverse breakdown higher than the maximum input voltage of the application. A Schottky diode is selected for its lower forward voltage. The Schottky diode is selected based on the appropriate power rating, which factors in the DC conduction losses and the AC losses because of the high switching frequencies. The power dissipation P_D is calculated with Equation 14.

$$P_D = I_O \times V_{FD} \times (1-D) + \frac{(V_I - V_{FD})^2 \times f_{SW} \times C_J}{2} = 1\ A \times 0.55\ V \times (1-0.389) + \frac{(14\ V - 0.55\ V)^2 \times 2.5\ MHz \times 30\ pF}{2} = 0.34\ W \quad (14)$$

where

- V_{FD} = forward conducting voltage of Schottky diode
- C_J = junction capacitance of the Schottky diode

8.2.2.5 Input Capacitor (C_I)

The TPS65301-Q1 requires an input ceramic decoupling capacitor type X5R or X7R and bulk capacitance to minimize input ripple voltage. The dc voltage rating of this input capacitance must be greater than the maximum input voltage. The capacitor must have an input ripple-current rating higher than the maximum input ripple current of the converter for the application. The input capacitors for power regulators are chosen to have reasonable capacitance-to-volume ratio and to be fairly stable over temperature. The value of the input capacitance is based on the input voltage desired (ΔV_I).

Use Equation 15 to calculate the input capacitance.

$$C_I = \frac{I_{O\_max} \times 0.25}{\Delta V_I \times f_{SW}} = \frac{1\ A \times 0.25}{0.3\ V \times 2.5\ MHz} = 0.33\ \mu F \quad (15)$$

Use Equation 16 to calculate the input-capacitor root-mean-square (RMS) ripple current I_{I_RMS}.

Because of variations in temperature and manufacture, use a 10-µF capacitor with a voltage rating greater than the maximum 45-V transient.

$$I_{I\_RMS} = I_O \times \sqrt{\frac{V_O}{V_{I\_min}} \times \left(\frac{V_{I\_min} - V_O}{V_{I\_min}} \right)} = 1\ A \times \sqrt{\frac{5.45\ V}{6\ V} \times \left(\frac{6\ V - 5.45\ V}{6\ V} \right)} = 0.29\ A \quad (16)$$

8.2.2.6 Loop Compensation

The double pole is due to the output-filter components inductor and capacitor. The calculations for the following equations use values taken from Figure 14.

8.2.2.6.1 Loop-Control Frequency Compensation

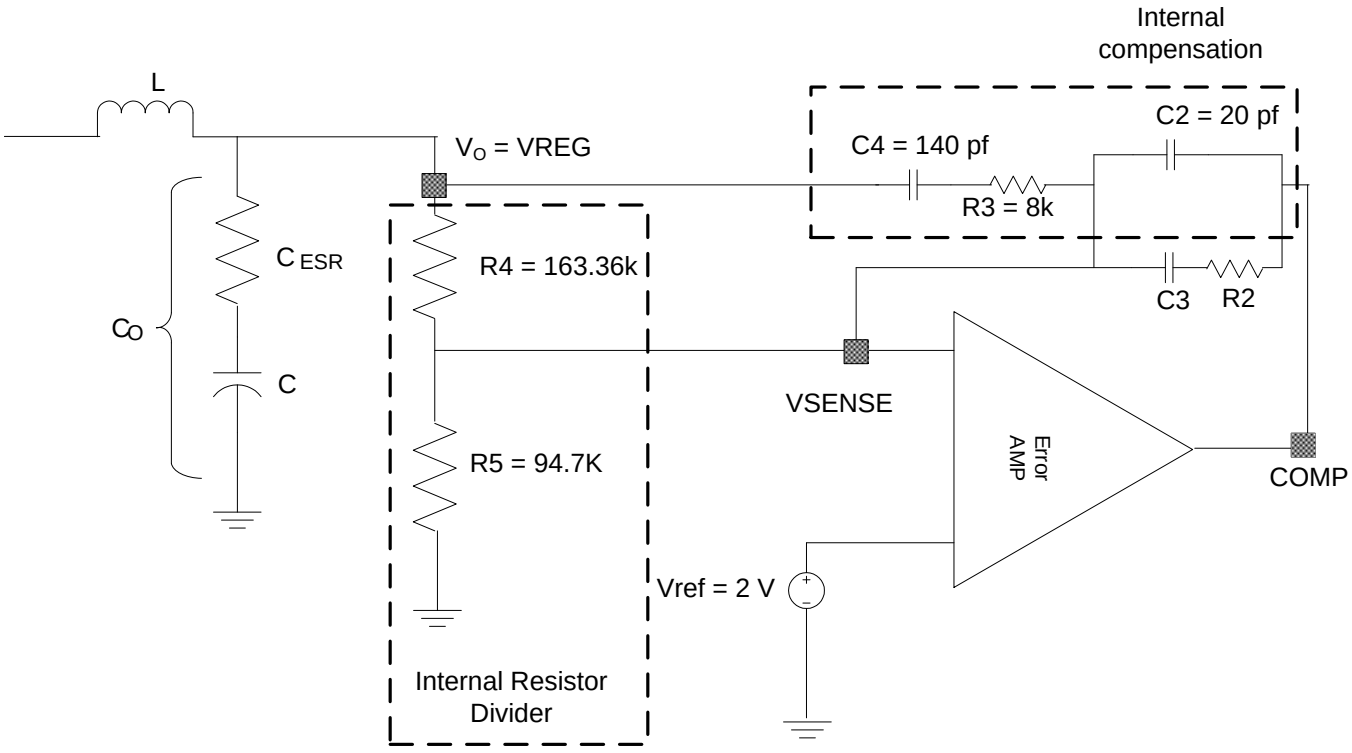


Figure 14. Loop-Control Frequency Compensation

8.2.2.6.1.1 Type III Compensation

$f_{CO} = f_{SW} \times 0.1$ (the cutoff frequency when the gain is 1 is called the unity-gain frequency).

f_{CO} is typically 1/5 to 1/10 of the switching frequency double-pole frequency response due to the LC output filter. The LC output filter gives a *double pole*, which has a –180° phase shift.

