TPS40210-EP 4.5-V to 52-V Input Current Mode Boost Controller

1 Features

- For Boost, Flyback, SEPIC, LED Drive Apps
- Wide Input Operating Voltage: 4.5 V to 52 V
- Adjustable Oscillator Frequency
- Fixed Frequency Current Mode Control
- Internal Slope Compensation
- Integrated Low-Side Driver
- Programmable Closed-Loop Soft-Start
- Overcurrent Protection
- External Synchronization Capable
- Reference 700 mV
- Low Current Disable Function

2 Applications

- LED Lighting
- Industrial Control Systems
- Battery Powered Systems

3 Description

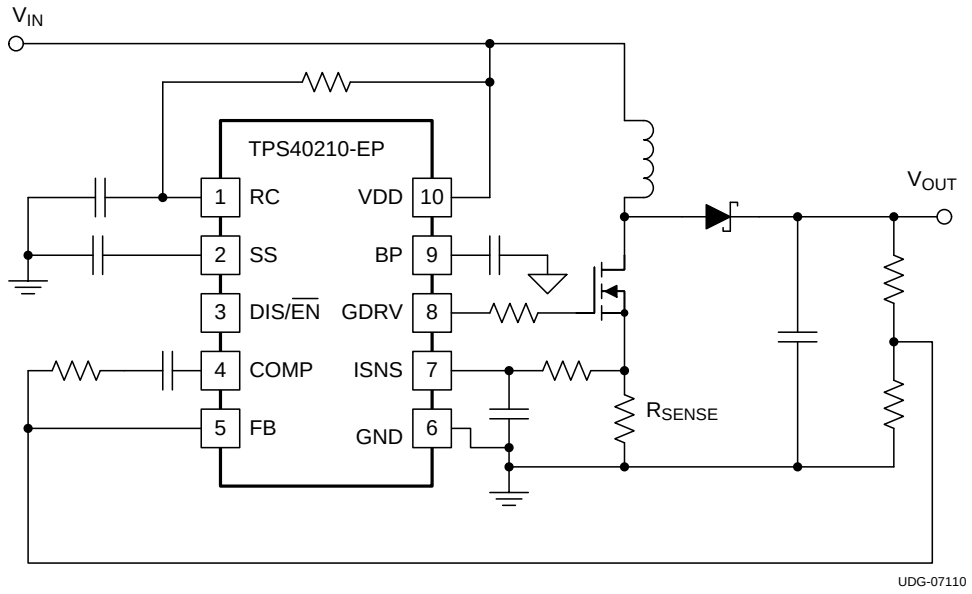
The TPS40210-EP is a wide-input voltage (4.5 to 52 V), nonsynchronous boost controller. The device is suitable for topologies which require a grounded source N-channel FET including boost, flyback, SEPIC, and various LED driver applications. The device features include programmable soft-start, overcurrent protection with automatic retry, and programmable oscillator frequency. Current mode control provides improved transient response and simplified loop compensation.

Device Information[1]

PART NUMBER	PACKAGE	BODY SIZE (NOM)
TPS40210-EP	VSON (10)	3.05 mm × 4.98 mm

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

Simplified Schematic



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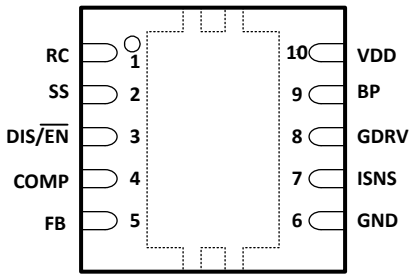
4 Revision History

DATE	REVISION	NOTES
November 2015	*	Initial release.

5 Pin Configuration and Functions

DRC Surface Mount Package
10-Pin VSON
Top View



Pin Functions

PIN		I/O	DESCRIPTION
NAME	NO.		
BP	9	O	Regulator output pin. Connect a 1.0-µF bypass capacitor from this pin to GND.
COMP	4	O	Error amplifier output. Connect control loop compensation network between COMP pin and FB pin.
DIS/$\overline{EN}$	3	I	Disable pin. Pulling this pin high, places the part into a shutdown mode. Shutdown mode is characterized by a very low quiescent current. While in shutdown mode, the functionality of all blocks is disabled and the BP regulator is shut down. This pin has an internal 1 MΩ pull-down resistor to GND. Leaving this pin unconnected enables the device.
FB	5	I	Error amplifier inverting input. Connect a voltage divider from the output to this pin to set output voltage. Compensation network is connected between this pin and COMP.
GDRV	8	O	Connect the gate of the power N channel MOSFET to this pin.
GND	6	—	Device ground.
ISNS	7	I	Current sense pin. Connect an external current sensing resistor between this pin and GND. The voltage on this pin is used to provide current feedback in the control loop and detect an overcurrent condition. An overcurrent condition is declared when ISNS pin voltage exceeds the overcurrent threshold voltage, 150 mV typical.
RC	1	I	Switching frequency setting pin. Connect a resistor from RC pin to VDD of the IC power supply and a capacitor from RC to GND.
SS	2	I	Soft-start time programming pin. Connect capacitor from SS pin to GND to program converter soft-start time. This pin also functions as a timeout timer when the power supply is in an overcurrent condition.
VDD	10	I	System input voltage. Connect a local bypass capacitor from this pin to GND. Depending on the amount of required slope compensation, this pin can be connected to the converter output. See *Application Information* for additional details.

6 Specifications

6.1 Absolute Maximum Ratings

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

			MIN	MAX	UNIT
	Input voltage	VDD	–0.3	52	V
		RC, SS, FB, DIS/$\overline{EN}$	–0.3	10	
		ISNS	–0.3	8	
	Output voltage	COMP, BP, GDRV	–0.3	9	
T_J	Operating junction temperature		–55	150	°C
T_{stg}	Storage temperature		–55	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.

6.2 ESD Ratings

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

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

6.3 Recommended Operating Conditions

		MIN	MAX	UNIT
V_{DD}	Input voltage	4.5	52	V
T_J	Operating junction temperature	–55	125	°C

6.4 Thermal Information

THERMAL METRIC(1)		TPS40210-EP DRC (VSON) 10 PINS	UNIT
$R_{\theta JA}$	Junction-to-ambient thermal resistance	67.2	°C/W
$R_{\theta JC(top)}$	Junction-to-case (top) thermal resistance	50.5	°C/W
$R_{\theta JB}$	Junction-to-board thermal resistance	41.0	°C/W
ψ_{JT}	Junction-to-top characterization parameter	2.4	°C/W
ψ_{JB}	Junction-to-board characterization parameter	40.7	°C/W
$R_{\theta JC(bot)}$	Junction-to-case (bottom) thermal resistance	15.6	°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 Electrical Characteristics

T_J = –55°C to 125°C, V_{DD}= 12V_{dc}, all parameters at zero power dissipation (unless otherwise noted)

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
VOLTAGE REFERENCE						
V_{FB}	Feedback voltage range	COMP = FB, 4.5 ≤ V_{DD} ≤ 52 V, T_J = 25°C	693	700	707	mV
		COMP = FB, 4.5 ≤ V_{DD} ≤ 52 V, –55°C ≤ T_J ≤ 125°C	686	700	714	
INPUT SUPPLY						
V_{DD}	Input voltage range		4.5		52	V
I_{DD}	Operating current	4.5 ≤ V_{DD} ≤ 52 V, no switching, V_{DIS} < 0.8		1.5	2.5	mA
		2.5 ≤ V_{DIS} ≤ 7 V		10	20	µA
		V_{DD} < $V_{UVLO(on)}$, V_{DIS} < 0.8			530	µA
UNDERVOLTAGE LOCKOUT						
$V_{UVLO(on)}$	Turn on threshold voltage		4.00	4.25	4.50	V
$V_{UVLO(hyst)}$	UVLO hysteresis		140	195	240	mV
OSCILLATOR						
f_{OSC}	Oscillator frequency range(1)		35		1000	kHz
	Oscillator frequency	R_{RC} = 182 kΩ, C_{RC} = 330 pF	260	300	340	
	Frequency line regulation	4.5 ≤ V_{DD} ≤ 52 V	-20%		7%	
V_{SLP}	Slope compensation ramp		520	620	720	mV
PWM						
$t_{ON(min)}$	Minimum pulse width	V_{DD} = 12 V(1)		275	400	ns
		V_{DD} = 30 V		90	200	
$t_{OFF(min)}$	Minimum off time			170	200	
V_{VLY}	Valley voltage			1.2		V
SOFT-START						
$V_{SS(ofst)}$	Offset voltage from SS pin to error amplifier input			700		mV
$R_{SS(chg)}$	Soft-start charge resistance		320	430	620	kΩ
$R_{SS(dchg)}$	Soft-start discharge resistance		840	1200	1600	
ERROR AMPLIFIER						
GBWP	Unity gain bandwidth product(1)		1.5	3.0		MHz
A_{OL}	Open loop gain(1)		60	80		dB
$I_{IB(FB)}$	Input bias current (current out of FB pin)			100	300	nA
$I_{COMP(src)}$	Output source current	V_{FB} = 0.6 V, V_{COMP} = 1 V	100	250		µA
$I_{COMP(snk)}$	Output sink current	V_{FB} = 1.2 V, V_{COMP} = 1 V	1.2	2.5		mA
OVERCURRENT PROTECTION						
$V_{ISNS(oc)}$	Overcurrent detection threshold (at ISNS pin)	4.5 ≤ V_{DD} < 52 V, –55°C ≤ T_J ≤ 125°C	120	150	180	mV
D_{OC}	Overcurrent duty cycle(1)				2%	
$V_{SS(rst)}$	Overcurrent reset threshold voltage (at SS pin)		100	150	350	mV
T_{BLNK}	Leading edge blanking(1)			75		ns

(1) Ensured by design. Not production tested.

Electrical Characteristics (continued)

T_J = –55°C to 125°C, V_{DD}= 12V_{dc}, all parameters at zero power dissipation (unless otherwise noted)

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
CURRENT SENSE AMPLIFIER						
A_{CS}	Current sense amplifier gain		4..2	5.6	7.4	V/V
$I_{B(ISNS)}$	Input bias current			1	3	µA
DRIVER						
$I_{GDRV(src)}$	Gate driver source current	V_{GDRV} = 4 V, T_J = 25°C	375	400		mA
$I_{GDRV(snk)}$	Gate driver sink current	V_{GDRV} = 4 V, T_J = 25°C	330	400		
LINEAR REGULATOR						
V_{BP}	Bypass voltage output	0 mA < I_{BP} < 15 mA	7	8	9	V
DISABLE/ENABLE						
$V_{DIS(en)}$	Turn-on voltage		0.7		1.3	V
$V_{DIS(hys)}$	Hysteresis voltage		25	130	220	mV
R_{DIS}	DIS pin pulldown resistance		0.7	1.1	1.6	MΩ

6.6 Typical Characteristics

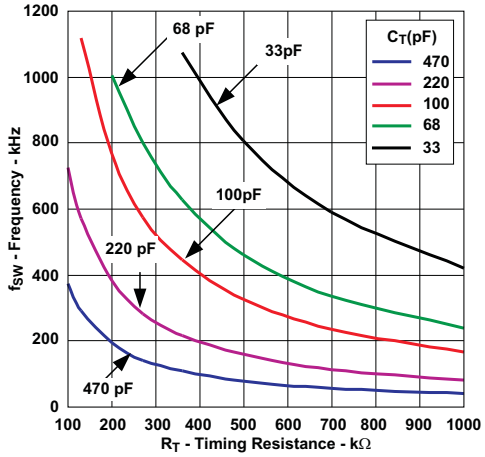


Figure 1. Frequency vs Timing Resistance

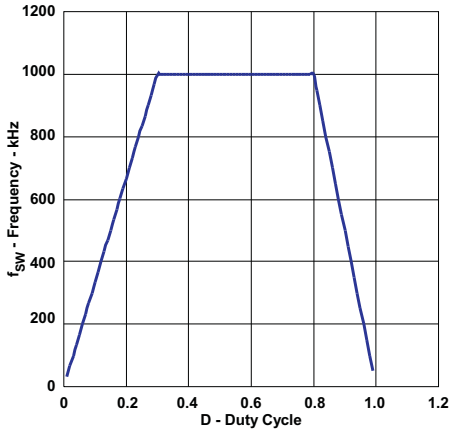


Figure 2. Switching Frequency vs Duty Cycle

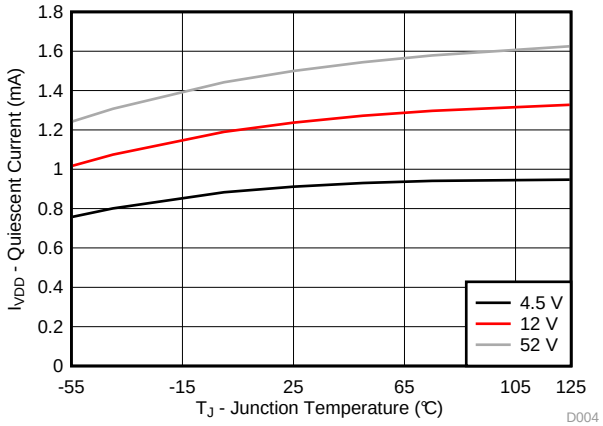


Figure 3. Quiescent Current vs Junction Temperature

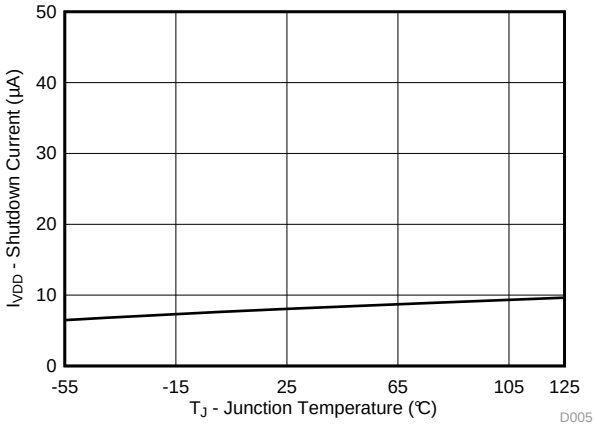


Figure 4. Shutdown Current vs Junction Temperature

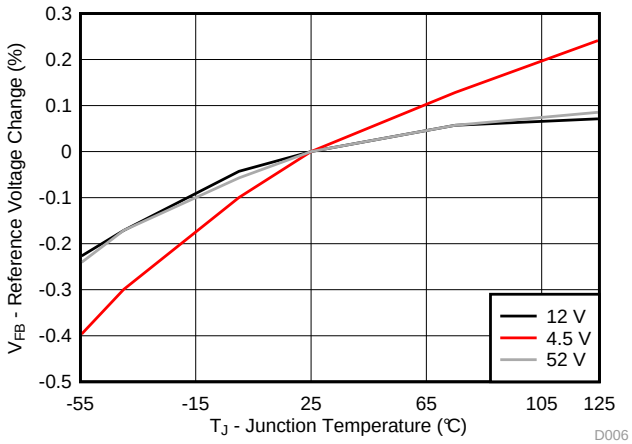


Figure 5. Reference Voltage Change vs Junction Temperature

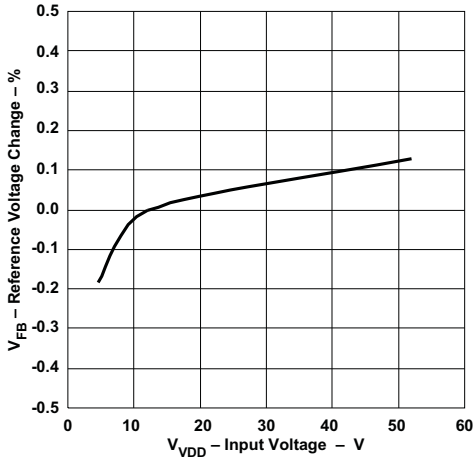


Figure 6. Reference Voltage Change vs Input Voltage

Typical Characteristics (continued)

