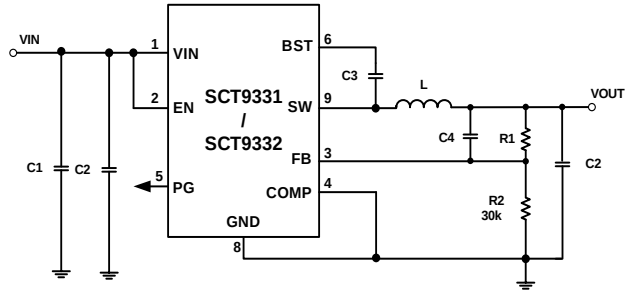
- 3.8V-32V Wide Input Voltage Range
- Up to 3.5A Continuous Output Load Current
- EMI Reduction
  - Proprietary Gate Design for Switching Node Ringing-free
  - Frequency Spread Spectrum (FSS)
- Pulse Skipping Mode (PSM) with 22uA Quiescent Current in Light Load Condition
  - SCT9331
  - Up to 80% Efficiency at 1mA Light Load
  - Up to 90% Efficiency at 10mA Light Load
- 0.8V ±1% Feedback Reference Voltage
- Fully Integrated 74m R_{dson} High Side MOSFET and 40m R_{dson} Low Side MOSFET
- 1uA Shut-down Current
- 450kHz Switching Frequency
- Precision Enable Threshold for Programmable UVLO Threshold and Hysteresis
- Output Over Voltage Protection
- Thermal Shutdown Protection at 170°C

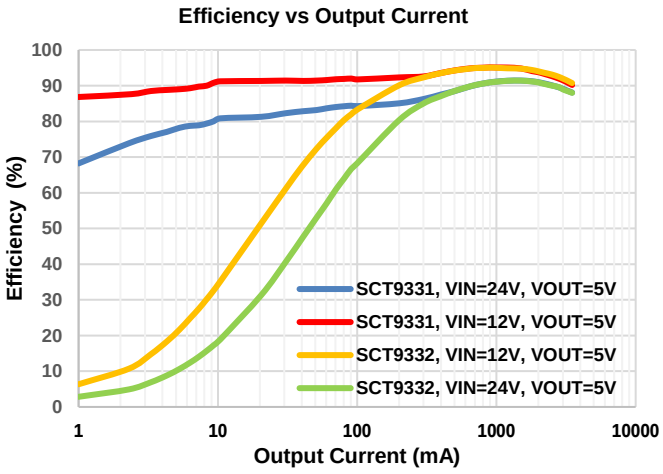
The SCT9331 and SCT9332 are 3.5A synchronous buck converters with up to 32V wide input voltage range, which fully integrates an 74m high-side MOSFET and a 40m low-side MOSFET to provide high efficiency step-down DCDC conversion. The SCT9331 and SCT9332 adopts peak current mode control with integrated compensation network. The SCT9331 supports the Pulse Skipping Modulation (PSM) with typical 22uA Ultra-Low Quiescent.

The SCT9331 and SCT9332 are optimized for Electromagnetic Interference (EMI) reduction. The converter has proprietary designed gate driver scheme to resist switching node ringing without sacrificing MOSFET turn-on and turn-off time, which further erases high frequency radiation EMI noise caused by the MOSFETs hard switching. The converter features Frequency Spread Spectrum (FSS) with a switching frequency jitter of ±6%, which reduces EMI by not allowing emitted energy to stay in any one frequency for a significant period of time.

The SCT9331 and SCT9332 offer output over-voltage protection, cycle-by-cycle peak current limit, and thermal shutdown protection. The device is available in a low-profile VDFN3020-13 3mm X 2mm package.

- White Goods, Home Appliance
- network systems
- Audio, WiFi Speaker
- Printer, Charging Station





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BST	6	Power supply for the high-side power MOSFET gate driver. Must connect a 0.1uF or greater ceramic capacitor between BST pin and SW node.
NC	7	NC
GND	8	Power ground. Must be soldered directly to ground plane.
SW	9	Switching node of the buck converter.

Over operating free-air temperature range unless otherwise noted

PARAMET 9 5p14/1

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SYMBOL	PARAMETER	TEST CONDITION	MIN	TYP	MAX	UNIT
V_{EN_H}	Enable high threshold		1.15	1.18	1.21	V
V_{EN_L}	Enable low threshold		1.02	1.1	1.14	V
I_{EN}	Enable pin input current	EN=1V	1	1.5	2	uA
I_{EN_HYS}	Enable pin hysteresis current	EN=1.5V		4		uA
Power MOSFETs						
R_{DSON_H}	High side FET on-resistance			74		m
R_{DSON_L}	Low side FET on-resistance			40		m
Feedback and Error Amplifier						
V_{FB}	Feedback Voltage		0.792	0.8	0.808	V
Current Limit						
I_{LIM_HSD}	HSD peak current limit		4.0	5	6	A
I_{LIM_LSD}	LSD valley current limit		3.2	4.2	5.2	A
Switching Frequency						
F_{SW}	Switching frequency	V_{IN}=12V, V_{OUT}=5V	400	450	500	kHz
t_{ON_MIN}	Minimum on-time			100	120	ns
Soft Start Time						
t_{SS}	Internal soft-start time			4		ms
Protection						
V_{OVP}	Output OVP threshold Hysteresis	V_{OUT} rising		110 5		% %
T_{SD}	Thermal shutdown threshold Hysteresis	T_J rising		170 25		°C

