

Output Short-Circuit Protection Solution for the TPS61088

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Low-Power DC/DC Converters

ABSTRACT

The TPS61088 boost converter implements a cycle-by-cycle current limit function to protect the device from overload conditions during boost switching. This current limit function is realized by detecting the current flowing through the low-side MOSFET. The current limit feature loses function in the output short-circuit conditions. This application note describes an output short-circuit-protection solution for the TPS61088 boost converter. When the output is shorted to ground or the load current is higher than a certain value, the TPS61088 is disconnected from the load. Similar solutions can be used on the TPS61089 as well.

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1 Boost Converter Block Diagram With Output Short-Circuit Protection

Figure 1 shows the simplified block diagram of a boost converter with output short-circuit protection.

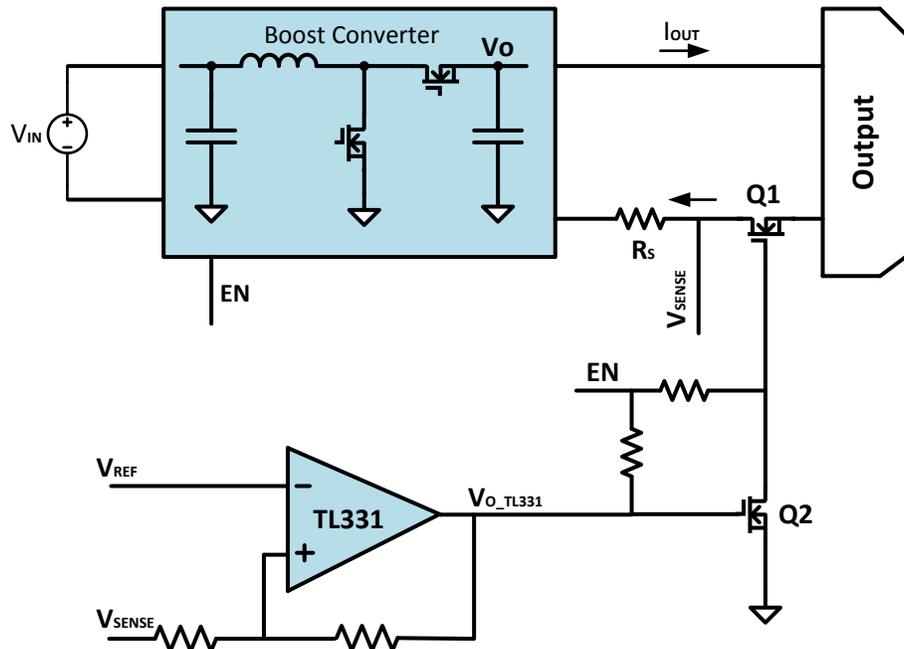


Figure 1. Boost Converter Block Diagram With Output Short-Circuit Protection

A shunt resistor, R_s , is placed in the output return. It converts the output current to a voltage signal V_{SENSE} . In the overload or output short-circuit condition, V_{SENSE} is higher than the reference voltage, V_{REF} . The output signal V_{O_TL331} of the comparator TL331 goes high, so Q2 turns on and Q1 turns off. Thus the boost converter is disconnected from the load. The hysteresis comparator circuit helps Q1 avoid being turned on and off frequently before fault clearing.

When the fault condition is cleared, the circuits can recover by toggling (disable, then enable) the EN pin of the boost IC. This pulls the EN signal down to GND, the boost converter turns off, V_{SENSE} is 0 mV, the output of the comparator TL331 is at logic low, and Q2 turns off. Next, pull the EN signal high, and the boost converter and Q1 turn on again. Thus, the circuit recovers.