Make the two zeroes close to the double pole (LC), for example, $f_{Z1} \approx f_{Z2} \approx \frac{1}{2}\pi(LC_{OUT})^{\frac{1}{2}}$.

1. Make the first zero below the filter double pole (approximately 50% to 75% of f_{LC})
2. Make the second zero at the filter double pole (f_{LC})

   Make the two poles above the crossover frequency f_{CO}.
3. Make the first pole at the ESR frequency (f_{ESR})
4. Make the second pole at 0.5 the switching frequency

The following compensation components are integrated in the device with the following typical values. Guidelines for compensation components:

R3 = 8 kΩ, C4 = 140 pF, C2 = 20 pF

Use Equation 17 to calculate the double pole to calculate the output filter components LC.

$$f_{LC} = \frac{1}{2\pi\sqrt{LC_O}} = \frac{1}{2\pi\sqrt{10\ \mu H \times 10\ \mu F}} = 15.9\ kHz \quad (17)$$

The ESR of the output capacitor C gives a zero that has a 90° phase shift. The ESR of the output capacitor should be in the range of 1 mΩ to 100 mΩ. Use Equation 18 to calculate the value of f_{ESR}.

$$f_{ESR} = \frac{1}{2\pi \times C_O \times ESR} = \frac{1}{2\pi \times 10\ \mu F \times 0.005\ \Omega} = 3.2\ MHz \quad (18)$$

8.2.2.6.1.2 PWM Modulator Gain K

$$K = \frac{V_I}{V_{ramp}}$$

where

- $V_{ramp} = V_I / 10$, V_I = Input operating voltage (19)

8.2.2.6.1.3 Resistor Values

In this design example, select a value of 94.7 kΩ for R5 and use Equation 20 to calculate the value of R4.

$$R4 = \frac{R5 \times (V_O - Vref)}{Vref} = \frac{94.7\ k\Omega \times (5.45\ V - 2\ V)}{2\ V} = 163.36\ k\Omega \quad (20)$$

Use Equation 21 to calculate the value of R2 for this design example.

$$R2 = \frac{f_{CO} \times V_{ramp} \times R4}{f_{LC} \times V_I} = \frac{250\ kHz \times 1.4\ V \times 163.36\ k\Omega}{15.9\ kHz \times 14\ V} = 256.9\ k\Omega \quad (21)$$

Calculate C3 based on placing a zero at 50% to 75% of the output-filter double-pole frequency (below set at 50%).

For this design example, use Equation 22 to calculate the value of C3 as 786 pF.

$$C3 = \frac{1}{\pi \times R2 \times f_{LC}} = \frac{1}{\pi \times 256.9\ k\Omega \times 15.9\ kHz} = 786\ pF \quad (22)$$

8.2.2.6.1.4 Gain of Amplifier

$$A_V = \frac{R2 \times (R4 + R3)}{(R4 \times R3)} \quad (23)$$

8.2.2.6.1.5 Poles and Zero Frequencies

The following equations were used in this design example:

$$f_{P1} = \frac{1}{2\pi \times R2 \times C2} = \frac{1}{2\pi \times 256.9\ k\Omega \times 20\ pF} = 30.98\ kHz$$

$$f_{P2} = \frac{1}{2\pi \times R3 \times C4} = \frac{1}{2\pi \times 8\ k\Omega \times 140\ pF} = 142.1\ kHz$$

$$f_{Z1} = \frac{1}{2\pi \times R2 \times C3} = \frac{1}{2\pi \times 264.2\ k\Omega \times 76\ pF} = 7.93\ kHz$$

$$f_{Z2} = \frac{1}{2\pi \times R4 \times C4} = \frac{1}{2\pi \times 168\ k\Omega \times 140\ pF} = 6.77\ kHz \quad (24)$$

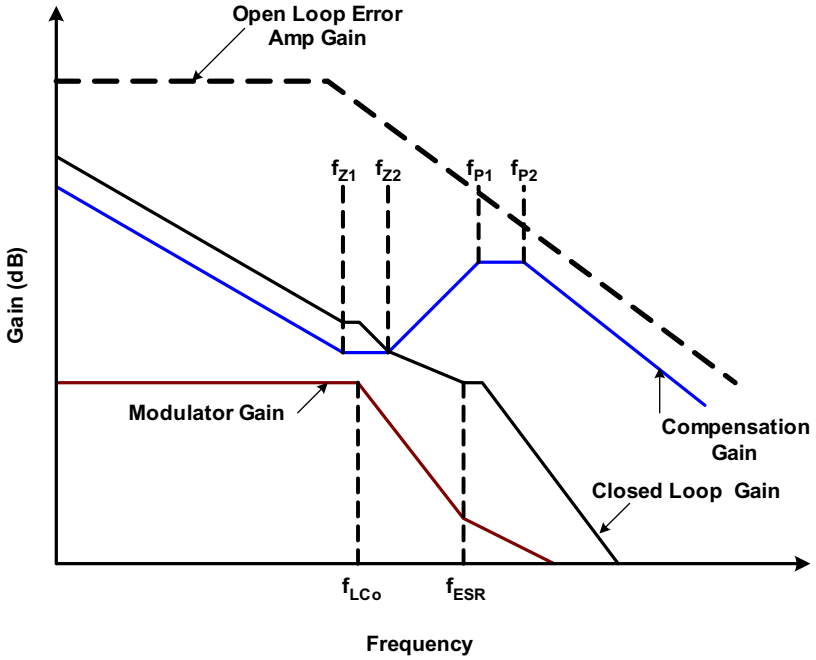


Figure 15. Typical Gain vs Frequency

8.2.2.7 Power Dissipation

8.2.2.7.1 Switch-Mode Power-Supply Losses

The power dissipation losses are applicable for continuous-conduction mode operation (CCM).

