11 Ways to Protect Your Power Path

Design Tips and Tradeoffs Using TI's Power Switches

Texas Instruments

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Introduction: Basics of Power Switches

Application Report

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Abstract

A *Power Switch* provides an electrical connection from a voltage source or ground to a load. It saves power across multiple voltage rails and protects subsystems from damage. It also provides enhanced component protection, inrush current protection, and minimizes printed-circuit board (PCB) size.

There are several power switch topologies with different functions that address different applications. *Load Switches* establish the power switch foundation by providing safe and reliable distribution of power. Applications typically using load switches include power distribution, power sequencing, inrush current control, and reduced current leakage. Integrated *Power MUX* devices are similar to load switches but allow for multiple input sources. This set of electronic switches is used to select and transition between two or more input power paths to a single output while also providing input power protection.

eFuses and *Hot Swap* controllers provide additional input power path protection functions such as current sense monitoring, current limiting, undervoltage and overvoltage protection, and thermal shutdown. This makes these devices ideal for hot-plug and transient events that would otherwise damage system components. These benefits help reduce system maintenance costs and maximize equipment uptime.

Ideal diode, ORing controllers provide protection against reverse-polarity conditions by monitoring an external FET, significantly reducing power loss, and blocking reverse current. Whenever a transient event occurs, the controller monitors and adjusts the external FET to prevent damage to upstream components.

Smart high-side switches are for off-board load protection. They provide additional diagnostic telemetry that monitors the output load current and detects short-circuit and open-load events. *Smart high-side switches* have adjustable current limits, allowing more reliable integration into applications with either large inrush current startup profiles or low peak currents. Adding a smart high-side switch to a design leads to a smarter and more robust solution for driving capacitive, inductive, and LED loads.

Low-side switches connect the load to ground instead of providing a connection between a power supply and the load. By including an integrated flyback diode, *low-side switches* help eliminate inductive load transients by dissipating current in a circular loop. This allows them to drive inductive loads such as solenoids, relays, and motors.

This application report highlights the different topologies within the power switch portfolio, and provides suggestions in choosing the correct solution for a faster design time.

Table 1. Power Switch Topology Table

	POWER DISTRIBUTION		INPUT POWER PROTECTION			OUTPUT POWER PROTECTION	
	Load Switch	Power MUX (2 input, 1 output)	eFuse (Internal FET)	Hot Swap (External FET)	Ideal Diode ORing Controller	Smart High-Side Switch	Low-Side Switch
Voltage Range	0 V to 18 V	2.8 V to 22 V	2.7 V to 60 V	±80 V	±75 V	6 V to 40 V	0 V to 100 V
Max Operating Current	15 A	4.5 A	15 A	N/A	N/A	12 A	1 A
Functions							
Inrush Current Control	✓	✓	✓	✓		✓	
Adjustable Current Limit		✓	✓	✓		✓	
Reverse Current Blocking	✓	✓	✓	✓	✓		
Current Sense Monitoring			✓	✓		✓	
Short-Circuit Protection	✓(1)	✓	✓	✓		✓	
Overvoltage Protection		✓	✓	✓			
Reverse Polarity Protection		✓	✓	✓	✓	✓	
Power Good Signal	✓		✓	✓			
Inductive Load Compatibility						✓	✓
Load-Dump Compatibility			✓	✓	✓	✓	✓
Thermal Shutdown	✓	✓	✓	✓		✓	

(1) Self protected load switch

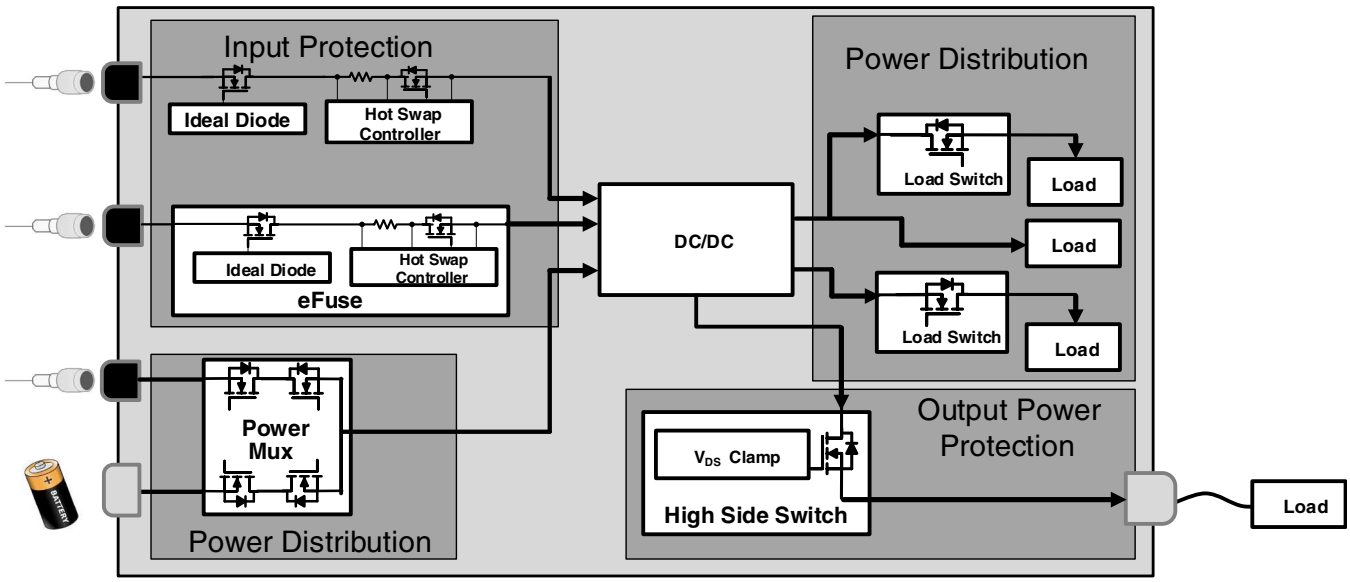

Figure 1. Typical Power Switch Use Cases

1 Load Switches

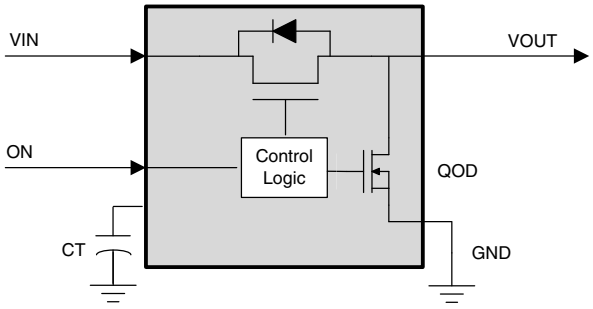

Figure 2. Load Switch Block Diagram

Integrated load switches are electronic switches that turn power rails on and off. When the internal FET turns on, current flows from the input to output and passes power to the downstream circuitry. When the device is enabled, the rise time of the output voltage (V_{OUT}) can be controlled by adjusting the capacitance on an external pin (CT pin). When the device is disabled, the fall time of V_{OUT} is controlled through the quick output discharge (QOD). QOD pulls the output to ground whenever the device is turned off, preventing the output from *floating* or entering an undetermined state.

Some common functions of load switches include power savings, power sequencing, and inrush current control. Power savings is important in applications looking to minimize current dissipation and maximize power efficiency. By disconnecting the supply from a load or subsystem, the switch minimizes power drawn from inactive loads. Power sequencing is important in applications where individual voltage rails need to be turned on and off in a specific order. By configuring the CT and QOD pins, the ramp-up and power-down timing can be adjusted. Inrush current control protects systems that contain large bulk capacitors near the load. When power is initially applied to the system, charging these capacitors can result in a large inrush current that exceeds the nominal load current. If left unaddressed, this can cause voltage rails to fall out of regulation due to the drop, resulting in the system entering an undesired state. Load switches can mitigate the inrush current by using the CT pin to manage the rise time of the power rail. This leads to a linear output slew rate with no voltage dips or external regulators required.

Table 2. Load Switch Examples(1)

DESCRIPTION	DEVICES	VOLTAGE RANGE	MAX CURRENT	TYPICAL R_{on}	PACKAGE
Adjustable rise time, adjustable QOD	TPS22918	1 V to 5.5 V	2 A	52 mΩ	SOT
	TPS22810	2.7 V to 18 V	2 A	79 mΩ	SOT
Space-constrained applications	TPS22915	1.05 V to 5.5 V	2 A	37 mΩ	CSP
	TPS22916	1 V to 5.5 V	2 A	60 mΩ	CSP
Self protected with controlled rise time	TPS22919	1.6 V to 5.5 V	1.5 A	90 mΩ	SOT
Lowest ON-resistance, Power Good indication	TPS22990	1 V to 5.5 V	10 A	3.9 mΩ	SON
Fast turn-on time (≤ 65 µs), Power Good indication, QOD, Thermal shutdown	TPS22971	0.65 V to 3.6 V	3 A	6.7 mΩ	DSBGA

(1) For more information about load switches, visit *TI.com/LoadSwitches*.

2 Power Multiplexing

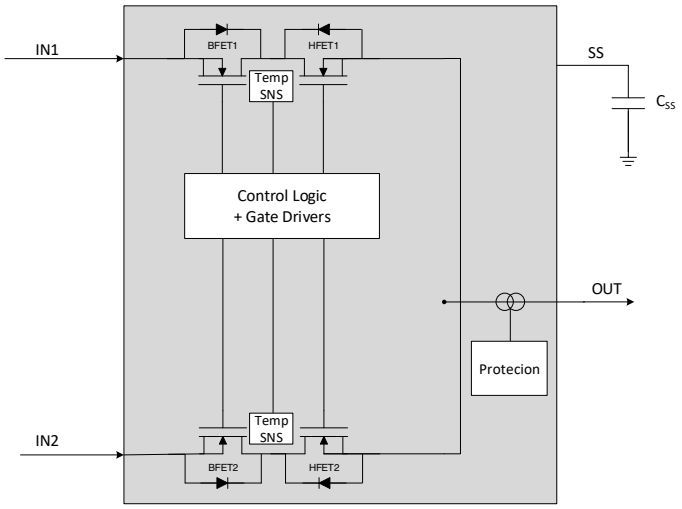

Figure 3. Power MUX Block Diagram

Integrated *Power MUX* devices allow a system to transition between different power sources seamlessly. If the main power supply fails, power multiplexing allows the system to switch to a backup power supply, such as a battery, to preserve operating conditions. Power multiplexing can also provide switching between two different voltage levels for subsystems that operate at two different voltages. In this scenario, to prevent reverse current flow from V_{OUT} into one of the V_{IN} channels, reverse current protection (RCP) blocks current from flowing back through the body diode. Power multiplexing also contains adjustable current limits. If the current exceeds the threshold set by the switch, the switch clamps the channel and prevent current from exceeding the limit. Furthermore, if the current limit forces the device to reach higher temperatures, thermal shutdown will turn off the switch until it can operate at safe conditions again. Similar to load switches, power MUX switches also contain inrush current control to prevent large transient current events.

Power MUX devices can switch between different power rails in three general ways: manually, automatically, or both. Manual switchovers occur with an external GPIO. Whenever the user wants to switch between power rails, the enable pin is toggled and the output is powered by the other power rail. Automatic switchover occurs whenever the primary power supply fails or is disconnected. When the device detects the voltage drop, it automatically switches to the backup power rail. There are some Power MUX solutions which offer the flexibility to be used in an automatic configuration and be controlled by a manual control signal. This method can have a default (automatic) priority, but then be overridden by an external microcontroller if needed.

Table 3. Integrated Power MUX Examples(1)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	MAX CURRENT	TYPICAL R_{on}	PACKAGE
Automatic priority and manual override, adjustable current limit	TPS2120	2.8 V to 22 V	3 A, each channel	62 mΩ	CSP
Automatic priority and manual override, fast output switchover, adjustable current limit	TPS2121	2.8 V to 22 V	4.5 A, each channel	56 mΩ	QFN

(1) For more information about power multiplexing, visit *TI.com/PowerMux*.

3 eFuses

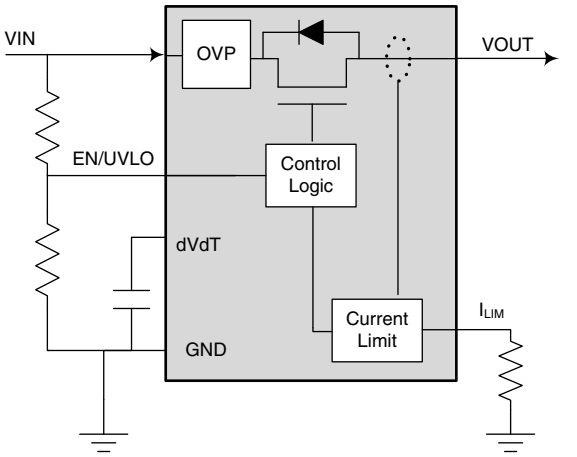

Figure 4. eFuse Block Diagram

eFuses are integrated power protection switches that provide voltage and current protection during fault events. These include short-circuit, overcurrent, overvoltage, undervoltage, and temperature events that might otherwise damage downstream loads. During a short-circuit transient event, the current through the *eFuse* increases very rapidly. The *eFuse* enables a fast-trip current threshold that terminates this rapid increase in less than 200 ns, protecting the supply. If an overvoltage event occurs on the input (VIN), the *eFuse* monitors the voltage across the internal FET and clamps the output voltage until the input falls below the overvoltage threshold. *eFuses* also come with built-in overtemperature protection that shuts down the FET if the junction temperature exceeds 150°C (typical). The *eFuse* either remains off (latch-off version) or attempts to restart (auto retry version) the device after the junction temperature decreases. *eFuses* offer many additional features similar to load switches including adjustable inrush current control and reverse current protection.

