

## 1A, 36V, 2.1MHz Synchronous Step-Down Converter

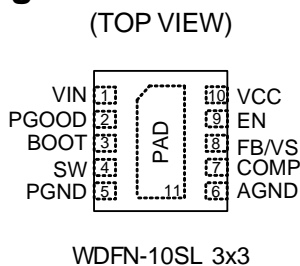
### General Description

The RTQ2131B is a high-efficiency, monolithic synchronous step-down DC-DC converter that can deliver up to 1A output current from a 3V to 36V widely input supply. The RTQ2131B current-mode control with external compensation allows the transient response to be optimized over various loads and output capacitors. Cycle-by-cycle current limit provides protection against short-circuit output, and soft-start can eliminate input current surge during startup. Input under-voltage lockout, output under-voltage protection, over-current protection and over-temperature protection offers completely safe and smooth operation in all applied conditions. The RTQ2131B is fully specified over the temperature range of  $T_A = -40^{\circ}\text{C}$  to  $125^{\circ}\text{C}$  and is available in WDFN-10SL 3x3 package.

### Applications

- Automotive Systems
- Car Camera Module and Car Cockpit Systems
- Connected Car Systems
- Point of Load Regulator in Distributed Power Systems
- Digital Set Top Boxes
- Broadband Communications

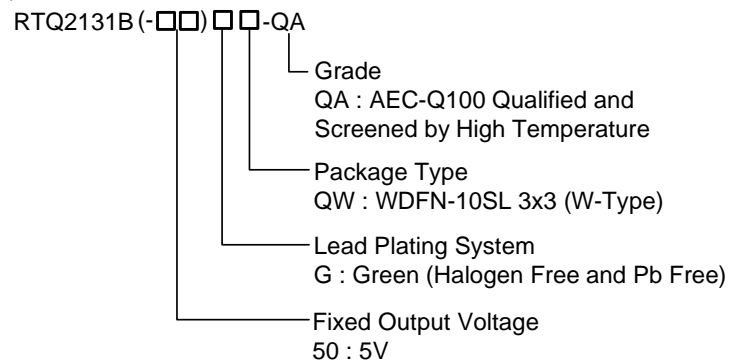
### Pin Configuration



### Features

- AEC-Q100 Grade 1 Qualified
- 3V to 36V Input Voltage Range
- Tight Switching Frequency Variation  $2.1\text{MHz} \pm 10\%$  Over Operating Ambient Temperature
- Built-In Spread-Spectrum Frequency Modulation
- 5V Fixed Output Voltage ( see Ordering Information for availability)
- 1A Output Current
- $-40^{\circ}\text{C}$  to  $125^{\circ}\text{C}$  Operating Ambient Temperature Range
- Current-Mode Control
- Integrated  $200\text{m}\Omega/160\text{m}\Omega$  MOSFETs
- Stable with Low-ESR Ceramic Output Capacitors
- Enable Control
- Over-Temperature Protection
- Output Under-Voltage Protection with Hiccup Mode
- Adjacent Pin Short-Circuit Protection
- Power-Good Indicator
- Cycle-by-Cycle Over-Current Protection
- Input Under-Voltage Lockout
- RoHS Compliant and Halogen Free

### Ordering Information



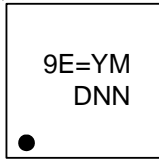
Note :

Richtek products are :

- ▶ RoHS compliant and compatible with the current requirements of IPC/JEDEC J-STD-020.
- ▶ Suitable for use in SnPb or Pb-free soldering processes.

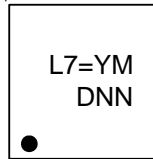
## Marking Information

RTQ2131BGQW-QA



9E= : Product Code  
YMDNN : Date Code

RTQ2131B-50GQW-QA



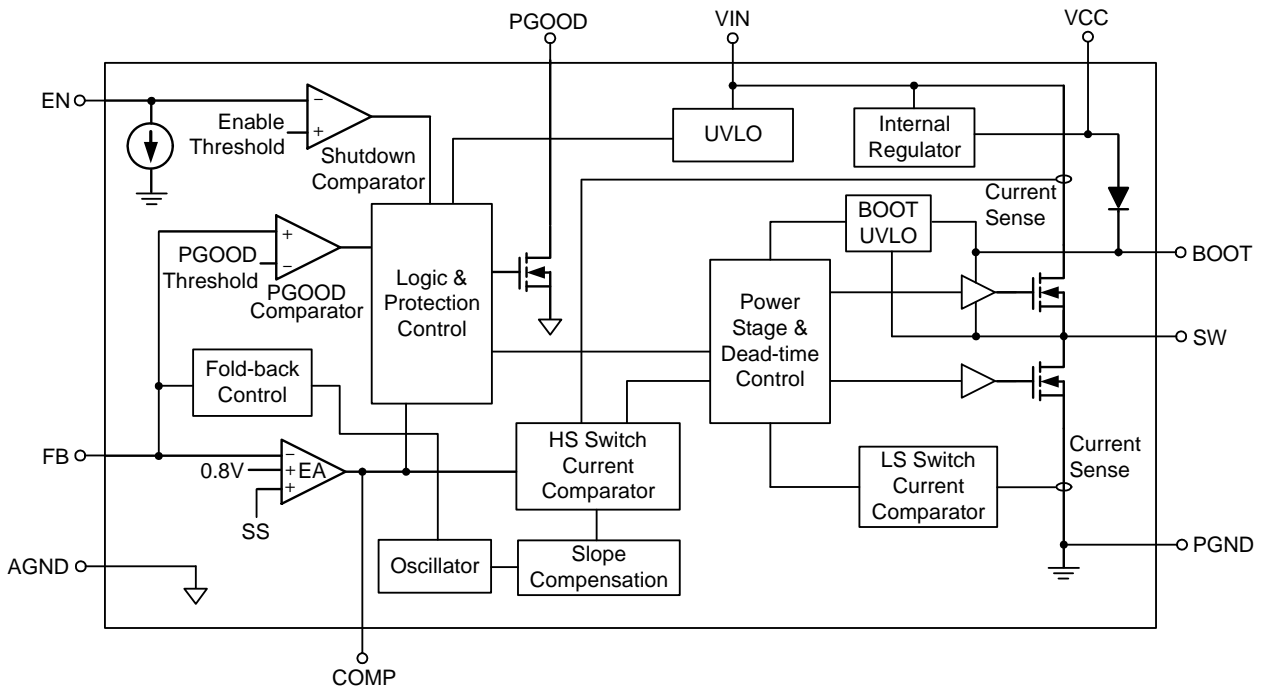
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## Functional Pin Description

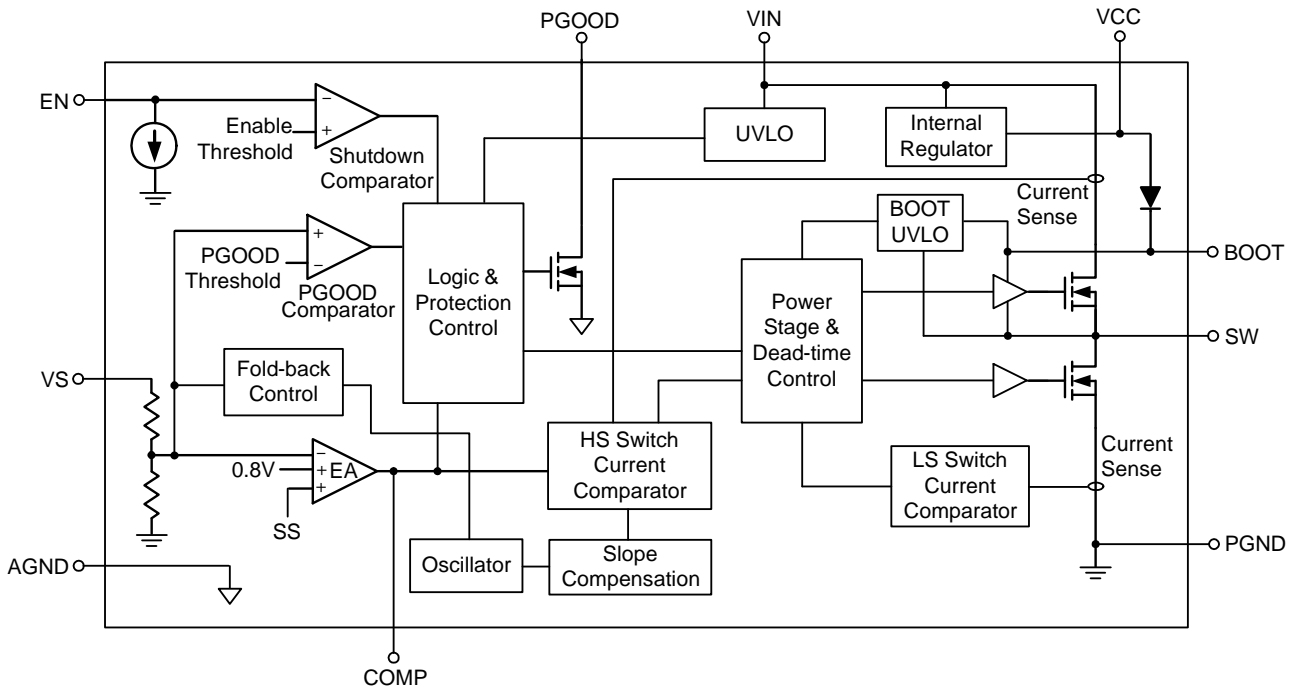
| Pin No.          | Pin Name | Pin Function   |
|------------------|----------|--|
| 1                | VIN      | Supply input. Supplies the power to the internal control circuit as well as the power switches of the device. Drive VIN with a 3V to 36V power source and bypass VIN to GND with a capacitor to eliminate noise on the input to the IC.  |
| 2                | PGOOD    | Power-good indicator output. This pin is an open-drain logic output that is pulled to ground when the output voltage is lower or higher than its specified threshold under the conditions of UVP, OTP, EN shutdown, or during slow start.  |
| 3                | BOOT     | Bootstrap supply for high-side gate driver. Connect a 100nF or greater capacitor from SW to BOOT to power the high-side switch.  |
| 4                | SW       | Switch node. SW is the switching node that supplies power to the output and connect the output LC filter from SW to the output load.   |
| 5                | PGND     | Power ground.  |
| 6                | AGND     | Analog ground.   |
| 7                | COMP     | Compensation node. The current comparator threshold increases with this control voltage. Connect external compensation elements to this pin to stabilize the control loop.   |
| 8                | FB/VS    | Output voltage sense. There are two output voltage setting options : one is that trimmed output voltage options for a fixed output voltage are available for the VS pin, and the other is through a resistive divider to sense the output voltage at the FB pin. The feedback reference voltage is 0.8V typically. |
| 9                | EN       | Enable control input. Connecting this pin to logic high can enable the device and connecting this pin to GND can disable the device.   |
| 10               | VCC      | Linear regulator output. VCC is the output of the internal 5V linear regulator powered by VIN. Decouple with a 1 $\mu$ F ceramic capacitor from VCC to ground for normal operation.  |
| 11 (Exposed Pad) | PAD      | Exposed pad. The exposed pad is internally unconnected and must be soldered to a large PCB copper area for maximum power dissipation.  |