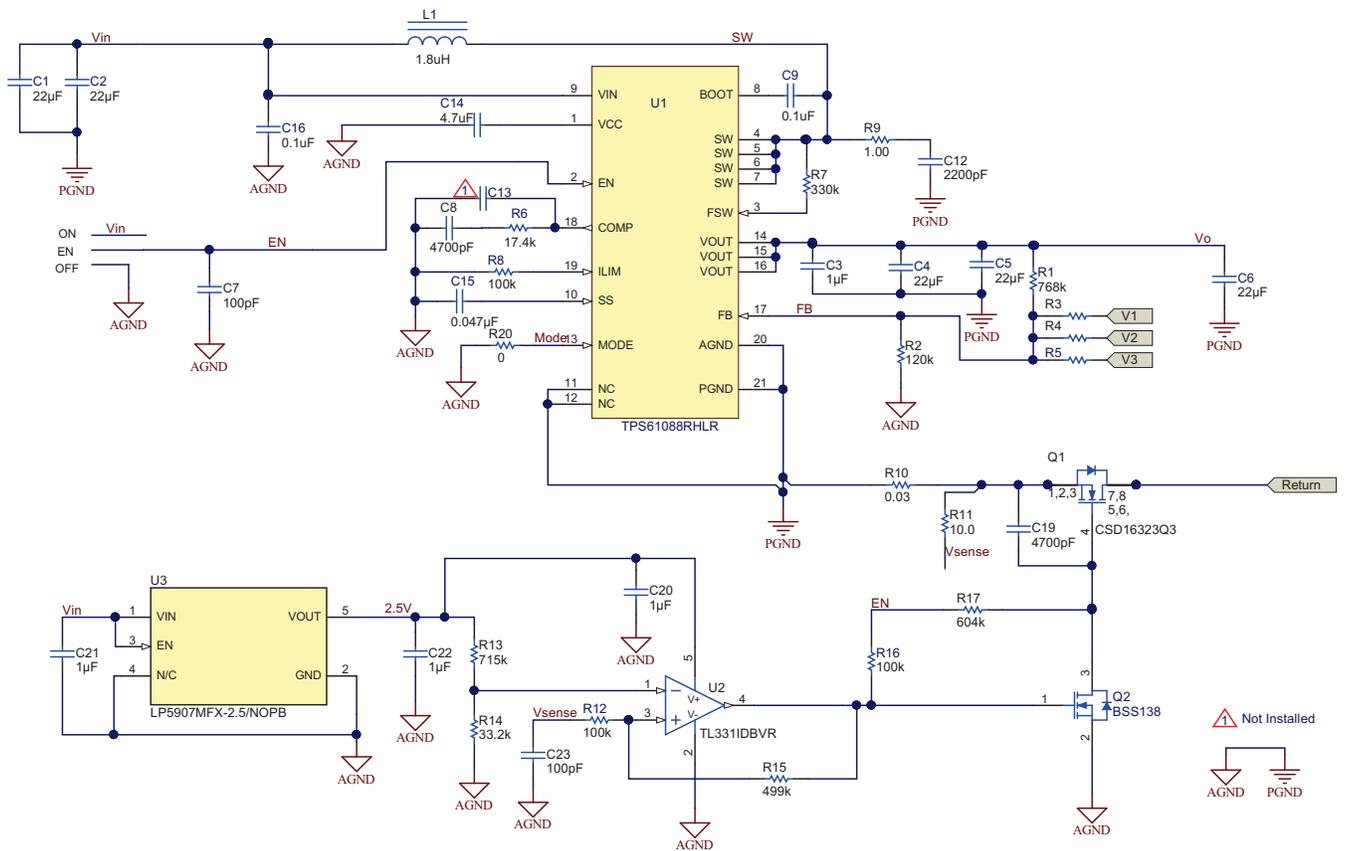
2 Design Example with the TPS61088

Table 1 lists the performance specification of this design example.

Table 1. Performance Specification

Input Voltage	Output Voltage/Output Current	Overcurrent Protection Point
3 V–4.2 V	5 V/3 A, 9 V/2 A, 12 V/1.5 A	$I_o \geq 4.4$ A

Figure 2 shows the schematic of this design example.



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Figure 2. Design Example Schematic

2.1 Setting Overcurrent Protection Point

The overcurrent protection point should be higher than the maximum output current. In this design example, the maximum output current is 3 A (occurs at $V_o = 5$ V condition). We set the overcurrent protection point at 4.4 A to avoid the protection circuit from false triggering at normal output current conditions.