Managing current flow from an active power bus to a subsidiary system can be a challenging task. As a device is inserted or removed from a live supply, it is possible to see a very large spike in current during the initial capacitor charging. An *eFuse* or Section 4 hot-swap controller ensures the safe insertion and operation of these systems. Unlike hot swap controllers, *eFuses* contain an integrated FET which minimizes total solution size. This allows *eFuses* to be used in applications such as power multiplexing. By using two *eFuses*, each *eFuse* can control a power rail while providing reverse current protection for its respective supply. *eFuses* are also UL 2367 certified, cutting down on system testing time.

Table 4. eFuse Examples(1)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	MAX CURRENT	TYPICAL R_{on}	PACKAGE
Overvoltage or undervoltage protection, QOD using FLT pin, adjustable current limit	TPS2595	2.7 V to 18 V	4 A	34 mΩ	SON
Lowest Ron circuit-breaker device, accurate load monitoring, adjustable transient fault management	TPS25982	2.5 V to 24 V	15 A	3 mΩ	QFN
Back to back FETs, status monitoring, thermal shutdown, internal reverse current blocking	TPS25942A	2.7 V to 18 V	5 A	42 mΩ	QFN
Reverse polarity protection, current sense output, adjustable current limit	TPS2660	4.2 V to 55 V	2 A	150 mΩ	SOP and QFN
Power limiting, overvoltage cut-off or voltage clamping functionality	TPS1663	4.5 V to 60 V	6 A	31 mΩ	SOP and QFN
Power limiting, reverse current blocking, reverse polarity protection	TPS2663	4.5 V to 60 V	6 A	31 mΩ	SOP and QFN

(1) For more information about eFuses, visit *www.ti.com/efuses*.

4 Hot Swap

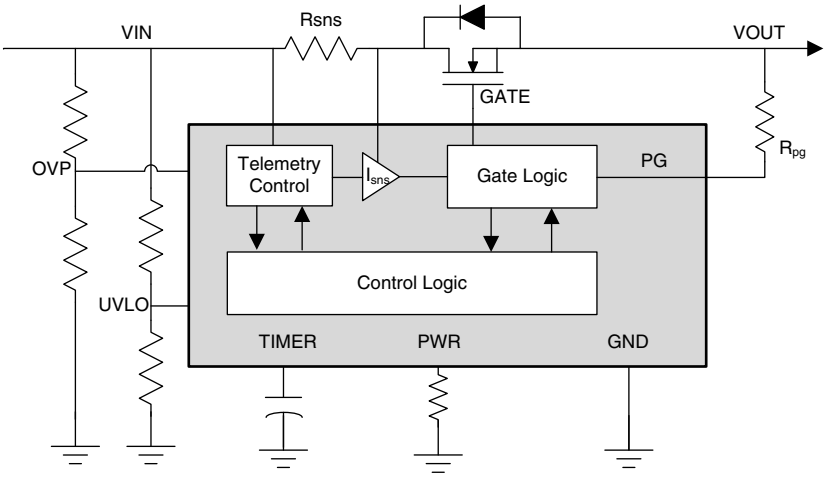

Figure 5. Hot Swap Block Diagram

Hot Swap controllers drive an external MOSFET that protects the system against hot swap events. *Hot Swap* controllers do not integrate a MOSFET as eFuses do. The external MOSFET allows *hot swap* controllers to operate at higher voltages and currents than eFuse devices. The controller monitors the gate voltage of the external FET and adjusts the voltage depending on the situation. When the device is inserted into a live power system, the controller measures the inrush current across R_{SNS}. If the value exceeds the programable current limit, the gate voltage is lowered and limits the current passing downstream. If the power dissipated across the FET exceeds the programmable power limit, then the gate voltage is reduced to lower the current flowing through R_{SNS}. The overvoltage and undervoltage pins also cut off the output voltage whenever the input voltage is not within specified thresholds.

To ensure that the external MOSFET remains within safe operating area (SOA), the *hot swap* controller regulates the current limit at higher V_{DS} voltages. The device also includes an assortment of telemetry that monitors the operating conditions. The Power Good (PG) signal turns on whenever the power rail reaches regulation, and some *hot swap* controllers contain PMBus monitoring that allows real-time feedback on the device status.

Since *hot swap* controllers operate by controlling an external R_{SNS} and MOSFET, they do not contain an innate current limit. The external components allow the user to customize the solution size and power requirements to fit their application.

Table 5. Hot Swap Examples(1)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	PACKAGE
18-V analog devices, small footprint, easy-to-use	TPS247xx	2.5 V to 18 V	SOP or QFN
Meets 240-V requirements for high-end applications, similar to TPS247xx	TPS2477x	2.5 V to 18 V	QFN
PMBus and I2C communication, balance between efficiency and accuracy	LM25066A, LM25066I	2.9 V to 17 V	QFN
Higher voltage applications, PMBus and I2C communication, external FET temperature and failure sensing	LM5066	10 V to 80 V	PWP
Higher voltage applications, SOA protection, current limit	LM5069	9 V to 80 V	SOP
Negative voltage support, SOA protection, PMBus and I2C communication	LM5064	–10 V to –80 V	SOP
Negative voltage support with dual current limit, soft-start disconnect, ORing support, –200 V maximum rating	TPS2352x	–10 V to –80 V	SOP
Circuit breaker function for severe current events, programmable fault timer, PG output	LM5067	–9 V to –80 V	SOP

(1) For more information about Hot Swap controllers, visit *www.ti.com/hotswap*.

5 Ideal Diode, ORing Controllers

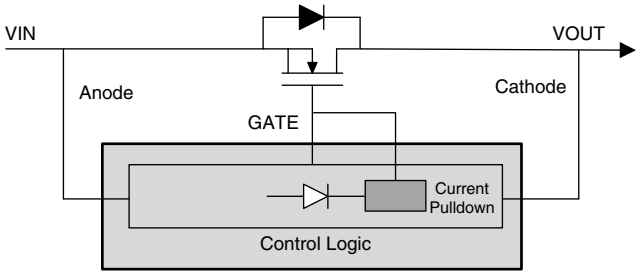

Figure 6. Ideal Diode Block Diagram

Ideal diode controllers control an external FET and, similar to a regular diode, can block reverse current whenever a reverse voltage event occurs. Whenever one of these events occur, the controller shuts off the FET and uses the body diode to prevent any transients from damaging upstream components. The controller can also prevent against ground shorts at the input (VIN) by using the same method.

Ideal diode controllers can also protect against reverse polarity conditions, commonly caused by connecting a battery incorrectly or mis-wiring a power supply. If the user accidentally switches the polarity on VIN, an additional diode from the controller to GND can be included to prevent damage to the IC or the power source. The controller also significantly lowers power dissipation normally found across diodes. By driving the external FET instead of a diode, the voltage drop typically found across diode solutions can be minimized.

Ideal diodes can also act as ORing controllers. Basic power redundancy architecture contains two or more power supplies connected to a single load. ORing solutions allow the system to switch between power sources if one were to fail, and can even connect power sources in parallel. This allows for uninterrupted power and saves on redundant power supply costs.

Table 6. Ideal Diode Controller Examples (1)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	TYPICAL QUIESCENT CURRENT	FORWARD VOLTAGE THRESHOLD	PACKAGE
Fast turnoff, voltage sensing	TPS241x	3 V to 16.5 V	N/A	10 mV	SOP and SOIC
Automotive qualified, low Iq, high efficiency	LM74700-Q1	3.2 V to 65 V	80 µA	20 mV	SOT
Low side ORing controller, FET diagnostics	LM5051	–6 V to –100 V	69 µA	45 mV	SOIC
Low Iq, reverse current protection, integrated FET	LM66100	1.5 V to 5.5 V	0.2 µA	79 mV (2)	SC-70

(1) For more information about *Ideal diode* controllers, visit *www.ti.com/idealdiode*.
(2) Typical forward voltage at Iout = 1 A.

6 Smart High-Side Switches

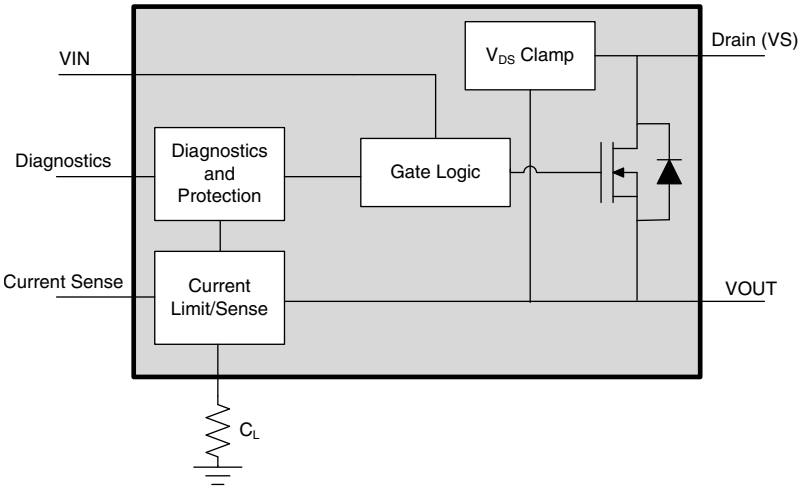

Figure 7. Smart High-side Switch Block Diagram

Smart high-side switches reliably drive off-board loads. These switches contain highly-adjustable and selectable current limits that enable a system to be optimally designed for specific loads. By connecting an external resistor to set the current-limit threshold, the switch protects the load and power supply from overstressing during short-circuits to GND events or power-up conditions. This enables more reliable designs by minimizing transient currents and supply droops. When the threshold is reached, a closed loop activates and clamps the output current to the set value. A fault is then reported on the CS pin.

These switches also offer highly-accurate current sensing to provide real-time diagnostics to the system. A current mirror sources current from VIN, reflecting this as voltage on the Current Sense (CS) pin. The CS pin does not need to be calibrated, and can serve as a diagnostics report pin. Whenever an open load or short happens, the voltage on the CS pin falls to 0 V. Whenever a current limit, thermal event, or an open load or short in the off state occurs, the voltage is pulled up to its maximum threshold. High-accuracy current monitoring and adjustable current limit are ideal for industrial applications like programmable logic controllers, motor valves, servo drives and control units.

Another functionality of *smart high-side* switches is load-dump compatibility, which allows these devices to connect directly to a 12-V battery without concerns about typical voltage and current transients. Additional protection includes mitigation of large inrush current events that would otherwise damage downstream components.

Smart high-side switches can be AEC-Q100 certified, allowing full integration into many automotive applications that require a low on-resistance and high voltage tolerances to accommodate voltage spikes and inrush current events. Some of these applications include front and rear lighting, seat heating, infotainment, cluster, powertrain, and ADAS.

Table 7. Smart High-Side Examples (1) (2)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	CURRENT SENSE ACCURACY	CONTINUOUS LOAD CURRENT	TYPICAL R_{on}	PACKAGE
Selectable current limit for design flexibility, low R_{ON}, small footprint, thermal sensing	TPS1HA08-Q1	3 V to 40 V	±5% at 1 A	0 A to 12 A	8 mΩ	SOP
Low standby current, highly accurate current sense, thermal shutdown	TPS1H100-Q1	3.5 V to 40 V	±3% at 1 A	0 A to 4 A	100 mΩ	SOP
	TPS27S100	3.5 V to 40 V	±3% at 1 A	0 A to 4 A	80 mΩ	SOP
Multi-channel support, fast hardware interrupts, low standby current, loss of GND diagnostics	TPSxH160-Q1	3.4 V to 40 V	±3% at 1 A	0 A to 1.8 A per channel	160 mΩ	SOP

(1) Additional devices in various typical R_{ON} values are available.
(2) For more information about smart high-side switches, see *ti.com/smarthighsideswitch*.

Table 7. Smart High-Side Examples [(1)] [(2)] (continued)

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	CURRENT SENSE ACCURACY	CONTINUOUS LOAD CURRENT	TYPICAL R_{on}	PACKAGE
Dual-channel, programmable current limit	TPS2HB08-Q1	3 V to 28 V	+3/-8% at 1 A	0 A to 8 A per channel	8 mΩ	SOP
Inductive load negative clamp with optimized slew rate, global fault report	TPSxH000-Q1	3.4 V to 40 V	-	0 A to 1A	1 Ω	SOP
Highly accurate current limit, supports full diagnostics with the digital status output	TPS1H200-Q1	3.4 V to 40 V	-	0 A to 2.5 A	200 mΩ	SOP

7 Low-Side Switches

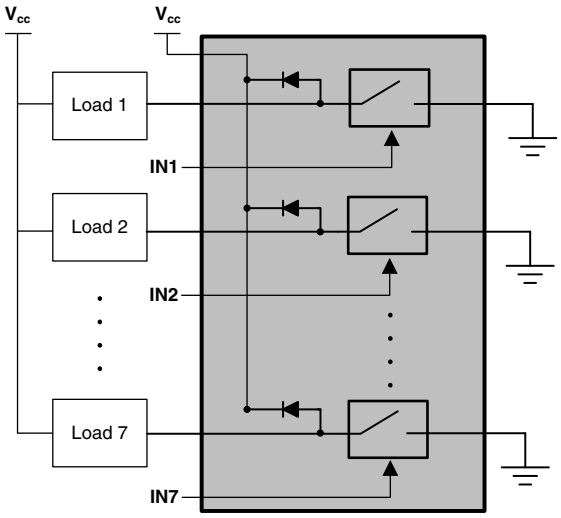

Figure 8. Low-Side Switch Block Diagram

Low-side switches are used to connect and disconnect ground from a load, unlike the rest of the power switch topologies. This configuration allows *low-side* switches to drive inductive loads; an internal flyback diode prevents inductive transients from damaging the circuit and components. Whenever the switch is opened, the inductive transients flow through the flyback diode and dissipate throughout the load. This makes these devices ideal for motors, solenoids, and relays.