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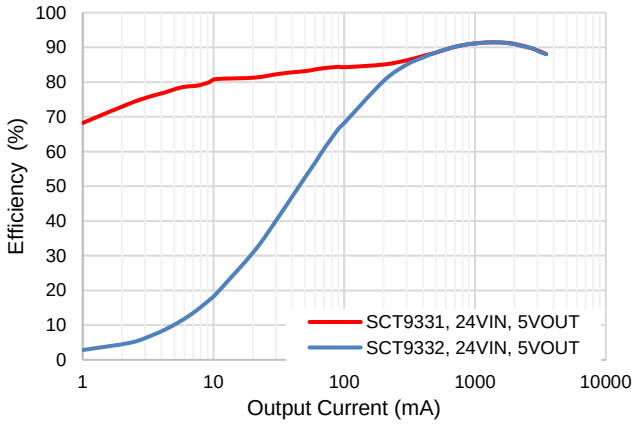

Figure 1. Efficiency vs Load Current

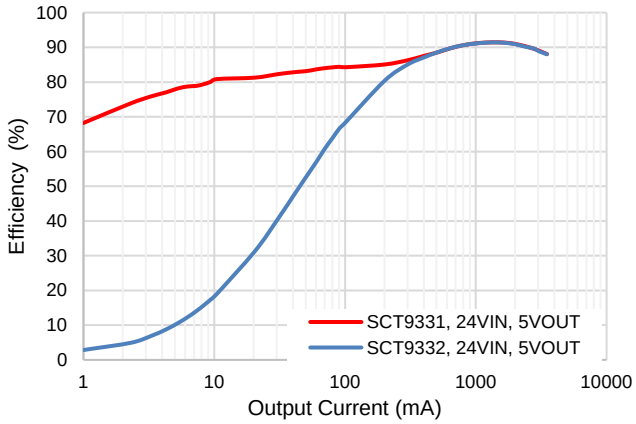

Figure 2. Efficiency vs Load Current

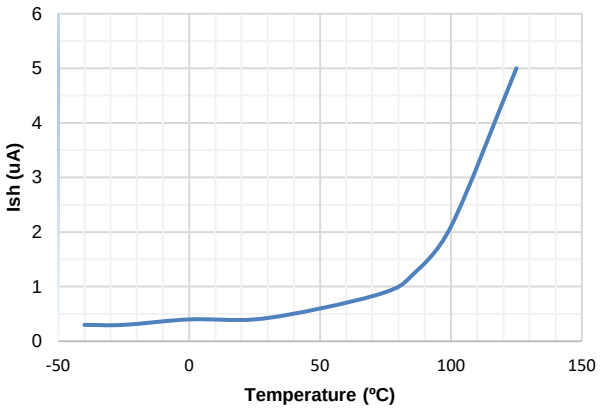

Figure 3. Shut-down Current vs Temperature

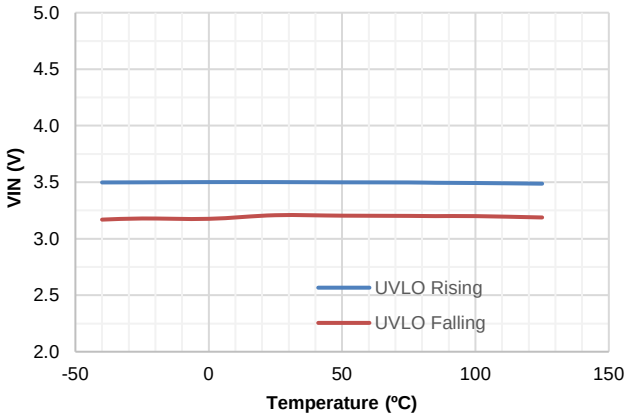

Figure 4. VIN UVLO vs Temperature

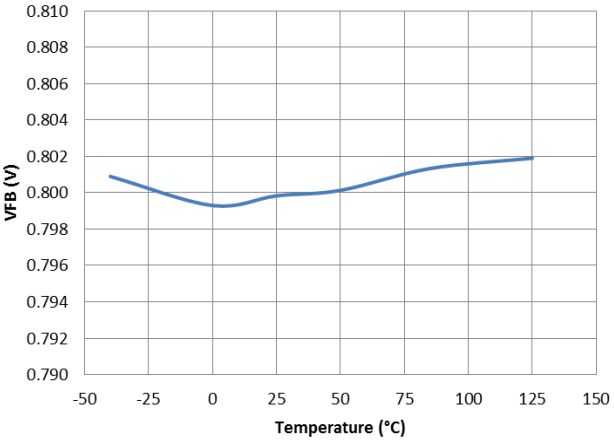

Figure 5. Reference Voltage vs Temperature

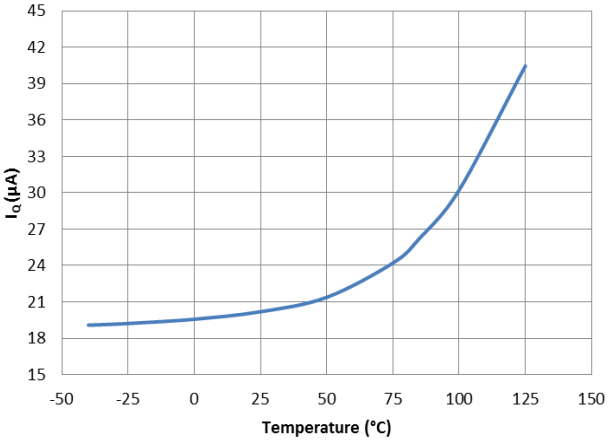

Figure 6. I_Q vs. Temperature, IOUT = 0A