2.2 Shunt Resistor Selection

A shunt resistor is the most versatile and cost-effective means of measuring the current. The voltage across it should be kept to a low value to reduce the power loss. In this design example, the shunt resistor value, R_S , is chosen as 30 mΩ. The maximum continuous current, I_{O_MAX} , flowing through the shunt resistor before OCP is 4.4 A. So the maximum shunt voltage V_{SENSE_MAX} can be calculated with Equation 1:

$$V_{SENSE_MAX} = R_S \times I_{O_MAX} \tag{1}$$

The minimum power rating of the shunt resistor can be calculated with Equation 2:

$$P_{\text{rating}} = V_{\text{SENSE_MAX}} \times I_{\text{O_MAX}} \quad (2)$$

So the minimum power rating of the shunt resistor is calculated as 0.58 W in this reference design. A general guideline is multiplying this minimum power rating by 2. That is, choose a ≥ 1 -W resistor in this design example to make it more robust in the overload or output short-circuit condition.

2.3 Comparator Circuit Design

Figure 3 illustrates the hysteresis comparator circuit in this reference design. The reference voltage, V_{REF} , is put at the inverting input of the TL331, $V_{\text{REF}} = 110$ mV. The current sense signal, V_{SENSE} , is connected to the non-inverting input of the TL331 through R12.

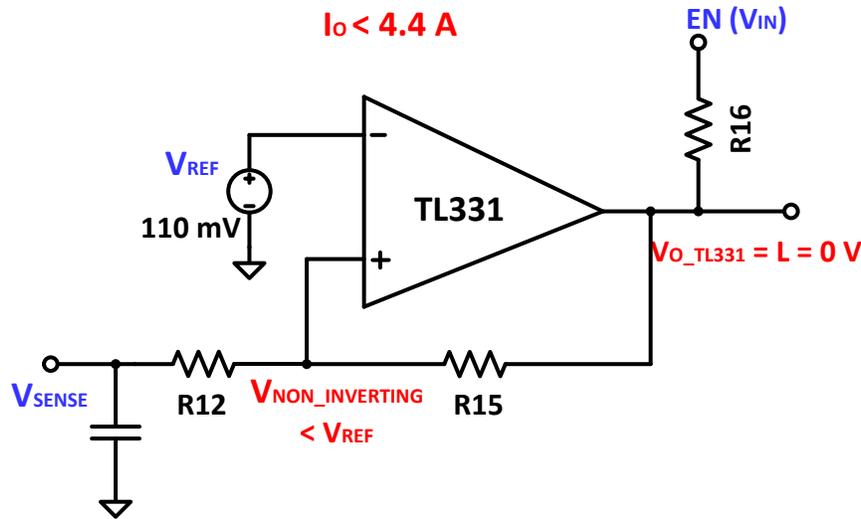


Figure 3. Comparator Circuit With Hysteresis

During the normal operation, the output current is lower than the overcurrent protection point, the output of the comparator is at logic low (0 V). So the voltage, $V_{\text{NON_INVERTING}}$, at the non-inverting input of the TL331 is:

$$V_{\text{NON_INVERTING}} = V_{\text{SENSE}} \times \frac{R_{15}}{R_{15} + R_{12}} \quad (3)$$

When the output current increases to the targeted transition point, we have:

$$V_{\text{SENSE}} = V_{\text{SENSE_MAX}} = 132 \text{ mV} \quad (4)$$

$$V_{\text{NON_INVERTING}} = V_{\text{REF}} = 110 \text{ mV} \quad (5)$$

Inserting Equation 4 and Equation 5 into Equation 3, calculates to:

$$R_{15} = 5 \times R_{12} \quad (6)$$

Select $R_{12} = 100 \text{ k}\Omega$, $R_{15} = 499 \text{ k}\Omega$ in this design example.

After the overcurrent protection, the output of the comparator is at logic high and Q1 turns off. The TPS61088 disconnects with the load. In order to avoid the Q1 from being turned on and off frequently before fault clearing, the voltage $V_{\text{NON_INVERTING}}$ at the positive input of the TL331 should always be higher than the reference voltage, V_{REF} , after OCP occurs. We can realize this by choosing an appropriate resistor, R16, under the minimum input voltage condition:

$$V_{\text{IN_MIN}} \times \frac{R_{12}}{R_{12} + R_{15} + R_{16}} \geq V_{\text{REF}} \tag{7}$$

$$R_{16} \leq \frac{V_{\text{IN_MIN}}}{V_{\text{REF}}} - R_{12} - R_{15}$$

where

- $V_{\text{IN_MIN}} = 3 \text{ V}$ (8)

In this design example, $R_{16} = R_{12} = 100 \text{ k}\Omega$, is chosen.