1. Conduction losses

   $P_{5.45V\_CON} = I_O^2 \times R_{ds(on)} \times (V_O/V_I)$

   where

   - I_O = Output current
   - V_O = VREG = Output voltage
   - V_I = Input voltage
   - f_{SW} = Switching frequency (25)

2. Switching losses

   $P_{5.45V\_SW} = ½ \times V_I \times I_O \times (t_r + t_{f)} \times f_{SW}$

   where

   - t_r = FET switching rise time (t_r max = 20 ns)
   - t_f = FET switching fall time (t_f max = 20 ns) (26)

3. Gate drive losses

   $P_{5.45V\_Gate} = V_{drive} \times Qg \times f_{sw}$

   where

   - V_{drive} = FET gate-drive voltage (typically V_{drive} = 6 V and V_{drive} max = 8 V)
   - $Qg = 1 \times 10^{-9}$ (nC) (typical) (27)

4. Supply losses

   $P_{IC} = V_I \times I_q$-normal (28)

Therefore:

$$P_{Total} = P_{CON} + P_{SW} + P_{Gate} + P_{5V\_Lin\ Reg} + P_{IC} \quad (29)$$

8.2.2.7.2 Linear Regulator (5V) and Sensor Supply (5VS)

$$P_{5V\_Lin\ Reg} = (VREG - 5\ V) \times I_{O\_5V} \quad (30)$$

$$P_{5VS\_Lin\ Reg} = (VREG - 5\ V) \times I_{O\_5VS} \quad (31)$$

Therefore, for this design, the following equations were used:

$$P_{5.45V\_CON} = I_O^2 \times Rds_{ON} \times (V_O / V_I) = (1\ A)^2 \times 0.5\ \Omega \times (5.45\ V / 14\ V) = 0.195\ W$$

$$P_{5.45\_SW} = 1/2 \times V_I \times I_O \times (tr + tf) \times f_{SW}$$
$$= 1/2 \times 14\ V \times 1\ A \times (20\ ns + 20\ ns) \times 2.5\ MHz = 0.7\ W$$

$$P_{5.45V\_Gate} = V_{drive} \times Q_g \times f_{SW} = 8\ V \times 1\ nC \times 2.5\ MHz = 0.02\ W$$

$$P_{5V\_Lin\_Reg} = (VREG - 5\ V) \times I_O = (5.45\ V - 5\ V) \times 0.2\ A = 0.09\ W$$

$$P_{5VS\_Lin\_Reg} = (VREG - 5\ V) \times I_O = (5.45\ V - 5\ V) \times 0.1\ A = 0.045\ W$$

$$P_{IC} = V_I \times I_{IC} = 14\ V \times 5.47\ mA = 0.08\ W$$

$$P_{Total} = P_{5.45V\_CON} + P_{5.45V\_SW} + P_{5.45V\_Gate} + P_{5V\_LinReg} + P_{5VS\_LinReg} + P_{IC}$$
$$= 0.195\ W + 0.7\ W + 0.02\ W + 0.09\ W + 0.045\ W + 0.08\ W = 1.13\ W \quad (32)$$

8.2.2.7.3 Total Power Dissipation

For given operating ambient temperature, T_A:

$$T_J = T_A + R_{th} \times P_{Total}$$

where

- T_J = Junction temperature in °C
- T_A = Ambient temperature in °C
- R_{th} = Thermal resistance of package in (°C/W)
- P_{Total} = Total power dissipation (watts) (33)

For a given max junction temperature $T_{J\text{-}Max}$ = 150°C

$$T_{A\text{-}Max} = T_{J\text{-}Max} - R_{th} \times P_{Total}$$

where

- $T_{A\text{-}Max}$ = Maximum ambient temperature in °C
- $T_{J\text{-}Max}$ = Maximum junction temperature in °C (34)

Other factors not included in the foregoing information which affect the overall efficiency and power losses are

- Inductor AC and DC losses
- Trace resistance and losses associated with the copper trace routing connection
- Schottky diode

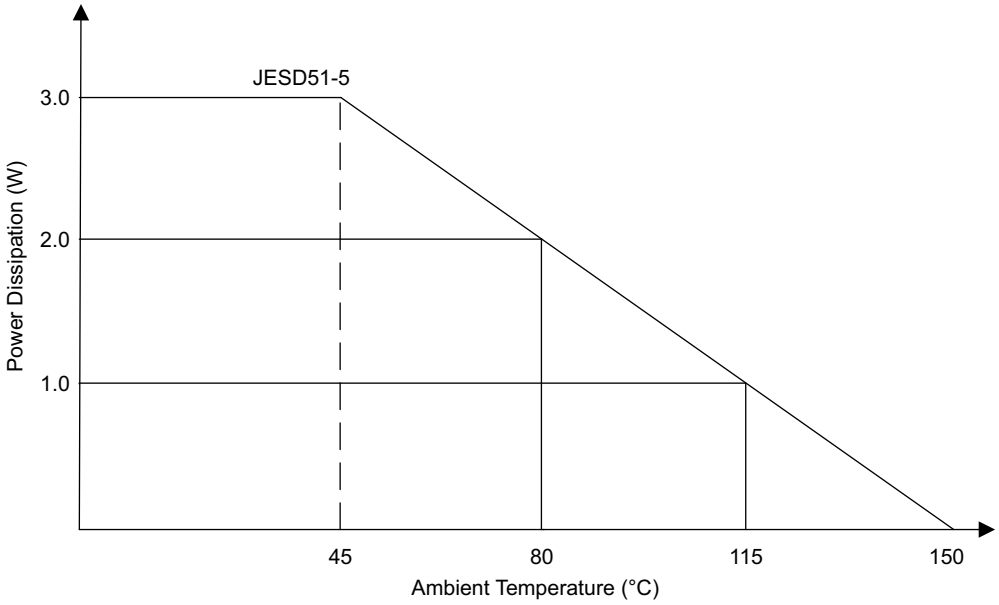


Figure 16. Power Dissipation Derating Profile, 24-Pin PWP Package With Thermal Pad