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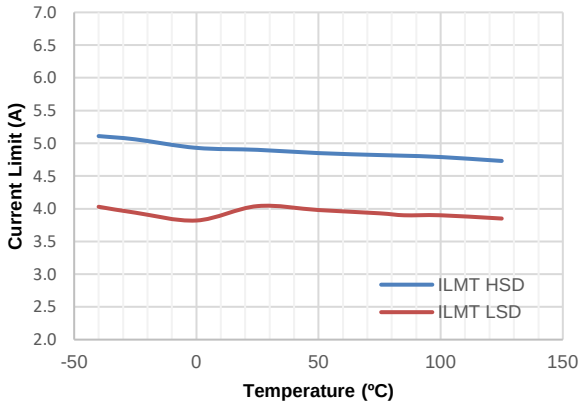

Figure 7. Peak Current Limit Vs Temperature

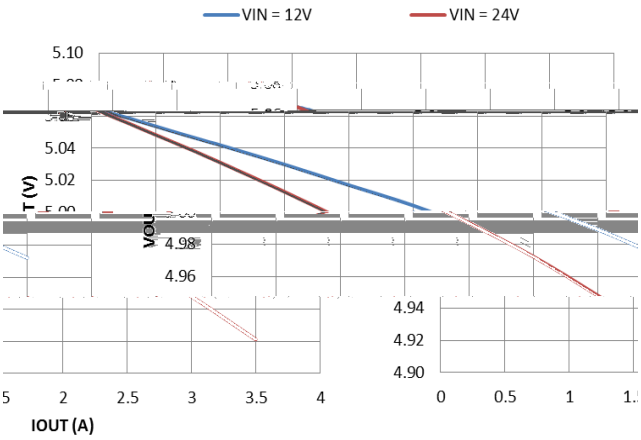

Figure 8. Load Regulation, SCT9332

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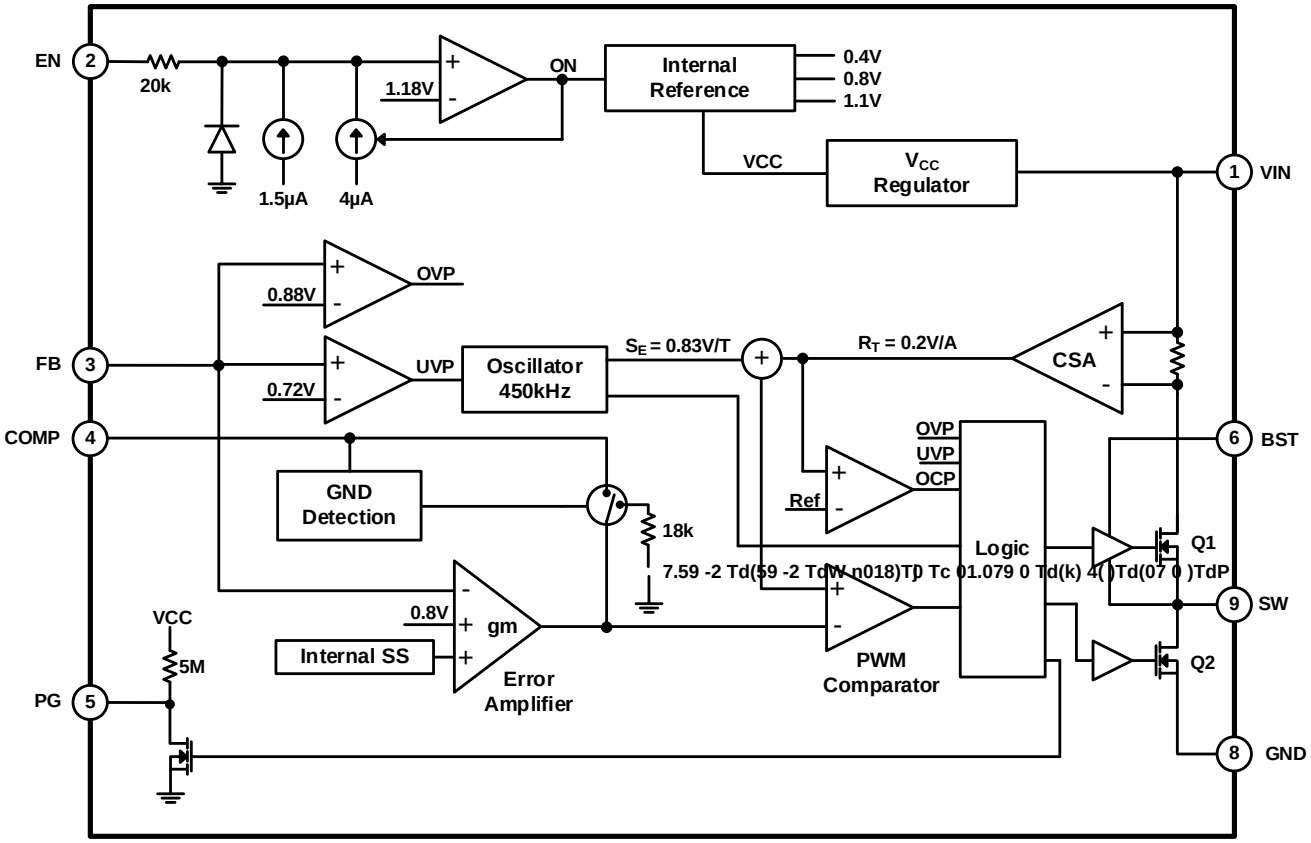
EN
2
20k
1.18V
ON
Internal Reference
0.4V
0.8V
1.1V
VCC
VCC Regulator
1
VIN
1.5μA
4μA
OVP
0.88V
FB
3
UVP
0.72V
Oscillator 450kHz
SE = 0.83V/T
RT = 0.2V/A
CSA
COMP
4
GND Detection
18k
OVP
UVP
OCP
Ref
Logic
6
BST
Q1
9
SW
Q2
0.8V
gm
Internal SS
Error Amplifier
PWM Comparator
VCC
5M
PG
5
8
GND