**Functional Block Diagram**

**Adjustable Output Voltage**



**Fixed 5V Output Voltage**



## Operation

### Switching Frequency and Operational Duty Cycle

With normal duty cycle, the RTQ2131B operates at fixed switching frequency = 2.1MHz. If the duty cycle is too high/low, in order to extend the operational input voltage range, the RTQ2131B is forced to lower its switching frequency by skipping on/off pulses.

The minimum on time of the RTQ2131B is 60ns (typ., at  $I_{OUT} = 0.5A$ ). If input voltage is too high and the required on pulse width becomes smaller than minimum on time, IC starts to skip on pulse once the  $V_{(COMP)}$  has reached its low clamped value.

The minimum off time of the RTQ2131B is 60ns (typ.). If input voltage drops and the required off pulse width becomes smaller than minimum off time, IC starts to skip off pulse and keeps H/S MOSFET on continuously. With this feature, IC can support > 99% duty cycle.

### Input Under-Voltage Lockout Threshold

The RTQ2131B includes an input under-voltage lockout protection (UVLO) circuit. If input voltage exceeds a UVLO rising threshold voltage ( $V_{UVLO}$ ), the converter will reset and prepare the PWM for operation. If the input voltage falls below a UVLO falling threshold voltage (i.e.  $V_{UVLO} - \Delta V_{UVLO}$ ) during normal operation, the converter will stop switching. There is some hysteresis between UVLO rising and falling threshold voltages to prevent a reset caused by noise.

Note that  $V_{IN} = 3V$  is only design for cold crank requirement. Normal input voltage should be larger than UVLO rising threshold to turn on.

### Chip Enable

The EN pin is the chip enable input. Pulling the EN pin low to 1.25V (typ) will disable output voltage and low to 0.4V will shut down the device. During shutdown mode, the shutdown current of the RTQ2131B is lower than 10 $\mu A$ . The EN pin has an internal pull-down current source. When VCC exceeds 5V, the current source typically sinks 1.2 $\mu A$  for  $V_{EN} < 4V$  and up to 70 $\mu A$  for  $V_{EN} > 4V$ .

### Internal Regulator

The internal regulator generates a 5V regulated voltage,  $V_{CC}$ , to drive internal circuitry. When  $V_{IN}$  is below 5V,  $V_{CC}$  will track with  $V_{IN}$ . A capacitor ( $= 1\mu F$ ), connected from  $V_{CC}$  to GND, is required.

### Internal Soft-Start Function

The RTQ2131B provides an internal soft-start function. The soft-start function is used to prevent large inrush current while the converter is being powered-up. The typical soft-start time (i.e. for the FB voltage to ramp from 0V to 0.8V) is 2ms.

When voltage of EN pin exceeds threshold voltage,  $V_{CC}$  will start up first and after 0.8ms output voltage ramp up during soft-start time as shown in Figure 1.

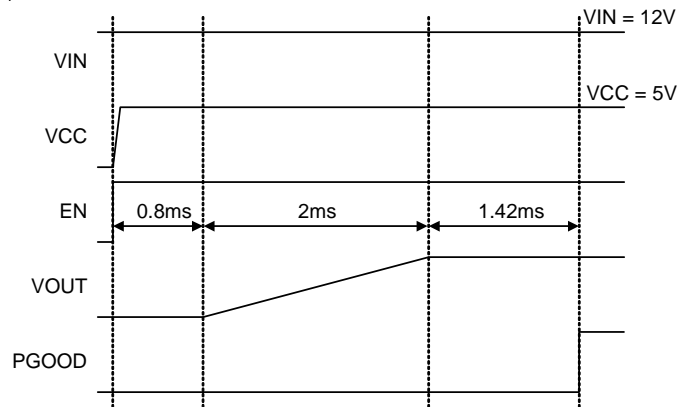


Figure 1. Soft-Start Sequence

### Power-Good Output

The RTQ2131B provides a Power-Good indicator Output function.

The PGOOD pin is an open-drain power-good output and requires an external pull-up resistor, connected to an external supply or the on-chip  $V_{CC}$  output. When the output voltage drops below 85% of its nominal voltage, PGOOD will be pulled low. It will be held low until the output voltage rises to 90% of the nominal voltage. During soft-start and in shutdown, PGOOD is actively held low. When the soft-start sequence is finished and the output voltage has reached 90% of its nominal voltage, PGOOD output driver is in high impedance and the PGOOD voltage will be pulled high by the external pull-up resistor.

**High-Side MOSFET Over-Current Limit**

IC detects the inductor current during the period high-side MOSFET turns on. When inductor current reaches the current limit value, IC will turn off high-side MOSFET, so it can provide cycle-by-cycle peak current limit protection to prevent IC from over current.

**Low-Side MOSFET Over-Current Limit**

The RTQ2131B not only implements the high-side switch current limit but also provides the low-side switch sourcing current limit and low-side switch sinking current limit for low-side MOSFET. With these current protections, the IC can easily control inductor current at both side power switches and avoid current runaway for short-circuit condition.

For the low-side switch sourcing current limit, there is a specific comparator in internal circuitry to compare the low-side MOSFET sourcing current to the internal set current limit at the end of every clock cycle. When the low-side sourcing current is higher than the set sourcing limit, the high-side power switch is not turned on and low-side power switch is kept on until the following clock cycle for releasing the above sourcing current to the load. It is allowed to turn on the high-side MOSFET again when the low-side current is lower than the set sourcing current limit at the beginning of a new cycle.

For the low-side switch sinking current limit protection, it is implemented by detecting the voltage across the low-side power switch. If the low-side reverse current exceeds the set sinking limit, both power switches are off immediately, and it is held to stop switching until the beginning of next cycle.

**Output Under-Voltage Protection**

The RTQ2131B includes output under-voltage protection (UVP) against over-load or short-circuited condition by constantly monitoring the feedback voltage  $V_{FB}$ . If  $V_{FB}$  drops below the under-voltage protection trip threshold, 50% (typ.) of the internal reference voltage, the UV comparator will go high to turn off the internal high-side MOSFET switches. If the output under-voltage condition continues for a period of time, the RTQ2131B will enter output under-voltage protection with hiccup mode. During

hiccup mode, the device remains shut down. After a period of time, a soft-start sequence for auto-recovery will be initiated. Upon completion of the soft-start sequence, if the fault condition is removed, the converter will resume normal operation; otherwise, such cycle for auto-recovery will be repeated until the fault condition is cleared. Hiccup mode allows the circuit to operate safely with low input current and power dissipation, and then resume normal operation as soon as the over-load or short-circuit condition is removed. The UVP profile is shown in Figure 2.

**BOOT UVLO**

To ensure that high-side switch can be controlled normally, the RTQ2131B implements BOOT UVLO function to prevent operation under low  $V_{(BOOT-SW)}$  condition. When  $V_{(BOOT-SW)}$  is lower than BOOT\_UVLO\_L level 2.5V (typ.), IC will disable 99% duty cycle operation, and the minimum off time will be enlarged to 100ns (typ.) to extend BOOT charging time. BOOT UVLO condition will be removed after  $V_{(BOOT-SW)}$  is higher than 2.8V (typ.).

**Over-Temperature Protection**

Over-temperature protection is implemented to prevent the chip from operating at excessively high temperatures. When the junction temperature is higher than 175°C, the OTP will shut down switching operation. The chip will automatically resume normal operation with a complete soft-start sequence once the junction temperature cools down by approximately 15°C.

**Adjacent Pin Short-Circuit Protection**

The RTQ2131B provide adjacent Pin short-circuit protection.