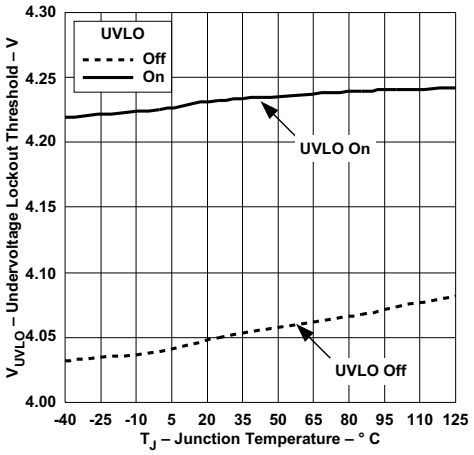


Figure 7. Undervoltage Lockout Threshold vs Junction Temperature

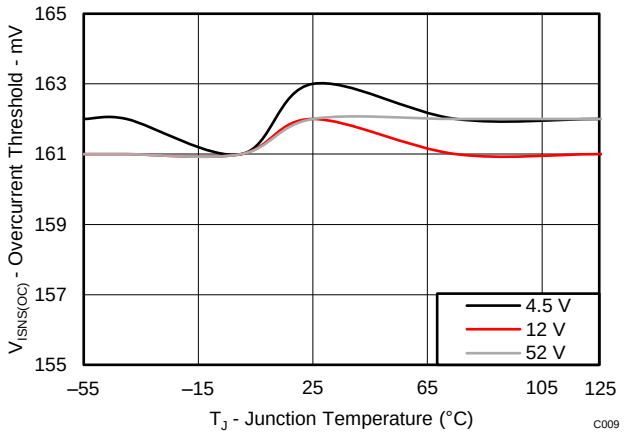


Figure 8. Overcurrent Threshold vs Junction Temperature

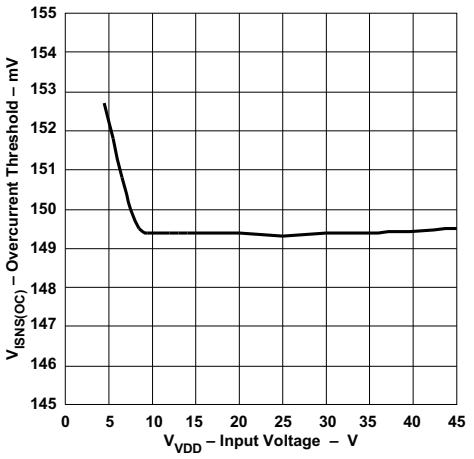


Figure 9. Overcurrent Threshold vs Input Voltage

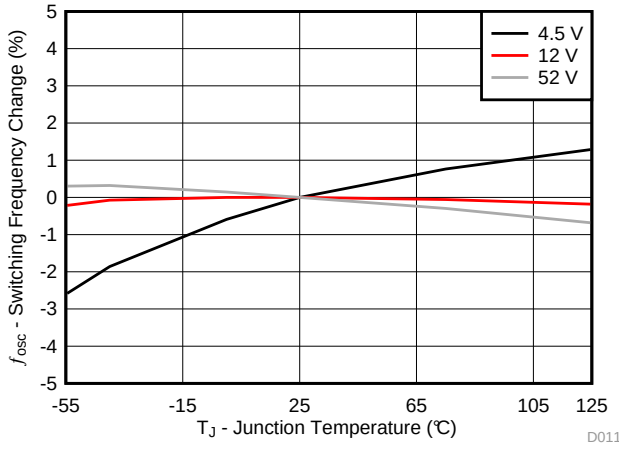


Figure 10. Switching Frequency Change vs Junction Temperature

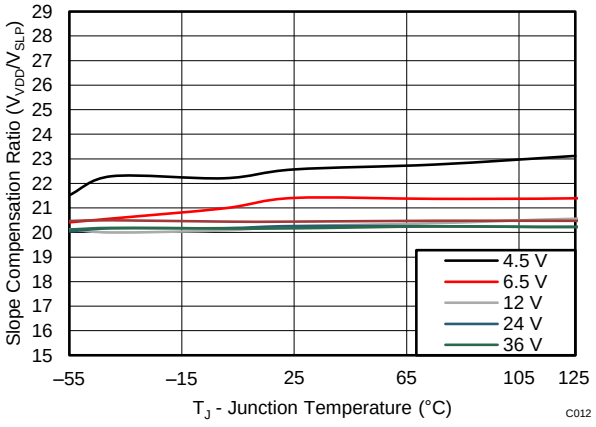


Figure 11. Oscillator Amplitude vs Junction Temperature

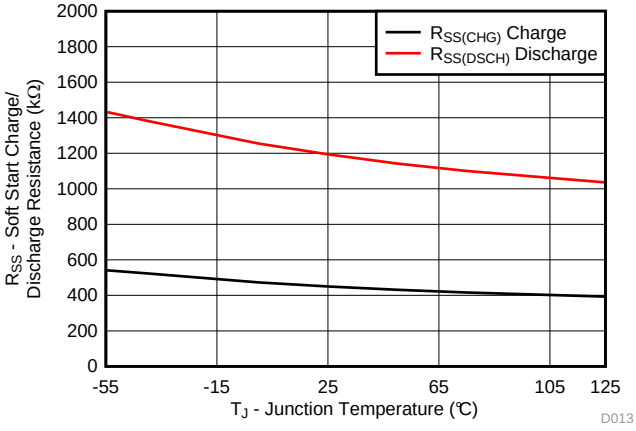


Figure 12. Soft-Start Charge/Discharge Resistance vs Junction Temperature

Typical Characteristics (continued)

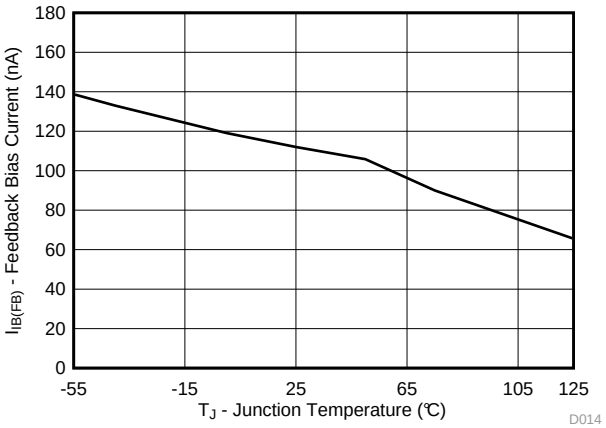


Figure 13. Fb Bias Current vs Junction Temperature

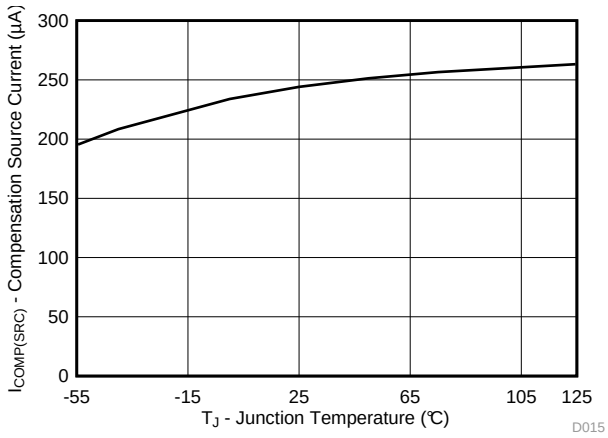


Figure 14. Compensation Source Current vs Junction Temperature

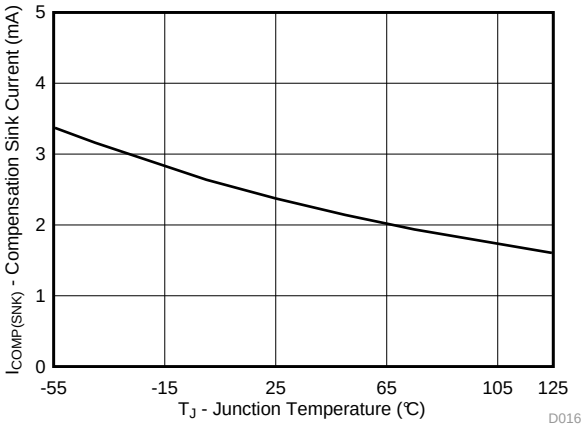


Figure 15. Compensation Sink Current vs Junction Temperature

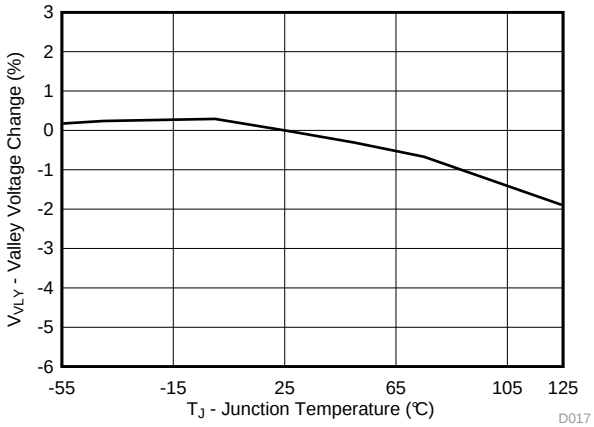


Figure 16. Valley Voltage Change vs Junction Temperature

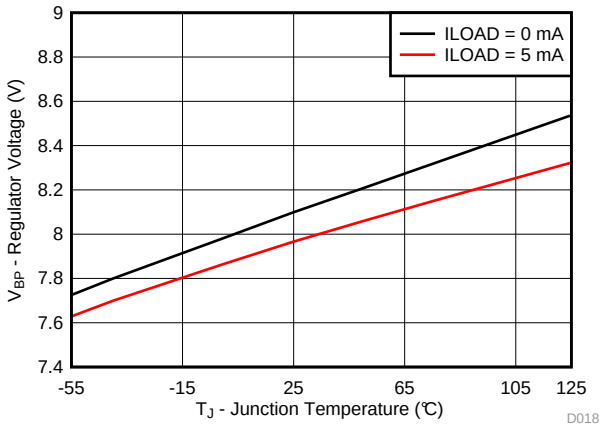


Figure 17. Regulator Voltage vs Junction Temperature

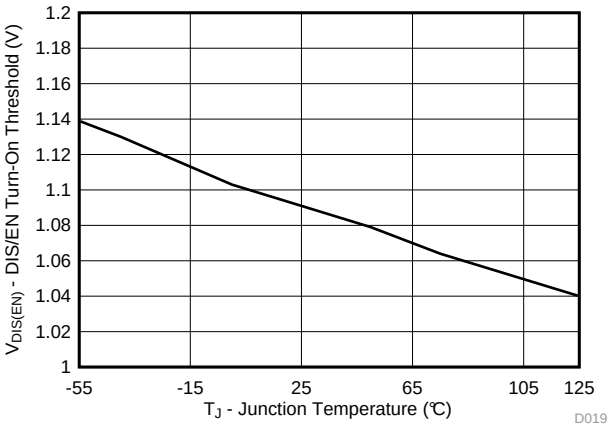


Figure 18. DIS/$\overline{EN}$ Turn-On Threshold vs Junction Temperature

Typical Characteristics (continued)

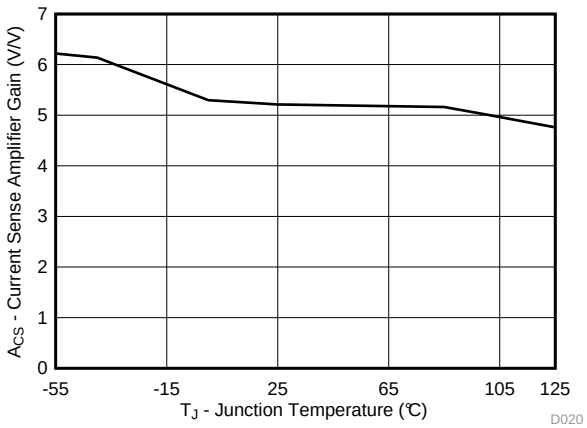


Figure 19. Current Sense Amplifier Gain vs Junction Temperature

7 Detailed Description

7.1 Overview

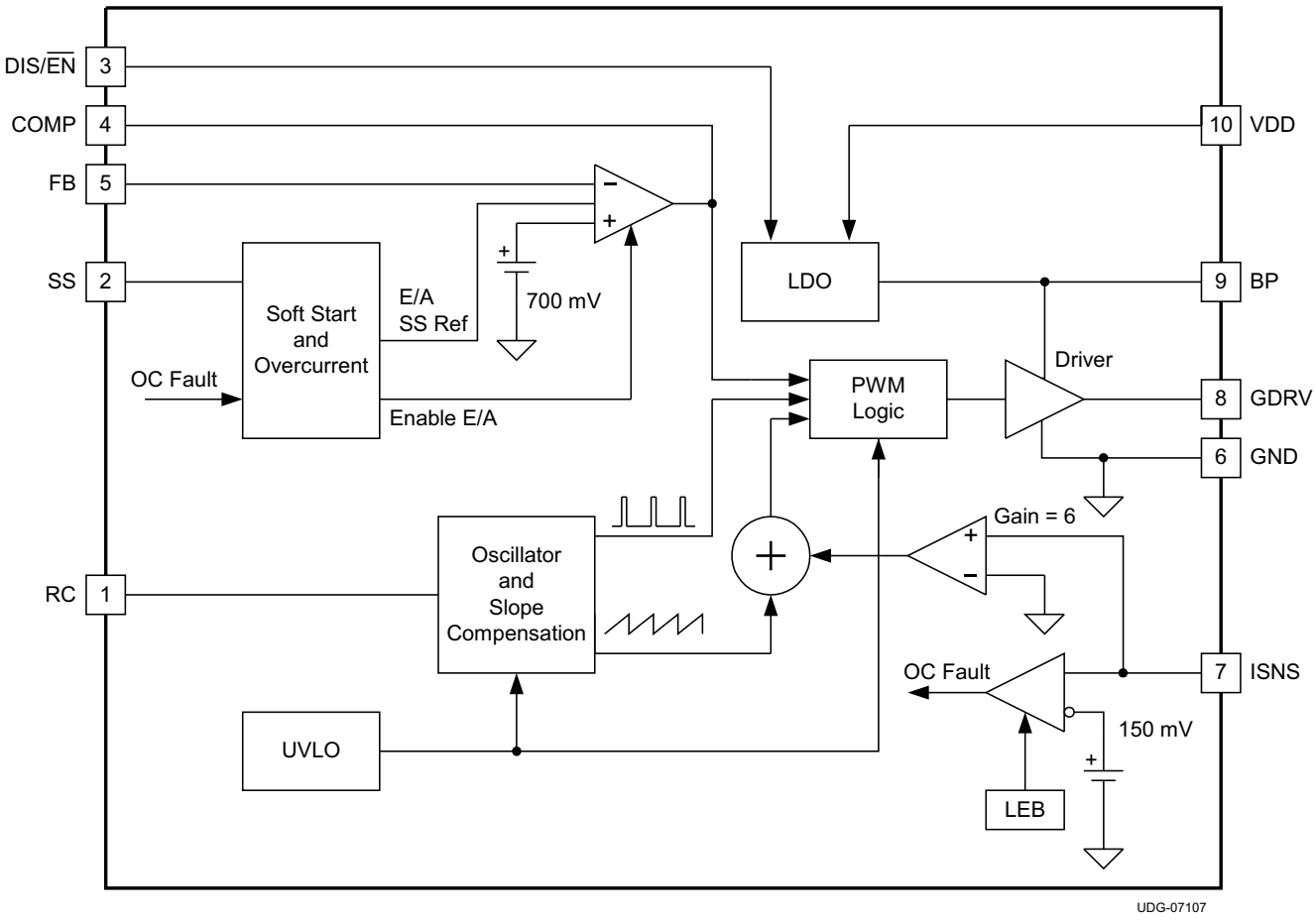
The TPS40210-EP is a peak current-mode control low-side controller with a built in 400-mA gate driver designed to drive N-channel MOSFETs at a fixed frequency. The frequency is adjustable from 35 kHz to 1000 kHz. Small size combined with complete functionality makes the part both versatile and easy to use.

The controller uses a low-value current-sensing resistor in series with the power MOSFET's source connection to detect switching current. When the voltage drop across this resistor exceeds 150 mV, the part enters an hiccup fault mode with a time period set by the external soft-start capacitor.

The TPS40210-EP uses voltage feedback to an error amplifier that is biased by a precision 700-mV reference. Internal slope compensation eliminates the characteristic sub-harmonic instability of peak current mode control with duty cycles of 50% or greater.