8.2.3 Application Curves

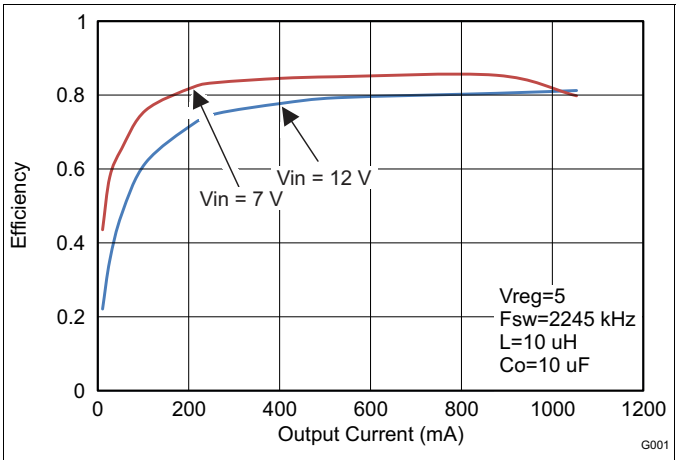


Figure 17. Efficiency vs Output Current on VREG

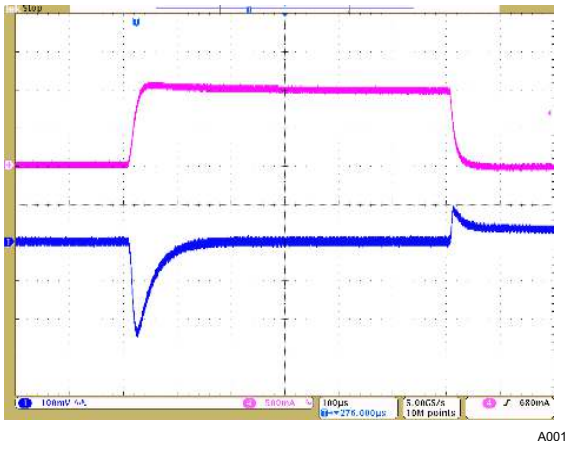
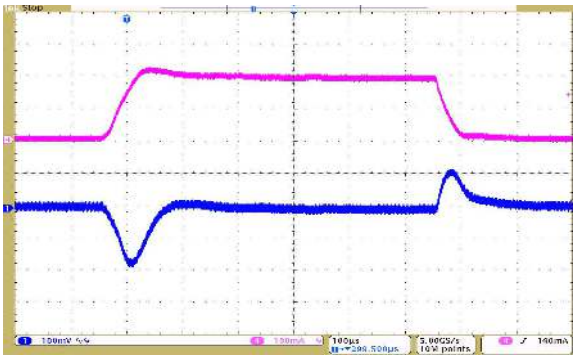
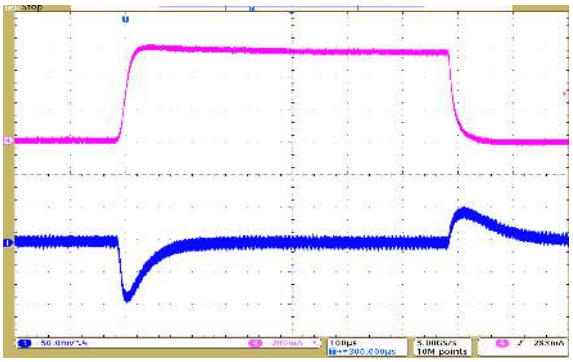


Figure 18. VREG Load Transient Response, 10 mA to 1 A



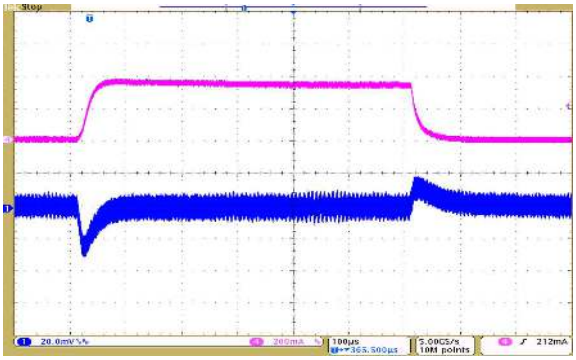
A002

Figure 19. VREG Load Transient Response, 10 mA to 200 mA



A003

Figure 20. VREG Load Transient Response, 10 mA to 550 mA



A004

Figure 21. VREG Load Transient Response, 10 mA to 350 mA

9 Power Supply Recommendations

The TPS65301-Q1 device is designed to operate using an input supply voltage range from 5.6 V to 40 V.

10 Layout

10.1 Layout Guidelines

The following guidelines are recommended for PCB layout of the TPS65301-Q1 device.

10.1.1 Inductor (L)

Use a low-EMI inductor with a ferrite-type shielded core. Other types of inductors may be used; however, they must have low-EMI characteristics and be located away from the low-power traces and components in the circuit.

10.1.2 Input Filter Capacitors (C_I)

Input ceramic filter capacitors should be located in close proximity to the VIN pin. Surface-mount capacitors are recommended to minimize lead length and reduce noise coupling.

10.1.3 Feedback

Route the feedback trace such that there is minimum interaction with any noise sources associated with the switching components. Recommended practice is to ensure placing the inductor away from the feedback trace to prevent a source of EMI noise.

10.1.4 Traces and Ground Plane

All power (high-current) traces should be thick and as short as possible. The inductor and output capacitors should be as close to each other as possible. This reduces EMI radiated by the power traces due to high switching currents.

In a two-sided PCB it is recommended to have ground planes on both sides of the PCB to help reduce noise and ground-loop errors. The ground connection for the input and output capacitors and IC ground should be connected to this ground plane.

In a multi-layer PCB, the ground plane is used to separate the power plane (where high switching currents and components are placed) from the signal plane (where the feedback trace and components are) for improved performance.

Also arrange the components such that the switching-current loops curl in the same direction. Place the high-current components such that during conduction the current path is in the same direction. This prevents magnetic field reversal caused by the traces between the two half-cycles, helping to reduce radiated EMI.