Low-side switches consists of two designs: Darlington pair arrays and low-side MOSFET solutions. Darlington pair solutions can support higher voltage applications due to the higher voltage ratings of the integrated BJTs, while the MOSFET solutions have lower on-resistances and lower leakage currents. Most of the *low-side* switches contain seven channels, which can be tied in parallel to support higher current operation.

Table 8. Low-side Switch Examples

DESCRIPTION	DEVICE	RECOMMENDED VOLTAGE RANGE	MAX CURRENT	NUMBER OF CHANNELS	PACKAGE
Darlington pair BJTs, higher voltage support	ULN2003A	0 V to 50 V	500 mA per channel	7	SOIC, SOP, and DIP Packages
Darlington pair BJTs, 8-channel support	ULN2803A	0 V to 50 V	500 mA per channel	8	SOIC
Low-side MOSFET solutions, low on-resistance and current leakage, power efficient	TPL7407LA	0 V to 30 V	600 mA per channel	7	SOIC or SOP

8 References

1. Texas Instruments, *Basics of Load Switches Application Report*
2. Texas Instruments, *What is an eFuse? Application Report*
3. Texas Instruments, *Robust Hot Swap Design Application Report*
4. Texas Instruments, *Adjustable Current Limit of Smart High Side Switch Application Report*

Revision History

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

Changes from Original (November 2018) to A Revision	**Page**
• Changed content in abstract.	1
• Changed headings, row 3, and row 7 in table 1.	2
• Added table notes to table 1.	2
• Changed row 3 in table 2.	3
• Changed content in section 2.	4
• Changed row 1 and 2 in table 3.	4
• Changed rows 2, 5, and 6 in table 4.	5
• Changed title and row 1, 2, and 4 in table 6.	7
• Changed content in section 6.	8
• Changed rows 4, 5, and 6 in table 7.	8

Chapter 1: Inrush Current Control

Author: Alex Triano

Abstract

When a subsystem turns on, current initially flows to any capacitors or inductance at the load. This current is often referred to as inrush current. If a switch turns on quickly (such as without slew-rate control), then the current may be large enough to interrupt other subsystems connected to the same input rail. In some cases, inrush current during a short period of time can irreversibly damage surrounding components or even the switch itself. It is possible to manage inrush current in several ways, although all techniques result in a reduction of the speed at which the output voltage rises.

The Dangers of Inrush Current

Inrush current may cause an upstream supply voltage to dip. If this voltage is supplying power to other subsystems, those components may experience misbehavior or a reset.

In **Figure 1**, the switch is enabled, which causes a fast transient voltage (dv/dt) and a large current spike on the input supply. This leads to a V_{SUPPLY} voltage dip, which is visible while V_{SW_OUT} continues to turn on. The voltage dip on V_{SUPPLY} causes DC/DC converters to shut down; V_{OUT} decays until V_{SUPPLY} recovers within the normal operating range. This scenario demonstrates how inrush current on one load can cause other loads to experience a reset.

Some systems have restrictions on how much current they can draw from the input. These restrictions may be set by safety standards, cable/connector ratings or the need to stay below the current rating of a fast-blow fuse. A failure to manage inrush current can exceed restrictions and lead to compliance failures or damaged cables, connectors or fuses.

A power switch that uses a metal-oxide semiconductor field-effect transistor (MOSFET) to turn power on and off can exceed the FET's safe operating area (SOA) and damage the switch itself. Most power FET manufacturers include a graph listing how much current the switch can handle at a given drain-to-source voltage (V_{DS}), ambient temperature and specified pulse duration. If a FET turns on quickly, into a large load capacitance, the total energy passing through the FET is one-half capacitance-voltage-squared (CV^2) during the turn on-time (t_{ON}) duration. This value can exceed the FET's SOA curve and damage the FET. If that same amount of energy is spread over a longer t_{ON}, then the FET has more time to dissipate that heat and could be designed to survive.

Common Techniques to Control Inrush Current

There are three traditional approaches and one newer approach to controlling inrush current. The first is to use a passive element such as a series resistor or a negative temperature coefficient (NTC) thermistor to slowly charge the output. The second is to use a switch and manage

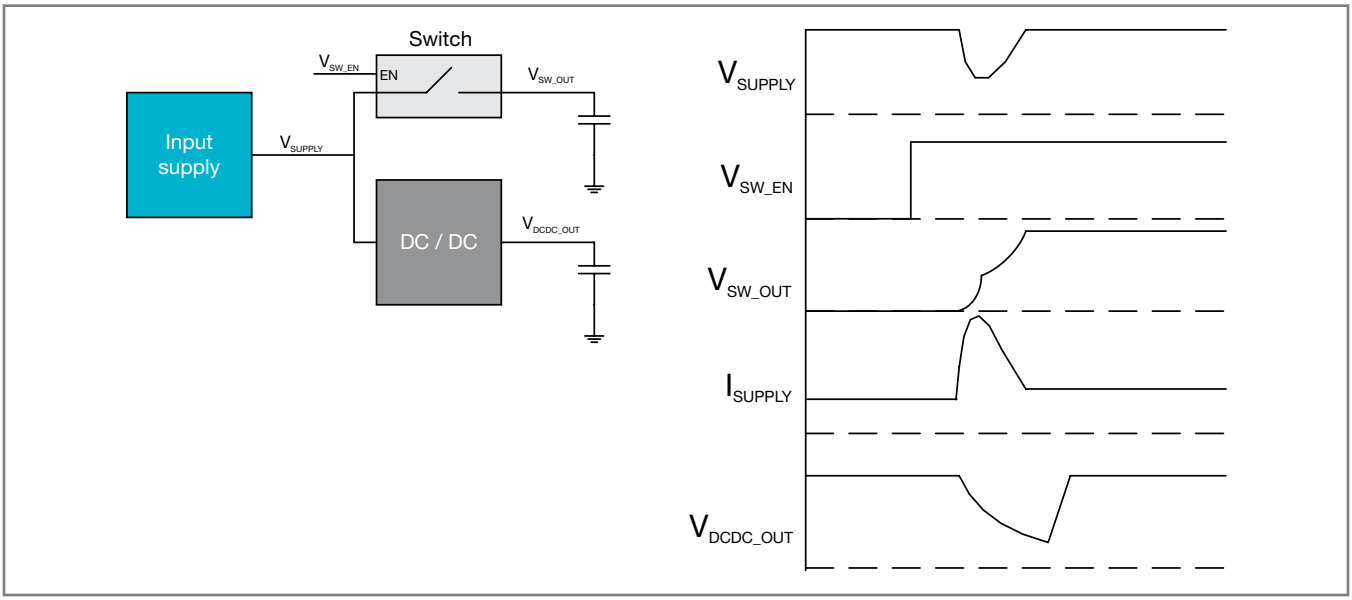

Figure 1: Inrush current on V_{SW_OUT}, causing a supply dip and reset on V_{DCDC_OUT}.

the slew rate at which the output voltage is rising. The third is to use a current limiter or a device designed to provide a constant current until the switch is fully on. The newest approach is to use thermal regulation, which provides maximum current while regulating the temperature of the switch to prevent failure.

Passive

Adding resistance in series with the supply limits current to the load. A pure resistor would limit the peak inrush current to V_{IN}/R, and then the current would begin decreasing as the output becomes fully charged. This technique will dissipate power through the resistor as well as limit the current that the load can draw once powered. But because of poor efficiency, this technique is usually limited to low-current applications (often much less than 1 A). **Figure 2** shows a series resistor to limit inrush current.

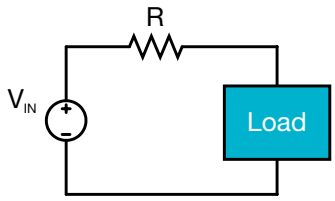

Figure 2: Series resistor to limit inrush current.

If adding an NTC, the resistance will start off high when the switch is cool and power has not flowed through it. As power is applied, the high resistance will allow a small amount of current through to the output. This will begin to self-heat the device and cause the resistance to drop, gradually allowing more and more current through to the load until the NTC is fully on. Since NTC performance relies heavily on ambient temperature, it may not be suitable for applications needing a wide operating temperature range. **Figure 3** shows how to use a series NTC thermistor to limit inrush current.

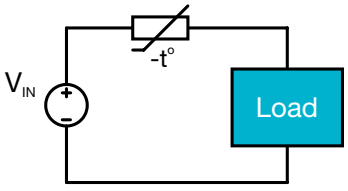

Figure 3: Using a series NTC thermistor to limit inrush current.

Slew-Rate Control

Controlling the rate at which the switch turns on directly controls the rate at which the output voltage rises. **Equation 1** calculates the inrush current as:

$$I_{INRUSH} = C_{LOAD} \times \frac{dV_{OUT}}{dt} \quad (1)$$

Slowing the switch speed (dV_{OUT}/dt) reduces I_{INRUSH} for a given, fixed C_{LOAD}.

RC Time Constant

For a discrete power switch, one common method to manage inrush current is to install a resistor-capacitor (RC) that slows down the MOSFET switching speed, as shown in **Figure 4**.

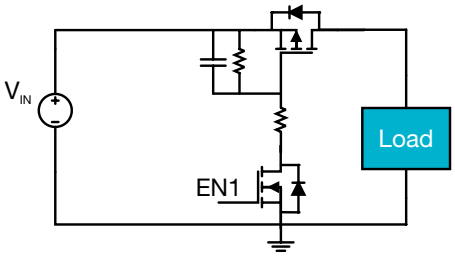

Figure 4: A discrete MOSFET with an RC rise time to limit inrush current.

This is a simple method to control the switch speed through an RC time constant. While the speed can be reduced and the total turn-on time increased, note that the dv/dt speed is nonlinear; therefore, the inrush current is nonlinear. The switch will start to turn on slowly and will exponentially increase as the switch becomes fully on. See **Figure 5** for RC-based inrush current.

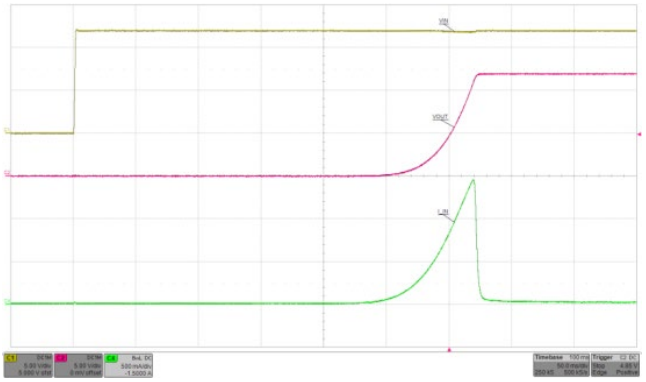

Figure 5: RC-based inrush current.

Linear Soft Start or dV/dt

Many integrated power switches have a method of controlling the output voltage rise time linearly. These switches may come with a fixed or adjustable rise time. In all cases, controlling the output voltage rise time linearly means controlling a constant dV_{OUT}/dt rate. In this case, if C_{LOAD} is constant and dV_{OUT}/dt is a constant speed, then I_{INRUSH} will also be a constant, as shown in **Figure 6**.

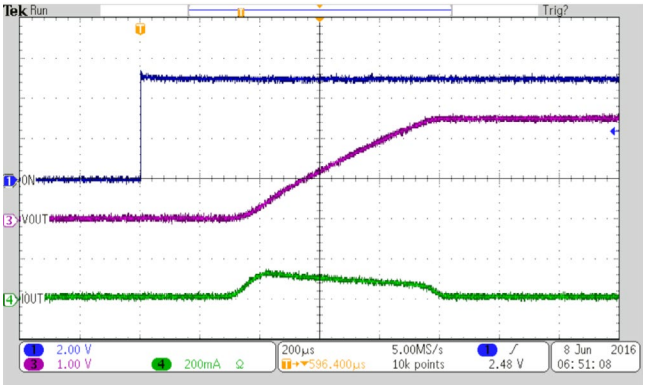

Figure 6: Linear dv/dt to limit inrush current using the Texas Instruments TPS22975 load switch.

Linear soft start has a couple of advantages over an RC time constant approach, including simple and precise calculations for inrush current. There may be situations where you have both a maximum inrush current restriction and a maximum turn-on time requirement. In these cases, if an RC time constant would either take too long to turn on or have too high of an inrush current, then a linear control scheme offers the best chance to satisfy both requirements. If you can't satisfy **Equation 2**, then there is no solution other than changing the design architecture to remove these restrictions, or moving to a smaller C_{LOAD} or a lower V_{OUT}:

$$I_{INRUSH\_MAX} = C_{LOAD} \times \frac{V_{OUT}}{t_{ON}} \quad (2)$$

Constant Current/Current Limit Regulation

Similar to a linear soft-start scheme, a constant current method of controlling inrush current would yield the same result if powering a purely capacitive load. If you charge with a constant I_{INRUSH}, then for a given C_{LOAD}, you will be charging at a constant dv/dt.

The difference is the linear soft start was regulating voltage, whereas a current limiter is regulating current. The approaches will differ once you begin introducing loads outside of a capacitor. If you have an integrated circuit that activates before the output voltage is fully on, then the linear soft-start method will have that current superimposed. The constant current method will change its slew rate, since some current is now going to the load instead of to the output capacitor.

In **Figure 7**, linear soft start is maintaining a constant slew rate for the output, whereas the constant current limit will change its slew rate when the load turns on at 2.5 V.

Constant current limit during startup can be helpful for situations where loads may turn on while the switch is still turning on. Constant-current-limit devices often provide protection against events such as an output short circuit. We will discuss current-limiting techniques in more detail in the next chapter, but it is important to understand that it is one form of managing inrush current during startup.