Overview

The SCT9331/SCT9332 device is 3.8V-32V input, 3.5A output, EMI friendly, fully integrated synchronous buck converters. The device employs fixed frequency peak current mode control. An internal clock with 450kHz frequency initiates turning on the integrated high-side power MOSFET Q1 in each cycle, then inductor current rises linearly and the converter charges output cap. When sensed voltage on high-side MOSFET peak current via the CSA block, rising above the voltage of internal COMP (see functional block diagram), the device turns off high-side MOSFET Q1 and turns on low-side MOSFET Q2. The inductor current decreases when MOSFET Q2 is ON. In the next rising edge of clock cycle, the low-side MOSFET Q2 turns off. This repeats on cycle-by-cycle based.

The the internal loop compensation network and the built-in 4ms soft-start simplify the SCT9331/SCT9332 footprints, and minimize the off-chip component counts. The quiescent current of SCT9331 is 22uA typical under no-load condition and no switching. When disabling the device, the shutdown current of SCT9331/SCT9332 is only 1 A. The SCT9331 works at Pulse Skipping Mode PSM to further increase the power efficiency in light load condition, hence the power efficiency can be achieved up to 88% at 5mA load condition.

The error amplifier serves the COMP node by comparing the voltage on the FB pin with an internal 0.8V reference voltage. When the load current increases, a reduction in the feedback voltage relative to the reference raises COMP voltage till the average inductor current matches the increased load current. This feedback loop well regulates the output voltage. The device also integrates an internal slope compensation circuitry to prevent sub-harmonic oscillation when duty cycle is greater than 50% for a fixed frequency peak current mode control. To provide the lower output ripple in light load condition, the SCT9332 works at the Force Pulse Width Modulation (FPWM) mode.

The SCT9331 device implements Frequency Spread Spectrum (FSS) with a switching frequency jitter of ±6%. FSS reduces EMI by not allowing emitted energy to stay in any one frequency for a significant period of time. The converter further dampens high frequency radiated EMI noise through the use of its proprietary gate driver scheme to achieve a ringing-free switching node voltage without sacrificing the MOSFET switching times.

In order to provide a small output ripple in light load conditions, the SCT9332 offers a fixed 450kHz switching frequency with FSS and PWM.

The hiccup mode minimizes power dissipation during prolonged output overcurrent or short conditions. The hiccup wait time is 512 cycles and the hiccup restart time is 8192 cycles. The SCT9331/SCT9332 device also features protections including cycle-by-cycle high-side MOSFET peak current limit, over-voltage protection, and over-temperature protection.

PSM Working Modes

In heavy load condition, the SCT9331 forces the device operating at PWM mode. When the load current decreasing, the internal COMP net voltage decreases as the inductor current down. With the load current further decreasing, the COMP net voltage decreases and be clamped at a voltage corresponding to the 600mA peak inductor current. When the load current approaches zero, the SCT9331 enter Pulse Skipping Mode (PSM) mode to increase the converter power efficiency at light load condition. When the inductor current decreases to zero, zero-cross detection circuitry on high-side MOSFET Q1 forces the Q1 off till the beginning of the next switching cycle. The buck converter does not sink current from the load when the output load is light and converter works in PSM mode.

FPWM Working Modes

To provide the lower output ripple in light load condition, the SCT9332 offers the fixed 450kHz switching frequency and works at the Force Pulse Width Modulation (FPWM) mode.

VIN Power

The SCT9331/SCT9332 is designed to operate from an input voltage supply range between 3.8V to 32V. At least a 0.1uF decoupling ceramic cap is recommended to bypass the supply noise. If the input supply locates more than a few inches from the converter, an additional electrolytic or tantalum bulk capacitor or with recommended 22uF may

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be required in addition to the local ceramic bypass capacitors.