2.4 Drive Circuit Design of Q1

The turn on speed of the MOSFET, Q1, is controlled by R17 and C19. It is turned on after the TPS61088 was enabled ($V_{\text{EN_HIGH}} = V_{\text{IN}}$). Choose an approximate RC time constant to avoid the big inrush current flowing through the drain and source of Q1 during its turn on.

The startup waveforms in Figure 4 show the Q1 gate drive voltage ($V_{\text{GS_Q1}}$) and the drain source current ($I_{\text{DS_Q1}}$) when the drive resistor $R_{17} = 10 \text{ k}\Omega$ ($C_{19} = 4700 \text{ pF}$). Observe that a big inrush current flows through Q1 and the sense resistor R10, forcing V_{SENSE} higher than V_{REF} . The output of the comparator goes high, Q2 turns on, $V_{\text{GS_Q1}}$ drops to 0 V, and Q1 turns off. Thus, the TPS61088 is disconnected from the load because of this high inrush current. So we need to slow down the turn on speed of the MOSFET Q1 to limit the inrush current and avoid the latch up problem during start up. That is to say, we should limit the inrush current ($I_{\text{DS_Q1}}$) below the overcurrent protection point during the maximum load current start up condition.

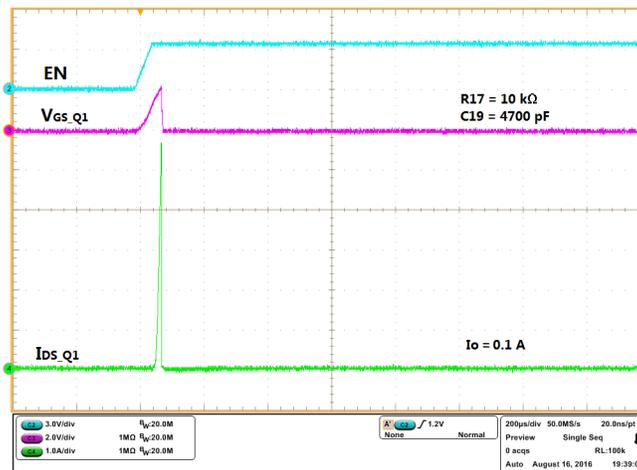


Figure 4. Drain Source Current $I_{\text{DS_Q1}}$ of Q1 With $R_{17} = 10 \text{ k}\Omega$

The startup waveform in Figure 5 shows the Q1 gate drive voltage ($V_{\text{GS_Q1}}$) and the drain source current ($I_{\text{DS_Q1}}$) when the drive resistor $R_{17} = 604 \text{ k}\Omega$ ($C_{19} = 4700 \text{ pF}$) at $I_{\text{O}} = 0.1 \text{ A}$ and $I_{\text{O}} = 2 \text{ A}$, independently. Observe that the maximum inrush current (3.1 A) is well below the OCP point (4.4 A) during the startup. So, if the maximum output current during the startup is 2 A, $R_{17} = 604 \text{ k}\Omega$, $C_{19} = 4700 \text{ pF}$ is fit for this design example.