The TPS40210-EP also incorporates a soft-start feature where the output follows a slowly rising soft-start voltage, preventing output-voltage overshoot. The DIS/EN disables the TPS40210-EP putting it in a low quiescent current shutdown mode.

7.2 Functional Block Diagram



7.3 Feature Description

7.3.1 Soft-Start

The soft-start feature of the TPS40210-EP is a closed-loop soft-start, meaning that the output voltage follows a linear ramp that is proportional to the ramp generated at the SS pin. This ramp is generated by an internal resistor connected from the BP pin to the SS pin and an external capacitor connected from the SS pin to GND. The SS pin voltage (V_{SS}) is level shifted down by approximately $V_{SS(ofst)}$ (approximately 700 mV) and sent to one of the "+" (the "+" input with the lowest voltage dominates) inputs of the error amplifier. When this level shifted voltage (V_{SSE}) starts to rise at time t_1 (see Figure 20), the output voltage the controller expects, rises as well. Since V_{SSE} starts at near 0 V, the controller attempts to regulate the output voltage from a starting point of zero volts. It cannot do this due to the converter architecture. The output voltage starts from the input voltage less the drop across the diode ($V_{IN} – V_D$) and rises from there. The point at which the output voltage starts to rise (t_2) is the point where the V_{SSE} ramp passes the point where it is commanding more output voltage than ($V_{IN} – V_D$). This voltage level is labeled $V_{SSE(1)}$. The time required for the output voltage to ramp from a theoretical zero to the final regulated value (from t_1 to t_3) is determined by the time it takes for the capacitor connected to the SS pin (C_{SS}) to rise through a 700-mV range, beginning at $V_{SS(ofst)}$ above GND.

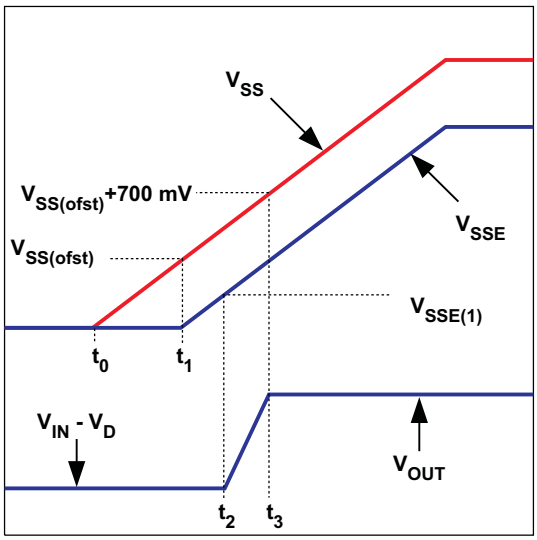


Figure 20. SS Pin Voltage and Output Voltage

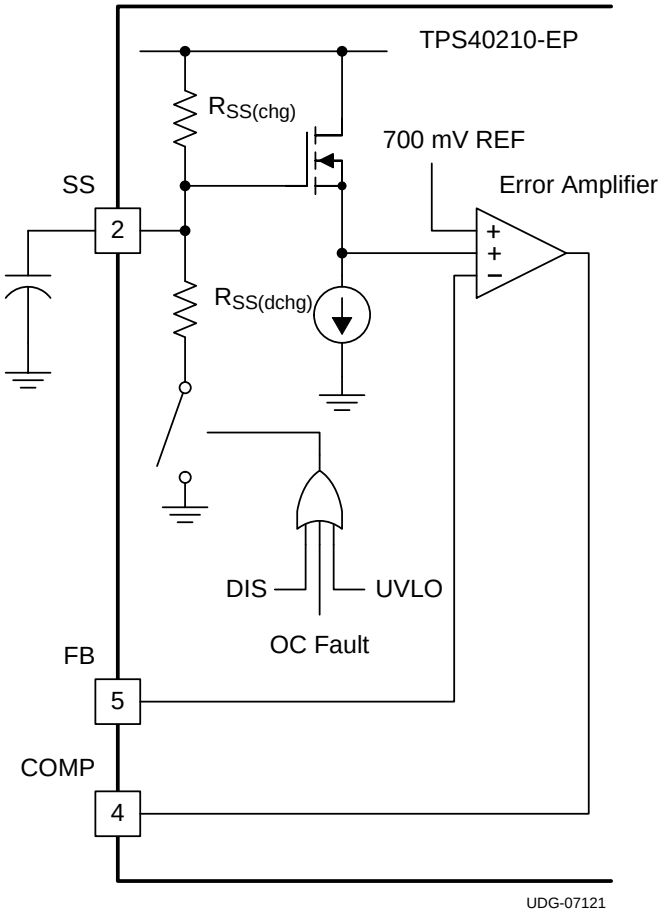


Figure 21. SS Pin Functional Circuit

Feature Description (continued)

The required capacitance for a given soft-start time $t_3 - t_1$ in Figure 20 is calculated in Equation 1.

$$C_{SS} = \frac{t_{SS}}{R_{SS} \times \ln\left(\frac{V_{BP} - V_{SS(ofst)}}{V_{BP} - \left(V_{SS(ofst)} + V_{FB}\right)}\right)} \quad (1)$$

where

- t_{SS} is the soft-start time, in seconds
- $R_{SS(chg)}$ is the SS charging resistance in Ω, typically 500 kΩ
- C_{SS} is the value of the capacitor on the SS pin, in F
- V_{BP} is the value of the voltage on the BP pin, in V
- $V_{SS(ofst)}$ is the approximate level shift from the SS pin to the error amplifier (~700 mV)
- V_{FB} is the error amplifier reference voltage, 700 mV typical

Note that t_{SS} is the time it takes for the output voltage to rise from 0 V to the final output voltage. Also note the tolerance on $R_{SS(chg)}$ given in the electrical specifications table. This contributes to some variability in the output voltage rise time and margin must be applied to account for it in design.

Also take note of V_{BP}. Its value varies depending on input conditions. For example, a converter operating from a slowly rising input initializes V_{BP} at a fairly low value and increases during the entire startup sequence. If the controller has a voltage above 8V at the input and the DIS pin is used to stop and then restart the converter, V_{BP} is approximately 8V for the entire startup sequence. The higher the voltage on BP, the shorter the startup time is and conversely, the lower the voltage on BP, the longer the startup time is.

The soft-start time (t_{SS}) must be chosen long enough so that the converter can start up without going into an overcurrent state. Since the over current state is triggered by sensing the peak voltage on the ISNS pin, that voltage must be kept below the overcurrent threshold voltage $V_{ISNS(oc)}$. The voltage on the ISNS pin is a function of the load current of the converter, the rate of rise of the output voltage and the output capacitance, and the current sensing resistor. The total output current that must be supported by the converter is the sum of the charging current required by the output capacitor and any external load that must be supplied during startup. This current must be less than the $I_{OUT(oc)}$ value used in Equation 15 or Equation 16 (depending on the operating mode of the converter) to determine the current sense resistor value. In these equations, the actual input voltage at the time that the controller reaches the final output voltage is the important input voltage to use in the calculations. If the input voltage is slowly rising and is at less than the nominal input voltage when the startup time ends, the output current limit is less than $I_{OUT(oc)}$ at the nominal input voltage. The output capacitor charging current must be reduced (decrease C_{OUT} or increase the t_{SS}) or $I_{OUT(oc)}$ must be increased and a new value for R_{ISNS} calculated.

$$I_{C(chg)} = \left(\frac{C_{OUT} \times V_{OUT}}{t_{SS}}\right) \quad (2)$$

$$t_{SS} > \left(\frac{C_{OUT} \times V_{OUT}}{(I_{OUT(oc)} - I_{EXT})}\right)$$

where

- $I_{C(chg)}$ is the output capacitor charging current in A
- C_{OUT} is the total output capacitance in F
- V_{OUT} is the output voltage in V
- t_{SS} is the soft-start time from Equation 1
- $I_{OUT(oc)}$ is the desired over current trip point in A
- I_{EXT} is any external load current in A (3)

Feature Description (continued)

The capacitor on the SS pin (C_{SS}) also plays a role in overcurrent functionality. It is used as the timer between restart attempts. The SS pin is connected to GND through a resistor, $R_{SS(dchg)}$, whenever the controller senses an overcurrent condition. Switching stops and nothing else happens until the SS pin discharges to the soft-start reset threshold, $V_{SS(rst)}$. At this point, the SS pin capacitor is allowed to charge again through the charging resistor $R_{SS(chg)}$, and the controller restarts from that point. The shortest time between restart attempts occurs when the SS pin discharges from $V_{SS(ofst)}$ (approximately 700 mV) to $V_{SS(rst)}$ (150 mV) and then back to $V_{SS(ofst)}$ and switching resumes. In actuality, this is a conservative estimate since switching does not resume until the V_{SSE} ramp rises to a point where it is commanding more output voltage than exists at the output of the controller. This occurs at some SS pin voltage greater than $V_{SS(ofst)}$ and depends on the voltage that remains on the output overvoltage the converter while switching has been halted. The fastest restart time can be calculated by using Equation 4, Equation 5, and Equation 6.

$$t_{DCHG} = R_{SS(dchg)} \times C_{SS} \times \ln\left(\frac{V_{SS(ofst)}}{V_{SS(rst)}}\right) \tag{4}$$

$$t_{CHG} = R_{SS(chg)} \times C_{SS} \times \ln\left(\frac{\left(V_{BP} - V_{SS(rst)}\right)}{\left(V_{BP} - V_{SS(ofst)}\right)}\right) \tag{5}$$

$$t_{RSTRT(min)} = t_{CHG} + t_{DCHG} \tag{6}$$

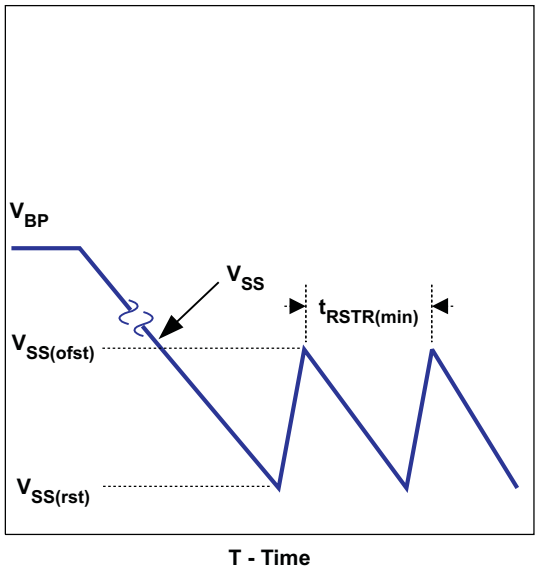


Figure 22. Soft-Start during Overcurrent

7.3.2 BP Regulator

The TPS40210-EP has an on board linear regulator the supplies power for the internal circuitry of the controller, including the gate driver. This regulator has a nominal output voltage of 8 V and must be bypassed with a 1-µF capacitor. If the voltage at the VDD pin is less than 8 V, the voltage on the BP pin will also be less and the gate drive voltage to the external FET is reduced from the nominal 8 V. This should be considered when choosing a FET for the converter.

Feature Description (continued)

Connecting external loads to this regulator can be done, but care must be taken to ensure that the thermal rating of the device is observed since there is no thermal shutdown feature in this controller. Exceeding the thermal ratings cause out of specification behavior and can lead to reduced reliability. The controller dissipates more power when there is an external load on the BP pin and is tested for dropout voltage for up to 5mA load. When the controller is in the disabled state, the BP pin regulator also shuts off so loads connected there power down as well. When the controller is disabled with the DIS/$\overline{EN}$ pin, this regulator is turned off.

The total power dissipation in the controller can be calculated as follows. The total power is the sum of P_Q, P_G, and P_E.

$$P_Q = V_{VDD} \times I_{VDD(en)} \tag{7}$$

$$P_G = V_{VDD} \times Q_g \times f_{SW} \tag{8}$$

$$P_E = V_{VDD} \times I_{EXT}$$

where

- P_Q is the quiescent power of the device in W
- V_{DD} is the VDD pin voltage in V
- $I_{DD(en)}$ is the quiescent current of the controller when enabled but not switching in A
- P_G is the power dissipated by driving the gate of the FET in W
- Q_g is the total gate charge of the FET at the voltage on the BP pin in C
- f_{SW} is the switching frequency in Hz
- P_E is the dissipation caused be external loading of the BP pin in W
- I_{EXT} is the external load current in A (9)

7.3.3 Shutdown (DIS/$\overline{EN}$ Pin)

The DIS/$\overline{EN}$ pin is an active high shutdown command for the controller. Pulling this pin above 1.2 V causes the controller to completely shut down and enter a low current consumption state. In this state, the regulator connected to the BP pin is turned off. There is an internal 1.1-MΩ pulldown resistor connected to this pin that keeps the pin at GND level when left floating. If this function is not used in an application, it is best to connect this pin to GND.

7.3.4 Minimum On-Time and Off-Time Considerations

The TPS40210-EP has a minimum off-time of approximately 200 ns and a minimum on-time of 300 ns. These two constraints place limitations on the operating frequency that can be used for a given input to output conversion ratio. See Figure 2 for the maximum frequency that can be used for a given duty cycle.

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, the duty cycle varies with changes to the load much more than it does when running in continuous conduction mode.

In continuous conduction mode, the duty cycle is related primarily to the input and output voltages.

$$\frac{V_{OUT} + V_D}{V_{IN}} = \frac{1}{1 - D} \tag{10}$$

$$D = \left(1 - \left(\frac{V_{IN}}{V_{OUT} + V_D}\right)\right) \tag{11}$$

In discontinuous mode the duty cycle is a function of the load, input and output voltages, inductance and switching frequency.

$$D = \frac{2 \times (V_{OUT} + V_D) \times I_{OUT} \times L \times f_{SW}}{(V_{IN})^2} \tag{12}$$

Feature Description (continued)

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 as follows.

$$I_{OUT(crit)} = \frac{(V_{OUT} + V_D - V_{IN}) \times (V_{IN})^2}{2 \times (V_{OUT} + V_D)^2 \times f_{SW} \times L} \quad (13)$$

For loads higher than the result of Equation 13, the duty cycle is given by Equation 11 and for loads less that the results of Equation 13, the duty cycle is given Equation 12. For Equations 1 through 4, 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

7.3.5 Setting the Oscillator Frequency

The oscillator frequency is determined by a resistor and capacitor connected to the RC pin of the TPS40210-EP. The capacitor is charged to a level of approximately $V_{DD}/20$ by current flowing through the resistor and is then discharged by a transistor internal to the TPS40210-EP. The required resistor for a given oscillator frequency is found from either Figure 1 or Equation 14.

$$R_T = \frac{1}{5.8 \times 10^{-8} \times f_{SW} \times C_T + 8 \times 10^{-10} \times f_{SW}^2 + 1.4 \times 10^{-7} \times f_{SW} - 1.5 \times 10^{-4} + 1.7 \times 10^{-6} \times C_T - 4 \times 10^{-9} \times C_T^2} \quad (14)$$

where

- R_T is the timing resistance in kΩ.
- f_{SW} is the switching frequency in kHz.
- C_T is the timing capacitance in pF.

For most applications a capacitor in the range of 68 pF to 120 pF gives the best results. Resistor values should be limited to between 100 kΩ and 1 MΩ as well. If the resistor value falls below 100 kΩ, decrease the capacitor size and recalculate the resistor value for the desired frequency. As the capacitor size decreases below 47 pF, the accuracy of Equation 14 degrades and empirical means may be needed to fine tune the timing component values to achieve the desired switching frequency.

7.3.6 Synchronizing the Oscillator

The TPS40210-EP can be synchronized to an external clock source. Figure 23 shows the functional diagram of the oscillator. When synchronizing the oscillator to an external clock, the RC pin must be pulled below 150 mV for 20 ns or more. The external clock frequency must be higher than the free running frequency of the converter as well. When synchronizing the controller, if the RC pin is held low for an excessive amount of time, erratic operation may occur. The maximum amount of time that the RC pin should be held low is 50% of a nominal output pulse, or 10% of the period of the synchronization frequency. If the external clock signal cannot operate with a low enough duty cycle to limit the amount of time the RC pin is held low, a resistor and capacitor can be added at the gate of the synchronization MOSFET. The capacitor should be added in series with the gate of the MOSFET to AC couple the rising edge of the synchronization signal. The resistor should be added from the gate of the MOSFET to ground to turn off the MOSFET. Typical values for the resistor and capacitor are 220 pF and 1 kΩ.