Under Voltage Lockout UVLO

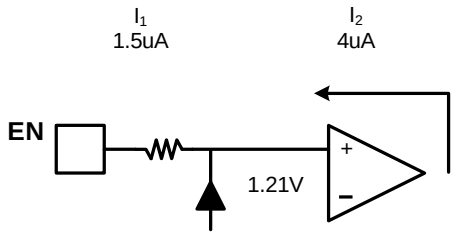
The SCT9331/SCT9332 Under Voltage Lock Out (UVLO) default startup threshold is typical 3.5V with VIN rising and shutdown threshold is 3.1V with VIN falling. The more accurate UVLO threshold can be programmed through the precision enable threshold of EN pin.

Enable and Start up

When applying a voltage higher than the EN high threshold (typical 1.18V/rise), the SCT9331/SCT9332 enables all functions and the device starts soft-start phase. The SCT9331/SCT9332 has the built in 4ms soft-start time to prevent the output overshoot and inrush current. When EN pin is pulled low, the internal SS net will be discharged to ground. Buck operation is disabled when EN voltage falls below its lower threshold (typically 1.1V/fall).

An internal 1.5uA pull up current source connected from internal LDO power rail to EN pin guarantees that floating EN pin automatically enables the device. For the application requiring higher VIN UVLO voltage than the default setup, there is a 4uA hysteresis pull up current source on EN pin which configures the VIN UVLO voltage with an off-chip resistor divider R3 and R4, shown in Figure 10. The resistor divider R3 and R4 are calculated by equation (1) and (2).

EN pin is a high voltage pin and can be directly connected to VIN to automatically start up the device with VIN rising to its internal UVLO threshold.



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- I_2=4uA
- V_{ENR}=1.18V
- V_{EMF}=1.1V

Power-GOOD (PG) Indicator

The SCT9331/SCT9332 has an open-drain output that is actively held low during soft start period until the output voltage reaches 90% of the target output. When the output voltage is outside of its regulation by -10%, the PG will pull low until the output returns to set value. The PG low to high transition is delayed by 2.5ms while the falling edge PG is delayed by 220 s to prevent false triggering.

Peak Current Limit and Hiccup Mode

The SCT9331/9332 have cycle-by-cycle peak current limit with sensing the internal high side MOSFET Q1 current during overcurrent condition. While the Q1 turns on, its conduction current is monitored by the internal sensing circuitry. Once the high-side MOSFET Q1 current exceeds the limit, it turns off immediately. If the Q1 over current time exceeds 512 switching cycles (hiccup waiting time), the buck converter enters hiccup mode and shuts down. After 8192 cycles off, the buck converter restarts to power up. The hiccup modes reduce the power dissipation in over current condition.

Over Voltage Protection and Minimum On-time

Both SCT9331/SCT9332 feature output over voltage protection (OVP). If the output feedback pin voltage exceeds 110% of feedback reference voltage, the converter stops switching immediately. When the output feedback pin voltage drops below 105% of feedback reference voltage, the converter resumes to switching. The OVP function prevents the connected output circuitry damaged from un-predictive overvoltage. Featured feedback overvoltage protection also prevents dynamic voltage spike to damage the circuitry at load during fast loading transient.

The high-side MOSFET Q1 has minimum on-time 100ns typical limitation. While the device operates at minimum on-time, further increasing VIN results in pushing output voltage beyond regulation point. With output feedback over voltage protection, the converter skips pulse by turning off high-side MOSFET Q1 and prevents output running away higher to damage the load.

Bootstrap Voltage Regulator

An external bootstrap capacitor between BST and SW pin powers floating high-side power MOSFET gate driver. The bootstrap capacitor voltage is charged from an integrated voltage regulator when high-side power MOSFET is off and low-side power MOSFET is on.

The floating supply (BST to SW) UVLO threshold is 2.7V rising and hysteresis of 350mV. When the converter operates with high duty cycle or prolongs in sleep mode for certain long time, the required time interval to recharging bootstrap capacitor is too long to keep the voltage at bootstrap capacitor sufficient. When the voltage across bootstrap capacitor drops below 2.35V, BST UVLO occurs. The SCT9331/SCT9332 intervenes to turn on low side MOSFET periodically to refresh the voltage of bootstrap capacitor to guarantee operation over a wide duty range.

EMI Reduction with Switching Node Ringing-free

In buck converter, the switching node ringing amplitude and cycles are critical especially related to the high frequency radiation EMI noise. The SCT9331/SCT9332 implements the multi-level gate driver speed technique to achieve the switching node ringing-free without scarifying the switching node rise/fall slew rate and power efficiency of the converter. The switching node ringing amplitude and cycles are damped by the built-in MOSFETs gate driving technique (SCT Patented Proprietary Design).

Thermal Shutdown

Once the junction temperature in the SCT9331/SCT9332 exceeds 170°C, the thermal sensing circuit stops converter switching and restarts with the junction temperature falling below 125°C. Thermal shutdown prevents the damage on device during excessive heat and power dissipation condition.