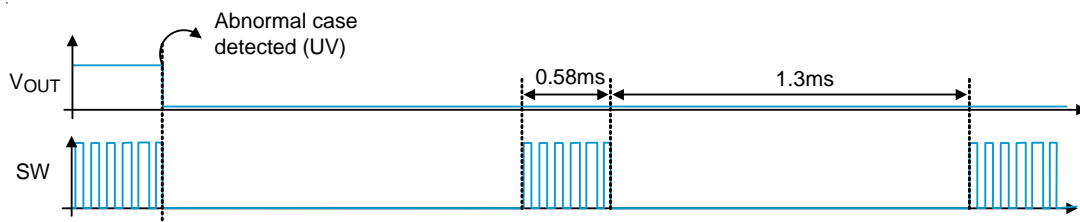


Figure 2. Output Under-Voltage Protection with Hiccup Mode

**Absolute Maximum Ratings** (Note 1)

- Supply Input Voltage,  $V_{IN}$  ----- -0.3V to 42V
- Switch Voltage,  $SW$  ----- -0.3V to 42V  
   <100ns ----- -5V to 46.3V
- BOOT to SW,  $V_{BOOT} - V_{SW}$  ----- -0.3V to 6V
- EN, PGOOD Voltage ----- -0.3V to 42V
- Other Pins ----- -0.3V to 6V
- Power Dissipation,  $P_D$  @  $T_A = 25^\circ C$   
   WDFN-10SL 3x3 ----- 3.22W
- Package Thermal Resistance (Note 2)  
   WDFN-10SL 3x3,  $\theta_{JA}$  ----- 38.8°C/W  
   WDFN-10SL 3x3,  $\theta_{JC}$  ----- 8.1°C/W
- Junction Temperature ----- 150°C
- Lead Temperature (Soldering, 10 sec.) ----- 260°C
- Storage Temperature Range ----- -65°C to 150°C
- ESD Susceptibility (Note 3)  
   HBM (Human Body Model) ----- 2kV

**Recommended Operating Conditions** (Note 4)

- Supply Voltage ----- 3V to 36V
- Output Voltage ----- 0.8V to 28V
- Junction Temperature Range ----- -40°C to 150°C
- Ambient Temperature Range ----- -40°C to 125°C

**Electrical Characteristics**

( $V_{IN} = 12V$ ,  $T_A = T_J = -40^\circ C$  to  $125^\circ C$ , unless otherwise specified)

| Parameter                                  | Symbol            | Test Conditions               | Min | Typ  | Max | Unit    |
|--|-------------------|-------------------------------|-----|------|-----|---------|
| <b>Supply Voltage</b>                      |                   |                               |     |      |     |         |
| Input Operating Voltage                    | $V_{IN}$          |                               | 3   | --   | 36  | V       |
| Under-Voltage Lockout Threshold            | $V_{UVLO}$        | $V_{IN}$ rising               | 3.6 | 3.8  | 4   | V       |
| Under-Voltage Lockout Threshold Hysteresis | $\Delta V_{UVLO}$ |                               | --  | 900  | --  | mV      |
| Shutdown Current                           | $I_{SHDN}$        | $V_{EN} = 0V$                 | --  | --   | 10  | $\mu A$ |
| Quiescent Current                          | $I_Q$             | $V_{EN} = 2V$ , not switching | --  | 1.1  | 1.3 | mA      |
| <b>Enable Voltage</b>                      |                   |                               |     |      |     |         |
| Enable Threshold Voltage                   | $V_{IH}$          | $V_{EN}$ rising               | 1.3 | 1.45 | 1.6 | V       |
|  | $V_{IL}$          | $V_{EN}$ falling              | 1.1 | 1.25 | 1.4 |         |
| <b>Output Voltage</b>                      |                   |                               |     |      |     |         |
| Output Voltage Sense (Note5)               | $V_S$             | $V_S = 5V$                    | 4.9 | 5    | 5.1 | V       |

| Parameter                               | Symbol                          | Test Conditions  | Min   | Typ  | Max   | Unit              |
|---|---------------------------------|--|-------|------|-------|-------------------|
| Reference Voltage                       | V <sub>REF</sub>                | 3V ≤ V <sub>IN</sub> ≤ 36V                                     | 0.788 | 0.8  | 0.812 | V                 |
| <b>Current Limit</b>                    |                                 |  |       |      |       |                   |
| High-Side Switch Current Limit          | I <sub>LIM_H</sub>              | V <sub>BOOT</sub> – V <sub>SW</sub> = 4.8V, minimum duty cycle | 1.36  | 1.6  | 1.84  | A                 |
| Low-Side Switch Sourcing Current Limit  | I <sub>sr_L</sub>               | From source to drain   | 1.08  | 1.35 | 1.62  | A                 |
| Low-Side Switch Sinking Current Limit   | I <sub>sk_L</sub>               | From drain to source   | --    | 0.8  | --    | A                 |
| <b>Switching</b>                        |                                 |  |       |      |       |                   |
| Switching Frequency                     | f <sub>sw</sub>                 |  | 1890  | 2100 | 2310  | kHz               |
| Minimum On-Time                         | t <sub>ON_MIN</sub>             |  | --    | 60   | 80    | ns                |
| <b>Internal MOSFET</b>                  |                                 |  |       |      |       |                   |
| High-Side On-Resistance                 | R <sub>DS(ON)_H</sub>           |  | --    | 200  | 360   | mΩ                |
| Low-Side On-Resistance                  | R <sub>DS(ON)_L</sub>           |  | --    | 160  | 288   |                   |
| <b>Soft-Start</b>                       |                                 |  |       |      |       |                   |
| Soft-Start Time                         | t <sub>ss</sub>                 |  | 1.3   | 2    | 2.7   | ms                |
| <b>Error Amplifier</b>                  |                                 |  |       |      |       |                   |
| Error Amplifier Trans-Conductance       | gm                              | -10μA < I <sub>COMP</sub> < 10μA                               | 665   | 950  | 1235  | μA/V              |
| COMP to Current Sense Trans-Conductance | gm <sub>CS</sub>                |  | 0.85  | 1.2  | 1.45  | A/V               |
| <b>Over-Temperature Protection</b>      |                                 |  |       |      |       |                   |
| Thermal Shutdown                        | T <sub>SD</sub>                 |  | --    | 175  | --    | °C                |
| Thermal Shutdown Hysteresis             | ΔT <sub>SD</sub>                |  | --    | 15   | --    |                   |
| <b>Power-Good</b>                       |                                 |  |       |      |       |                   |
| Power-Good Rising Threshold             | V <sub>FB</sub> rising (Good)   |  | 85    | 90   | 95    | %V <sub>REF</sub> |
|   | V <sub>FB</sub> rising (Fault)  |  | --    | 120  | --    |                   |
| Power-Good Falling Threshold            | V <sub>FB</sub> falling (Fault) |  | 80    | 85   | 90    | %V <sub>REF</sub> |
|   | V <sub>FB</sub> falling (Good)  |  | --    | 117  | --    |                   |

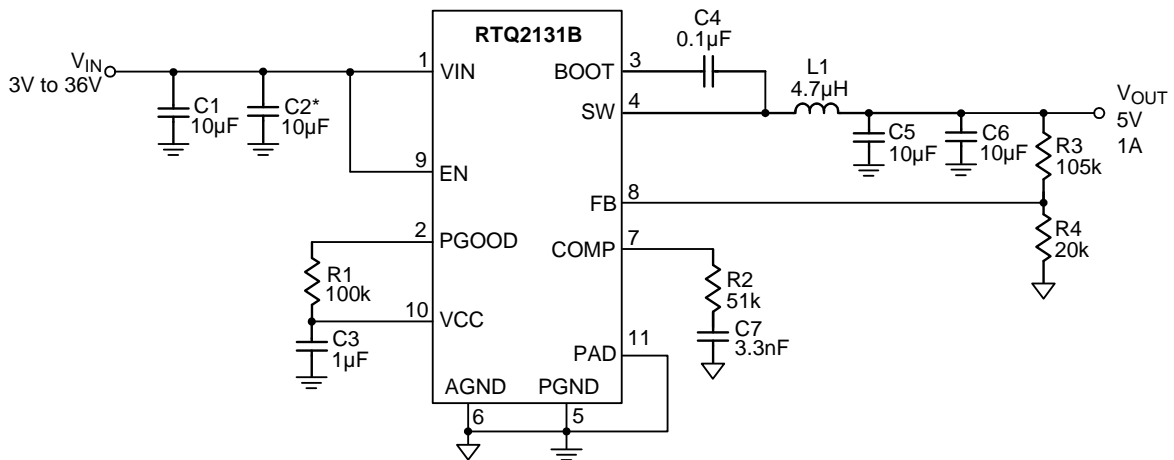


| Parameter                          | Symbol | Test Conditions  | Min | Typ | Max | Unit |
|------------------------------------|--------|--|-----|-----|-----|------|
| Power-Good Leakage Current         |        | PGOOD signal good, $V_{FB} = V_{REF}$ , $V_{PGOOD} = 5.5V$ | --  | 30  | --  | nA   |
| Power-Good Sink Current Capability |        | PGOOD signal fault, $I_{PGOOD}$ sinks 0.2mA                | --  | --- | 0.3 | V    |
| Spread Spectrum                    |        |  |     |     |     |      |
| Spread-Spectrum Rang               | SS     |  | --  | +6  | --  | %    |

- Note 1.** Stresses beyond those listed “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 in the operational sections of the specifications is not implied. Exposure to absolute maximum rating conditions may affect device reliability.
- Note 2.**  $\theta_{JA}$  is measured under natural convection (still air) at  $T_A = 25^\circ C$  with the component mounted on a high effective-thermal-conductivity four-layer test board on a JEDEC 51-7 thermal measurement standard. The first layer is filled with copper.  $\theta_{JC}$  is measured at the exposed pad of the package.
- Note 3.** Devices are ESD sensitive. Handling precaution is recommended.
- Note 4.** The device is not guaranteed to function outside its operating conditions.
- Note 5.** There are two output voltage setting options : one is that trimmed output voltage options for a fixed output voltage are available for the VS pin, and the other is through a resistive divider to sense the output voltage at the FB pin.