Feature Description (continued)

Under circumstances where the duty cycle is less than 50%, a Schottky diode connected from the RC pin to an external clock may be used to synchronize the oscillator. The cathode of the diode is connected to the RC pin. The trip point of the oscillator is set by an internal voltage divider to be 1/20 of the input voltage. The clock signal must have an amplitude higher than this trip point. When the clock goes low, it allows the reset current to restart the RC ramp, synchronizing the oscillator to the external clock. This provides a simple, single-component method for clock synchronization.

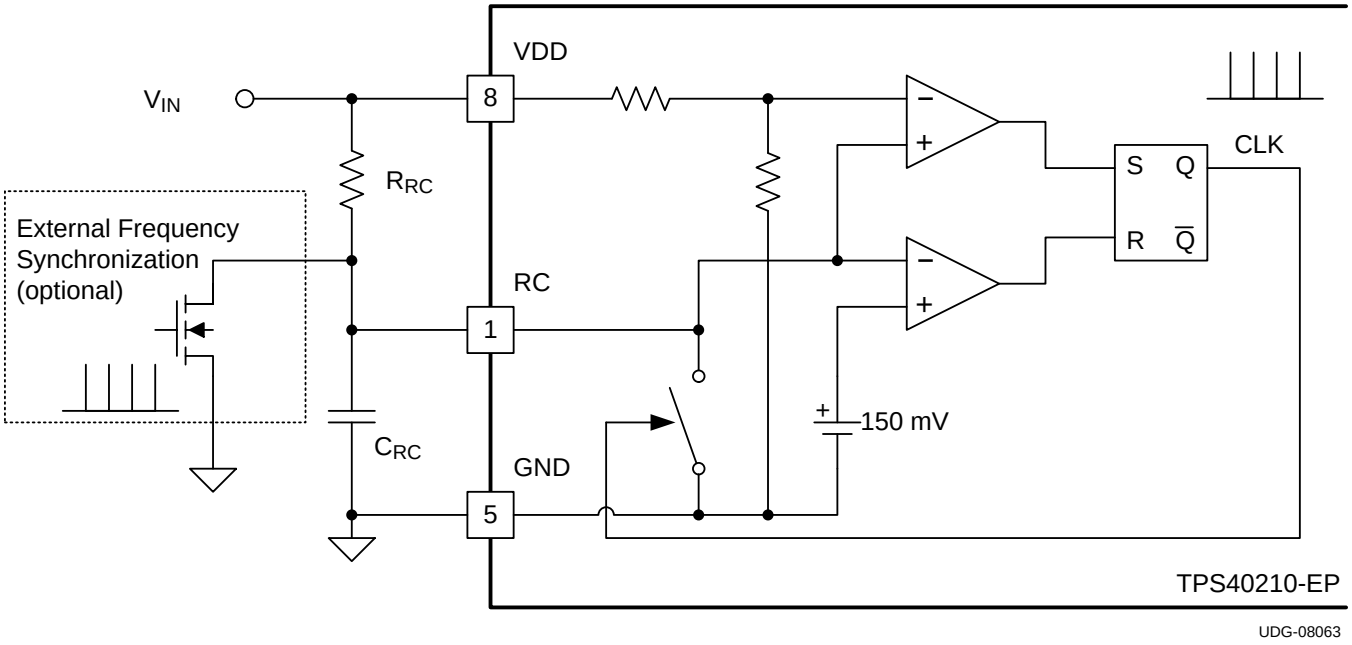


Figure 23. Oscillator Functional Diagram

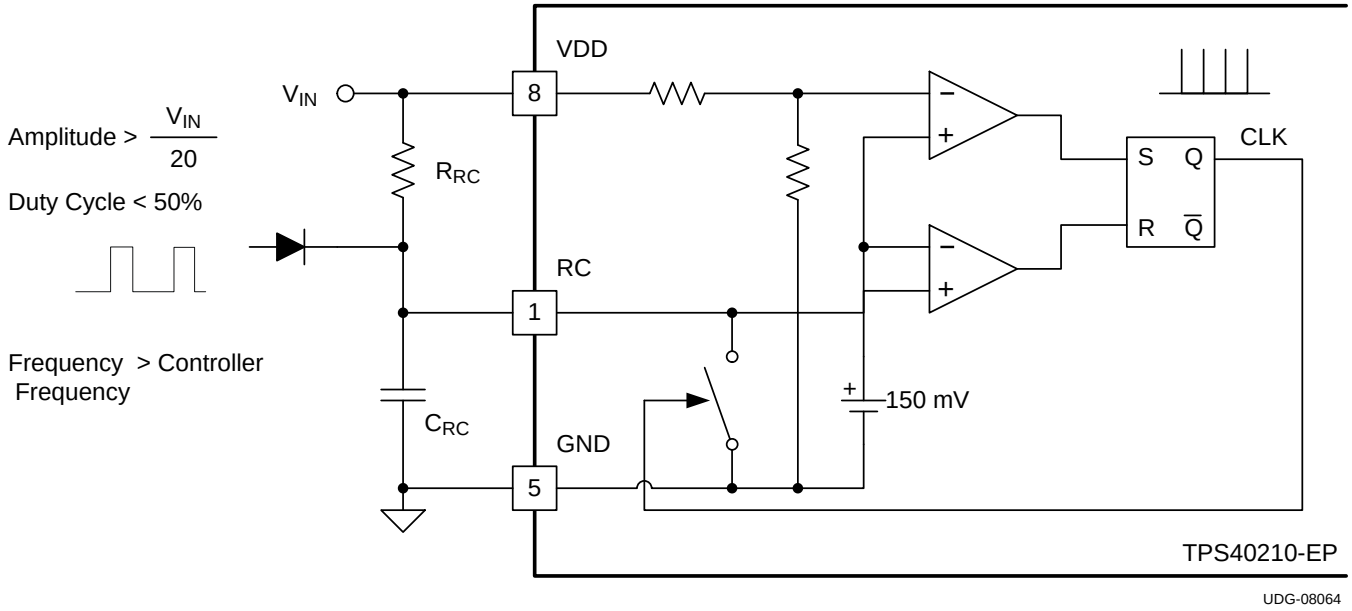


Figure 24. Diode Connected Synchronization

Feature Description (continued)

7.3.7 Current Sense and Overcurrent

The TPS40210-EP is a current mode controller that uses a resistor in series with the source terminal power FET to sense current for both the current mode control and overcurrent protection. The device enters a current limit state if the voltage on the ISNS pin exceeds the current limit threshold voltage $V_{ISNS(oc)}$ from the electrical specifications table. When this happens the controller discharges the SS capacitor through a relatively high impedance and then attempt to restart. The amount of output current that causes this to happen is dependent on several variables in the converter.

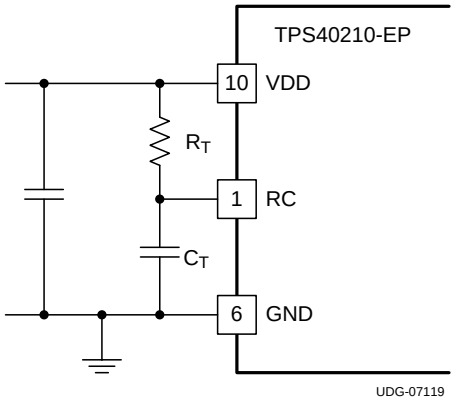


Figure 25. Oscillator Components

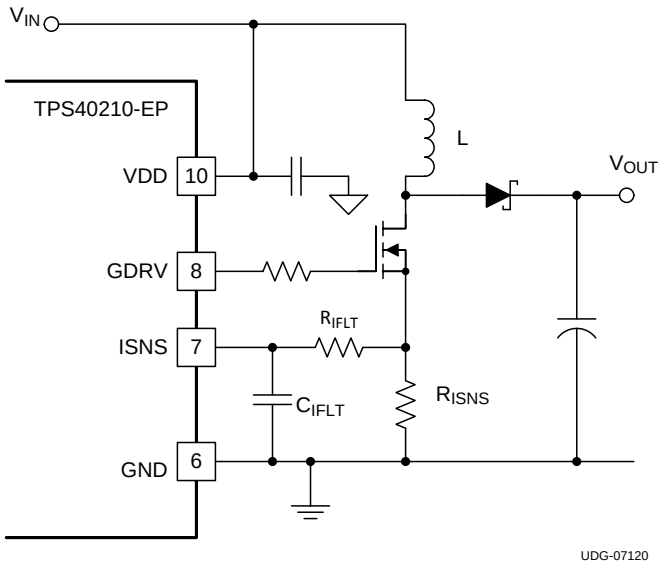


Figure 26. Current Sense Components

The load current overcurrent threshold is set by proper choice of R_{ISNS}. If the converter is operating in discontinuous mode the current sense resistor is found in Equation 15.

$$R_{ISNS} = \frac{f_{SW} \times L \times V_{ISNS(oc)}}{\sqrt{2 \times L \times f_{SW} \times I_{OUT(oc)} \times (V_{OUT} + V_D - V_{IN})}} \quad (15)$$

If the converter is operating in continuous conduction mode R_{ISNS} can be found in Equation 16.

$$R_{ISNS} = \frac{V_{ISNS}}{\left(\frac{I_{OUT}}{1-D}\right) + \left(\frac{I_{RIPPLE}}{2}\right)} = \frac{V_{ISNS}}{\left(\frac{I_{OUT}}{(1-D)}\right) + \left(\frac{D \times V_{IN}}{2 \times f_{SW} \times L}\right)} \quad (16)$$

where

- R_{ISNS} is the value of the current sense resistor in Ω.
- $V_{ISNS(oc)}$ is the overcurrent threshold voltage at the ISNS pin (from electrical specifications).
- D is the duty cycle (from Equation 11).
- f_{SW} is the switching frequency in Hz.
- V_{IN} is the input voltage to the power stage in V (see text).
- L is the value of the inductor in H.
- I_{OUT}(oc) is the desired overcurrent trip point in A.
- V_D is the drop across the diode in Figure 26.

Feature Description (continued)

The TPS40210-EP has a fixed undervoltage lockout (UVLO) that allows the controller to start at a typical input voltage of 4.25 V. If the input voltage is slowly rising, the converter might have less than its designed nominal input voltage available when it has reached regulation. As a result, this may decreases the apparent current limit load current value and must be taken into consideration when selecting R_{ISNS}. The value of V_{IN} used to calculate R_{ISNS} must be the value at which the converter finishes startup. The total converter output current at startup is the sum of the external load current and the current required to charge the output capacitors. See the *Soft-Start* for information on calculating the required output capacitor charging current.

The topology of the standard boost converter has no method to limit current from the input to the output in the event of a short circuit fault on the output of the converter. If protection from this type of event is desired, it is necessary to use some secondary protection scheme, such as a fuse, or rely on the current limit of the upstream power source.

7.3.8 Current Sense and Subharmonic Instability

A characteristic of peak current mode control results in a condition where the current control loop can exhibit instability. This results in alternating long and short pulses from the pulse width modulator. The voltage loop maintains regulation and does not oscillate, but the output ripple voltage increases. The condition occurs only when the converter is operating in continuous conduction mode and the duty cycle is 50% or greater. The cause of this condition is described in SLUA101, available at www.ti.com. The remedy for this condition is to apply a compensating ramp from the oscillator to the signal going to the pulse width modulator. In the TPS40210-EP the oscillator ramp is applied in a fixed amount to the pulse width modulator. The slope of the ramp is given in Equation 17.

$$s_e = f_{SW} \times \left(\frac{V_{VDD}}{20} \right) \tag{17}$$

To ensure that the converter does not enter into sub-harmonic instability, the slope of the compensating ramp signal must be at least half of the down slope of the current ramp signal. Since the compensating ramp is fixed in the TPS40210-EP, this places a constraint on the selection of the current sense resistor.

The down slope of the current sense wave form at the pulse width modulator is described in Equation 18.

$$m2 = \frac{A_{CS} \times R_{ISNS} \times (V_{OUT} + V_D - V_{IN})}{L} \tag{18}$$

Since the slope compensation ramp must be at least half, and preferably equal to the down slope of the current sense waveform seen at the pulse width modulator, a maximum value is placed on the current sense resistor when operating in continuous mode at 50% duty cycle or greater. For design purposes, some margin should be applied to the actual value of the current sense resistor. As a starting point, the actual resistor chosen should be 80% or less that the value calculated in Equation 19. This equation calculates the resistor value that makes the slope compensation ramp equal to one half of the current ramp downslope. Values no more than 80% of this result would be acceptable.

$$R_{ISNS(max)} = \frac{V_{VDD} \times L \times f_{SW}}{60 \times (V_{OUT} + V_D - V_{IN})} \tag{19}$$

where

- S_e is the slope of the voltage compensating ramp applied to the pulse width modulator in V/s.
- f_{SW} is the switching frequency in Hz.
- V_{DD} is the voltage at the VDD pin in V.
- m2 is the down slope of the current sense waveform seen at the pulse width modulator in V/s.
- R_{ISNS} is the value of the current sense resistor in Ω.
- V_{OUT} is the converter output voltage V_{IN} is the converter power stage input voltage.
- V_D is the drop across the diode in Figure 26.

Feature Description (continued)

It is possible to increase the voltage compensation ramp slope by connecting the VDD pin to the output voltage of the converter instead of the input voltage as shown in Figure 26. This can help in situations where the converter design calls for a large ripple current value in relation to the desired output current limit setting.

NOTE

Connecting the VDD pin to the output voltage of the converter affects the startup voltage of the converter since the controller undervoltage lockout (UVLO) circuit monitors the VDD pin and senses the input voltage less the diode drop before startup. The effect is to increase the startup voltage by the value of the diode voltage drop.

If an acceptable R_{ISNS} value is not available, the next higher value can be used and the signal from the resistor divided down to an acceptable level by placing another resistor in parallel with C_{IFLT}.

7.3.9 Current Sense Filtering

In most cases, a small filter placed on the ISNS pin improves performance of the converter. These are the components R_{IFLT} and C_{IFLT} in Figure 26. The time constant of this filter should be approximately 10% of the nominal pulse width of the converter. The pulse width can be found using Equation 20.

$$t_{ON} = \frac{D}{f_{SW}} \tag{20}$$

The suggested time constant is then

$$R_{IFLT} \times C_{IFLT} = 0.1 \times t_{ON} \tag{21}$$

The range of R_{IFLT} should be from about 1 kΩ to 5 kΩ for best results. Higher values can be used but this raises the impedance of the ISNS pin connection more than necessary and can lead to noise pickup issues in some layouts. C_{IFLT} should be located as close as possible to the ISNS pin as well to provide noise immunity.

7.3.10 Control Loop Considerations

There are two methods to design a suitable control loop for the TPS40210-EP. The first and preferred if equipment is available is to use a frequency response analyzer to measure the open loop modulator and power stage gain and to then design compensation to fit that. The usage of these tools for this purpose is well documented with the literature that accompanies the tool and is not be discussed here.

The second option is to make an initial guess at compensation, and then evaluate the transient response of the system to see if the compensation is acceptable to the application or not. For most systems, an adequate response can be obtained by simply placing a series resistor and capacitor (R_{FB} and C_{FB}) from the COMP pin to the FB pin as shown in Figure 27. The initial compensation selection can be done more accurately with aid of WEBENCH® to select the components or the average Spice model to simulate the open loop modulator and power stage gain.