10.2 Layout Example

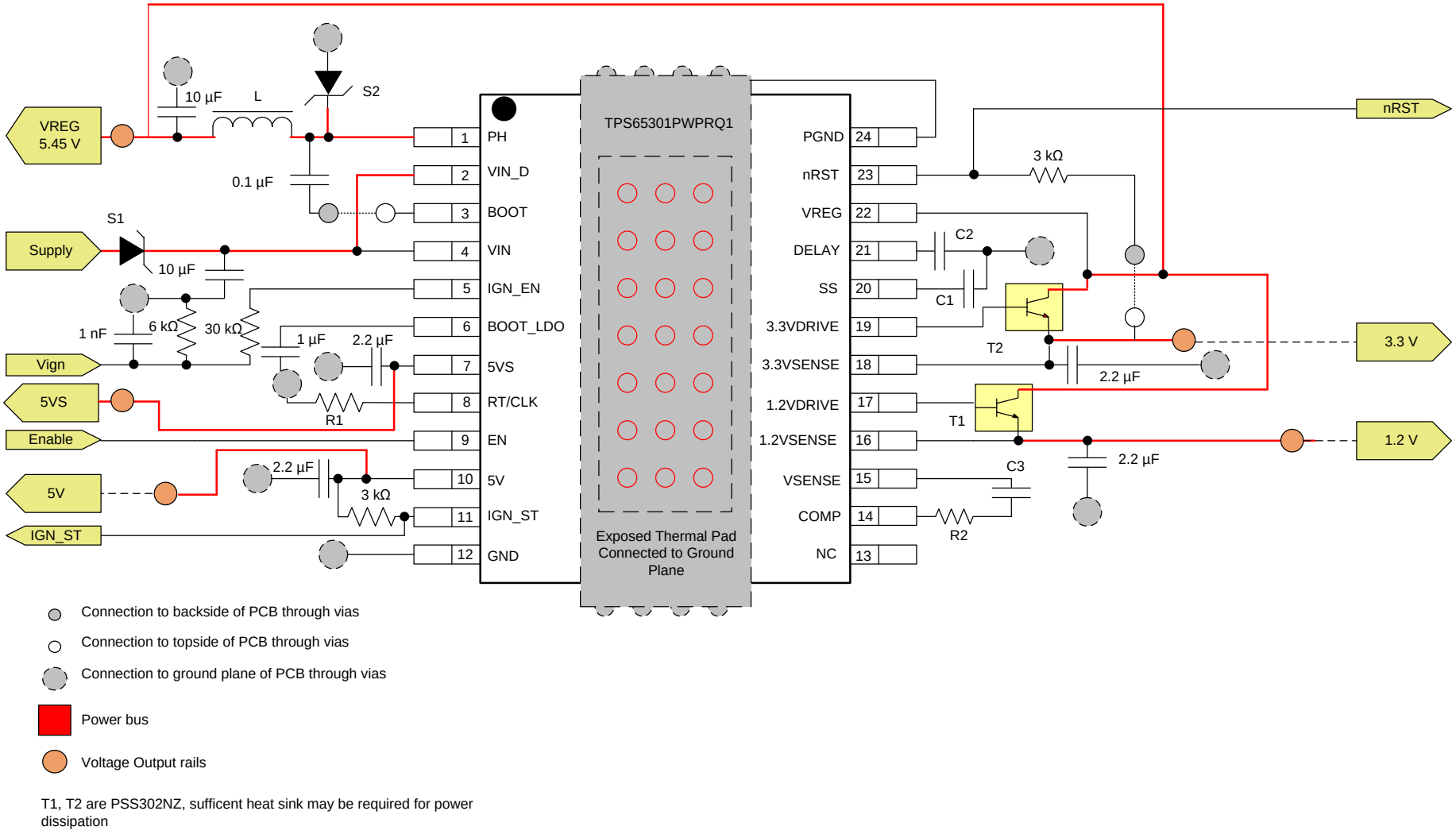


Figure 22. PCB Layout

11 Device and Documentation Support

11.1 Documentation Support

11.1.1 Related Documentation

For related documentation see the following:

TPS65301EVM User's Guide, SLVU929

11.2 Community Resource

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 Trademarks

E2E is a trademark of Texas Instruments.
All other trademarks are the property of their respective owners.

11.4 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.

11.5 Glossary

SLYZ022 — *TI Glossary*.

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

12 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.

PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead finish/ Ball material (6)	MSL Peak Temp (3)	Op Temp (°C)	Device Marking (4/5)	Samples
TPS65301QPWPRQ1	ACTIVE	HTSSOP	PWP	24	2000	RoHS & Green	NIPDAU	Level-3-260C-168 HR	-40 to 125	TPS65301	Samples

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

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

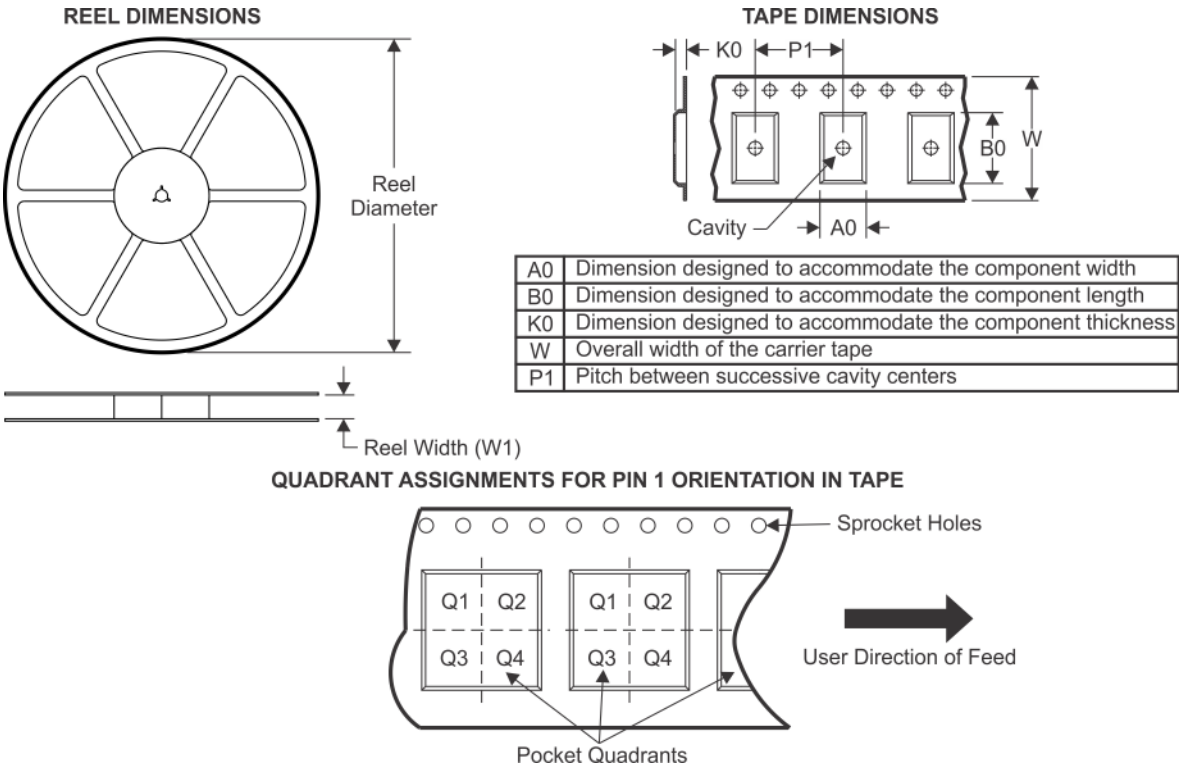
(3) MSL, Peak Temp. - The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