## Typical Application Circuit

### Adjustable Output Voltage

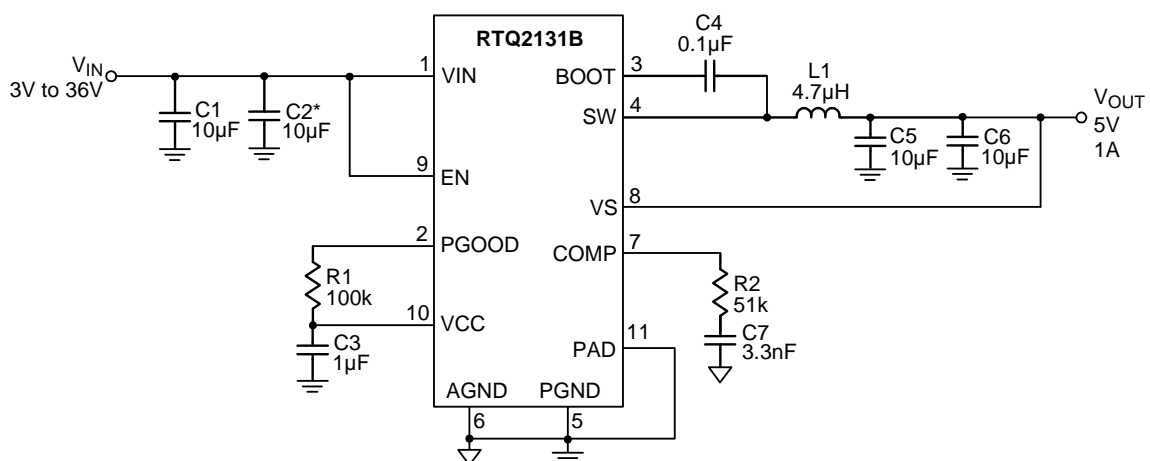


\* : if it is needed for FMEA, C2 can be added.

**Table 1. Suggested Component Values for Adjustable Output Voltage Version**

| V <sub>OUT</sub> (V) | R3 (kΩ) | R4 (kΩ) | L1 (µH) | C <sub>OUT</sub> (µF) | R2 (kΩ) | C7 (nF) |
|----------------------|---------|---------|---------|-----------------------|---------|---------|
| 1                    | 4.99    | 20      | 2.2     | 20                    | 10      | 3.3     |
| 1.2                  | 10      | 20      | 2.2     | 20                    | 12      | 3.3     |
| 1.8                  | 30      | 24      | 2.2     | 20                    | 17.4    | 3.3     |
| 2.8                  | 49.9    | 20      | 2.8     | 20                    | 33      | 3.3     |
| 3.3                  | 75      | 24      | 3.3     | 20                    | 39      | 3.3     |
| 5                    | 105     | 20      | 4.7     | 20                    | 51      | 3.3     |

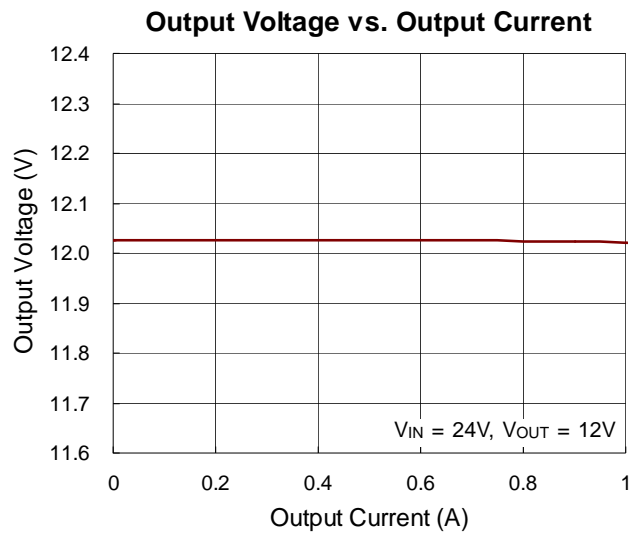
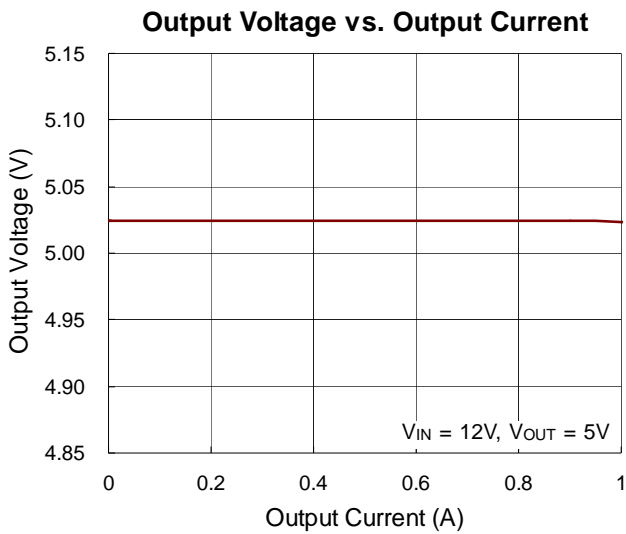
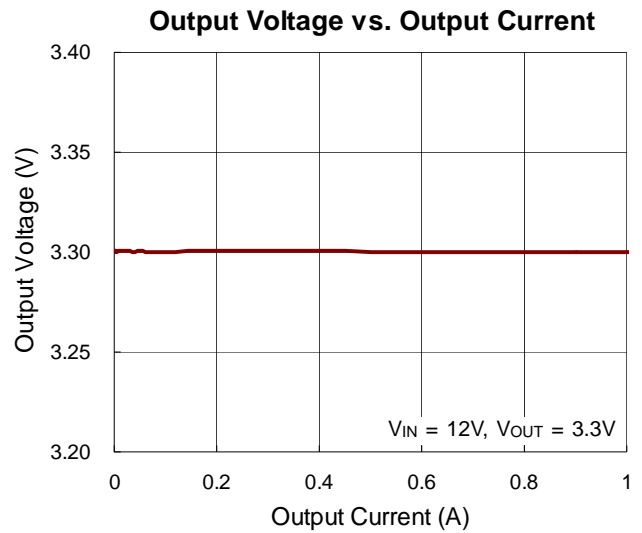
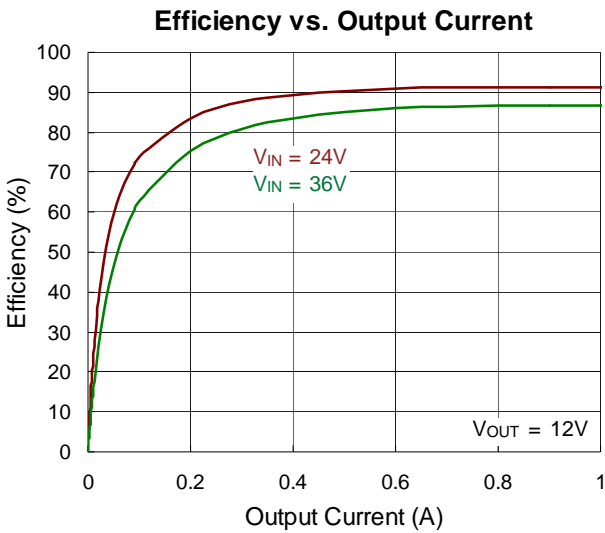
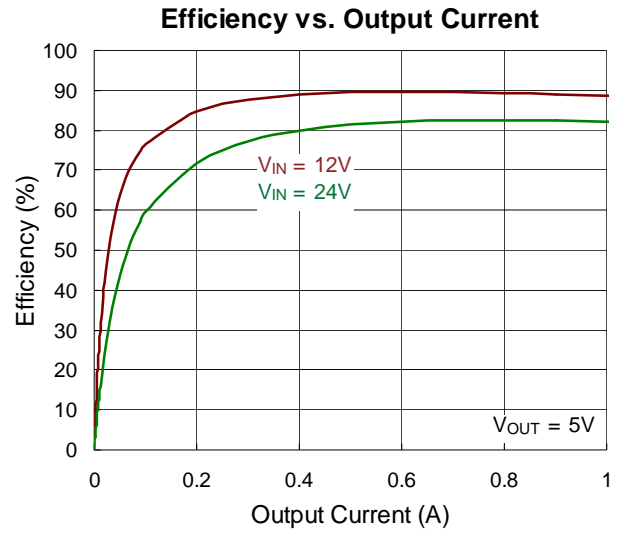
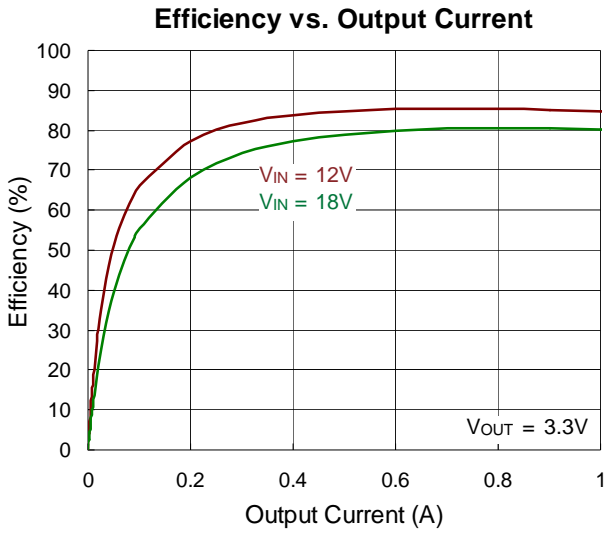
### Fixed 5V Output Voltage