Thermal Regulation

In order to turn on as quick as possible but maintain protection for the switch, there is a final technique using thermal regulation. This can be useful in situations with

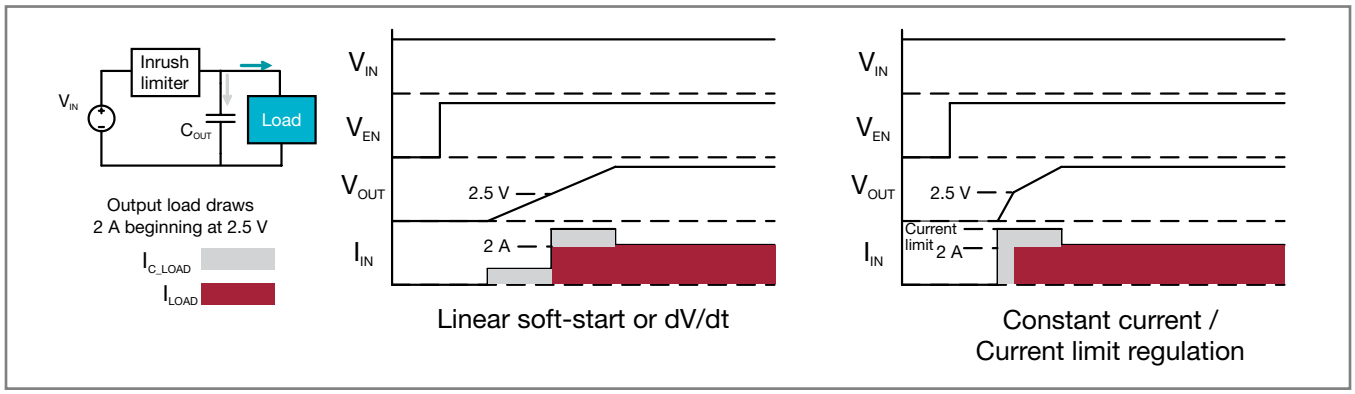

Figure 7: Linear dv/dt versus constant-current-limit regulation.

a large or unknown output capacitance (such as driving an offboard load). Once the junction temperature rises to a specified level, the device starts controlling the inrush current profile in order to maintain a regulated junction temperature. This will continue until the load capacitors are fully charged, or until a timeout. To learn more about a switch with a thermal regulation loop, see the TPS2663 eFuse (or electronic fuse), also highlighted in **Figure 8**.

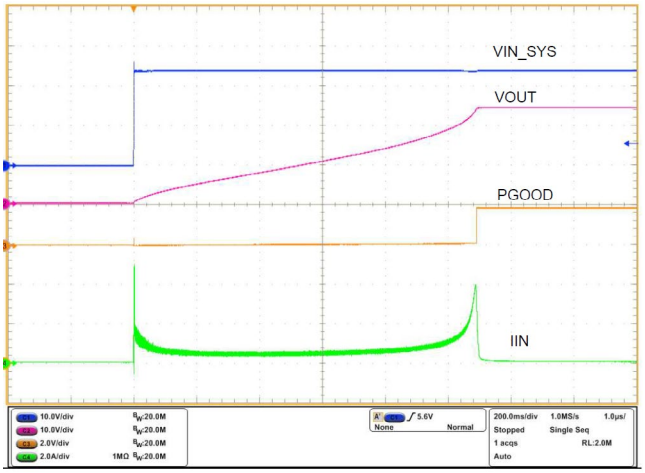

***Figure 8:** Thermal regulation loop using the TPS2663 with 24-V V_{IN} and 30 mF of output capacitance.*

Chapter 2 Adjustable Current Limit

Authors: Rakesh Panguloori and Lokesh Ghulyani

Abstract

Current limiting is essential in many systems to restrict transient overload currents to permissible levels. Traditional current-limiting solutions such as discrete resettable fuse circuits are inferior in performance due to their inaccuracy, slower response, and a lack of configurability and repeatability. This chapter discusses how integrated power switch solutions overcome the limitations of discrete current-limiting solutions.

Integrated power switch solutions provide accurate and adjustable current limit that helps limit overload currents precisely, thereby reducing front-stage power budgets and the size of passive components, cables and printed circuit board (PCB) traces. In addition, their accurate current-limiting (or output power-limiting) function simplifies system design by complying with critical safety standards.

The Need for Current Limiting

The semiconductor components, passive filters, PCB traces and interfacing cables found inside electronic equipment all have an inherent maximum current capability that when exceeded will cause them to overheat, possibly to an unrecoverable state. To overcome this, some kind of current-limiting device restricts fault currents within the maximum current-rating limits and brings the system back to normal operation once the fault is cleared.

In the example shown in **Figure 1**, a DC/DC converter is supplying power to various internal subsystems, each requiring a specific voltage for proper operation. An overload event in any one of these subsystems can affect the operation of system-critical loads connected to the same power distribution bus, leading to an erroneous system function or system reset.

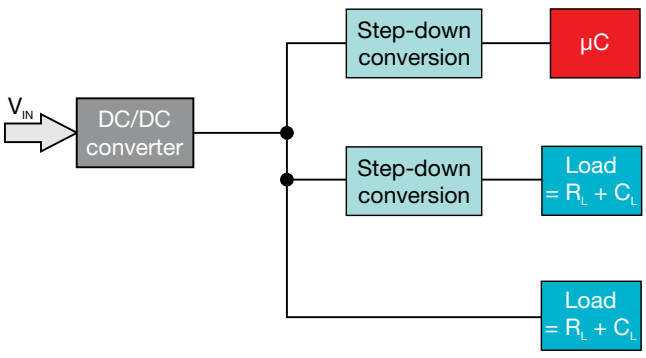

Figure 1: Power distribution in typical electronic equipment.

To overcome such a scenario, you can use current-limiting devices in each of the power paths to manage overload and short-circuit events, and to guard sensitive circuitry for reliable system operation.

These parameters determine the performance of a current-limiting device:

- Accuracy.
- Adjustability.
- Repeatability.
- Response time.

Traditional Ways to Implement Current Limiting

Fuses are traditionally considered protection devices, isolating overload or short-circuit faults from the main system. While fuses are an inexpensive solution, they can fall short of many of the protection requirements imposed on modern electronic equipment. The overload current needs to be exceedingly higher (500%) than the rated fuse current in order to produce a response within milliseconds. This makes it extremely difficult to predict the precise overcurrent level at which the fuse will open. A conservative fuse current rating selection may lead to a fuse blowup during inrush current events. In addition, once the fuse blows during an overload event, it has to be physically replaced, which increases system downtime and maintenance costs. **Figure 2** shows such a scenario.

Figure 2: A blown fuse needs to be physically replaced every time in order to resume normal operation.

A resettable fuse is a positive temperature coefficient (PTC) device that increases its on-resistance with temperature. During an overload event, the excessive load current increases power loss, thereby increasing

its on-resistance. The higher on-resistance helps limit the overload current and protects the circuit. Unlike a physical fuse, PTCs allow current to flow after the fault is cleared without replacing the device.

Because PTCs are actuated by the heating effect of an overcurrent load, their reaction time is limited to several milliseconds, as they have a natural dependency on the ambient temperature, as shown in **Figure 3**. Another property of a resettable fuse is that its on-resistance increases after every reset, which raises concerns about achieving repeatable performance over time.

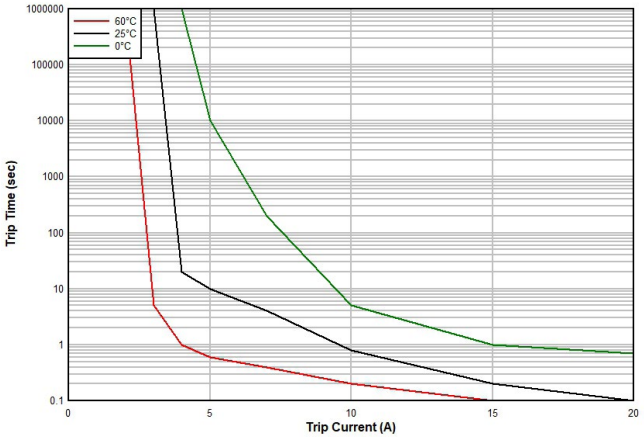

Figure 3: Trip time vs. trip current temperature dependency.

Discrete Current-Limiting Circuits

Figure 4 shows an example of a current-limit circuit using discrete components. The resistor-capacitor components around discrete power switch Q1 slow down the metal-oxide semiconductor field-effect transistor switching speed and provide inrush current control. In this implementation, the combination of series sense resistor Rs and P-channel N-channel P-channel transistor Q2 provide active current limiting.

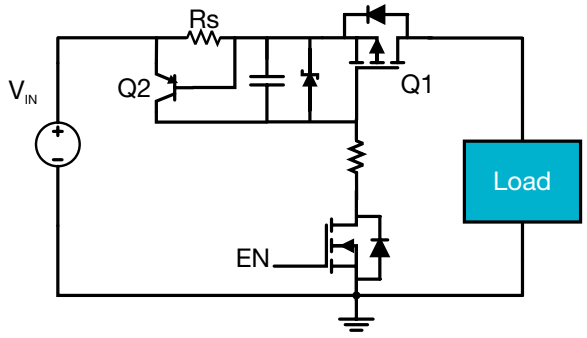

Figure 4: Active current-limiting circuit using discrete components.

In steady-state operation, Q1 is fully turned on and load current flows through the series sense resistor Rs and discrete power switch Q1. The voltage drop across Rs generates bias voltage VBE for Q2, which is used to control Q1 in a closed-loop form. When the load current reaches high enough to bias Q2 on, Q2 reduces the gate-to-source voltage of Q1 to limit the load current to VBE/Rs.

The response time of this discrete approach is much better than a fuse or PTC, but still suffers from poor accuracy. Current-limiting accuracy can be in the range of 30% considering the variation of VBE voltage over temperature. Another disadvantage is the power loss in the sense resistor; for a 5-A application, dissipation would be as high as 3.25 W (0.65 V × 5 A = 3.25 W). It is possible to reduce the power loss by using either a current-sense amplifier or a fast comparator to disable pass switch Q1 in the event of overload, but this adds complexity and cost. Discrete current-limiting circuits do not incorporate thermal protection, so these solutions need careful selection of the pass field-effect transistor (FET) and prudent thermal design to keep the pass FET within safe operating area limits under extreme fault conditions. All of these limitations often lead to a trade-off between accuracy, cost, complexity, PCB size and power loss.

eFuses

An eFuse is an active current protection device with an integrated FET used to limit currents to safe levels during fault conditions. The common elements of an eFuse are a power switch to modulate the load current, a current-sensing element and control logic. As shown in **Figure 5** on the folllowing page, the current-limit threshold can be set to a desired value through external resistor R_{LIM} at the I_{LIM} pin of the eFuse, providing system flexibility in adjusting the current limit to a wide range.

In a steady-state condition, the integrated power switch is fully turned on (just like a discrete switch) to minimize voltage drop in the power path. If at any point the load current reaches the set current-limit level, I_{LIM}, the gate control transitions the integrated power switch into linear mode. This increases the on-resistance and corresponding voltage drop across the eFuse to provide a constant output current.

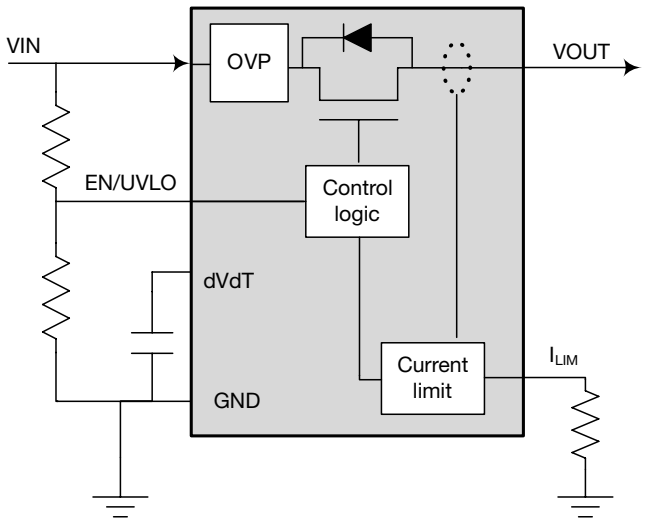

Figure 5: Block diagram of an eFuse.

Figure 6 shows the overload response of an eFuse with the current limit set at 4 A. As you can see, the eFuse responds quickly to an overload event (typically hundreds of microseconds) and regulates the output current to I_{LIM}. Once the fault is removed, the eFuse recovers back to normal operation smoothly.

An eFuse uses a current mirror circuit to measure the path current without the need for an external sense resistor (like Rs in **Figure 4**). This approach not only saves board space but also eliminates the loss in sense resistors and provides better current-limiting accuracy over discrete approaches—in the range of ±5% to ±8%.

Another major advantage of eFuse devices is their integrated overtemperature protection. The device shuts off the integrated power switch for junction temperatures above 150°C to ensure robust protection.

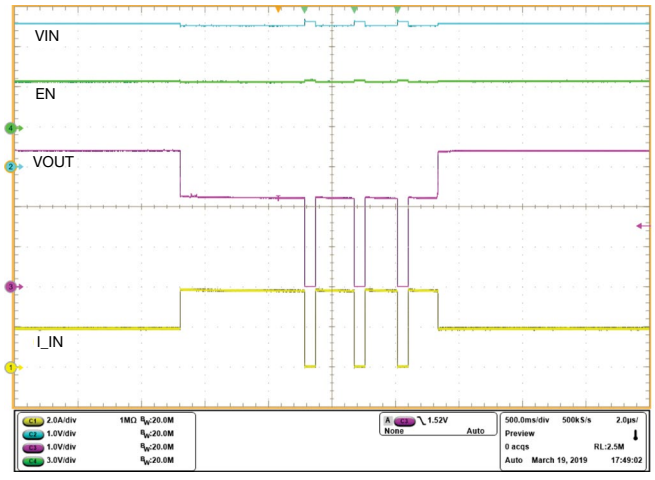

Figure 6: Transient overload current limiting and recovery with the TPS2595 eFuse.