Feature Description (continued)

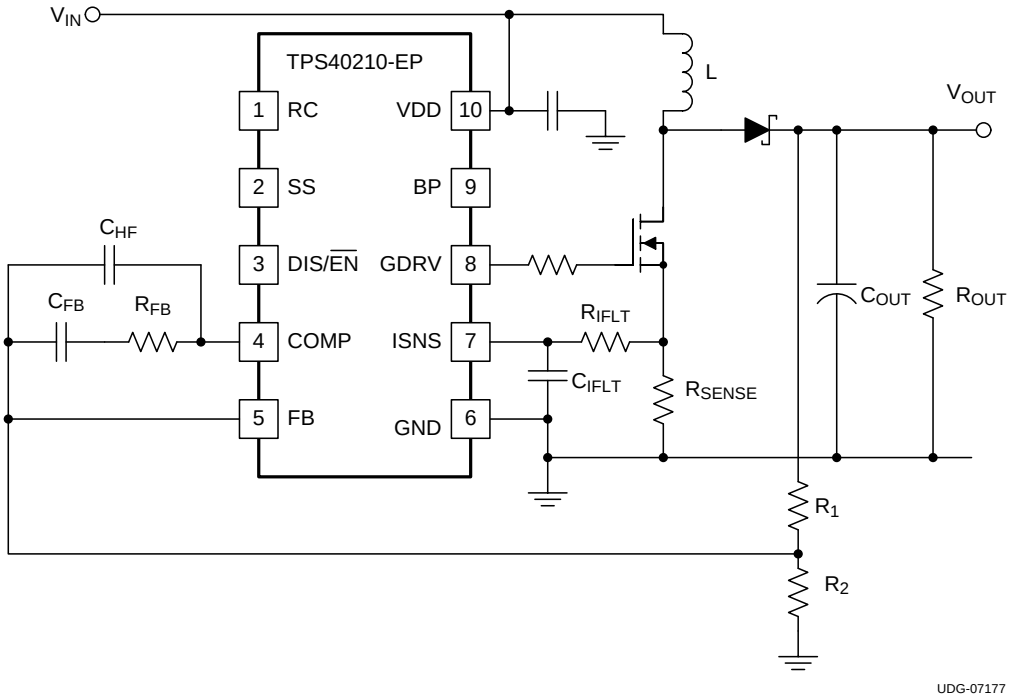


Figure 27. Basic Compensation Network

The natural phase characteristics of most capacitors used for boost outputs combined with the current mode control provide adequate phase margin when using this type of compensation. To determine an initial starting point for the compensation, the desired crossover frequency must be considered when estimating the control to output gain. The model used is a current source into the output capacitor and load.

When using these equations, the loop bandwidth should be no more than 20% of the switching frequency, f_{SW}. A more reasonable loop bandwidth would be 10% of the switching frequency. Be sure to evaluate the transient response of the converter over the expected load range to ensure acceptable operation.

$$\left|K_{CO}\right| = g_M \times \left|Z_{OUT}\left(f_{CO}\right)\right| \tag{22}$$

$$g_M = \frac{0.13 \times \sqrt{L \times \frac{f_{SW}}{R_{OUT}}}}{\left(R_{ISNS}\right)^2 \times \left(120 \times R_{ISNS} + L \times f_{SW}\right)} \tag{23}$$

$$\left|Z_{OUT}\right| = R_{OUT} \times \sqrt{\frac{\left(1 + \left(2\pi \times f_L \times R_{ESR} \times C_{OUT}\right)^2\right)}{1 + \left(\left(R_{OUT}\right)^2 + 2 \times R_{OUT} \times R_{ESR} + \left(R_{ESR}\right)^2\right) \times \left(2\pi \times f_L \times C_{OUT}\right)^2}}$$

where

- K_{CO} is the control to output gain of the converter, in V/V
- g_M is the transconductance of the power stage and modulator, in S
- R_{OUT} is the output load equivalent resistance, in Ω
- Z_{OUT} is the output impedance, including the output capacitor, in Ω
- R_{ISNS} is the value of the current sense resistor, in Ω
- L is the value of the inductor, in H
- C_{OUT} is the value of the output capacitance, in F

Feature Description (continued)

- R_{ESR} is the equivalent series resistance of C_{OUT}, in Ω
- f_{SW} is the switching frequency, in Hz
- f_L is the desired crossover frequency for the control loop, in Hz (24)

These equations assume that the operation is discontinuous and that the load is purely resistive. The gain in continuous conduction can be found by evaluating Equation 23 at the resistance that gives the critical conduction current for the converter. Loads that are more like current sources give slightly higher gains than predicted here. To find the gain of the compensation network required for a control loop of bandwidth f_L, take the reciprocal of Equation 22.

$$K_{COMP} = \frac{1}{|K_{CO}|} \tag{25}$$

The GBWP of the error amplifier is only guaranteed to be at least 1.5MHz. If K_{COMP} multiplied by f_L is greater than 750 kHz, reduce the desired loop crossover frequency until this condition is satisfied. This ensures that the high-frequency pole from the error amplifier response with the compensation network in place does not cause excessive phase lag at f_L and decreased phase margin in the loop.

The RC network connected from COMP to FB places a zero in the compensation response. That zero should be approximately 1/10th of the desired crossover frequency, f_L. With that being the case, R_{FB} and C_{FB} can be found from Equation 26 and Equation 27

$$R_{FB} = \frac{R1}{|K_{CO}|} = R1 \times K_{COMP} \tag{26}$$

$$C_{FB} = \frac{10}{2\pi \times f_L \times R_{FB}}$$

where

- R1 is the high side feedback resistor in Figure 27, in Ω
- f_L is the desired loop crossover frequency, in Hz (27)

Thought not strictly necessary, it is recommended that a capacitor be added between COMP and FB to provide high-frequency noise attenuation in the control loop circuit. This capacitor introduces another pole in the compensation response. The allowable location of that pole frequency determines the capacitor value. As a starting point, the pole frequency should be 10 × f_L. The value of C_{HF} can be found from Equation 28.

$$C_{HF} = \frac{1}{20\pi \times f_L \times R_{FB}} \tag{28}$$

While the error amplifier GBWP will usually be higher, it can be as low as 1.5 MHz. If 10 × K_{Comp} × f_L > 1.5 MHz, the error amplifier gain-bandwidth product may limit the high-frequency response below that of the high-frequency capacitor. To maintain a consistent high-frequency gain roll-off, C_{HF} can be calculated by Equation 29.

$$C_{HF} = \frac{1}{2\pi \times 1.5 \times (10)^6 \times R_{FB}}$$

where

- C_{HF} is the high-frequency roll-off capacitor value in F
- R_{FB} is the mid band gain setting resistor value in Ω (29)

7.3.11 Gate Drive Circuit

Some applications benefit from the addition of a resistor connected between the GDRV pin and the gate of the switching MOSFET. In applications that have particularly stringent load regulation (under 0.75%) requirements and operate from input voltages above 5 V, or are sensitive to pulse jitter in the discontinuous conduction region, this resistor is recommended. The recommended starting point for the value of this resistor can be calculated from Equation 30.

Feature Description (continued)

$$R_G = \frac{105}{Q_G} \quad (30)$$

where

- Q_G is the MOSFET total gate charge at 8V, V_{GS} in nC
- R_G is the suggested starting point gate resistance in Ω

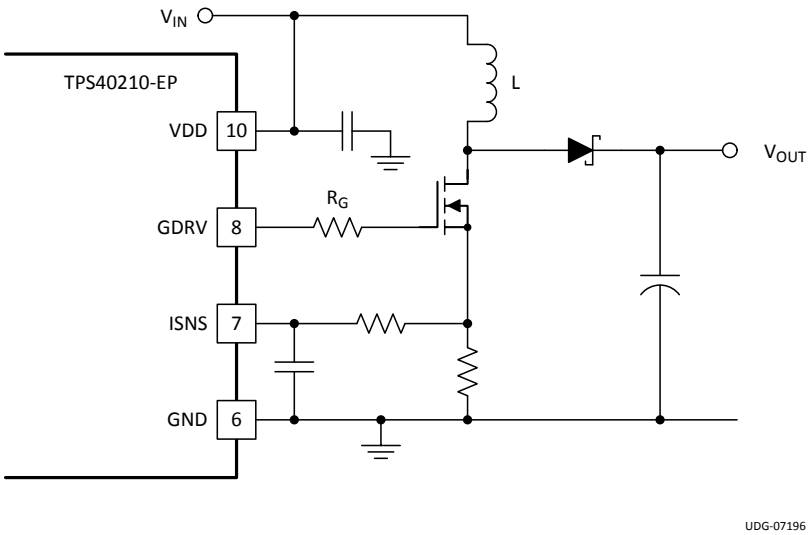


Figure 28. Gate Drive Resistor

7.4 Device Functional Modes

7.4.1 Operation Near Minimum Input Voltage

The TPS40210-EP is designed to operate with input voltages above 4.5 V. The typical VDD UVLO threshold is 4.25 V and the device may operate at input voltages down to the UVLO voltage. At input voltages below the actual UVLO voltage, the device will not switch. When V_{VDD} passes the UVLO threshold the device will become active. Switching is enabled and the soft-start sequence is initiated. The TPS40210-EP will ramp up the output voltage at the rate determined by the external capacitor at the SS pin.

7.4.2 Operation With DIS/$\overline{EN}$ Pin

The DIS/$\overline{EN}$ pin has a 1.2-V typical threshold which can be used to disable the TPS40210-EP. With DIS/$\overline{EN}$ forced above this threshold voltage the device is disabled and switching is inhibited even if V_{VDD} is above its UVLO threshold. Hysteresis on the DIS/$\overline{EN}$ pin threshold gives a typical turn-on threshold of 1.05 V. If the DIS/$\overline{EN}$ is left floating or is pulled below the 1.05-V threshold while V_{VDD} is above its UVLO threshold, the device becomes active. Switching is enabled and the soft-start sequence is initiated. The TPS40210-EP will ramp up the output voltage at the rate determined by the external capacitor at the soft-start pin.

8 Application and Implementation

NOTE

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

8.1 Application Information

The TPS40210-EP is a 4.5-V to 52-V low-side controller with an integrated gate driver for a low-side N-channel MOSFET. This device is typically used in a boost topology to convert a lower DC voltage to a higher DC voltage with a peak current limit set by an external current sense resistor. It can also be configured in a SEPIC, Flyback and LED drive applications. In higher current applications, the maximum current can also be limited by the thermal performance of the external MOSFET and rectifying diode switch. Use the following design procedure to select external components for the TPS40210-EP. The design procedure illustrates the design of a typical boost regulator with the TPS40210-EP. Alternatively, use the WEBENCH software to generate a complete design. The WEBENCH software uses an iterative design procedure and accesses a comprehensive database of components when generating a design.

8.2 Typical Application

The following example illustrates the design process and component selection for a 12-V to 24-V nonsynchronous boost regulator using the TPS40210-EP controller.

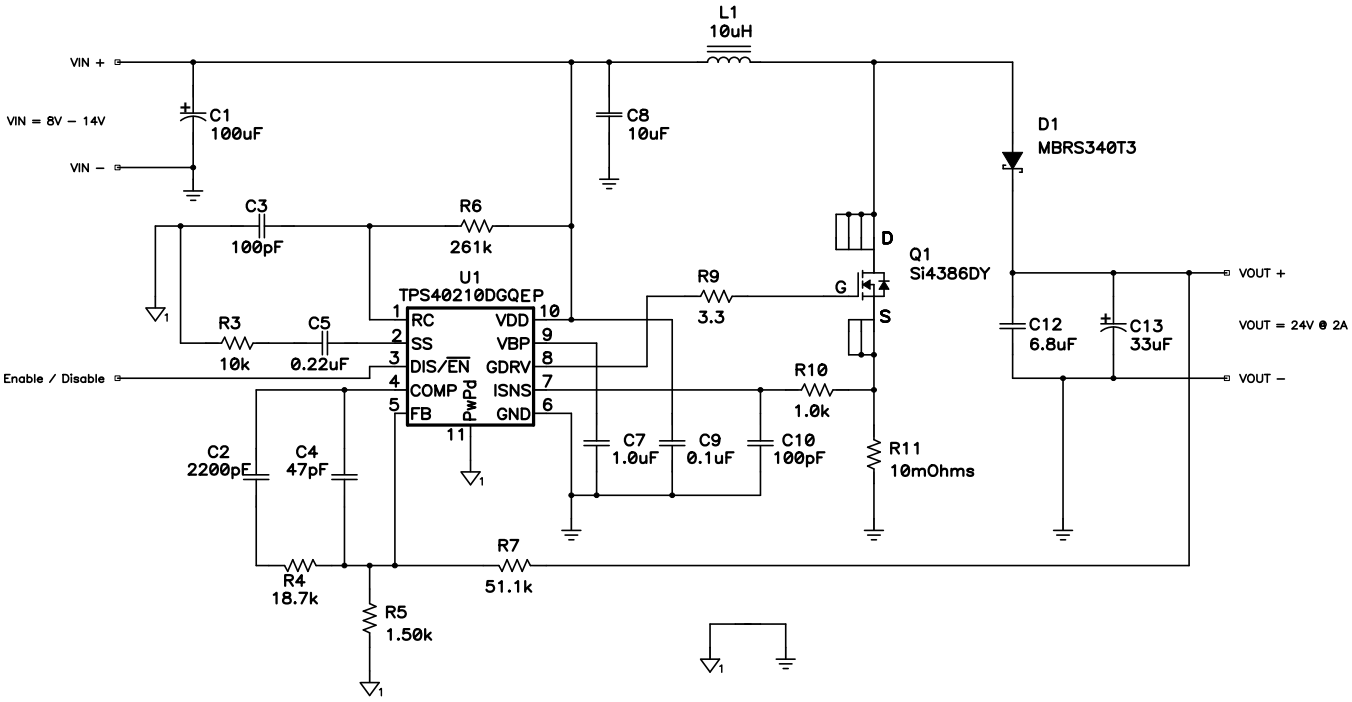


Figure 29. TPS40210-EP Design Example – 12-V to 24-V at 2 A

Typical Application (continued)

8.2.1 Design Requirements

Table 1. TPS40210-EP Design Example Specifications

PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
INPUT CHARACTERISTICS						
V_{IN}	Input voltage		8	12	14	V
I_{IN}	Input current			4.4		A
	No load input current				0.05	
$V_{IN(UVLO)}$	Input undervoltage lockout			4.5		V
OUTPUT CHARACTERISTICS						
V_{OUT}	Output voltage		23.5	24.0	24.5	V
	Line regulation				1%	
	Load regulation				1%	
$V_{OUT(ripple)}$	Output voltage ripple				500	mV_{PP}
I_{OUT}	Output current	8 V ≤ V_{IN} ≤ 14 V	0.1	1	2.0	A
I_{OCP}	Output overcurrent inception point		3.5			
	Transient response					
ΔI	Load step			1		A
	Load slew rate			1		A/µs
	Overshoot threshold voltage			500		mV
	Settling time			5		ms
SYSTEM CHARACTERISTICS						
f_{SW}	Switching frequency			600		kHz
η_{PK}	Peak efficiency	V_{IN} = 12 V		95%		
η	Full load efficiency	V_{IN} = 12 V, I_{OUT} = 2 A		94%		
T_{OP}	Operating temperature range	8 V ≤ V_{IN} ≤ 14 V, I_{OUT} ≤ 2 A		25		°C
MECHANICAL DIMENSIONS						
W	Width			1.5		inch
L	Length			1.5		
h	Height			0.5		

8.2.2 Detailed Design Procedure

8.2.2.1 Duty Cycle Estimation

The duty cycle of the main switching MOSFET is estimated using Equation 31 and Equation 32.

$$D_{MIN} \approx \frac{V_{OUT} - V_{IN(max)} + V_{FD}}{V_{OUT} + V_{FD}} = \frac{24\,V - 14\,V + 0.5\,V}{24\,V + 0.5\,V} = 42.9\% \quad (31)$$

$$D_{MAX} \approx \frac{V_{OUT} - V_{IN(min)} + V_{FD}}{V_{OUT} + V_{FD}} = \frac{24\,V - 8\,V + 0.5\,V}{24\,V + 0.5\,V} = 67.3\% \quad (32)$$

Using an estimated forward drop (V_{FD}) of 0.5V for a schottky rectifier diode, the approximate duty cycle is 42.9% (minimum) to 67.3% (maximum).