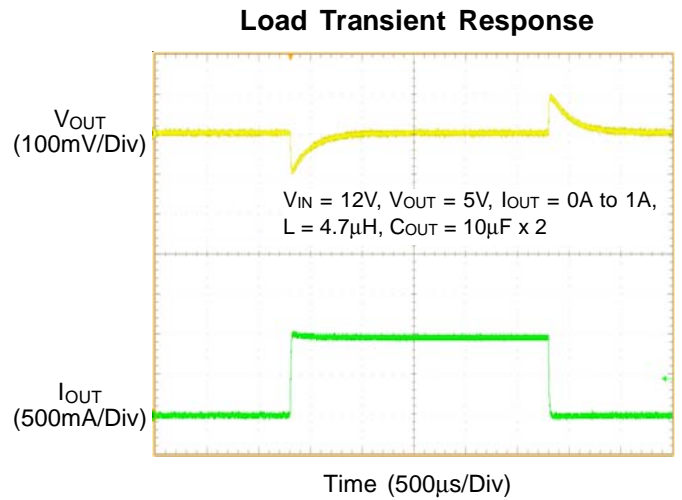
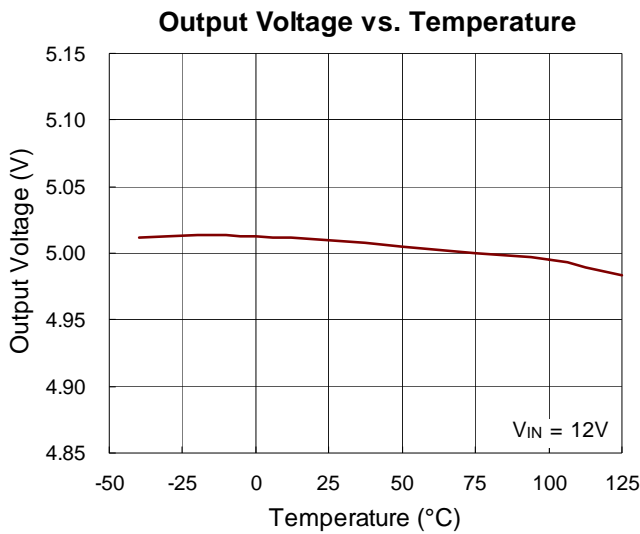
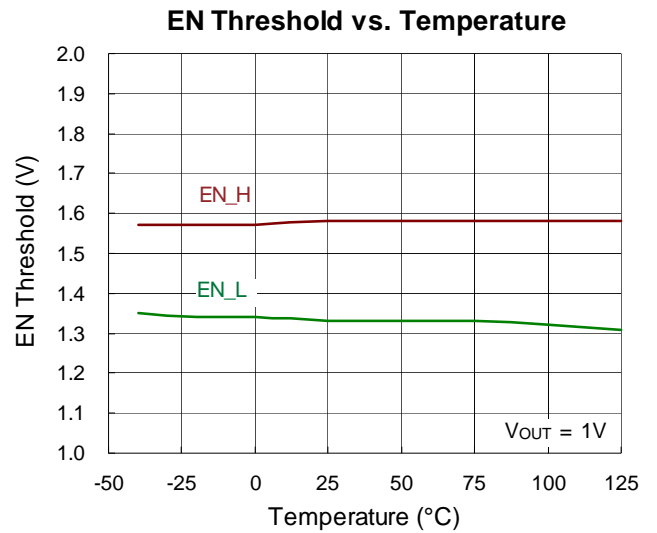
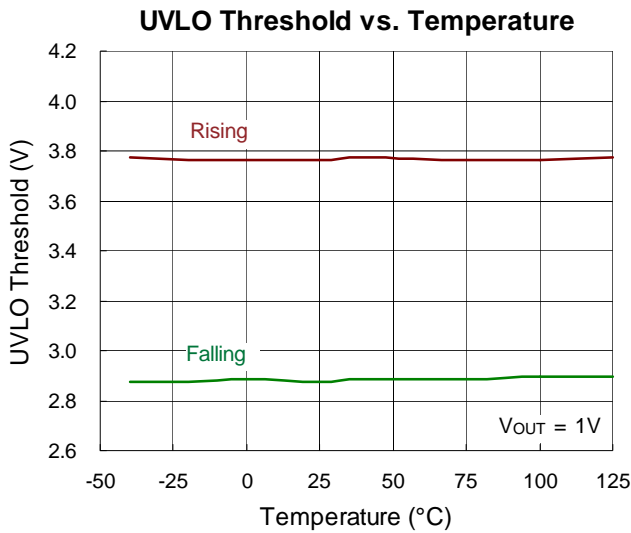
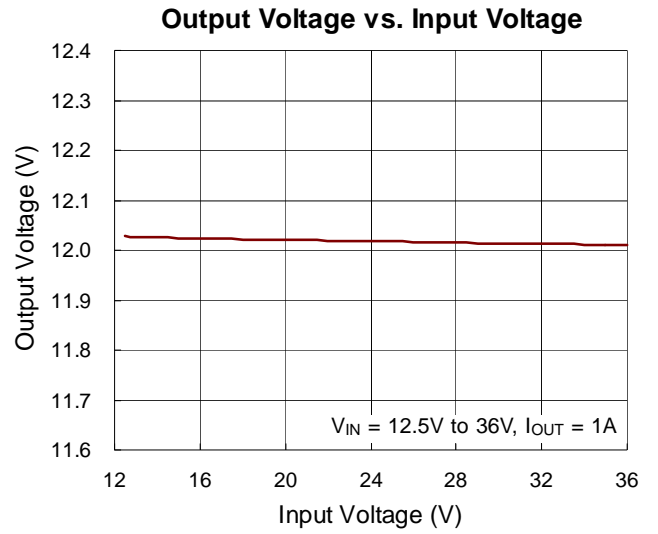
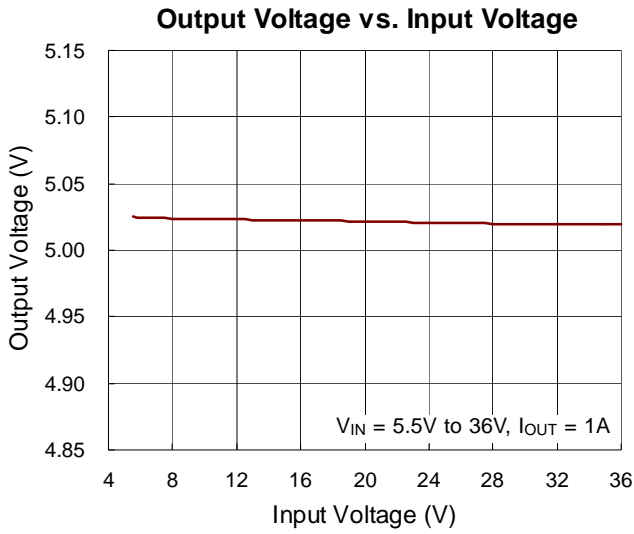


\* : if it is needed for FMEA, C2 can be added.

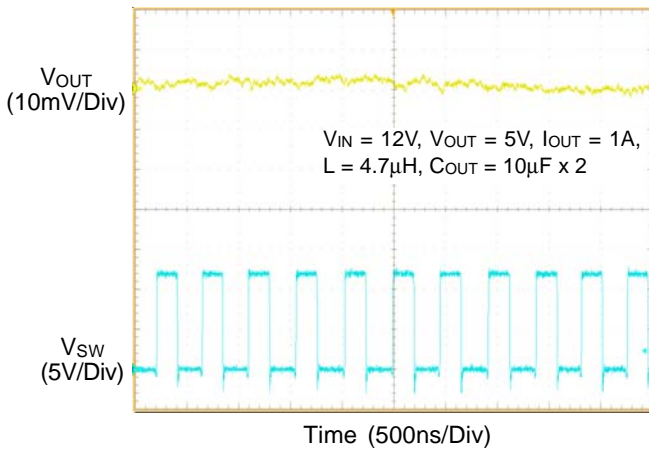
Note : All the input and output capacitors are the suggested values, referring to the effective capacitances, subject to any de-rating effect, like a DC Bias.