High-Side Switches

Smart power switches offer robust protection against overload and short-to-ground events through overcurrent protection and are available in Automotive Electronics Council (AEC)-Q100 options. These devices provide a current-limiting function similar to eFuses but differ in their ability to drive inductive offboard loads.

As shown in **Figure 7**, smart power switches have options to use either an internal, fixed high current limit (when the C_L pin is ground) or an external lower current-limit value (when the C_L pin is tied to ground through external resistor R_{CL}). The adjustable low current limit greatly reduces the fault energy that the internal power switch has to handle during a fault, in turn reducing stress on the power switch and improving system reliability.

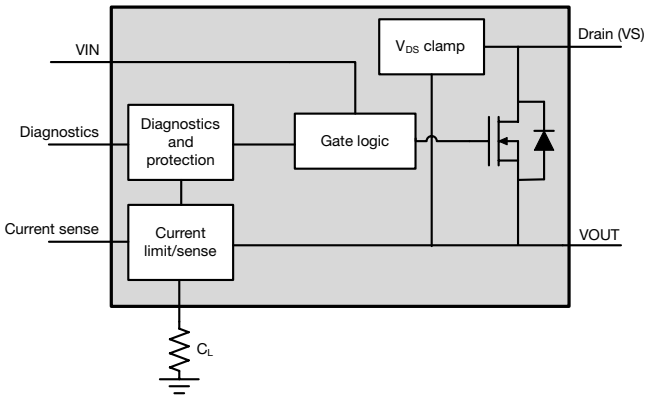

Figure 7: Block diagram of a high-side switch.

The lower current limit also helps clamp the inrush currents to a much lower value, as shown in **Figure 8** on the following page. Lower fault currents save system-level costs by minimizing PCB trace widths and cable sizes, along with lower connector ratings and component tolerances. For more information on the benefits of adjustable current limiting with smart power switches, see the application report, "Adjustable Current Limit of Smart High-Side Switches."

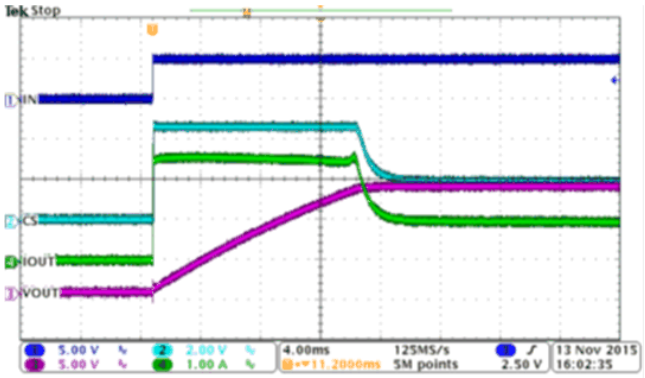

Figure 8: Inrush current clamping with the TPS1H100-Q1 high-side switch.

Power Limiting

Power limiting is usually required to limit the power delivered to the load. For power supplies with a narrow voltage range, a current-limiting device would be sufficient to limit power. But for power supplies with a wider voltage range, a power-limiting device is required. **Figure 9** illustrates a typical implementation for limiting power.

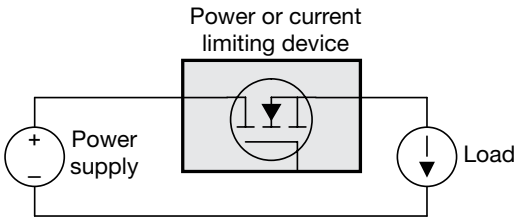

Figure 9: Typical power limiting.

The Need for Power Limiting

In most applications, it's possible to achieve power limiting with current limiting, but certain applications need to precisely limit the energy or power delivered into the load. Industrial systems such as programmable logic controllers have a wide operating voltage range, from 18 V to 36 V. In telecommunication systems, the operating voltage range varies from 36 V to 72 V, thus may require power limiting as well as current limiting.

Limiting current for a fixed output voltage or a smaller variation in voltage can achieve power limiting. But larger variations in voltage require an additional control loop in order to limit power. This additional current loop adjusts the current limit with voltage to maintain constant power. TI's Precision Power Limiting Solution reference design provides an example of this additional control loop.

There are safety and compliance standards such as International Electrotechnical Commission (IEC) 61010-1 for industrial equipment power and energy limiting. Complying with these standards means that precise power limiting is also required.

Certain power supplies are inherently power-limited; these power supplies come under National Electrical Code (NEC) Class 2 or IEC 60950 standards. Achieving power limiting in these power supplies requires an additional power-limiting circuit.

Implementing Power Limiting Using an eFuse

To precisely limit power, such as in industrial applications, TI's TPS2663 and TPS1663 eFuses offer an integrated solution. These eFuses have an additional power-limiting control loop, and the output power limit is configurable through the P_{LIM} pin. The TPS2663 and TPS1663 can limit output power with an accuracy of ±6% up to 150 W. These devices also support load transients by enabling current twice the programmed current limit (two times the pulse current) to pass to the transient load.

Figure 10 shows the power-limiting response of the TPS2663. The output power limit is set to 100 W by the resistor on the P_{LIM} pin; the device allows a pulse current of 12 A and a pulse power of 282 W for a duration of 25 ms to support the load transients. For more details on power limiting with TPS2663 devices, see the data sheet.

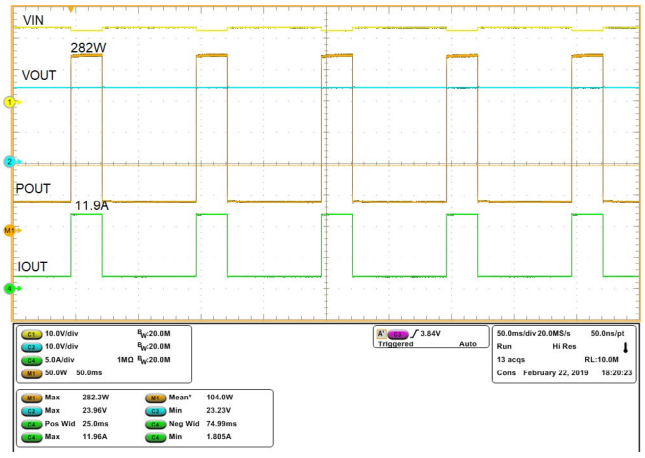

Figure 10: Power limiting with the TPS26633 (two times the pulse current support, I_{LIM} = 6 A, V_{IN} = 24 V).

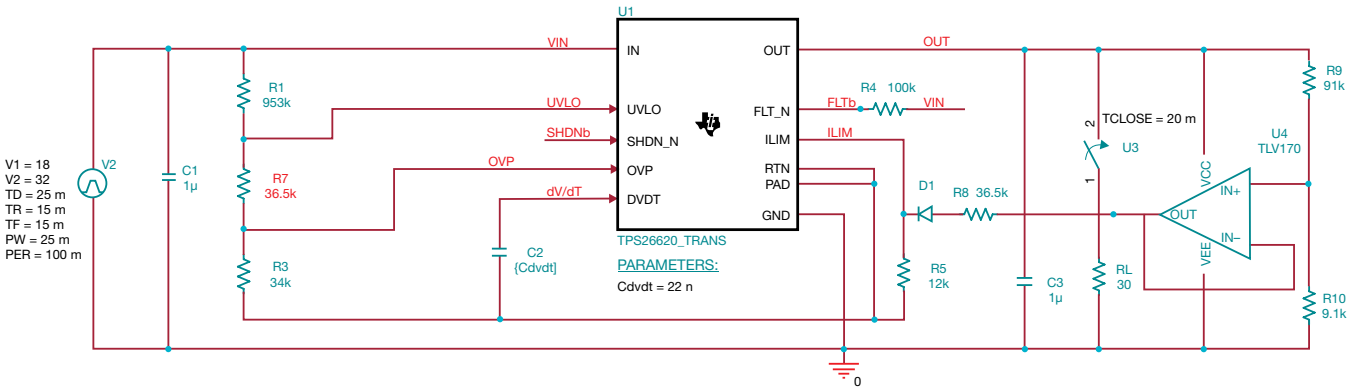

Figure 11: Power limiting with the TPS2662, with an additional control loop.

Implementing Power Limiting with an External Control Loop

It is possible to use current limiting to achieve power limiting for power supplies with a narrow input-voltage range. Achieving power limiting in wider-voltage-range power supplies requires an additional control loop for precise power limiting that adjusts the current limit based on the output voltage.

Figure 11 shows an implementation of the TPS2662 for power limiting. The additional control loop uses a TLV170 amplifier to inject current proportional to the output voltage into the I_{LIM} pin for power limiting. **Figure 12** provides the PSpice simulation results for the circuit shown in **Figure 11**. For this simulation, the input voltage varies from 18 V to 32 V. The TLV170 adjusts its current limit and the output power is limited to less than 10 W.

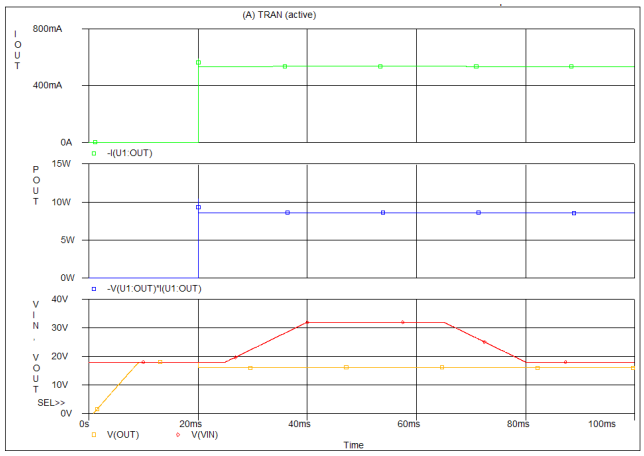

Figure 12: Simulation results for power limiting with the TPS2662.

Conclusion

Current limiting is a basic protection requirement in most modern electronic equipment. Although discrete components work, they require more board space, are less efficient and can generally be less cost effective. TI integrated solutions, eFuses and high-side switches, provide accurate current limiting, faster response times and can repair themselves without user intervention.

Chapter 3 Short-Circuit Protection

Author: Alec Forbes

Abstract

Short-circuit protection is critical to ensuring robust reliability for output power protection, as short-circuit failures can cause supply failures or even damage to systems. The most common implementation of short-circuit protection is through the use of a current-limited output; however, it is important to understand the impact of current limit on short-circuit reliability.

Short-circuit behavior and protection requirements vary based on the output impedance, ranging from terminal shorts to a partial resistive short. Depending on the short-circuit impedance and load profile, the current and energy profile will take different forms.

What is a Short Circuit?

In personal electronics, industrial, automotive and communication applications, short-circuit protection is a safety requirement for all power distribution outputs. Short circuits occur whenever a charged conductor makes physical contact to a grounded element in the system. During a short circuit, the electrical short between power supply and ground can create a very low impedance path, causing a high, uncontrolled sustained current flow. This type of flow is extremely dangerous for systems and can cause integrated circuit (IC) failure, as seen in **Figure 1**. Thus, any systems with a risk of short-circuit events must implement robust protection.

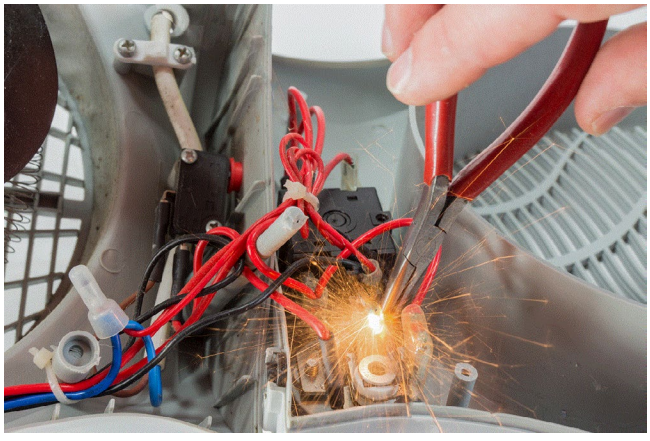

Figure 1: Example of IC failure.

The risk of a short circuit becomes higher in systems that contain some or all of these characteristics:

- **Output cables.** Over time, cabling can wear down and present exposed metal, which can make contact with connectors.
- **Debris-filled environments.** A common cause of short circuits is metallic debris shorting pins or traces.
- **Maintenance requirements.** Any possibility of exposure to metal tools during maintenance increases the risk of an accidental short-circuit event.
- **Strong voltage supplies.** The more energy that a power supply can provide, the higher the risk of a short-circuit event causing a health hazard.

Automotive systems, for example, regulate output short-circuit protection through the Automotive Electronics Council (AEC)-Q100-012 standard, which mandates that systems should be capable of surviving thousands of repetitive short-circuit events. The application report, "AEC-Q100-012 Short-Circuit Reliability Test Results for Smart High-Side Switches," documents TI's test process for high-side switches.

Protection Against Short Circuits

The primary danger during a short-circuit event is the sustained high current flow caused by low output impedance. So in order to protect against short circuits, the system must have output power protection that can recognize a short-circuit event and then shut the system off quickly and safely.

The best way to implement this type of protection is to use a current-limiting switch on the output. This switch will provide protection against a sustained uncontrolled short-circuit current and shut off, protecting the output circuitry. Because the output impedance is unknown and often has an inductive element, the switch must be able to safely demagnetize the inductance, and thus needs a flyback mechanism. **Figure 2** on the following page shows the typical setup that has potential for short circuit.

There are three primary short-circuit conditions that a system must be able to withstand: terminal short circuits, cable short circuits and resistive short circuits.