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

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

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

TAPE AND REEL INFORMATION

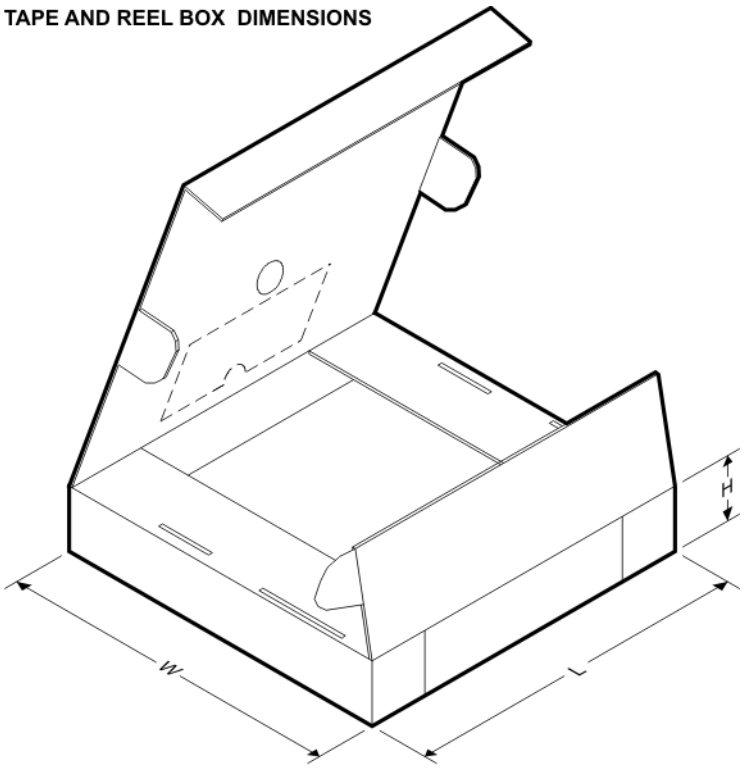


A0	Dimension designed to accommodate the component width
B0	Dimension designed to accommodate the component length
K0	Dimension designed to accommodate the component thickness
W	Overall width of the carrier tape
P1	Pitch between successive cavity centers

*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
TPS65301QPWPRQ1	HTSSOP	PWP	24	2000	330.0	16.4	6.95	8.3	1.6	8.0	16.0	Q1

TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS65301QPWPRQ1	HTSSOP	PWP	24	2000	350.0	350.0	43.0