**Typical Operating Characteristics**

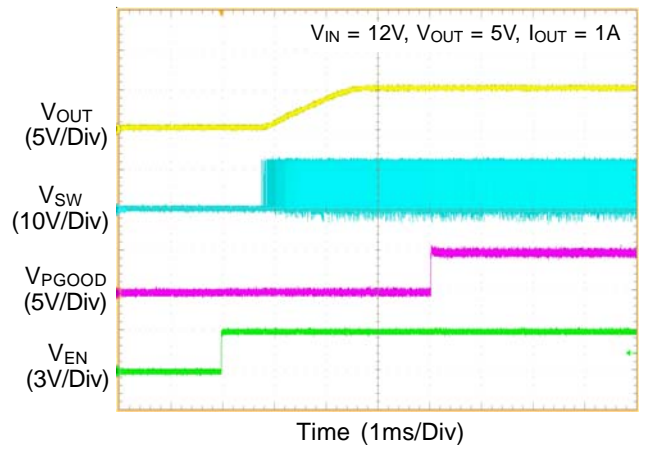




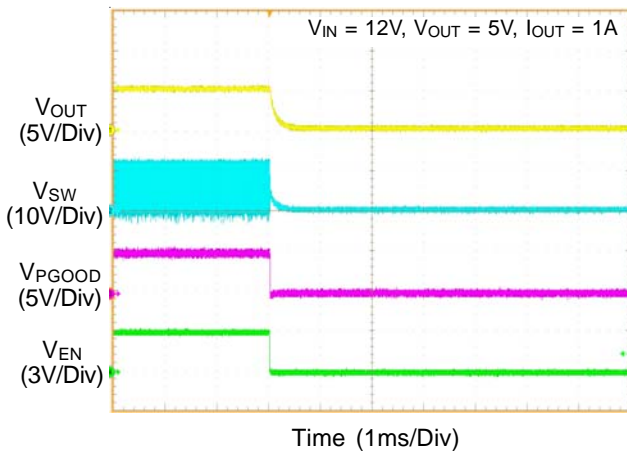
**Output Ripple Voltage**



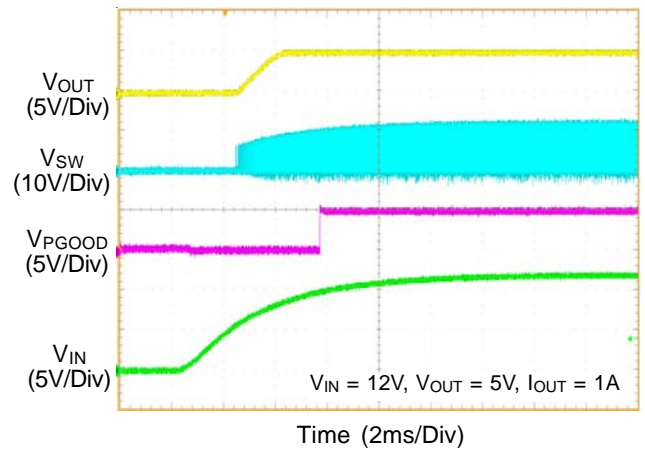
**Power On from EN**



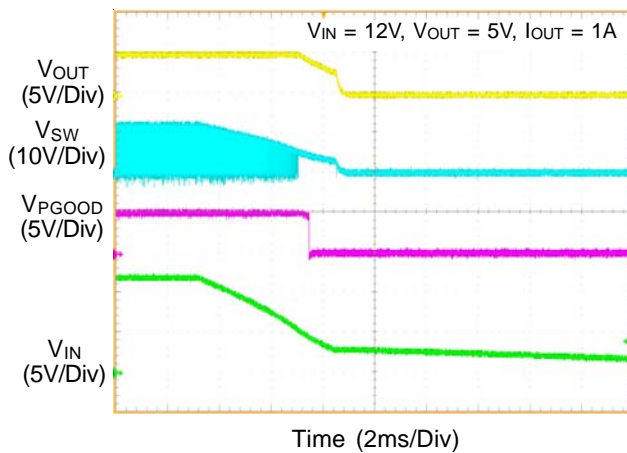
**Power Off from EN**



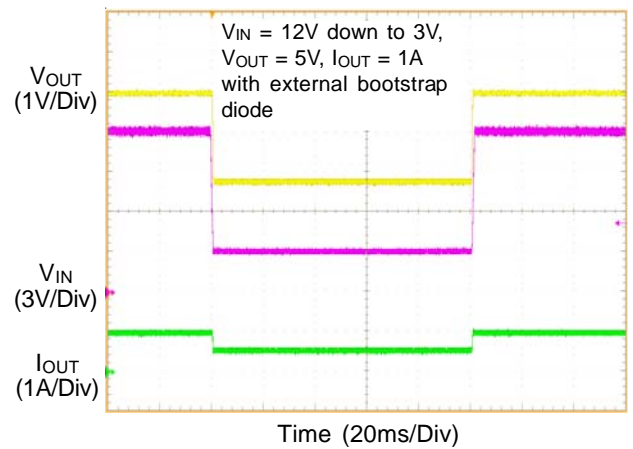
**Power On from VIN**



**Power Off from VIN**



**Momentary Drop**



## Application Information

The output stage of a synchronous buck converter is composed of an inductor and capacitor, which stores and delivers energy to the load, and forms a second-order low-pass filter to smooth out the switch node voltage to maintain a regulated output voltage. The output voltage can be up to 28V at proper operation condition.

### Output Voltage Setting

For the RTQ2131B, there are two output voltage setting options: one is that trimmed output voltage options for a fixed output voltage are available for the VS pin, and the other is through a resistive divider to sense a fraction of the output voltage at the FB pin as shown in Figure 3.

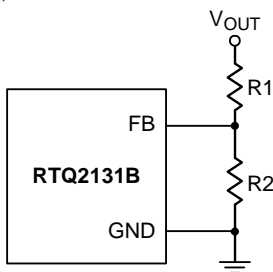


Figure 3. Output Voltage Setting

The output voltage is set by the external resistive voltage divider, and can be calculated by the following equation :

$$V_{OUT} = 0.8V \times \left(1 + \frac{R1}{R2}\right)$$

### Spread-Spectrum Operation

Due to the periodicity of the switching signals, the energy concentrates in one particular frequency and also in its odds harmonics. These levels or energy is radiated and therefore this is where a potential EMI issue arises. The RTQ2131B have spread-spectrum function to overcome EMI issue. The switching frequency varies by +6% relative to the internally generated operating frequency (typically 2.1MHz). By varying the frequency 6% only in the positive direction, the RTQ2131B still guarantees that the 2.1MHz switching frequency does not drop into the AM band limit of 1.8MHz.

### External Bootstrap Diode

Connect a 0.1μF low-ESR ceramic capacitor between the BOOT and SW pins. This capacitor provides the supply

voltage for the high-side MOSFET gate driver. It is recommended to add an external bootstrap diode from an external 5V supply voltage to the BOOT pin to improve efficiency when the input voltage  $V_{IN}$  is lower than 5.5V, or duty cycle is higher than 65%. A low-cost bootstrap diode can be used, such as IN4148 or BAT54. The external 5V supply voltage can be a fixed voltage supply coming from the system, or the 5V output voltage generated in the system. Note that the BOOT voltage must be lower than 5.5V.

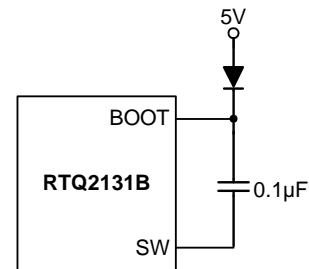


Figure 4. External Bootstrap Diode

### Chip Enable Operation

For automatic start-up, the EN pin can be connected to  $V_{IN}$ , directly. The inherent hysteresis makes EN useful as a simple time delay. To add an additional time delay, the EN pin can be connected to  $V_{IN}$  through a resistor  $R_{EN}$  and to GND through a capacitor  $C_{EN}$ , as shown in Figure 5. The additional time delay for switching operation to start can be calculated with the EN's internal logic threshold, 1.45V (typ.).

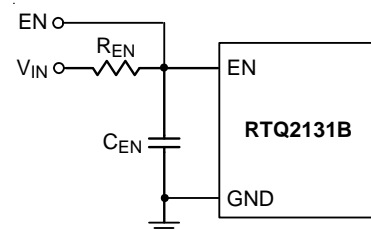


Figure 5. Enable Timing Control

An external MOSFET can be added to implement a logic-controlled EN pin, as shown in Figure 6. In this case, a 100kΩ pull-up resistor, R<sub>EN</sub>, is connected between V<sub>IN</sub> and the EN pin. The MOSFET Q1 can provide the logic control on the EN pin, pulling it down.

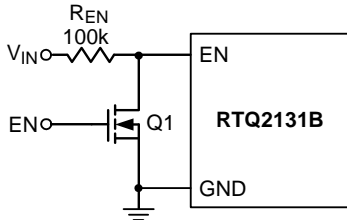


Figure 6. Digital Enable Control Circuit

**Inductor Selection**

For a given input and output voltage, the inductor value and operating frequency can determine the inductor ripple current. The ripple current ΔI<sub>L</sub> increases with higher V<sub>IN</sub> and decreases with higher inductance.

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

A lower inductor ripple current reduces not only ESR losses in output capacitors but also output ripple voltage. Higher efficiency operation can be achieved at higher frequency with smaller ripple current. This, however, requires a larger inductor.

For the ripple current selection, the value of ΔI<sub>L</sub>, which is I<sub>MAX</sub> multiplied by 0.3 will be a reasonable starting point. The largest ripple current occurs at the highest V<sub>IN</sub>. To guarantee that the ripple current stays below the specified maximum, the inductor value should be chosen according to the following equation :

$$L = \left( \frac{V_{OUT}}{f_{SW} \times \Delta I_L(MAX)} \right) \times \left( 1 - \frac{V_{OUT}}{V_{IN}(MAX)} \right)$$

To ensure the converter can operate properly, the maximum inductor peak current must always meet the inductor saturation current rating and temperature rating.

**C<sub>IN</sub> and C<sub>OUT</sub> Selection**

The input capacitance, C<sub>IN</sub>, is needed to filter the trapezoidal current at the source of the high-side MOSFET. To prevent large ripple current, a low-ESR input capacitor sized for the maximum RMS current should be used. The approximate RMS current equation is given as below :

$$I_{RMS} = I_{OUT} \times \frac{V_{OUT}}{V_{IN}} \times \sqrt{\frac{V_{IN}}{V_{OUT}} - 1}$$

This formula has a maximum at V<sub>IN</sub> = 2 x V<sub>OUT</sub>, where I<sub>RMS</sub> = I<sub>OUT</sub> / 2. This simple worst-case condition is commonly used for design because even significant deviations do not offer much relief.

Choose a capacitor rated at a higher temperature than required. Several capacitors may also be paralleled to meet size or height requirements in the design. For the input capacitor, two 10μF low-ESR ceramic capacitors are suggested.

The selection of C<sub>OUT</sub> is determined by the required ESR to minimize the ripple voltage. Moreover, the bulk capacitance is also a key for C<sub>OUT</sub> selection to ensure that the control loop is stable. Loop stability can be checked by viewing the load transient response as described in a later section.