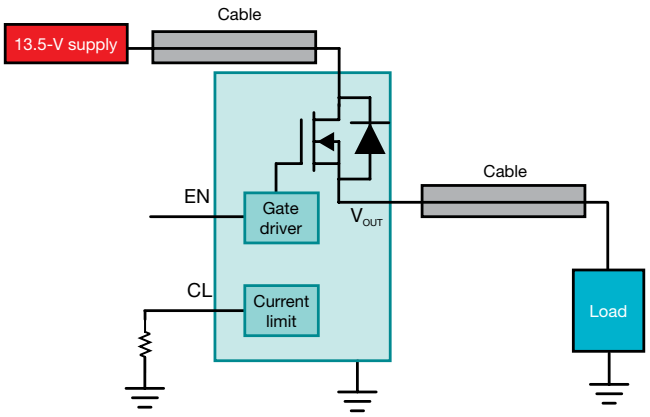

Figure 2: Typical setup with potential for short circuit.

Terminal Short Circuit

The simplest type of short circuit is the terminal short circuit, illustrated in **Figure 3** with a TI high-side switch providing an output protection stage. The output current increases quickly until it hits a 90-A current limit, at which point the switch recognizes a short-circuit event and turns the switch off, preventing dangerous current flow.

Because this is a terminal short circuit, there is minimal series impedance, and the current-limiting output is easily able to protect the system from the short circuit. The rise and fall times of the short-circuit current are very quick, and the entire event is over in less than 50 µs.

To understand the short circuit, take a look at the total cumulative fault energy, which occurs in two portions:

- As the short-circuit current ramps up, the current causes resistive power losses in the field-effect transistor (FET); however, they are low, as the FET is in saturation mode and the resistance is very small.
- After the switch hits its current limit and turns off, small trace inductive elements in the short circuit create a negative voltage spike that causes the voltage clamp in the switch to engage and the drain-to-source voltage (V_{DS}) to rise to 40 V. While the inductance demagnetizes, this high V_{DS} requires the dissipation of more fault energy.

Because of the short length of the pulse, the total fault energy is kept to a relatively safe 18 mJ, despite the high current limit. This case is the easiest short circuit to survive, as there are no series impedances that will impact the output protection.

Cable Short Circuit

In the event of a short circuit through a cable, accounting for the cable series impedance increases the difficulty of protection. The waveform in **Figure 4** on the following page shows a TI high-side switch short-circuited; however, this time the short circuit occurs at the load side of a 5-µH cable (a standard value for a 5-m cable).

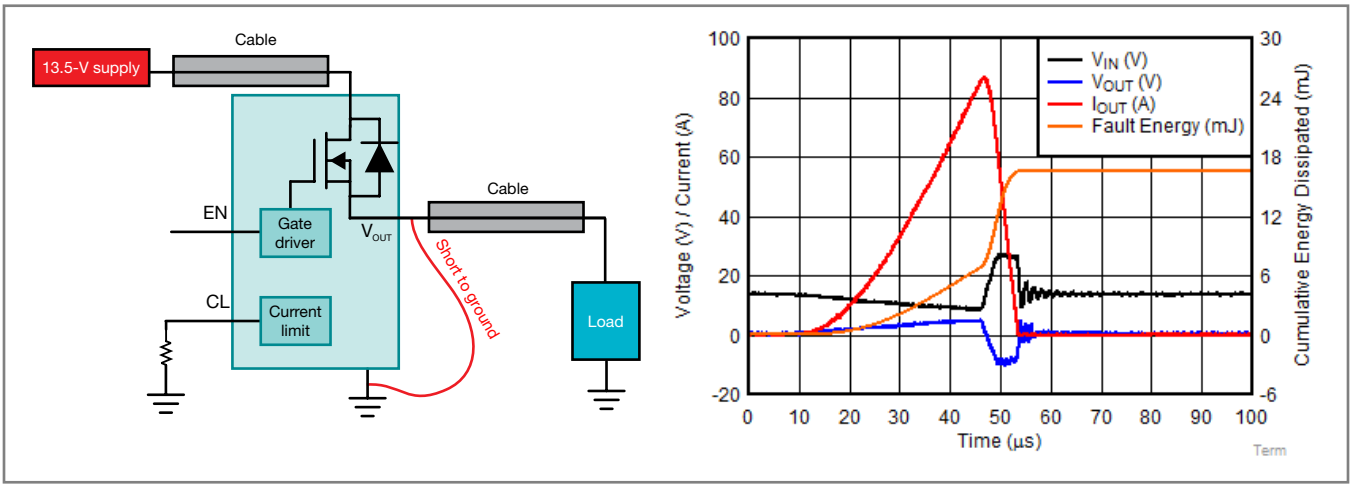

Figure 3: Terminal short on output.

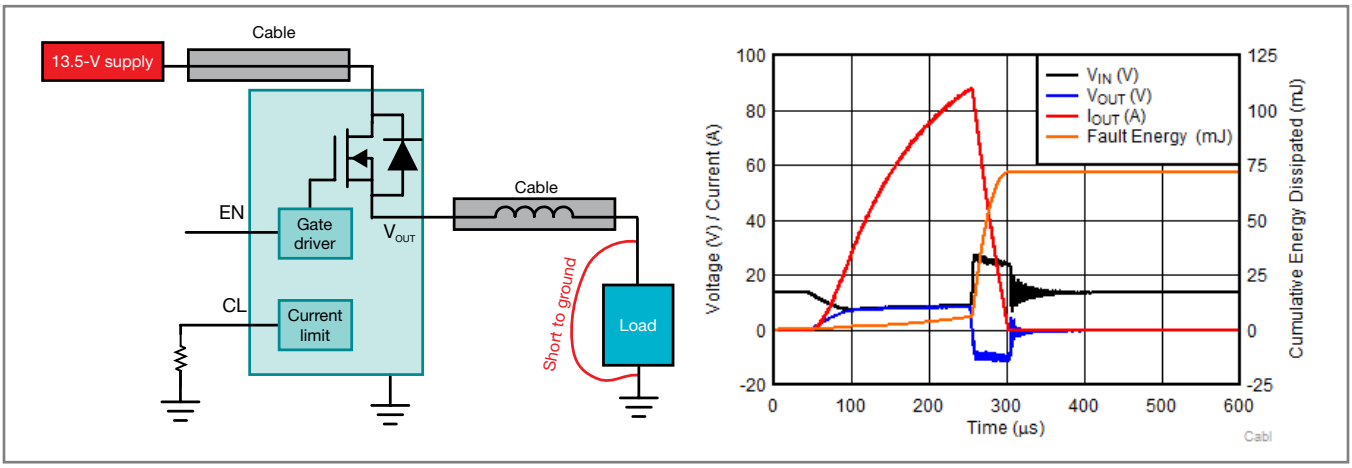

Figure 4: Cable short circuit across the load.

The extra series inductance significantly slows down the current rise and fall times, so the pulse takes 10 times longer and the total fault energy is 75 mJ—four times higher than the terminal short circuit. Energy dissipation occurs in two portions similar to the terminal short-circuit case—but now that the output inductance is 5 µH rather than just parasitics (typically less than 10 nH), the inductor demagnetization contains a much higher percentage of the cumulative fault energy.

The energy dissipated in the fault can be very large due to the high current, so the inductive clamp must be rated high enough to dissipate this energy. **Equation 1** calculates the energy dissipated while discharging an inductive element through a series clamp:

$$E = \frac{1}{2} \times L \times I_{PK}^2 \times \left(\frac{V_{IN}}{V_{CLAMP} - V_{IN}} \right) \quad (1)$$

The high-side switch parameter, V_{CLAMP}, is device-dependent, but it's usually 40 V to 60 V. Therefore, with even a small 5-µH cable inductance, a high current limit can create a large demagnetization energy that the output protection stage must absorb. The input voltage can also increase during demagnetization because of the input inductance, increasing the total fault energy that must be dissipated.

Most high-side switches list the inductive load energy peak, so make sure that the calculated energy from the maximum cable length and maximum current is below the rated energy. **Figure 5** below shows these values for a TI high-side switch.

Resistive Short Circuit

A resistive short circuit occurs when debris or corrosion causes a partial short circuit or a load failure. In the short circuits described above, a low series resistance causes the current to increase until it hits the current limit. When there is some series resistance with a high current limit, there is a risk that the current will saturate before hitting the current limit. In the event of a series resistance, the output protection must have a low-enough current limit to register the short circuit as a fault event.

Figure 6 on the following page shows a case where the series resistance limits the short-circuit current to below the output protection current limiter, so the device does not turn off the output and current flows indefinitely.

During this time, there is a high, uncontrolled current flow until the system hits thermal shutdown or a secondary method of current monitoring turns the system off. For reliability against resistive short circuits, you must either design the overall system to manage the transient

Energy dissipation during turnoff, E_{TOFF}	Single pulse, L_{OUT} = 5 mH, $T_{J,start}$ = 125°C	200 mJ
Energy dissipation during turnoff, E_{TOFF}	Repetitive pulse, L_{OUT} = 5 mH, $T_{J,start}$ = 125°C	80 mJ

Figure 5: Inductive energy dissipation of a TI high-side switch.

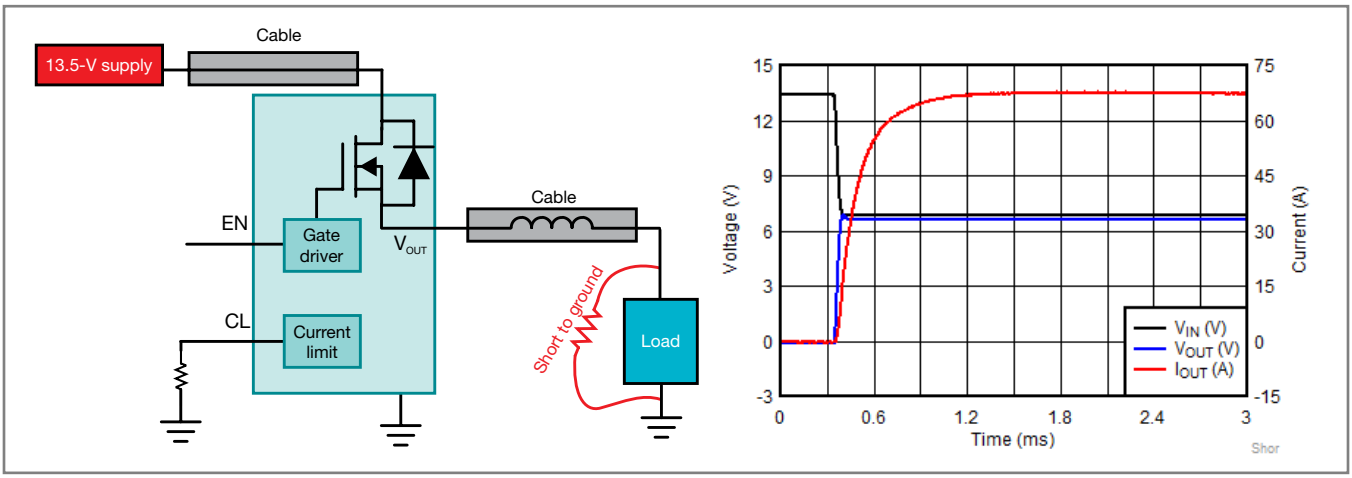

Figure 6: Resistive short circuit.

current pulse or ensure that it will shut off by minimizing the output protection current limit.

The "Improved Automotive Short Circuit Reliability Through Adjustable Current Limiting" application note describes this event in greater detail.

System Interruption

The previous sections covered how to ensure that the system will survive a short-circuit event. But even if the system does survive without damage, a short-circuit event should cause as minimal of an interruption to the system as possible.

During a short-circuit event, high current can overload the input supply and create an undervoltage lockout event for the entire system. To prevent this, it's important that the short circuit is registered as quickly as possible so to facilitate the implementation of short-circuit protection with solutions that have a very fast response time.

To minimize the response time, many protection devices implement a fast-trip response that triggers within nanoseconds of an overcurrent event. If the fast-trip threshold is hit, the device instantly turns off to protect the supply, and then slowly ramps up the output current safely to the chosen current limit. This keeps overshoot during the short circuit from having an adverse effect on the supply.

Protection Considerations

To protect against short circuits, it's best practice to use an output protection stage with an integrated current limit and inductive clamp. Typically, this protection is implemented through a high-side switch with an integrated output clamp, or through an eFuse or load switch IC with an external flyback diode.

When designing short-circuit protection, consider:

- Minimizing the current-limiting value. Lowering the current limit minimizes the energy that must be dissipated in a short circuit. Because this energy is a function of the square of current, minimizing the peak current significantly decreases the risk of failure. Keep the current limit as close to the maximum working current as possible.
- Confirming that the output protection can dissipate the required fault energy. Calculate the maximum short-circuit energy expected based on cable lengths and current limits, and check that the output protection stage will be able to safely dissipate that energy.
- Limiting input and output inductances. Inductances in series with the short circuit significantly increase the demagnetization time and total fault energy that must be dissipated.
- Guaranteeing the registration of a resistive short circuit. In the event of a resistive short circuit, ensure that either the current limit is low enough to recognize a partial fault or that there is an alternative method

of overcurrent detection (like a microcontroller) continuously monitoring the load current.

- Understanding protection response times. In the event that a short circuit causes an overcurrent, there is a response time before the I_{LIM} circuitry is able to limit the current, during which there is some overshoot. Understand how the response time will impact the maximum possible current.

By designing an output protection stage around these principles, you can safely protect an output from a short circuit and minimize the risk of damage.

Chapter 4 Reverse Current Blocking

Author: Alek Kaknevicius

Abstract

Reverse current occurs when current travels from output to input rather than from input to output. This chapter explains where reverse current comes from, why it can be harmful to systems and how to modify a design to protect against reverse current.

What is Reverse Current?

Reverse current is when there is a higher voltage at the output of a system than at the input, causing current to flow backwards through the system. There are two common sources of reverse voltage: the first is when power is disconnected from a system and the input voltage suddenly drops. During this time, a higher voltage can be left at the output, which temporarily produces reverse current. See **Figure 1**.