8.2.2.2 Inductor Selection

The peak-to-peak ripple is chosen to be 30% of the maximum input current.

$$I_{RIPPLE(max)} = 0.3 \times \frac{I_{OUT(max)}}{1 - D_{MIN}} = 0.3 \times \frac{2}{1 - 0.429} = 1.05\,A \quad (33)$$

The minimum inductor size can be estimated using Equation 34.

$$L_{MIN} \approx \frac{V_{IN(max)}}{I_{RIPPLE(max)}} \times D_{MIN} \times \frac{1}{f_{SW}} = \frac{14\,V}{1.05\,A} \times 0.429 \times \frac{1}{600\,kHz} = 9.5\,\mu H \quad (34)$$

The next higher standard inductor value of 10 µH is selected. The ripple current for nominal and minimum V_{IN} is estimated by Equation 35 and Equation 36.

$$I_{RIPPLE(V_{in}typ)} \approx \frac{V_{IN}}{L} \times D \times \frac{1}{f_{SW}} = \frac{12\,V}{10\,\mu H} \times 0.50 \times \frac{1}{600\,kHz} = 1.02\,A \quad (35)$$

$$I_{RIPPLE(Vinmin)} \approx \frac{V_{IN}}{L} \times D \times \frac{1}{f_{SW}} = \frac{8\,V}{10\,\mu H} \times 0.673 \times \frac{1}{600\,kHz} = 0.90\,A \quad (36)$$

The worst case peak-to-peak ripple current occurs at 50% duty cycle (V_{IN} = 12.25 V) and is estimated as 1.02 A. Worst case RMS current through the inductor is approximated by Equation 37.

$$I_{Lrms} = \sqrt{\left(I_{L(avg)}\right)^2 + \left(\tfrac{1}{12} I_{RIPPLE}\right)^2} \approx \sqrt{\left(\frac{I_{OUT(max)}}{1-D_{MAX}}\right)^2 + \left(\tfrac{1}{12} I_{RIPPLE(VINmin)}\right)^2} = \sqrt{\left(\frac{2}{1-0.673}\right)^2 + \left(\left(\tfrac{1}{12}\right) \times 0.90A\right)^2} = 6.13\,Arms \quad (37)$$

The worst case RMS inductor current is 6.13 A. The peak inductor current is estimated by Equation 38.

$$I_{Lpeak} \approx \frac{I_{OUT(max)}}{1-D_{MAX}} + \left(\tfrac{1}{2}\right) I_{RIPPLE(Vinmin)} = \frac{2}{1-0.673} + \left(\tfrac{1}{2}\right) 0.90 = 6.57\,A \quad (38)$$

A 10-µH inductor with a minimum RMS current rating of 6.13 A and minimum saturation current rating of 6.57 A must be selected. A TDK RLF12560T-100M-7R5 7.5-A 10-µH inductor is selected.

This inductor power dissipation is estimated by Equation 39.

$$P_L \approx \left(I_{Lrms}\right)^2 \times DCR \quad (39)$$

The TDK RLF12560T-100M-7R5 12.4-mΩ DCR dissipates 466-mW of power.

8.2.2.3 Rectifier Diode Selection

A low forward voltage drop schottky diode is used as a rectifier diode to reduce its power dissipation and improve efficiency. Using 80% derating on V_{OUT} for ringing on the switch node, the rectifier diode minimum reverse breakdown voltage is given by Equation 40.

$$V_{(BR)R(min)} \geq \frac{V_{OUT}}{0.8} = 1.25 \times V_{OUT} = 1.25 \times 24\,V = 30\,V \quad (40)$$

The diode must have reverse breakdown voltage greater than 30 V. The rectifier diode peak and average currents are estimated by Equation 41 and Equation 42.

$$I_{D(avg)} \approx I_{OUT(max)} = 2\,A \quad (41)$$

$$I_{D(peak)} = I_{L(peak)} = 6.57\,A \quad (42)$$

The power dissipation in the diode is estimated by Equation 43.

$$P_{D(max)} \approx V_{FD} \times I_{D(avg)} = 0.5\,V \times 2\,A = 1W \quad (43)$$

For this design, the maximum power dissipation is estimated as 1 W. Reviewing 30-V and 40-V schottky diodes, the MBRS340T3, 40-V, 3-A diode in an SMC package is selected. This diode has a forward voltage drop of 0.48 V at 6 A, so the conduction power dissipation is approximately 960 mW, less than half its rated power dissipation.

8.2.2.4 Output Capacitor Selection

Output capacitors must be selected to meet the required output ripple and transient specifications.

$$C_{OUT} = 8\frac{I_{OUT} \times D}{V_{OUT(ripple)}} \times \frac{1}{f_{SW}} = 8\left(\frac{2\,A \times 0.673}{500\,mV}\right) \times \frac{1}{600\,kHz} = 36\,\mu F \quad (44)$$

$$ESR = \frac{7}{8} \times \frac{V_{OUT(ripple)}}{I_{L(peak)} - I_{OUT}} = \frac{7}{8} \times \frac{500\,mV}{6.57\,A - 2\,A} = 96\,m\Omega \quad (45)$$

A Panasonic EEEFC1V330P 35-V 33-µF, 120-mΩ bulk capacitor and a 6.8-µF ceramic capacitor are selected to provide the required capacitance and ESR at the switching frequency. The combined capacitance of 39.8 µF and ESR of 60 mΩ are used in compensation calculations.

8.2.2.5 Input Capacitor Selection

Since a boost converter has continuous input current, the input capacitor senses only the inductor ripple current. The input capacitor value can be calculated by Equation 46 and Equation 47 .

$$C_{IN} > \frac{I_{RIPPLE}}{4 \times V_{IN(ripple)} \times f_{SW}} = \frac{1.02\,A}{4 \times 60\,mV \times 600\,kHz} = 7.1\,\mu F \quad (46)$$

$$ESR < \frac{V_{IN(ripple)}}{2 \times I_{RIPPLE}} = \frac{60\,mV}{2 \times 1.02\,A} = 29\,m\Omega \quad (47)$$

For this design to meet a maximum input ripple of 60mV (1/2% of V_{IN} nominal), a minimum 7.1-µF input capacitor with ESR less than 29mΩ is needed. A 10-µF, X7R ceramic capacitor is selected.

8.2.2.6 Current Sense and Current Limit

The maximum allowable current sense resistor value is limited by both the current limit and sub-harmonic stability. These two limitations are given by Equation 48 and Equation 49.

$$R_{ISNS} < \frac{V_{ISNS(OC)MIN}}{1.1 \times \left(I_{L(peak)} + I_{Drive}\right)} = \frac{120\,mV}{1.1 \times (6.57\,A + 0.50\,A)} = 15.4\,m\Omega \quad (48)$$

$$R_{ISNS} < \frac{V_{IN(MAX)} \times L \times f_{SW}}{60 \times (V_{OUT} + V_{FD} - V_{IN})} = \frac{14\,V \times 10\,\mu H \times 600\,kHz}{60 \times (24\,V + 0.48\,V - 14\,V)} = 134\,m\Omega \quad (49)$$

With 10% margin on the current limit trip point (the 1.1 factor) and assuming a maximum gate drive current of 500 mA, the current limit requires a resistor less than 15.4 mΩ and stability requires a sense resistor less than 134 mΩ. A 10-mΩ resistor is selected. Approximately 2 mΩ of routing resistance is added in compensation calculations.

The power dissipation in R_{ISNS} is calculated by Equation 50.

$$P_{R_{ISNS}} = (I_{LRMS})^2 \cdot R_{ISNS} \cdot D \quad (50)$$

At maximum duty cycle, this is 0.253 W.

8.2.2.7 Current Sense Filter

To remove switching noise from the current sense, an RC filter is placed between the current sense resistor and the ISNS pin. A resistor with a value between 1 kΩ and 5 kΩ is selected and a capacitor value is calculated by Equation 51.

$$C_{IFLT} = \frac{0.1 \times D_{MIN}}{f_{SW} \times R_{IFLT}} = \frac{0.1 \times 0.429}{600\,kHz \times 1k\Omega} = 71pF \quad (51)$$

For a 1-kΩ filter resistor, 71pF is calculated and a 100-pF capacitor is selected.

8.2.2.8 Switching MOSFET Selection

The TPS40210-EP drives a ground referenced N-channel FET. The $R_{DS(on)}$ and gate charge are estimated based on the desired efficiency target.

$$P_{DISS(total)} \approx P_{OUT} \times \left(\frac{1}{\eta} - 1\right) = V_{OUT} \times I_{OUT} \times \left(\frac{1}{\eta} - 1\right) = 24\,V \times 2\,A \times \left(\frac{1}{0.95} - 1\right) = 2.526\,W \quad (52)$$

For a target of 95% efficiency with a 24-V Input voltage at 2 A, maximum power dissipation is limited to 2.526 W. The main power dissipating devices are the MOSFET, inductor, diode, current sense resistor and the integrated circuit, the TPS40210-EP.

$$P_{FET} < P_{DISS(total)} - P_L - P_D - P_{Risns} - V_{IN(max)} \times I_{VDD(max)} \quad (53)$$

This leaves 812 mW of power dissipation for the MOSFET. This can likely cause an SO-8 MOSFET to get too hot, so power dissipation is limited to 500 mW. Allowing half for conduction and half for switching losses, we can determine a target $R_{DS(on)}$ and Q_{GS} for the MOSFET by Equation 54 and Equation 55.

$$Q_{GS} < \frac{3 \times P_{FET} \times I_{DRIVE}}{2 \times V_{OUT} \times I_{OUT} \times f_{SW}} = \frac{3 \times 0.50\,W \times 0.50\,A}{2 \times 24\,V \times 2\,A \times 600\,kHz} = 13.0\,nC \quad (54)$$

A target MOSFET gate-to-source charge of less than 13.0 nC is calculated to limit the switching losses to less than 250 mW.

$$R_{DS(on)} < \frac{P_{FET}}{2 \times (I_{RMS})^2 \times D} = \frac{0.50\,W}{2 \times 6.13^2 \times 0.673} = 9.9\,m\Omega \quad (55)$$

A target MOSFET $R_{DS(on)}$ of 9.9 mΩ is calculated to limit the conduction losses to less than 250 mW. Reviewing 30-V and 40-V MOSFETs, an Si4386DY 9-mΩ MOSFET is selected. A gate resistor was added per Equation 30. The maximum gate charge at V_{GS}= 8V for the Si4386DY is 33.2 nC, this implies R_G = 3.3 Ω.

8.2.2.9 Feedback Divider Resistors

The primary feedback divider resistor (R_{FB}) from V_{OUT} to FB should be selected between 10 kΩ and 100 kΩ to maintain a balance between power dissipation and noise sensitivity. For a 24-V output a high feedback resistance is desirable to limit power dissipation so R_{FB} = 51.1 kΩ is selected.

$$R_{BIAS} = \frac{V_{FB} \times R_{FB}}{V_{OUT} - V_{FB}} = \frac{0.700\,V \times 51.1k\Omega}{24\,V - 0.700\,V} = 1.53\,k\Omega \quad (56)$$

R_{BIAS} = 1.50kΩ is selected.

8.2.2.10 Error Amplifier Compensation

Compensation selection can be done with aid of WEBENCH to select compensation components or with the aid of the average Spice model to simulate the open loop modulator and power stage gain. Alternatively the following procedure gives a good starting point.

While current mode control typically only requires Type II compensation, it is desirable to layout for Type III compensation to increase flexibility during design and development. Current mode control boost converters have higher gain with higher output impedance, so it is necessary to calculate the control loop gain at the maximum output impedance, estimated by Equation 57.

$$R_{OUT(max)} = \frac{V_{OUT}}{I_{OUT(min)}} = \frac{24\,V}{0.1A} = 240\,\Omega \quad (57)$$

The transconductance of the TPS40210-EP current mode control can be estimated by Equation 58.

$$g_M = \frac{0.13 \times \sqrt{L \times \frac{f_{SW}}{R_{OUT}}}}{(R_{ISNS})^2 \times (120 \times R_{ISNS} + L \times f_{SW})} = \frac{0.13 \times \sqrt{10\,\mu H \times \frac{600\,kHz}{240\,\Omega}}}{(12\,m\Omega)^2 \times (120 \times 12\,m\Omega + 10\,\mu H \times 600\,kHz)} = 19.2\ \frac{A}{V} \quad (58)$$

The maximum output impedance Z_{OUT}, can be estimated by Equation 59.

$$|Z_{OUT}(f)| = R_{OUT} \times \sqrt{\frac{\left(1 + (2\pi \times f \times R_{ESR} \times C_{OUT})^2\right)}{1 + \left((R_{OUT})^2 + 2 \times R_{OUT} \times R_{ESR} + (R_{ESR})^2\right) \times (2\pi \times f \times C_{OUT})^2}} \quad (59)$$

$$|Z_{OUT}(f_L)| = 240\,\Omega \times \sqrt{\frac{\left(1 + (2\pi \times 30\,kHz \times 60\,m\Omega \times 39.8\,\mu F)^2\right)}{1 + \left((240\,\Omega)^2 + 2 \times 240\,\Omega \times 60\,m\Omega + (60\,m\Omega)^2\right) \times (2\pi \times 30\,kHz \times 39.8\,\mu F)^2}} = 0.146\,\Omega \quad (60)$$

At the desired crossover frequency (f_L) of 30 kHz, Z_{OUT} becomes 0.146 Ω.

The modulator gain at the desired cross-over can be estimated by Equation 61.

$$|K_{CO}| = g_M \times |Z_{OUT}(f_{CO})| = 19.2 \frac{A}{V} \times 0.146\,\Omega = 2.80 \quad (61)$$

The feedback compensation network needs to be designed to provide an inverse gain at the cross-over frequency for unity loop gain. This sets the compensation mid-band gain at a value calculated in Equation 62.

$$K_{COMP} = \frac{1}{|K_{CO}|} = \frac{1}{2.80} = 0.357 \quad (62)$$

To set the mid-band gain of the error amplifier to K_{COMP} use Equation 63.

$$R4 = R7 \times K_{COMP} = \frac{R7}{|K_{CO}|} = \frac{51.1\,k\Omega}{2.80} = 18.2\,k\Omega \quad (63)$$

R4 = 18.7 kΩ selected.

Place the zero at 1/10th of the desired cross-over frequency.

$$C2 = \frac{10}{2\pi \times f_L \times R4} = \frac{10}{2\pi \times 30\,kHz \times 18.7\,k\Omega} = 2837\,pF \quad (64)$$

C2 = 2200 pF selected.

Place a high-frequency pole at about 5 times the desired cross-over frequency and less than one-half the unity gain bandwidth of the error amplifier:

$$C4 \approx \frac{1}{10\pi \times f_L \times R4} = \frac{1}{10\pi \times 30\,kHz \times 18.7\,k\Omega} = 56.74\,pF \quad (65)$$

$$C4 > \frac{1}{\pi \times GBW \times R4} = \frac{1}{\pi \times 1.5\,MHz \times 18.7\,k\Omega} = 11.35\,pF \quad (66)$$

C4 = 47 pF selected.

8.2.2.11 RC Oscillator

The RC oscillator calculation is given as shown in Equation 14 in the datasheet, substituting 100 for C_T and 600 for f_{SW}. For a 600-kHz switching frequency, a 100pF capacitor is selected and a 262-kΩ resistor is calculated (261-kΩ selected).

8.2.2.12 Soft-Start Capacitor

Since VDD > 8 V, the soft-start capacitor is selected by using Equation 67 to calculate the value.

$$C_{SS} = 20 \times T_{SS} \times 10^{-6} \quad (67)$$

For T_{SS} = 12 ms, C_{SS} = 240 nF. A 220-nF capacitor is selected.

8.2.2.13 Regulator Bypass

A regulator bypass (BP) capacitor of 1.0 µF is selected per the datasheet recommendation.