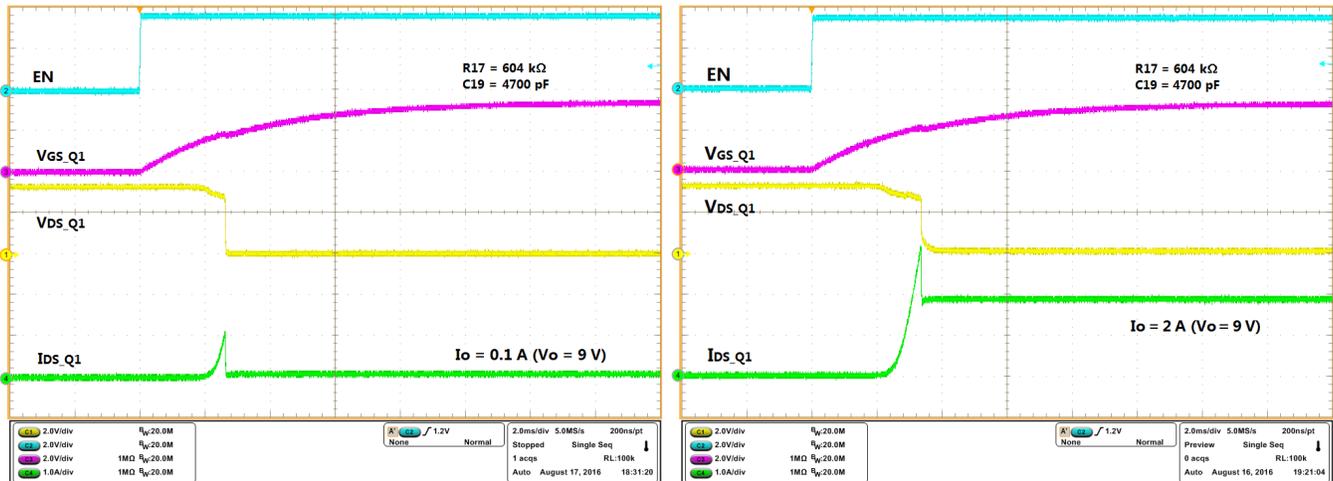


Figure 5. Drain Source Current I_{DS_Q1} of Q1 With $R_{17} = 604 \text{ k}\Omega$

2.5 Q1 Selection

In this application, the switching aspects of a MOSFET matter little because the MOSFET is on almost 100% of the time. So, the on resistance, $R_{DS(on)}$, is the key factor. We choose CSD16323Q3C in this design example. When the gate drive voltage is 3 V, the $R_{DS(on)}$ is only 5.4 m Ω . So the total conversion efficiency is nearly unchanged with this extra load disconnect switch.

The turn on speed of Q1 is very slow in this design example. It will work in the linear mode region for about 1 ms (the linear mode operation starts when the V_{GS} voltage reaches the threshold voltage $V_{GS(th)}$ and ends with the drain-source voltage reaches zero) before being fully turned on. Maximum safe operating area (SOA) must be checked to verify the MOSFET can withstand the thermal stress. In this design example, the minimum battery voltage is 3 V, the linear mode operation time is smaller than 1 ms. If Q1 can stay within the SOA (pink shaded area) shown in Figure 3 during turn on, then the MOSFET is fit for the application.

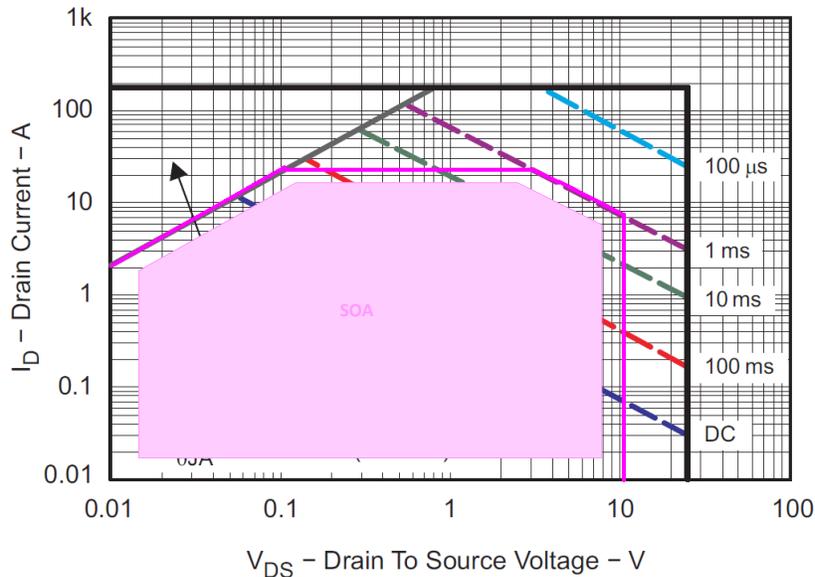


Figure 6. Maximum Safe Operating Area of MOSFET Q1

2.6 Other Points

In this design example, the reference voltage V_{REF} is generated from the LDO LP2907. Most applications already have reference voltages in the system, so this LDO can be removed.

3 Conclusion

This application note introduces an output short-circuit solution for the boost converter. This solution requires an additional single comparator and a power MOSFET. With this small amount of circuitry, the TPS61088 boost converter can be disconnected from the load within 5 μ s in the output short-circuit condition. Similar solutions can be used to TPS61089 as well.

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