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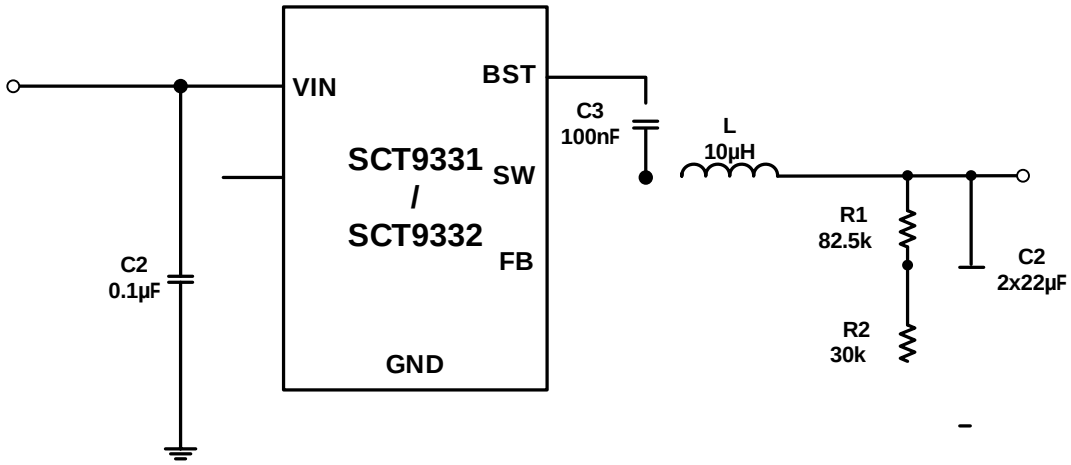
Typical Application



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Input Capacitor Selection

For good input voltage filtering, choose low-ESR ceramic capacitors. A ceramic capacitor 10 F is recommended for the decoupling capacitor and a 0.1 F ceramic bypass capacitor is recommended to be placed as close as possible to the VIN pin of the SCT9331/SCT9332.

Use Equation (3) to calculate the input voltage ripple:

$$\Delta = \frac{}{\times} \times \frac{V_{OUT}}{V_{IN}} \times \left(1 - \frac{}{}\right) \quad (3)$$

Where:

- C_{IN} is the input capacitor value
- f_{sw} is the converter switching frequency
- I_{OUT} is the maximum load current

Due to the inductor current ripple, the input voltage changes if there is parasitic inductance and resistance between the power supply and the VIN pin. It is recommended to have enough input capacitance to make sure the input voltage ripple less than 100mV. Generally, a 35V/10uF input ceramic capacitor is recommended for most of applications. Choose the right capacitor value carefully with considering high-capacitance ceramic capacitors DC bias effect, which has a strong influence on the final effective capacitance.

Inductor Selection

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different core loss. For a certain inductor, larger current ripple generates higher DCR and ESR conduction losses and higher core loss.

Output Capacitor Selection

For buck converter, the output capacitor value determines the regulator pole, the output voltage ripple, and how the regulator responds to a large change in load current. The output capacitance needs to be selected based on the most stringent of these three criteria.

For small output voltage ripple, choose a low-ESR output capacitor like a ceramic capacitor, for example, X5R and X7R family. Typically, 1 2x 22 F ceramic output capacitors work for most applications. Higher capacitor values can be used to improve the load transient response. Due to a capacitor's de-rating under DC bias, the bias can significantly reduce capacitance. Ceramic capacitors can lose most of their capacitance at rated voltage. Therefore, leave margin on the voltage rating to ensure adequate effective capacitance.

From the required output voltage ripple, use the equation (6) to calculate the minimum required effective capacitance, C_{OUT}.

$$= \frac{\Delta}{8 \times \quad \times} \tag{6}$$

Where

- $V_{OUTRipple}$ is output voltage ripple caused by charging and discharging of the output capacitor.
- I_{LPP} is the inductor peak to peak ripple current, equal to k_{IND} * I_{OUT}.
- f_{SW} is the converter switching frequency.

The allowed maximum ESR of the output capacitor is calculated by the equation (7).

$$= \frac{}{\Delta} \tag{7}$$

The output capacitor affects the crossover frequency f_C. Considering the loop stability and effect of the internal loop compensation parameters, choose the crossover frequency less than 45 kHz ($\frac{1}{10} \times f_{SW}$) without considering the feed-forward capacitor. A simple estimation for the crossover frequency without feed forward capacitor is shown in equation (8), assuming C_{OUT} has small ESR.

$$> \frac{18 \times \quad \times \quad \times 0.8}{2 \times \quad \times} \tag{8}$$

Where

- G_M is the transfer conductance of the error amplifier (300uS).
- G_{MP} is the gain from internal COMP to inductor current, which is 5A/V.
- f_C is the cross over frequency.

Additional capacitance de-rating for aging, temperature and DC bias should be factored in which increases this minimum value. Capacitors generally have limits to the amount of ripple current they can handle without failing or producing excess heat. An output capacitor that can support the inductor ripple current must be specified. The capacitor data sheets specify the RMS (Root Mean Square) value of the maximum ripple current. Equation (9) can be used to calculate the RMS ripple current the output capacitor needs to support.

$$= \frac{\cdot (\quad - \quad)}{\sqrt{12} \cdot \quad \cdot \quad \cdot} \tag{9}$$

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Output Feed-Forward Capacitor Selection

The SCT9331/SCT9332 has the internal integrated loop compensation as shown in the function block diagram. The compensation network includes a 18k resistor and a 7.6nF capacitor. Usually, the type II compensation network has a phase margin between 60 and 90 degree. However, if the output capacitor has ultra-low ESR, the converter results in low phase margin. To increase the converter phase margin, a feed-forward cap C_{ff} is used to boost the phase margin at the converter cross-over frequency f_c. Equation (10) is used to calculate the feed-forward capacitor.

$$= \frac{1}{2 \cdot \times _1} \quad (10)$$

Output Feedback Resistor Divider Selection

The SCT9331/SCT9332 features external programmable output voltage by using a resistor divider network R1 and R2 as shown in the typical application circuit Figure11. Use equation (11) to calculate the resistor divider values.

$$_1 = \frac{(\quad - \quad) \times \ _2}{} \quad (11)$$

Set the resistor R2 value to be approximately 30k. Slightly increasing or decreasing R1 can result in closer output voltage matching when using standard value resistors.
Table 1 is the recommend external components for the application with integrated loop compensation for SCT9331 and SCT9332.