The output ripple voltage, ΔV<sub>OUT</sub>, is determined by :

$$\Delta V_{OUT} \leq \Delta I_L \left( ESR + \frac{1}{8 \times f_{SW} \times C_{OUT}} \right)$$

The highest output ripple voltage will be at the maximum input voltage since ΔI<sub>L</sub> increases with input voltage. Multiple capacitors placed in parallel may be needed to meet the ESR and RMS current handling requirements. Ceramic capacitors of larger capacitance and lower cost are now available in smaller case sizes. The high ripple current, and voltage ratings, and low-ESR characteristics of such capacitors make them ideal for switching regulator applications. However, care must be taken when these capacitors are used at input and output. When a ceramic capacitor is used at the input and the power supply is from a wall adapter through a long wire, the ceramic capacitor and the wire inductance can result in significant input voltage ringing, triggered by a load step at the output. At best, this ringing will be coupled to the output and be mistaken as loop instability. At worst, an inrush current through the long wire may cause a large voltage spike at V<sub>IN</sub> to damage the converter.

## Compensation Network Design

The purpose of loop compensation is to ensure stable operation while maximizing the dynamic performance. An undercompensated system may result in unstable operations. Typical symptoms of an unstable power supply include: audible noise from the magnetic components or ceramic capacitors, jittering in the switching waveforms, oscillation of output voltage, overheating of power FETs and so on.

In most cases, the peak current mode control architecture used in the RTQ2131B only requires two external components to achieve a stable design as shown in Figure 7. The compensation can be selected to accommodate any capacitor type or value. The external compensation also allows the user to set the crossover frequency and optimize the transient performance of the device. Around the crossover frequency the peak current mode control (PCMC) equivalent circuit of Buck converter can be simplified as shown in Figure 8. The method presented here is easy to calculate and ignores the effects of the slope compensation that is internal to the device. Since the slope compensation is ignored, the actual cross over frequency will usually be lower than the crossover frequency used in the calculations. It is always necessary to make a measurement before releasing the design for final production. Though the models of power supplies are theoretically correct, they cannot take full account of circuit parasitic and component nonlinearity, such as the ESR variations of output capacitors, then on linearity of inductors and capacitors, etc. Also, circuit PCB noise and limited measurement accuracy may also cause measurement errors. A Bode plot is ideally measured with a network analyzer while RICHTEK application note [AN038](#) provides an alternative way to check the stability quickly and easily. Generally, follow the following steps to calculate the compensation components :

1. Set up the crossover frequency,  $f_c$ . For stability purposes, our target is to have a loop gain slope that is  $-20\text{dB/decade}$  from a very low frequency to beyond the crossover frequency. Do "NOT" design the crossover frequency over  $90\text{kHz}$  with the RTQ2131B. For dynamic purposes, the higher the bandwidth, the faster the load transient response. The downside to

high bandwidth is that it increases the regulators susceptibility to board noise which ultimately leads to excessive falling edge jitter of the switch node voltage.

2.  $R_{\text{COMP}}$  can be determined by :

$$R_{\text{COMP}} = \frac{2\pi \times f_c \times V_{\text{OUT}} \times C_{\text{OUT}}}{g_m \times V_{\text{REF}} \times g_{m\_CS}} = \frac{2\pi \times f_c \times C_{\text{OUT}}}{g_m \times g_{m\_CS}} \times \frac{R_{\text{UP}} + R_{\text{DOWN}}}{R_{\text{DOWN}}}$$

where

$g_m$  is the error amplifier gain of trans-conductance ( $950\mu\text{A/V}$ )

$g_{m\_CS}$  is COMP to current sense ( $1.2\text{ A/V}$ ).

3. A compensation zero can be placed at or before the dominant pole of buck which is provided by output capacitor and maximum output loading ( $R_L$ ). Calculate

$C_{\text{COMP}}$ :

$$C_{\text{COMP}} = \frac{R_L \times C_{\text{OUT}}}{R_{\text{COMP}}}$$

Output capacitor and its ESR provide a zero and optional  $C_{\text{COMP}2}$  can be used to cancel this zero.

$$C_{\text{COMP}2} = \frac{R_{\text{ESR}} \times C_{\text{OUT}}}{R_{\text{COMP}}}$$

Note : Generally,  $C_{\text{COMP}2}$  is not necessary for the output using ceramic capacitor due to the ESR is extremely low.

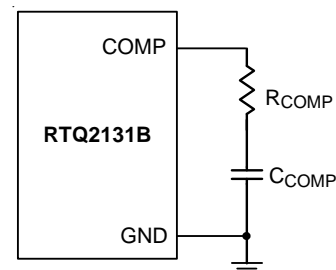


Figure 7. External Compensation Components



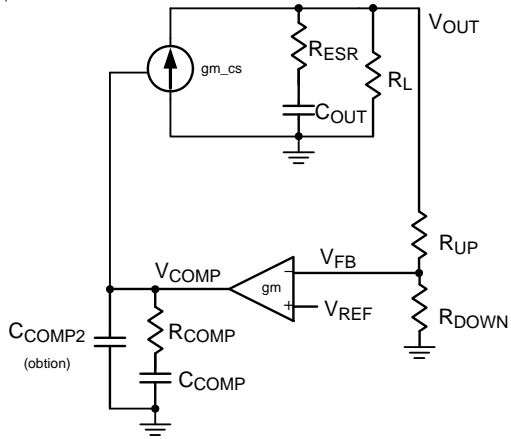


Figure 8. Simplified Equivalent Circuit of Buck with PCMC

**Thermal Considerations**

The junction temperature should never exceed the absolute maximum junction temperature  $T_{J(MAX)}$ , listed under Absolute Maximum Ratings, to avoid permanent damage to the device. The maximum allowable power dissipation depends on the thermal resistance of the IC package, the PCB layout, the rate of surrounding airflow, and the difference between the junction and ambient temperatures. The maximum power dissipation can be calculated using the following formula :

$$P_{D(MAX)} = (T_{J(MAX)} - T_A) / \theta_{JA}$$

where  $T_{J(MAX)}$  is the maximum junction temperature,  $T_A$  is the ambient temperature, and  $\theta_{JA}$  is the junction-to-ambient thermal resistance.

For continuous operation, the maximum operating junction temperature indicated under Recommended Operating Conditions is 150°C. The junction-to-ambient thermal resistance,  $\theta_{JA}$ , is highly package dependent. For a WDFN-10SL 3x3 package, the thermal resistance,  $\theta_{JA}$ , is 38.8°C/W on a standard JEDEC 51-7 high effective-thermal-conductivity four-layer test board. The maximum power dissipation at  $T_A = 25^\circ\text{C}$  can be calculated as below :

$$P_{D(MAX)} = (150^\circ\text{C} - 25^\circ\text{C}) / (38.8^\circ\text{C/W}) = 3.22\text{W for a WDFN-10SL 3x3 package.}$$

The maximum power dissipation depends on the operating ambient temperature for the fixed  $T_{J(MAX)}$  and the thermal

resistance,  $\theta_{JA}$ . The derating curves in Figure 9 allows the designer to see the effect of rising ambient temperature on the maximum power dissipation.

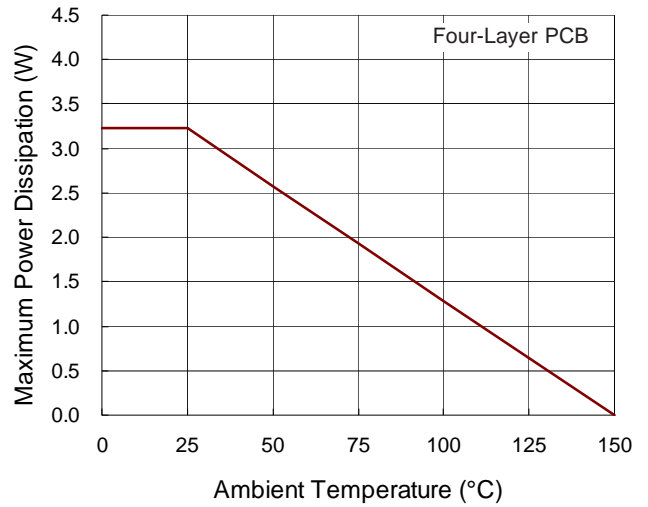
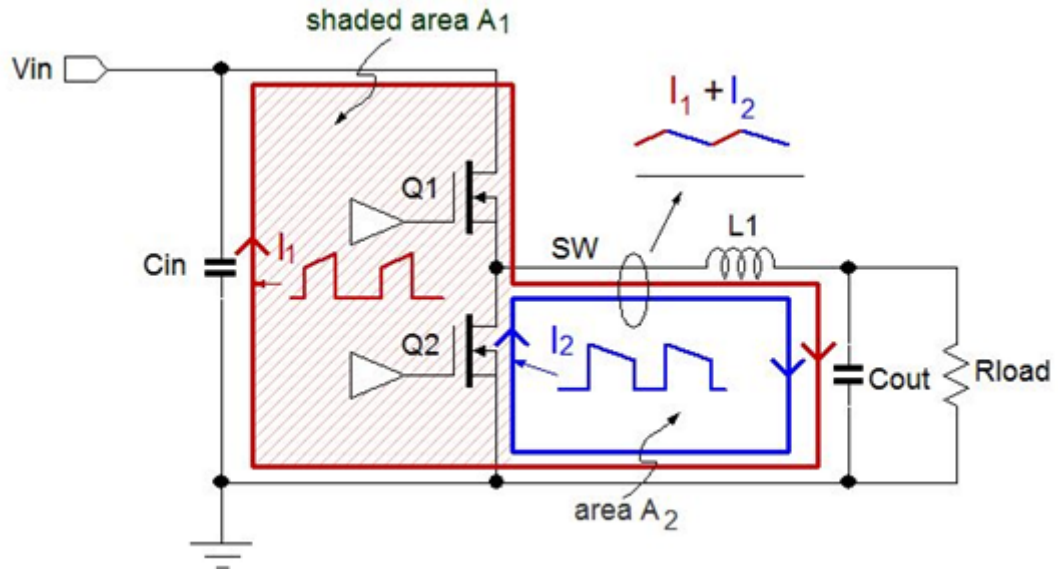


Figure 9. Derating Curve of Maximum Power Dissipation

## Layout Considerations

The below figure shows the two main loops in the buck converter, where the shaded loop with area  $A_1$  is the most critical loop because the current  $I_1$  in this loop consists of discontinuous current pulses with high  $dl/dt$ . Since radiation

is proportional to the loop area, it is most important to make the loop area  $A_1$  as small as possible. Loop area  $A_2$  is less critical, because the current in this loop ( $I_1 + I_2$ ) is a continuous triangle shaped waveform with much lower  $dl/dt$  than  $I_1$ .