8.2.2.14 Bill of Materials

Table 2. Bill of Materials

REFERENCE DESIGNATOR	DESCRIPTION	SIZE	PART NUMBER	MANUFACTURER
C1	100 µF, aluminum capacitor, SM, ± 20%, 35 V	0.406 x 0.457	EEEFC1V101P	Panasonic
C2	2200 pF, ceramic capacitor, 25 V, X7R, 20%	0603	Std	Std
C3	100 pF, ceramic capacitor, 16 V, C0G, 10%	0603	Std	Std
C4	47 pF, ceramic capacitor, 16V, X7R, 20%	0603	Std	Std
C5	0.22 µF, ceramic capacitor, 16 V, X7R, 20%	0603	Std	Std
C7	1.0 µF, ceramic capacitor, 16 V, X5R, 20%	0603	Std	Std
C8	10 µF, ceramic capacitor, 25 V, X7R, 20%	0805	C3225X7R1E106M	TDK
C9	0.1 µF, ceramic capacitor, 50 V, X7R, 20%	0603	Std	Std
C10	100 pF, ceramic capacitor, 16 V, X7R, 20%	0603	Std	Std
D1	Schottky diode, 3 A, 40 V	SMC	MBRS340T3	On Semi
L1	10 µH, inductor, SMT, 7.5 A, 12.4 mΩ	0.325 x 0.318 inch	RLF12560T-100M-7R5	TDK
Q1	MOSFET, N-channel, 40 V, 14 A, 9mΩ	SO-8	Si4840DY	Vishay
R3	10 kΩ, chip resistor, 1/16 W, 5%	0603	Std	Std
R4	18.7 kΩ, chip resistor, 1/16 W, 1%	0603	Std	Std
R5	1.5 kΩ, chip resistor, 1/16 W, 1%	0603	Std	Std
R6	261 kΩ, chip resistor, 1/16 W, 1%	0603	Std	Std
R7	51.1 kΩ, chip resistor, 1/16 W, 1%	0603	Std	Std
R9	3.3 Ω, chip resistor, 1/16 W, 5%	0603	Std	Std
R10	1.0 kΩ, chip resistor, 1/16 W, 5%	0603	Std	Std
R11	10 mΩ, chip resistor, 1/2 W, 2%	1812	Std	Std
U1	IC, 4.5 V-52 V I/P, current mode boost controller	DGQ10	TPS40210-EPDGQ	TI

8.2.3 Application Curves

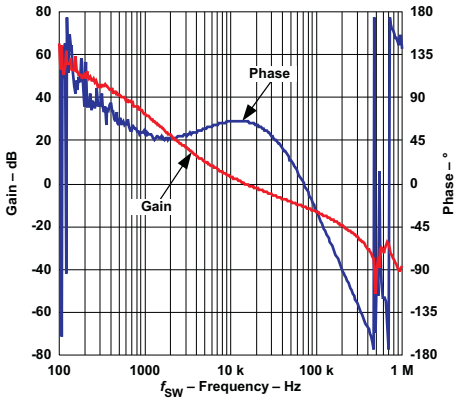


Figure 30. Gain and Phase vs Frequency

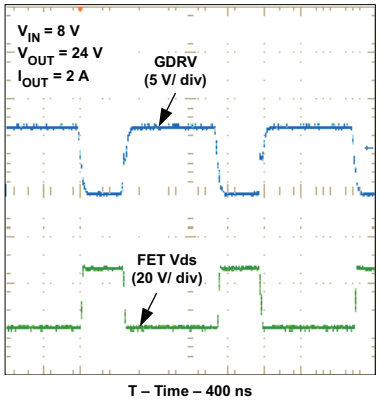


Figure 31. FET VDS and VGS Voltages vs Time

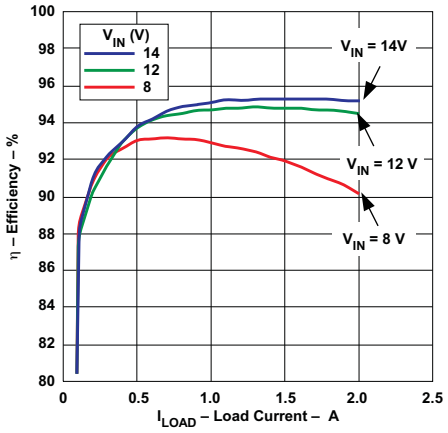


Figure 32. Efficiency vs Load Current

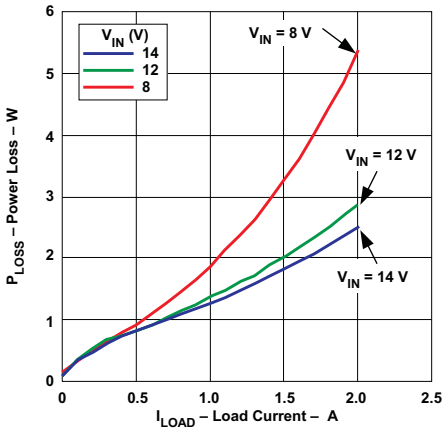


Figure 33. Power Loss vs Load Current

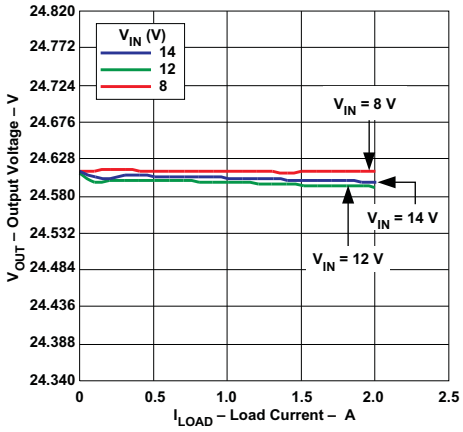


Figure 34. Output Voltage vs Load Current

9 Power Supply Recommendations

The TPS40210-EP is designed to operate from an input voltage supply range between 4.5 V and 52 V. This input supply should remain within the input voltage range of the TPS40210-EP. If the input supply is located more than a few inches from the buck power stage controlled by the TPS40210-EP, additional bulk capacitance may be required in addition to ceramic-bypass capacitors. An electrolytic capacitor with a value of 100 uF is a typical choice.

10 Layout

10.1 Layout Guidelines

- For the maximum effectiveness from C9, place it near the VDD pin of the controller. Excessive high frequency noise on VDD during switching degrades overall regulation as the load increases.
- Keep the output loop (Q1-D1-C12-R11) as small as possible. A larger loop can degrade current limit accuracy and increase rediated emissions.
- For best current limit accuracy keep the ISNS filter components C10 and R10 near the ISNS and GND pins.
- Avoid connecting traces carrying large AC currents through a ground plane. Instead, use PCB traces on the top layer to conduct the AC current and use the ground plane as a noise shield.
- Split the ground plane as necessary to keep noise away from the TPS40210-EP and noise sensitive areas such as components connected to the RC pin, FB pin, COMP pin and SS pin. Also keep these noise sensitive components close to the TPS40210-EP IC.
- Keep C7 near the BP and GND pins to provide good bypass for the BP regulator.
- The GDRV trace should be as close as possible to the power FET gate to minimize parisitic resistance and inductance in the trace. The parasitics should also be minimized in the return path from the source of the MOSFET, through the sense resistor and back to the GND pin.
- Keep the SW node as physically small as possible to minimize parasitic capacitance and radiated emissions.
- For good output voltage regulation, Kelvin connections should be brought from the load to the top FB resistor R7.

10.2 Layout Example

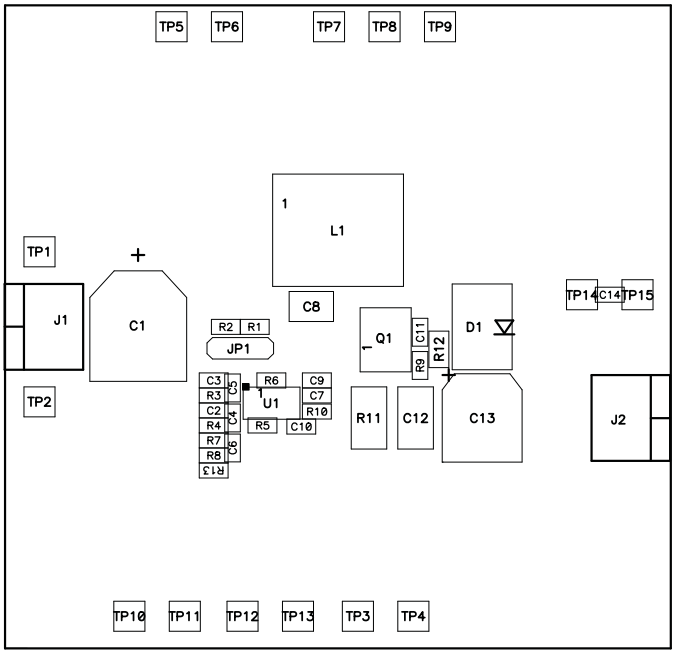


Figure 35. Component Placement

Layout Example (continued)

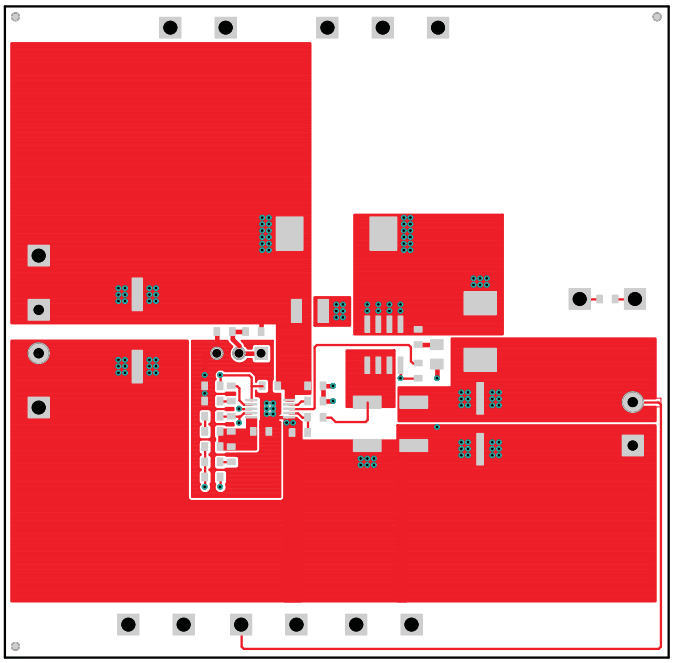


Figure 36. Top Copper

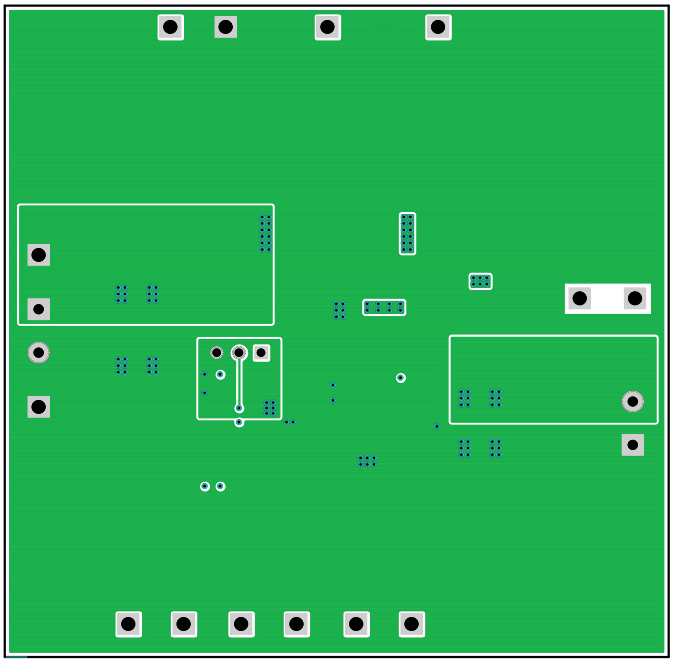


Figure 37. Bottom Copper Viewed from Top

Layout Example (continued)

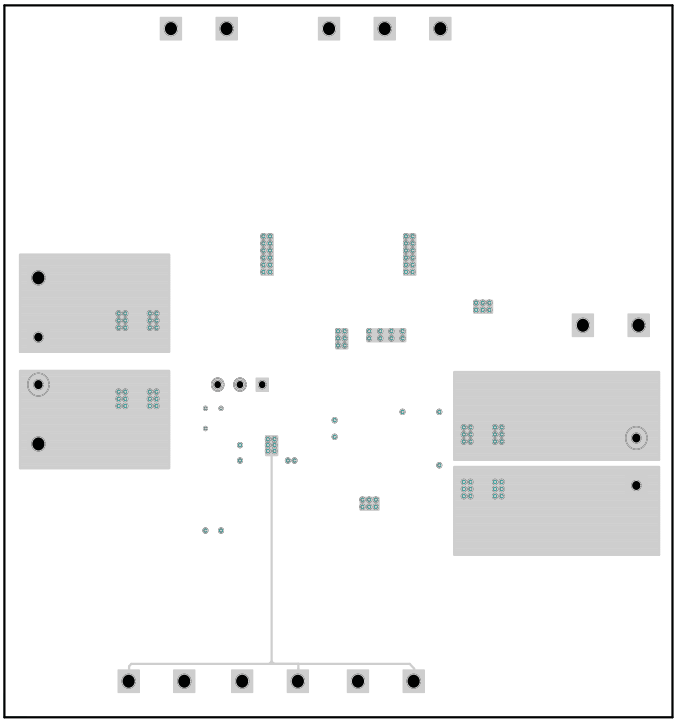


Figure 38. Internal 1 Copper Viewed from Top

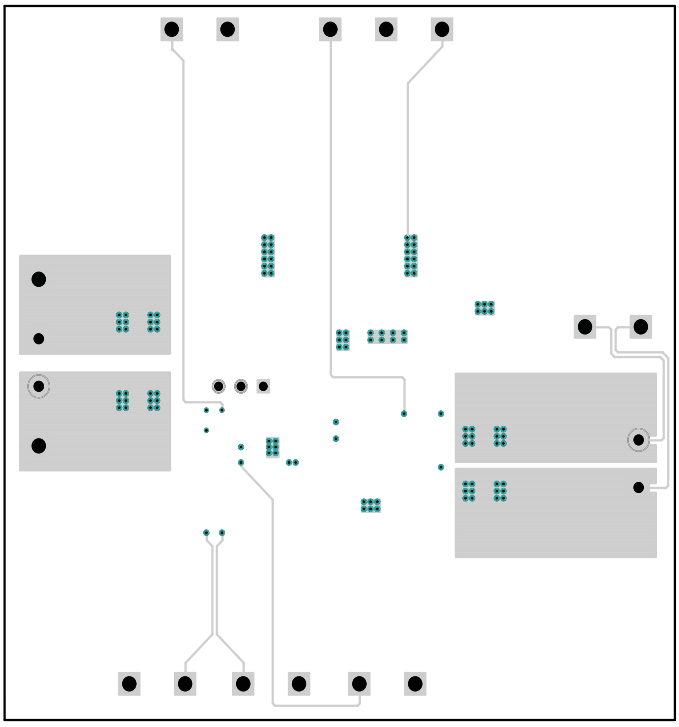


Figure 39. Internal 2 Copper Viewed from Top

11 Device and Documentation Support

11.1 Device Support

11.1.1 Third-Party Products Disclaimer

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11.1.2 Related Devices

The following devices have characteristics similar to the TPS40210-EP and may be of interest.

Table 3. Related Parts

DEVICE	DESCRIPTION
TPS6100x	Single- and dual-cell boost converter with start-up into full load
TPS6101x	High-efficiency 1-cell and 2-cell boost converters
TPS6300x	High-efficiency single inductor buck-boost converter with 1.8-A switches

11.2 Documentation Support

11.2.1 References

These references may be found on the web at www.power.ti.com under Technical Documents. Many design tools and links to additional references, may also be found at www.power.ti.com

1. *Design and Application Guide for High Speed MOSFET Gate Drive Circuits*, SEM 1400, 2001 Seminar Series
2. *Designing Stable Control Loops*, SEM 1400, 2001 Seminar Series
3. Additional PowerPAD™ information may be found in Applications Briefs SLMA002 and SLMA004
4. QFN/SON PCB Attachment, SLUA271, June 2002

11.3 Community Resources

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.