Table 1. Recommended External Components

Vout	L1	COUT	R1	R2	Rt	C4
3.3V	6.5uH	3*22uF	93.5k	30k	2k	22p
5V	10uH	3*22uF	158k	30k	2k	100p
12V	22uH	3*22uF	422k	30k	2k	330p

Loop Response Design

When the COMP pin is not connected to GND, the COMP pin is active for external loop compensation. The regulator uses constant frequency peak current mode control architecture to achieve a fast loop response. The inductor is not considered as a state variable since its peak current is constant. The system becomes a single order system. It is much easier to design a type II compensator (Figure 13) to stabilize the loop than in case of a voltage mode control. Peak current mode control has an inherent input voltage feed-forward function to achieve good line regulation. Figure 12 shows the small signal model of the synchronous buck regulator.

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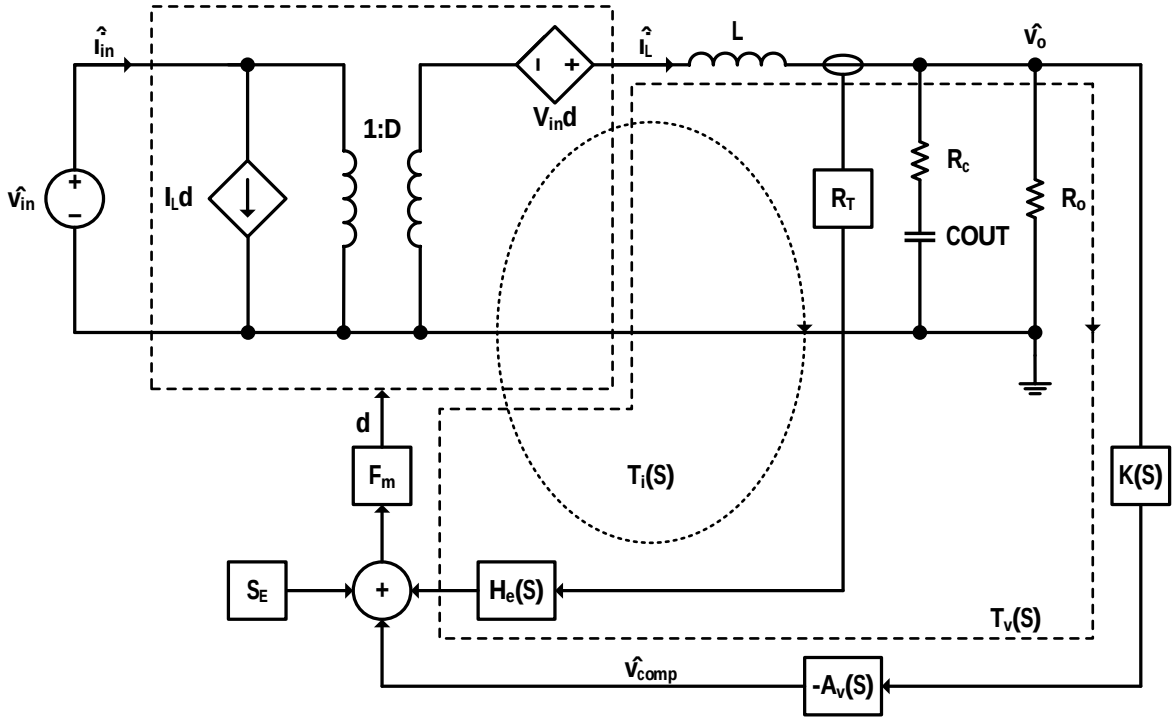

Figure 12. Small Signal Model of Buck Regulator

Where:

- $T_v(S)$ is the voltage loop
- $T_i(S)$ is the current loop
- K(S) is the voltage sense gain
- $-A_v(S)$ is the feedback compensation gain
- $H_e(S)$ is the current sampling function
- F_m is the PWM comparator gain
- V_{in} is the DC input voltage
- D is the duty cycle
- R_c is the ESR of the output capacitor, COUT
- R_o is the output load resistance $_{in}$ is the AC small-signal input voltage
- $_{in}$ is the AC small-signal input current
-
- $_L$ is the AC small signal of the inductor current
- $_o$ is the AC small signal of output voltage
- $_{comp}$ is the AC small signal voltage of the compensation network

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- Co is the effective output capacitance.

Be cautioned that most ceramic will degrade with voltage stress or temperature extremes.

The compensator capacitor C5 and C6 are then equal to:

$$5 = \frac{\cdot}{\cdot\ 5}, \quad 6 = \frac{\cdot}{5}, \frac{1}{\cdot\ \cdot\ 5} \tag{14}$$

Where :

- Io is the output load current
- Rc is the ESR equivalent of the Co
- Fs is the switching frequency. In most cases, C6 can omit.