Follow the PCB layout guidelines for optimal performance of the device.

- ▶ Keep the traces of the input and output current paths as short and wide as possible. To keep small main loop area,  $C_{IN} \rightarrow IC\_VIN \rightarrow IC\_SW \rightarrow Inductor (L1) \rightarrow C_{OUT} \rightarrow PGND$ .
- ▶ To minimize area  $A_1$  in the layout, the input capacitors should be placed as close as possible to the IC, and make a short connection with IC VIN and PGND.
- ▶ Place high frequency decoupling capacitor  $C_{IN1}$  as close as possible to the IC to reduce the loop impedance and minimize switch node ringing.
- ▶ To further reduce loop impedance, the input current return path should be placed underneath the input capacitors and let the return path overlap the top input power line to get small AC return path.
- ▶ To minimize distance between the layer of main power loop to layer of current return path to get small AC return path. Minimizing the parasitic loop impedance will minimize switch node ringing and EMI.
- ▶ SW node is with high frequency voltage swing and should be connected to inductor by short trace.

- ▶ BOOT node is with high frequency voltage swing and place  $C_{BOOT}$  capacitor with smallest parasitic loop.
- ▶ Keep analog components away from the SW and BOOT node to prevent stray capacitive noise pickup and minimize EMI.
- ▶ Connect feedback network behind the output capacitors.
- ▶ Place the feedback components near the device.
- ▶ Connect all analog grounds to a common node and then connect the common node to the power ground behind the output capacitors.
- ▶ Vias must be placed to inner ground layers to help reduce ground impedance and to serve as heat removal from the IC thermal pad.
- ▶ Place capacitor for VCC and should be placed as close as possible to the IC.
- ▶ The example of the RTQ2131B PCB layout guide are shown as below

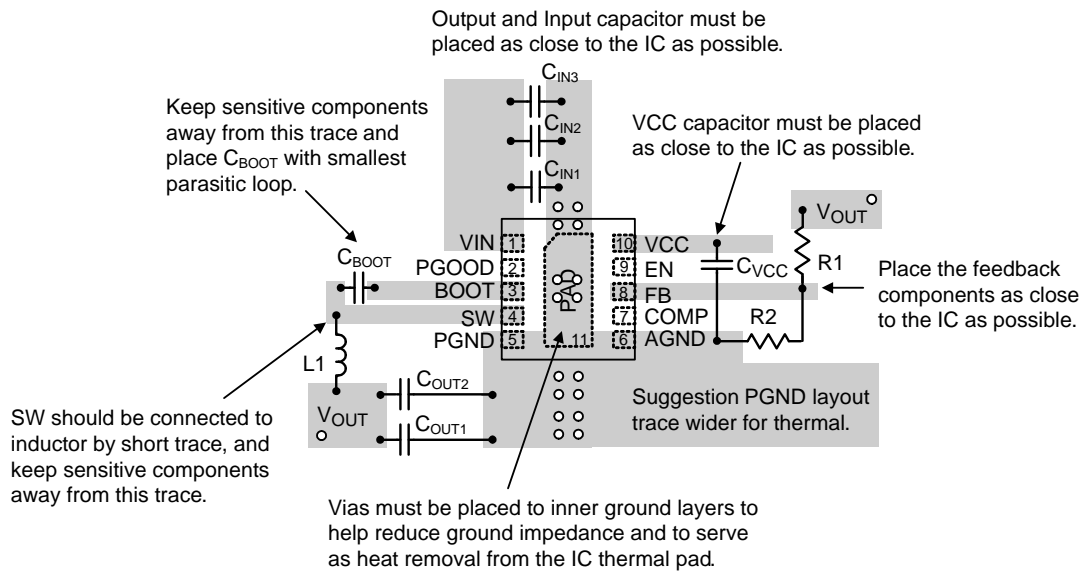
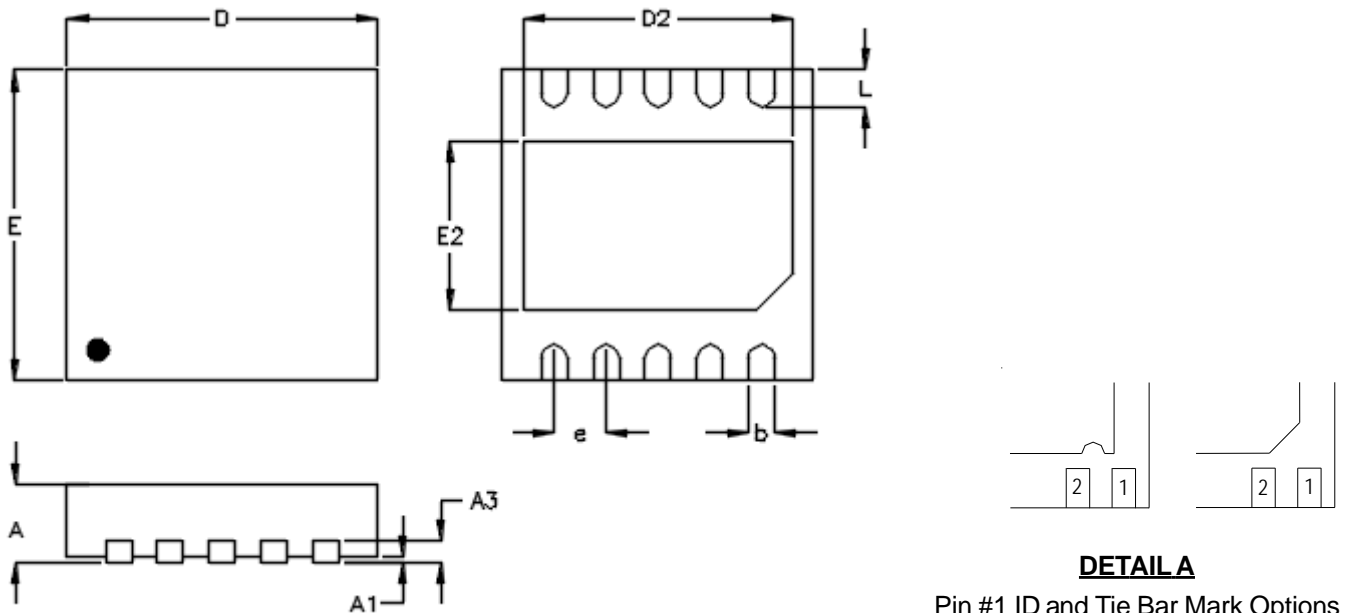


Figure 10. PCB Layout Guide

Outline Dimension



**DETAIL A**

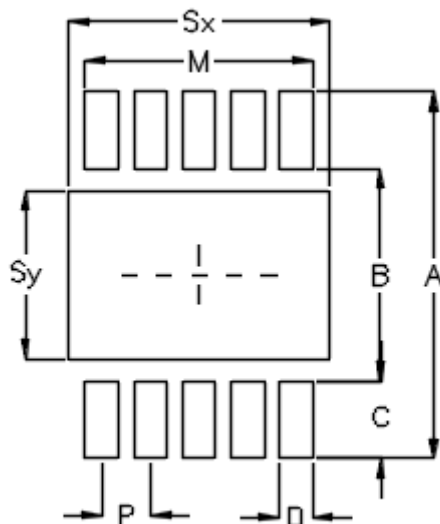
Pin #1 ID and Tie Bar Mark Options

Note : The configuration of the Pin #1 identifier is optional, but must be located within the zone indicated.

| Symbol | Dimensions In Millimeters |       | Dimensions In Inches |       |
|--------|---------------------------|-------|----------------------|-------|
|        | Min                       | Max   | Min                  | Max   |
| A      | 0.700                     | 0.800 | 0.028                | 0.031 |
| A1     | 0.000                     | 0.050 | 0.000                | 0.002 |
| A3     | 0.175                     | 0.250 | 0.007                | 0.010 |
| b      | 0.200                     | 0.300 | 0.008                | 0.012 |
| D      | 2.900                     | 3.100 | 0.114                | 0.122 |
| D2     | 2.550                     | 2.650 | 0.100                | 0.104 |
| E      | 2.900                     | 3.100 | 0.114                | 0.122 |
| E2     | 1.590                     | 1.690 | 0.063                | 0.067 |
| e      | 0.500                     |       | 0.020                |       |
| L      | 0.300                     | 0.400 | 0.012                | 0.016 |

W-Type 10SL DFN 3x3 Package

**Footprint Information**



| Package           | Number of Pin | Footprint Dimension (mm) |      |      |      |      |      |      |      | Tolerance |
|-------------------|---------------|--------------------------|------|------|------|------|------|------|------|-----------|
|                   |               | P                        | A    | B    | C    | D    | Sx   | Sy   | M    |           |
| V/W/U/XDFN3*3-10S | 10            | 0.50                     | 3.80 | 2.20 | 0.80 | 0.35 | 2.70 | 1.74 | 2.35 | ±0.05     |

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