11.4 Trademarks

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11.5 Electrostatic Discharge Caution

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

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

11.6 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 part number	Status (1)	Material type (2)	Package | Pins	Package qty | Carrier	RoHS (3)	Lead finish/ Ball material (4)	MSL rating/ Peak reflow (5)	Op temp (°C)	Part marking (6)
TPS40210MDRCTEP	Active	Production	VSON (DRC) | 10	250 | SMALL T&R	Yes	NIPDAU	Level-2-260C-1 YEAR	-55 to 150	4210EP
TPS40210MDRCTEP.A	Active	Production	VSON (DRC) | 10	250 | SMALL T&R	Yes	NIPDAU	Level-2-260C-1 YEAR	-55 to 150	4210EP
V62/16602-01XE	Active	Production	VSON (DRC) | 10	250 | SMALL T&R	Yes	NIPDAU	Level-2-260C-1 YEAR	-55 to 150	4210EP

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

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

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

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

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

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

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

OTHER QUALIFIED VERSIONS OF TPS40210-EP :

- Catalog : TPS40210

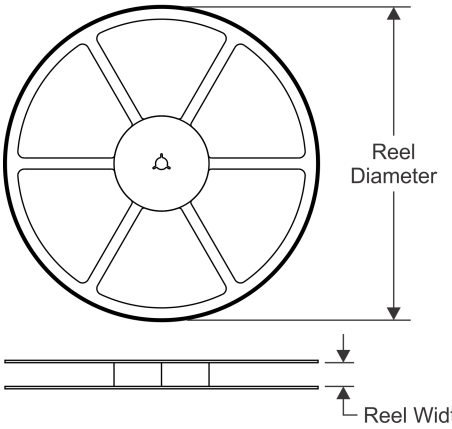
- Automotive : TPS40210-Q1

NOTE: Qualified Version Definitions:

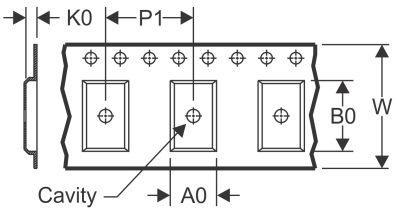
- Catalog - TI's standard catalog product
- Automotive - Q100 devices qualified for high-reliability automotive applications targeting zero defects

TAPE AND REEL INFORMATION

REEL DIMENSIONS

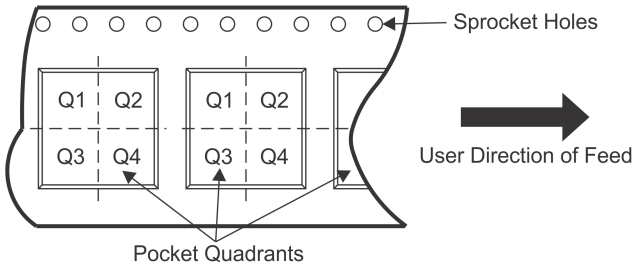


TAPE DIMENSIONS



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

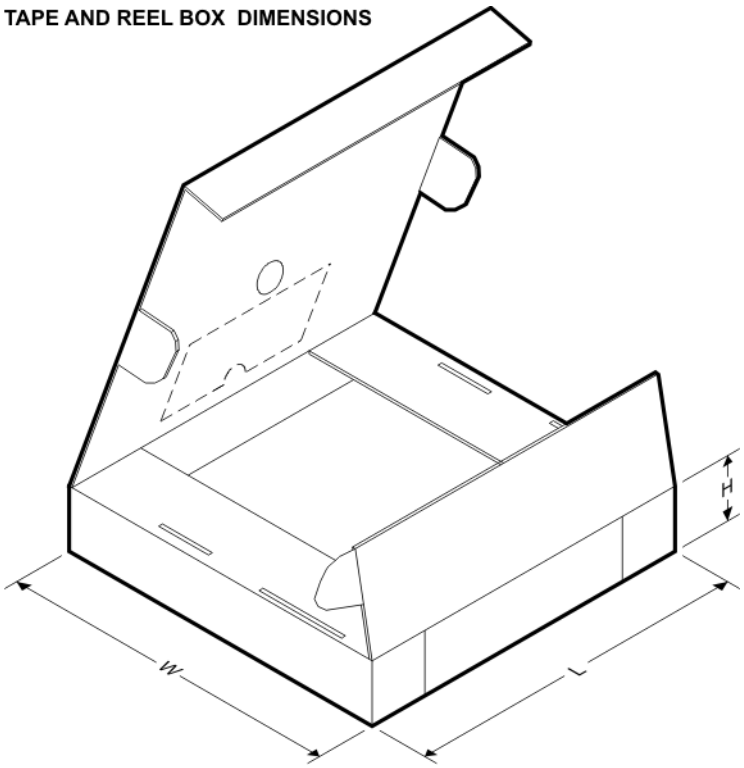
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE



*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
TPS40210MDRCTEP	VSON	DRC	10	250	180.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2

TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
TPS40210MDRCTEP	VSON	DRC	10	250	210.0	185.0	35.0

GENERIC PACKAGE VIEW

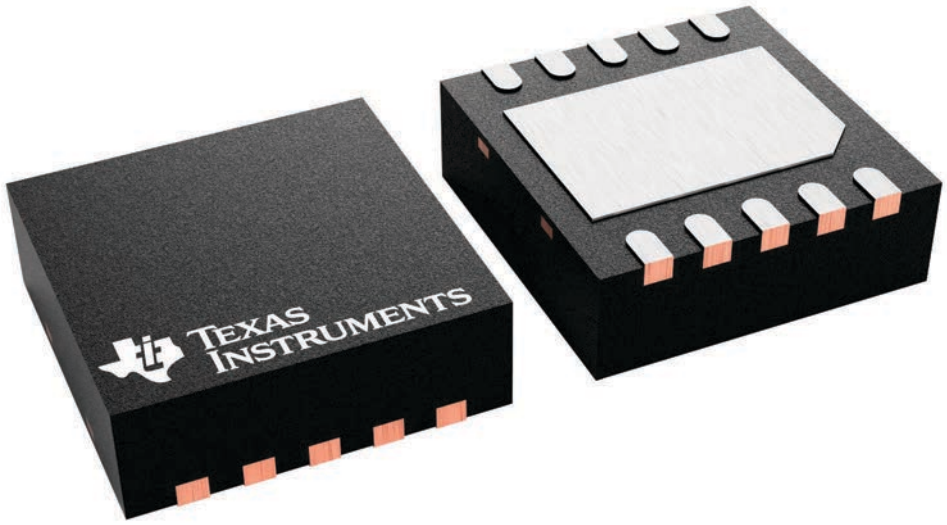
DRC 10

VSON - 1 mm max height

3 x 3, 0.5 mm pitch

PLASTIC SMALL OUTLINE - NO LEAD

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



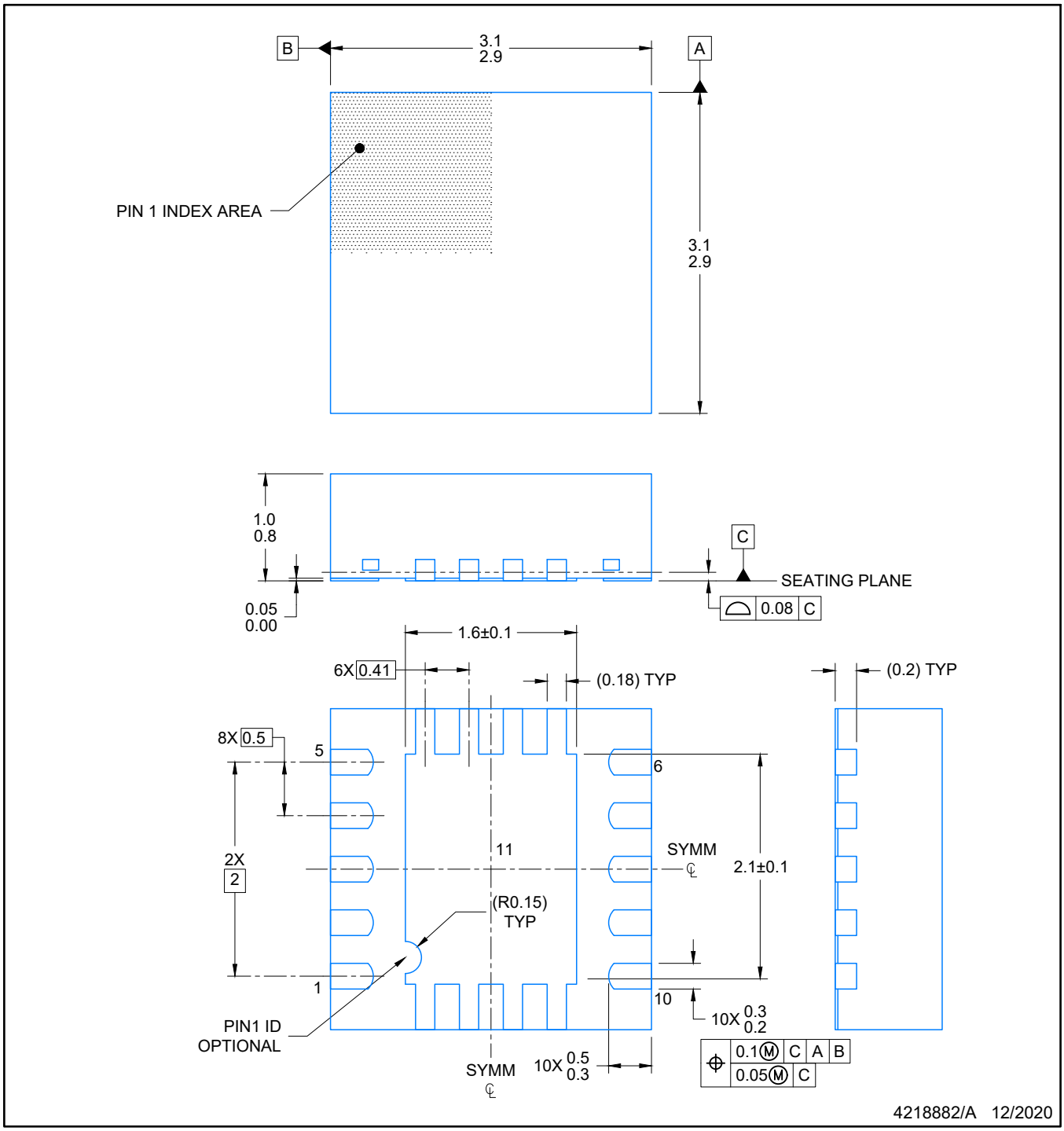
4226193/A

PACKAGE OUTLINE

DRC0010G

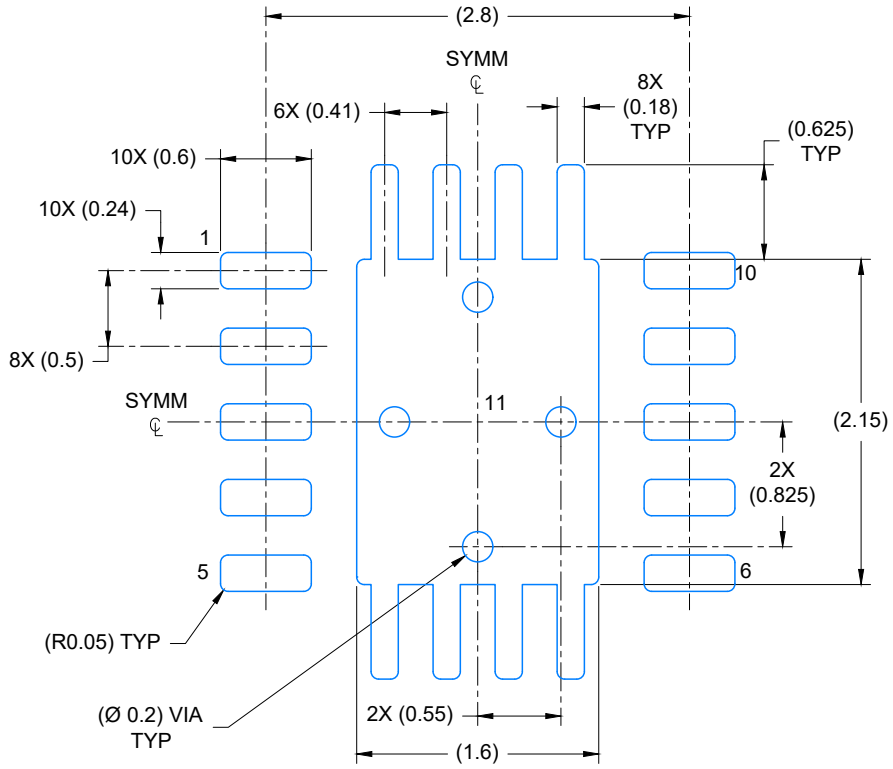
VSON - 1 mm max height

PLASTIC QUAD FLATPACK-NO LEAD

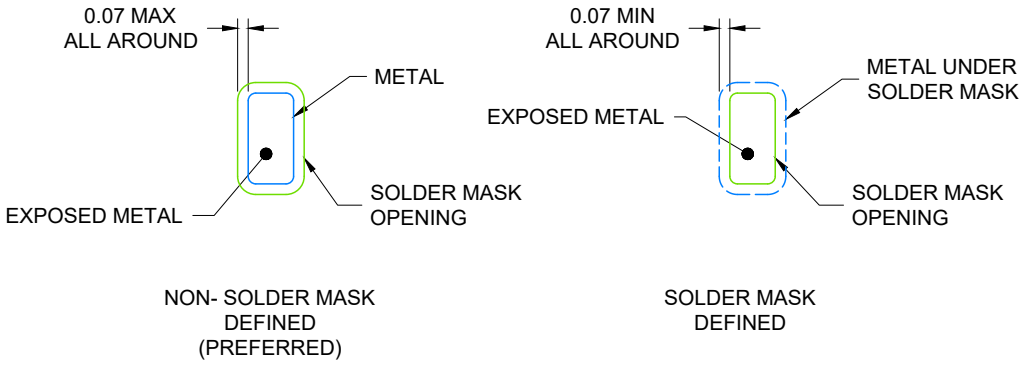


NOTES:

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



LAND PATTERN EXAMPLE
EXPOSED METAL SHOWN
SCALE: 20X



SOLDER MASK DETAILS

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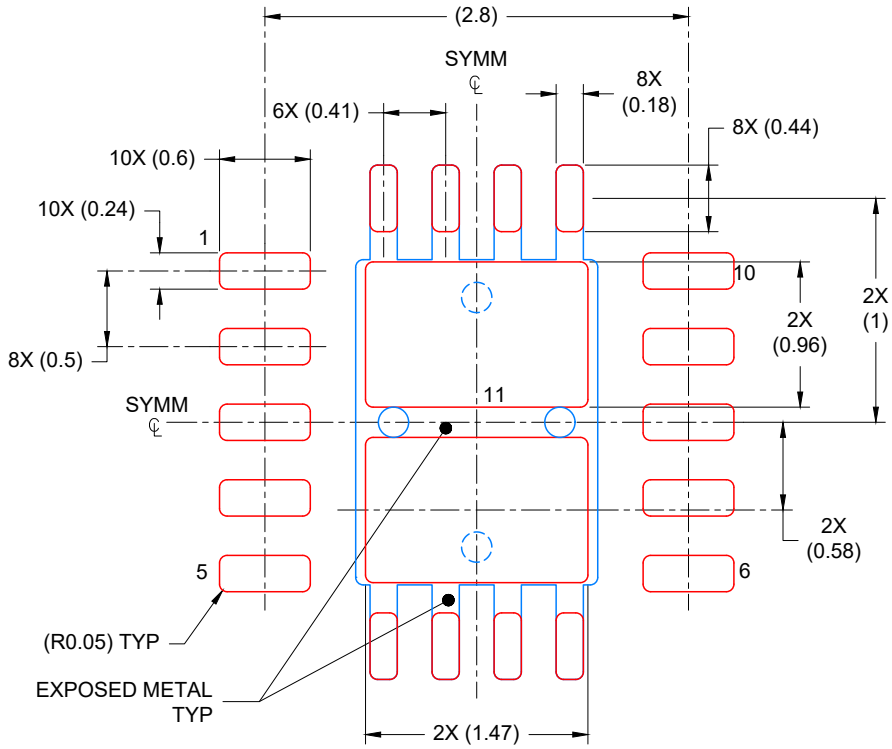
NOTES: (continued)

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



SOLDER PASTE EXAMPLE
BASED ON 0.1 mm THICK STENCIL

EXPOSED PAD
79% PRINTED COVERAGE BY AREA
SCALE: 20X

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NOTES: (continued)

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

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Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265