An optional zero, ωz2, can boost the phase margin but it can also increase the gain crossover. Place this zero at 2 to 5 times the fc. Then C4 is equal to:

$$4 = \frac{1}{10\ \cdot\ \cdot\ 1} \quad \frac{1}{4\ \cdot\ \cdot\ 1} \tag{15}$$

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Application Waveforms

Figure 14. SW node waveform and Output Ripple
VIN=12V, IOUT=3.5A

Figure 15. SW node Waveform and Output Ripple
VIN=12V, IOUT=10mA

Figure 16 5Qq2 -0.004 Tw -4.0 0 9.

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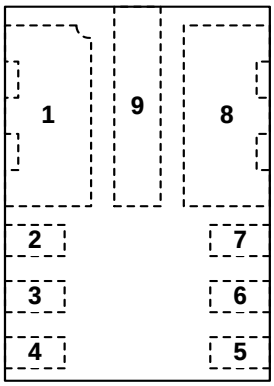
Layout Guideline

Figure 20 and Figure 21 are the recommended PCB layout of SCT9331 and SCT9332 with or without external compensation network.

1. The SCT9331 and SCT9332 works at 3.5A load current so heat dissipation is a major concern in the layout of the PCB. 2oz copper for both the top and bottom layers is recommended.
2. Place the input capacitors as closely across VIN and GND as possible.
3. Place the inductor as close to SW as possible.
4. Place the output capacitors as close to GND as possible.
5. Place the feedback components as close to FB as possible.
6. If using four or more layers, use at least the 2nd and 3rd layers as GND to maximize thermal performance.
7. Add as many vias under both the thermally exposed GND pad and GND plane for heat dissipation to all the GND layers.
8. Add as many vias under both the thermally exposed VIN pad and VIN plane for heat dissipation to all the VIN layers.



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Thermal Considerations

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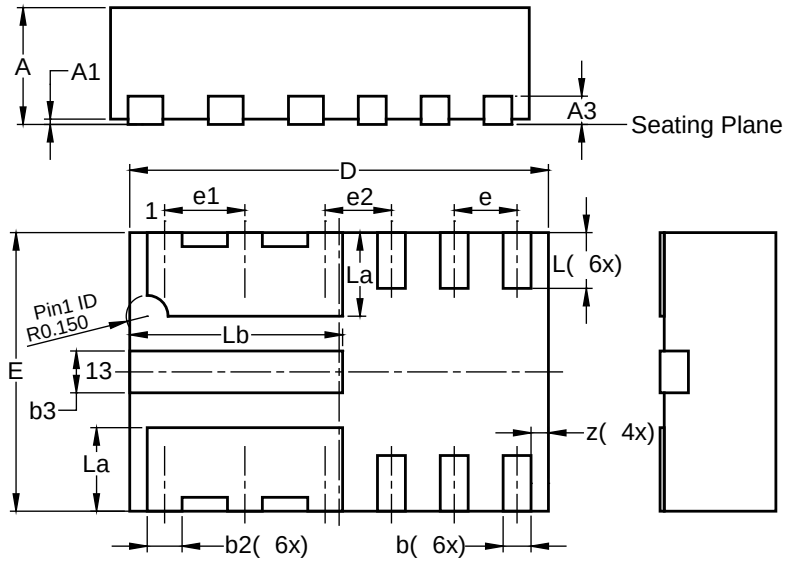
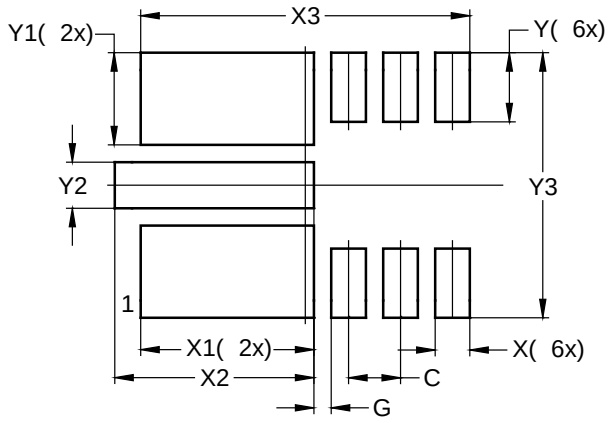
V-DFN3020-13 (Type A)



V-DFN3020-13			
Dim	**Min**	**Max**	**Typ**
A	0.75	0.85	0.80
A1	0.00	0.05	0.02
A3	-	-	0.203
b	0.15	0.25	0.20
b2	0.20	0.30	0.25
b3	0.25	0.35	0.30
D	2.95	3.05	3.00
E	1.95	2.05	2.00
e	0.45 BSC		
e1	0.575 BSC		
e2	0.475 BSC		
L	0.35	0.45	0.40
La	0.55	0.65	0.60
Lb	1.475	1.575	1.525
z	-	-	0.125
All Dimensions in mm			

Suggested Pad Layout

V-DFN3020-13 (Type A)



Dimensions	Value (in mm)
C	0.45
G	0.15
X	0.30
X1	1.50
X2	1.73
X3	2.85
Y	0.60
Y1	0.80
Y2	0.40
Y3	2.30

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