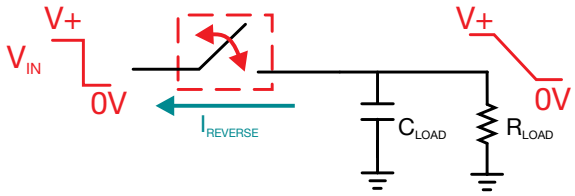

Figure 1: Reverse current caused by a sudden loss of power.

The other cause of reverse current is when a metal-oxide semiconductor field-effect transistor (MOSFET) is used for a load-switching application and the body diode becomes forward-biased.

The output of a switch with a higher voltage than the input causes a reverse voltage, which is what causes reverse current. Don't confuse this with negative voltage, also known as reverse polarity. A negative voltage occurs in situations where the positive and negative terminals of a power supply are switched. In this case, what should be connected to ground actually has a voltage from the positive input to the system, as shown in **Figure 2**. This causes a different current phenomenon through the device than reverse polarity. For more information on reverse polarity protection, see Chapter 5.

Why Do You Need to Block Reverse Current?

Reverse current can damage internal circuitry and power supplies. Depending on the path from output to input, reverse current spikes can also damage cables and connectors. If you are using a MOSFET for load-switching applications, reverse current can travel backwards through its body diode if it becomes forward-biased, because the FET output voltage is greater than the input voltage. This causes a linear rise in power dissipation across the body diode, expressed by **Equation 1**:

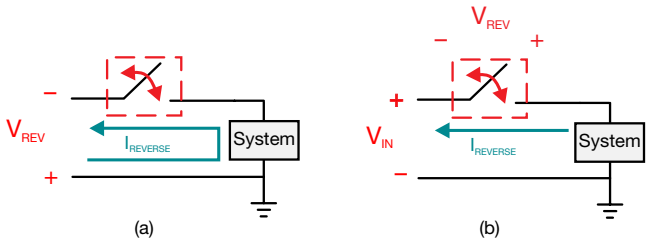

Figure 2: Reverse polarity (a); reverse voltage (b).

$$P = IV: P_D = I_{REV} \times V_{DROP} \quad (1)$$

where P_D is the power dissipated across the body diode, I_{REV} is reverse current through the device and V_{DROP} is the voltage drop across the body diode.

If the heat generated by the power dissipation through the device exceeds the thermal rating of the device, then combustion can occur. Therefore, it is essential to limit reverse current flow, or reverse voltage.

When Do You Need to Block Reverse Current?

There are several applications where blocking reverse current is necessary.

Power Multiplexing

Power multiplexing is the practice of using a switching circuit to choose one of multiple power supplies for a system, with the ability to switch between them. **Figure 3** on the following page shows this configuration.

If one of the power-supply voltages is higher than the other, then it's possible for reverse current to occur even when the other power rail has an "open" switch—one example is when using FETs to switch the power supplies. The higher voltage at the output of the open FET causes reverse current to flow from the higher-voltage power supply, through the FET body diode and into the lower-voltage power supply. **Figure 4** on the following page shows the application of 5 V to a system while the switch for the 3.3-V rail is open.

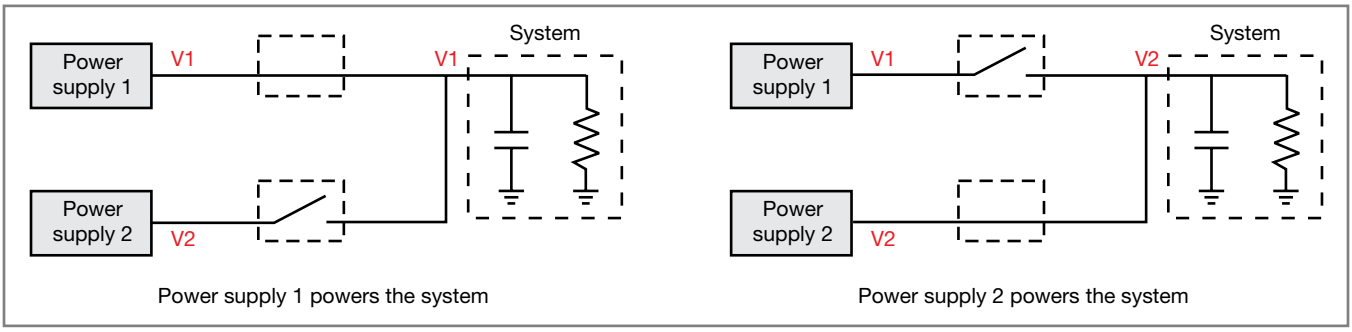

***Figure 3:** Power multiplexing configuration.*

You can see in **Figure 4** that with a simple FET solution for the switch, reverse current is able to flow through the FET body diode, even when the switch is open.

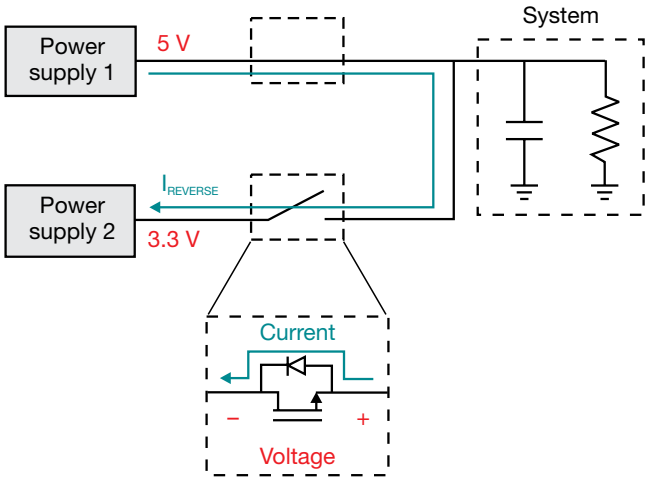

***Figure 4:** Reverse current caused by power multiplexing.*

ORing

ORing is similar to power multiplexing, except that rather than selecting one supply to power the system, the highest voltage always powers the system. Reverse current blocking is also necessary here, because each ORing switch sees reverse current when the other switch is closed.

Sudden Loss of Input Power

When a closed switch suddenly loses power on its input, there is the potential for reverse current. If the capacitance on the output of the switch is larger than the input, then the voltage on the output will decay more slowly. This means that while power is decaying, the voltage on the output of the switch will fall more slowly than the input. During this period, the voltage on the output of the switch will be larger than the input, so reverse current will flow across the switch. To avoid this, you should have a switch with reverse current blocking, or a larger input than output capacitance.

In some systems, a supercapacitor holds up the output when the input supply drops. This is sometimes referred to as a "last gasp" circuit, which allows the system to safely power down. It's also a good idea to have reverse current protection here in order to protect any upstream supplies or components.

How Do You Block Reverse Current?

There are several ways to block reverse current.

Diodes

Diodes are great for high-voltage, low-current applications. However, diodes cause a forward-voltage drop that increases the total power dissipation in the system and drops the power supply by 0.6 V to 0.8 V. This can cause decreased efficiency in the system and a shortened battery life. A popular alternative is the use of a Schottky diode. They have lower forward-voltage drops, but are more expensive and have higher reverse current leakage, which could cause problems for the system such as supply or battery damage.

Back-to-Back MOSFETs

Using back-to-back MOSFETs is a powerful option, since it offers current blocking in both directions when the MOSFETs are turned off. In comparison to the diode solution, there is a lower voltage drop from the power supply to the load. However, this implementation takes up a larger amount of space on the board, requiring several components to build, as shown in **Figure 5** on the following page.

Aside from the large solution size and bill-of-materials count, this solution also doubles the on-resistance of a single MOSFET solution and does not provide

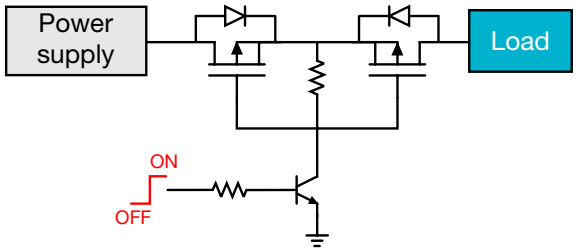

Figure 5: Reverse current blocking with back-to-back MOSFETs disabled.

any reverse current blocking when the MOSFETs are turned on.

Backwards MOSFET

If a MOSFET is positioned so that the body diode is facing from input to output, then there will be no reverse current flow when the MOSFET is switched off (open) (see Figure 6).

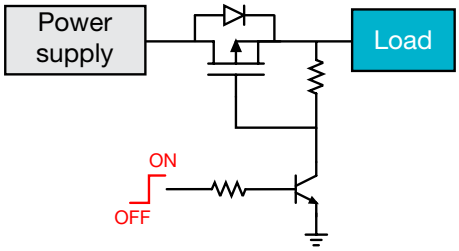

Figure 6: Reverse current blocking using a single MOSFET disabled.

The disadvantage with this solution is that it is not possible to switch the supply off, since there is always a diode path between the source and the load. This is a preferable status for diode ORing but not so much for power multiplexing because you may need to prioritize a lower voltage supply, such as when the main supply for a system is a lower voltage but the backup battery is a higher voltage.

With an ORing circuit, the battery will power the system until the voltage has been drained to below the main supply voltage. A power multiplexing circuit will prevent the battery from being used until the system decides to switch over to the backup battery.

Switching the MOSFET Body Terminal

To block reverse current with a single MOSFET, the body terminal must either be biased to the highest supply or ground, depending on if the FET is P-channel or N-channel. This is usually not possible for a discrete MOSFET, since the body terminal is not typically accessible.

Power Switches that Block Reverse Current

Several types of power switches are capable of blocking reverse current.

Load Switches

Load switches integrate reverse current blocking by switching the body terminal of the internal MOSFET. This allows the device to have a low on-resistance while still providing protection to the supply. If the load switch is off, then reverse current is always blocked.

Some devices offer always-on reverse current blocking, which will turn off the device if the output voltage is greater than the input voltage by a certain voltage threshold. The disadvantage with an always-on reverse current blocking scheme is that the output voltage needs to be greater than the input voltage before the device turns off, so some amount of reverse current will flow through the device before this occurs. **Figure 7** shows an example of the expected reverse current through a load switch before reverse current blocking is enabled.

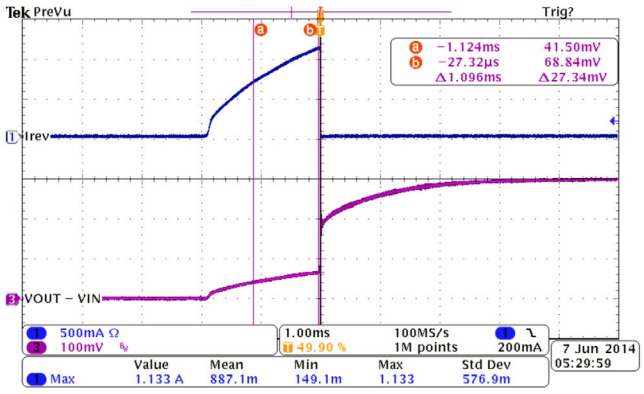

Figure 7: Full-time reverse current protection (V_{IN} = 3.0 V, V_{OUT} ramp up from 3.0 V to 3.3 V).

eFuses

With eFuses, a back-to-back MOSFET implementation is most common due to the MOSFETs' higher voltage capability. In some devices, both MOSFETs are integrated into the device, but in others, there is a pin for driving an external second FET to create a back-to-back configuration. Similar to load switches, there are some eFuses that come with always-on reverse current

blocking by monitoring the difference between V_{OUT} and V_{IN}, while others only provide reverse current blocking when turned off.

Power Multiplexers

Integrated power multiplexers safely handle power multiplexing applications so that reverse current does not flow from one supply into the other. Current power multiplexers adopt both the back-to-back and body terminal configurations, and all power multiplexers have reverse current blocking.

Ideal Diodes

Ideal diodes are controllers that drive an external MOSFET with the body diode facing from supply to load. These devices will automatically detect the input and output voltage difference to always enable reverse current blocking while maintaining a low on-resistance.

Conclusion

Applications such as power multiplexing or a loss of power can lead to a reverse voltage event that will induce reverse current. This can cause damage to both power supply and system. A TI power switch is both a size- and cost-effective solution helping prevent damage from reverse current. The TI power-switch portfolio has a variety of devices with different specifications for reverse current protection, making them a good fit for a wide variety of applications.

Chapter 5 Reverse Polarity Protection

Author: Karikalan Selvaraj

Abstract

Many front-end power system designs require protection from damage caused by a reverse-connected input power supply, a reversed battery connection or miswiring of field power-supply lines.

Schottky diodes are the traditional choice for protection against reverse polarity conditions, but the associated power loss from the forward conduction requires careful thermal management, leading to increased system cost and space. Plus, increased demand for higher power density necessitates better and more efficient reverse polarity protection methods.

Reverse Polarity Protection Overview

Power system modules or subsystems powered from a field power supply or battery power require protection from reverse polarity connections caused by miswiring during system maintenance or reinstallation. Most front-end power systems need protection from dynamic reverse polarity conditions that can occur during a surge event or an inductive load disconnect from the battery. Reverse-connected supplies can cause damage to connected subsystems, circuits and components.

Many automotive battery-powered subsystems require protection from reverse battery connections during a repair or jump-start from another vehicle, and from transient reverse input conditions that can occur during a vehicle's operating lifetime.

Reverse Input Polarity Protection

Figure 1 shows battery input lines connected in reverse. When this occurs, huge current flows through the electrostatic discharge (ESD) diode of microcontrollers (MCUs), DC/DC converters or other integrated circuits, and causes severe damage to battery-connected subsystems. A reverse-connected battery can damage polarized components such as electrolytic capacitors, as shown in **Figure 2**.