GENERIC PACKAGE VIEW

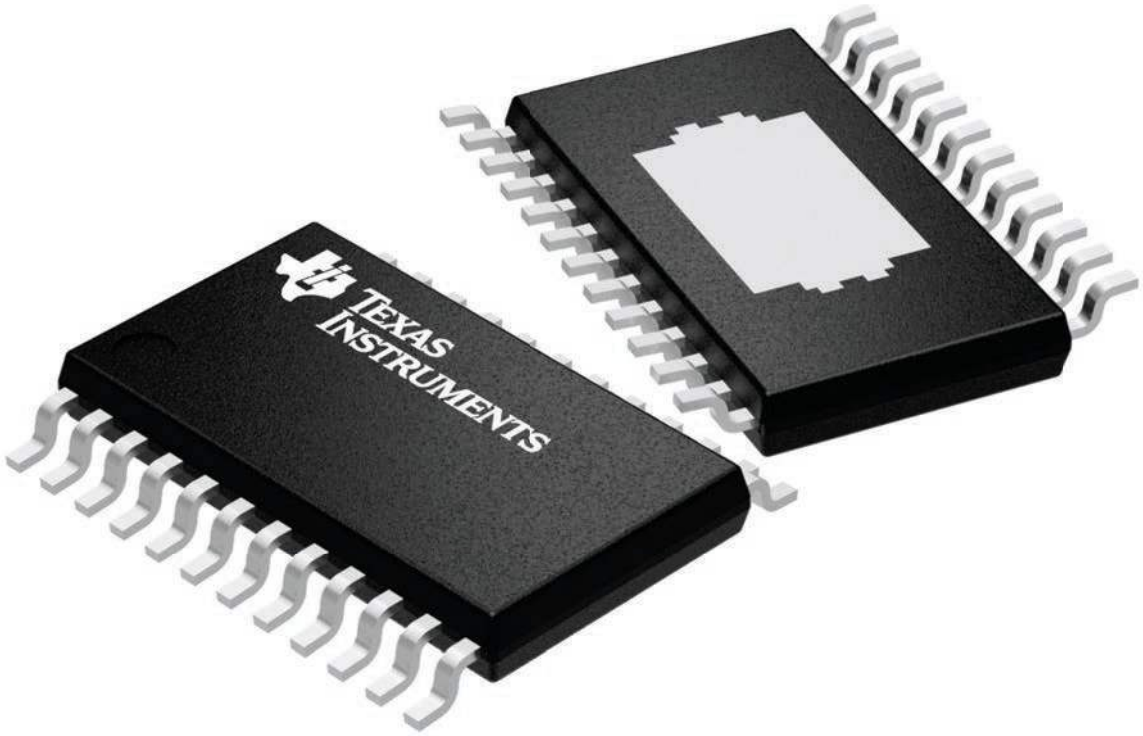
PWP 24

PowerPAD™ TSSOP - 1.2 mm max height

4.4 x 7.6, 0.65 mm pitch

PLASTIC SMALL OUTLINE

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



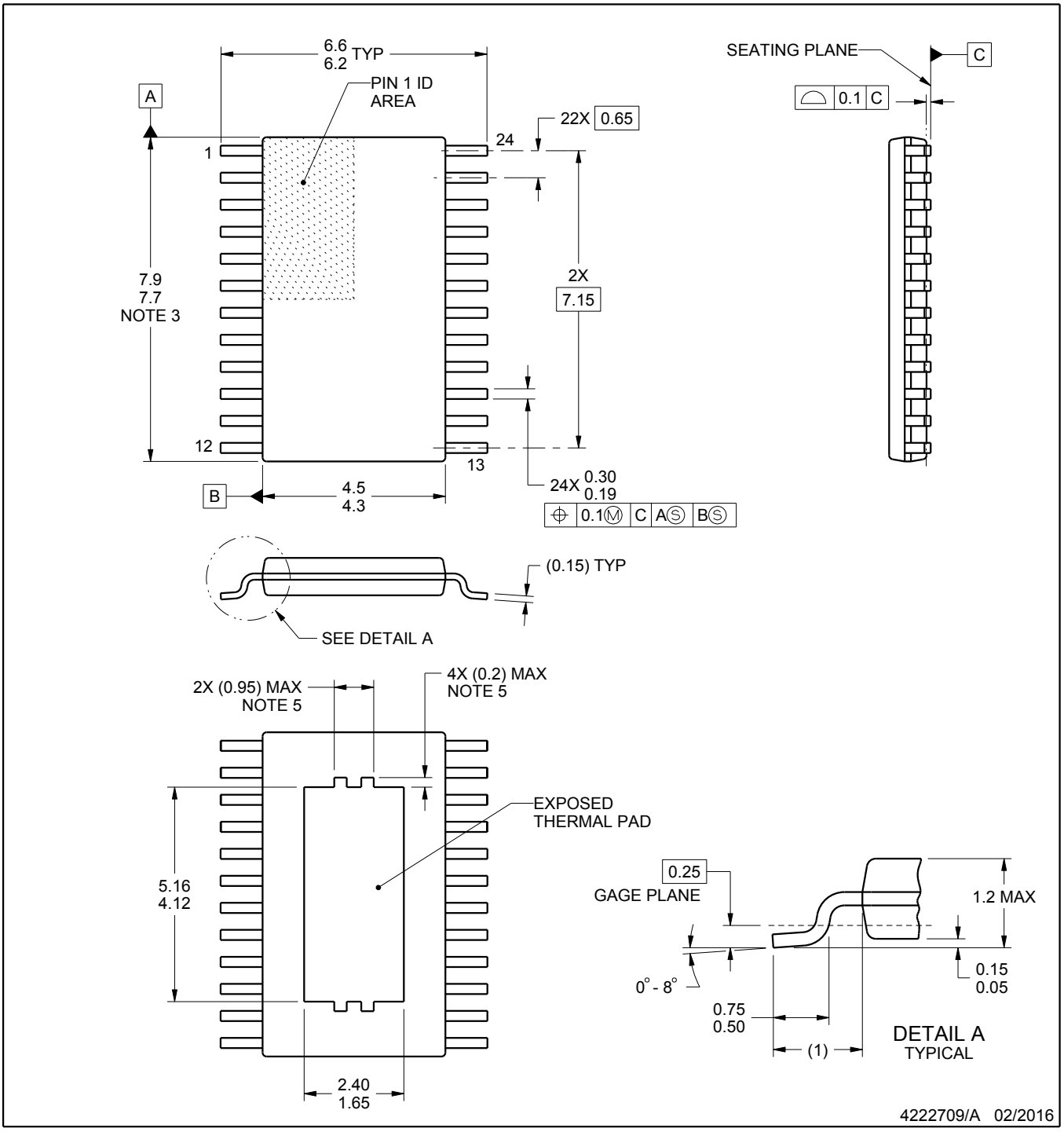
4224742/B

PACKAGE OUTLINE

PWP0024B

PowerPAD™ TSSOP - 1.2 mm max height

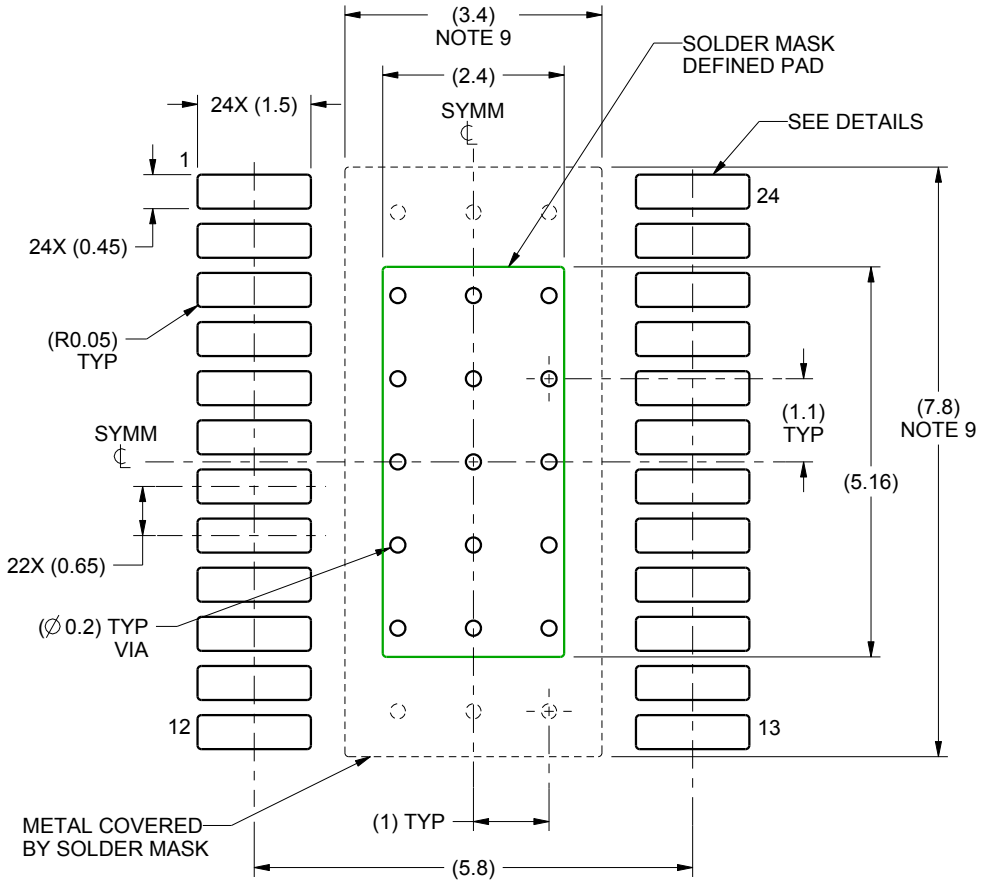
PLASTIC SMALL OUTLINE



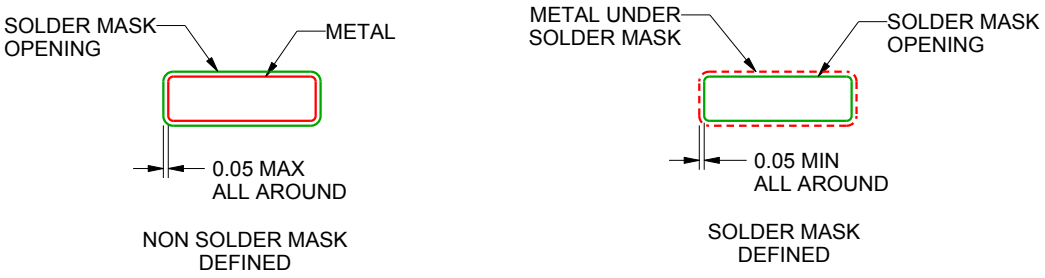
PowerPAD is a trademark of Texas Instruments.

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. This dimension does not include mold flash, protrusions, or gate burrs. Mold flash, protrusions, or gate burrs shall not exceed 0.15 mm per side.
4. Reference JEDEC registration MO-153.
5. Features may not be present and may vary.



LAND PATTERN EXAMPLE
SCALE:10X

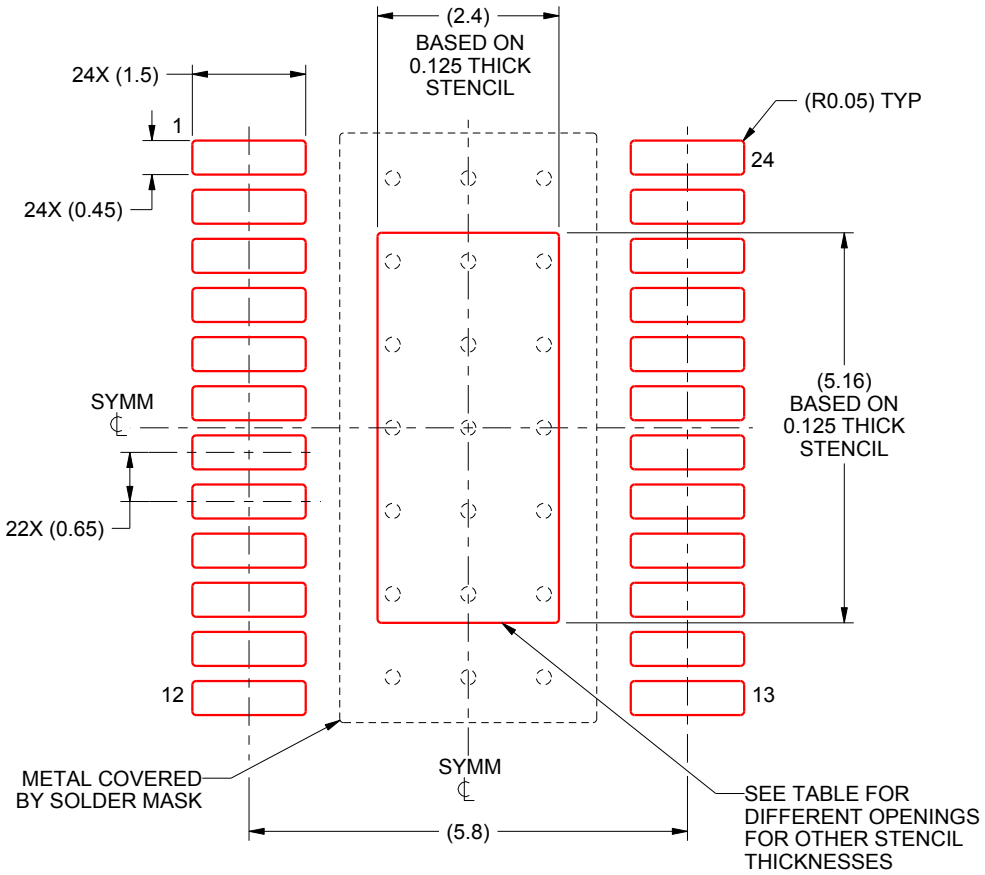


SOLDER MASK DETAILS
PADS 1-24

4222709/A 02/2016

NOTES: (continued)

6. Publication IPC-7351 may have alternate designs.
7. Solder mask tolerances between and around signal pads can vary based on board fabrication site.
8. This package is designed to be soldered to a thermal pad on the board. For more information, see Texas Instruments literature numbers SLMA002 (www.ti.com/lit/slma002) and SLMA004 (www.ti.com/lit/slma004).
9. Size of metal pad may vary due to creepage requirement.



SOLDER PASTE EXAMPLE
EXPOSED PAD
100% PRINTED SOLDER COVERAGE BY AREA
SCALE:10X

STENCIL THICKNESS	SOLDER STENCIL OPENING
0.1	2.68 X 5.77
0.125	2.4 X 5.16 (SHOWN)
0.15	2.19 X 4.71
0.175	2.03 X 4.36

4222709/A 02/2016

NOTES: (continued)

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

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