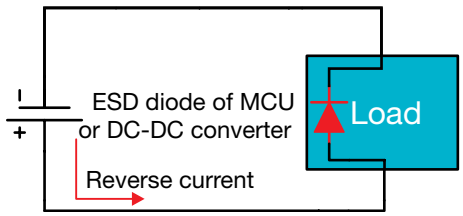

***Figure 1:** Reverse-connected battery: MCU or DC/DC converter.*

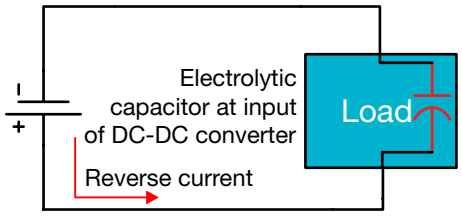

***Figure 2:** Reverse-connected battery: polarized components.*

Common Reverse Polarity Protection Methods

Popular methods to protect against reverse input polarity employ either a Schottky diode or a P-channel metal-oxide semiconductor field-effect transistor (MOSFET) with discrete components.

Reverse Polarity Protection Using Schottky Diodes

The simplest method of reverse battery protection is to add a series diode at input of the system power path. **Figure 3** shows a reverse battery protection scheme using a Schottky diode. When the battery is installed correctly, load current flows in the forward direction of the diode. If the battery is installed with the wrong polarity, the diode is reverse-biased and blocks reverse current, thereby protecting the load from a negative voltage.

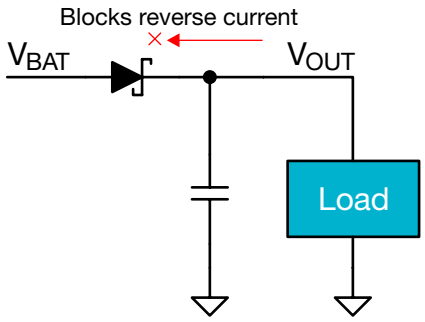

***Figure 3:** Reverse input protection using a Schottky diode.*

Bulk capacitance placed at the output holds the output from falling immediately and can supply the load for a short time before the input supply recovers.

Drawbacks of using a Schottky diode for reverse battery protection include:

- **Power dissipation:** A forward conduction loss due to a forward voltage drop results in significant efficiency loss at higher load currents.
- **Thermal management:** Managing the power dissipation requires a heat sink, increasing total cost and space.
- **Higher forward voltage drop:** In systems with lower voltage bus (3.3 V), a typical voltage drop of 0.4 V in diode will result in lower voltage headroom for the downstream loads to operate.
- **Reverse leakage current:** Reverse leakage current of high-voltage Schottky diodes increases dramatically with junction temperatures, resulting in higher power dissipation during reverse conduction.

Reverse Polarity Protection Using Discrete MOSFETs

You can reduce the forward drop of the diode by replacing the Schottky diode with a P-channel MOSFET and orienting its body diode in the same direction as the Schottky diode (shown in **Figure 4**). During normal battery operation, the body diode of the MOSFET is forward-biased and conducts for a very short time until the MOSFET turns on when the gate voltage pulls below the source. When the battery polarity is reversed, the gate-source voltage swings positive and the MOSFET turns off, protecting the downstream circuits from a negative voltage.

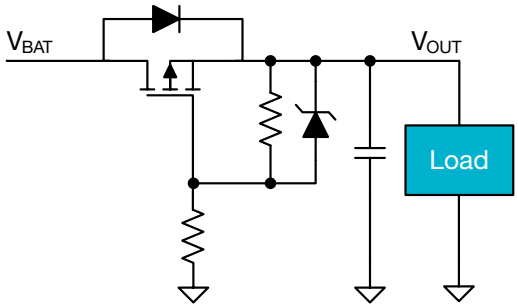

Figure 4: Reverse input protection using a P-channel MOSFET.

An alternate method of reverse battery protection is using an N-channel MOSFET on the low side, such as on the ground return path. The operating principle is similar to the P-channel MOSFET in **Figure 4**. During normal operation, the body diode of the MOSFET is forward-biased and conducts until the MOSFET turns on. The MOSFET turns on quickly as the battery input charges the gate through the current-limiting resistors. The MOSFET turns off during static reverse battery or dynamic reverse battery conditions after the battery input starts to swing negative, as the gate-source voltage starts to go below MOSFET threshold voltage (Vth) and swings negative.

Reverse Polarity Protection vs. Reverse Current Blocking

Reverse polarity protection, also known as reverse hookup protection, prevents damage to a load from a negative voltage at the input during a reverse-connected battery or an inductive load disconnect from a battery. Reverse polarity protection does not necessarily block reverse current flowing into the battery from the load or downstream DC/DC converters.

In many systems, large holdup capacitors provide sufficient backup power during a short interruption of the supply line or a shorted battery input so that the subsystem can function uninterrupted or perform maintenance housekeeping tasks such as memory dump before turning off. Reverse current blocking prevents reverse current from flowing back into the battery from the load, and allows holdup capacitors to provide additional backup time for the subsystem to function during various dynamic reverse battery conditions or short interruptions.

The reverse input protection shown in **Figure 4** or **Figure 5** below does not block reverse current from flowing back into the battery, since the MOSFETs are on. In **Figure 6** on the following page, the input power goes from a positive to negative value. The P-channel MOSFET remains on for a short duration of time, since

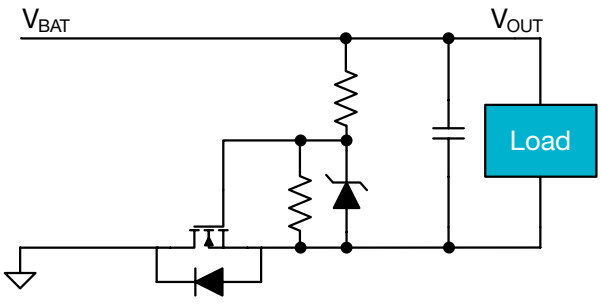

Figure 5: Reverse input protection using an N-channel MOSFET.

its V_{GS} is still transitioning from a negative to positive value. This results in a temporary negative voltage spike on the output. In contrast, the Schottky diode constantly blocks reverse current from flowing into the supply and inherently provides both reverse polarity protection and reverse current blocking.

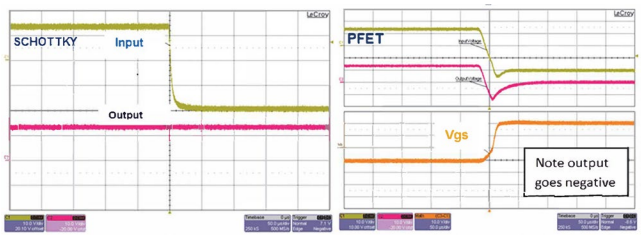

Figure 6: Schottky diode vs. a P-channel MOSFET.

Drawbacks of using discrete MOSFETs include:

- **Lack of reverse current blocking.** MOSFETS are not suitable for applications with a large holdup capacitance. Reverse current blocking is possible with an external comparator, but will increase cost and space.
- **Power dissipation at low input voltages.** The P-channel MOSFET's drain-to-source on-resistance ($R_{DS(on)}$) increases drastically as the input voltage reduces because the maximum gate-to-source voltage (V_{GS}) is limited by the input voltage range. Power dissipation due to increased $R_{DS(on)}$ reduces efficiency in applications with a wide input voltage range. For example, a 12-V automotive battery voltage can vary from 3.5 V during cold-crank conditions to 28 V during a jump-start scenario.
- **P-channel MOSFET size and cost.** For higher-power loads, the size and cost of a P-channel MOSFET can increase considerably.
- **Low-side N-channel MOSFET.** All systems may not be able to tolerate a jump in the system ground voltage during turn on and turn off or load current transients.

Reverse Polarity Protection Using TI Power Switches

TI's ideal diode controllers and eFuse devices have integrated reverse polarity protection, reverse current blocking capability and low forward conduction loss. This enables them to provide more efficient reverse polarity protection than a Schottky diode or discrete MOSFET solution.

Reverse Protection Using Ideal Diode Controllers

An ideal diode controller drives an external N-channel MOSFET to work as an ideal diode with a very low forward voltage drop and no reverse current. **Figure 7** shows reverse polarity protection using TI's LM74700-Q1 controller.

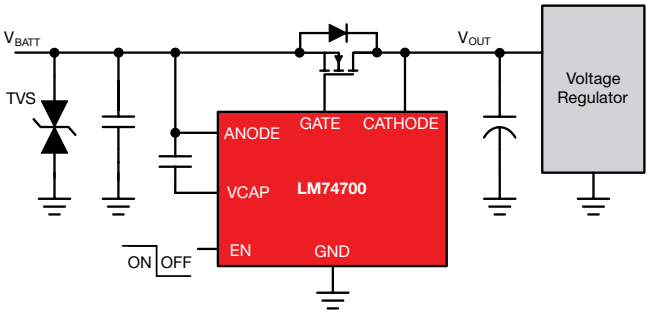

Figure 7: Reverse polarity protection using the LM74700-Q1.

The power MOSFET is connected in such way that its body diode blocks reverse current when the MOSFET turns off. The forward voltage drop and power dissipation are reduced significantly as the MOSFET turns on during forward conduction. Ideal diode controllers sense the reverse current flowing through the MOSFET and turn it off very quickly, allowing the body diode to block reverse current.

An ideal diode controller offers a low regulated forward voltage, true reverse current blocking, fast reverse current response, and very low shutdown current and operating quiescent current.

These features enable the ideal diode controller to emulate an ideal diode in reverse polarity protection applications. **Figure 8** on the following page shows the response of the LM74700-Q1 to a reverse-connected input supply. The LM74700-Q1 blocks reverse current completely by turning the MOSFET off when the input supply lines are connected in reverse polarity. The output does not dip negative, but instead remains at 0 V. The load is isolated from the reverse input supply connection and protected from damage caused by reverse current.

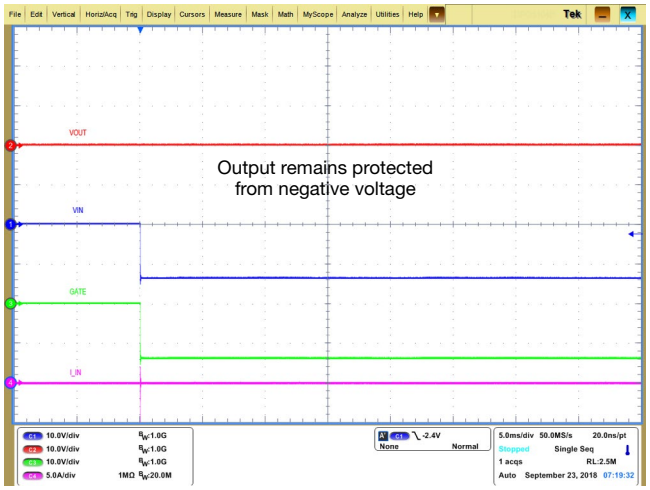

Figure 8: The LM74700-Q1's response to reverse input polarity.

Dynamic Reverse Polarity

Automotive subsystems powered through a car battery are subjected to various electrical transients on their power-supply lines and are expected to function without interruption. One such transient condition is the dynamic reverse polarity specified in International Organization for Standardization (ISO) 7637-2 pulse 1, where a negative transient voltage as low as –150 V is applied to the 12-V battery supply line with 10-Ω generator impedance for 2 ms.

Figure 9 shows the response of the LM74700-Q1 to dynamic reverse polarity applied at its input. Before the test pulse is applied, the MOSFET is on and allows the load current to pass through. When applying the ISO 7637-2 test pulse 1 at the battery input, the load current starts to reverse quickly and tries to pull the output voltage negative. The LM74700-Q1 detects the reverse current and turns the MOSFET off within 0.75 µs to block reverse current and prevent the output from going negative. It also prevents discharge of the bulk holdup capacitors.

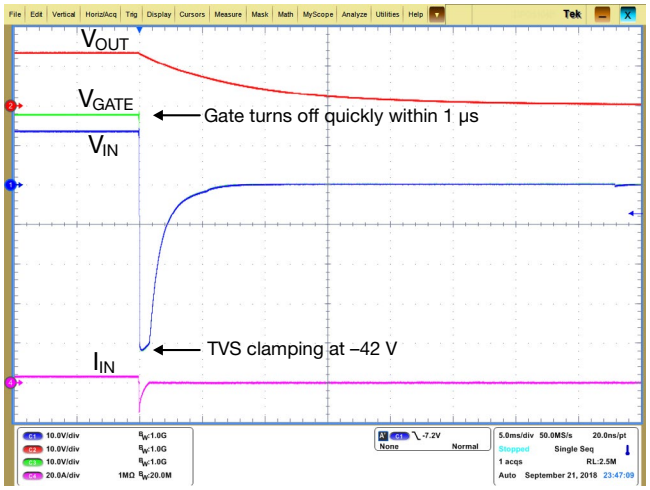

Figure 9: The LM74700-Q1's response to ISO 7637-2 pulse 1.

Reverse Polarity Protection Using eFuses

The TPS2660 family of eFuse devices feature fully integrated reverse input supply protection without the need for additional components. These devices can withstand a -60-V reverse voltage without damage.

Figure 10 compares the reverse input polarity protection functionality of the TPS2660 eFuse against an external Schottky diode. An integrated MOSFET replaces an external blocking diode, providing a fully integrated reverse current blocking and reverse polarity solution.

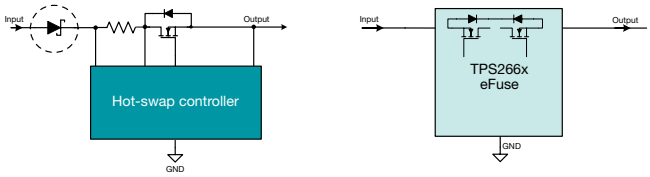

Figure 10: Integrated reverse polarity protection using the TPS2660 eFuse.

Figure 11 shows the response of the TPS2660 eFuse to a reverse input supply (miswiring) of –57 V. The TPS2660 blocks reverse current completely by turning the MOSFET off in the event of a reverse input supply and prevents damage to downstream circuits and components.

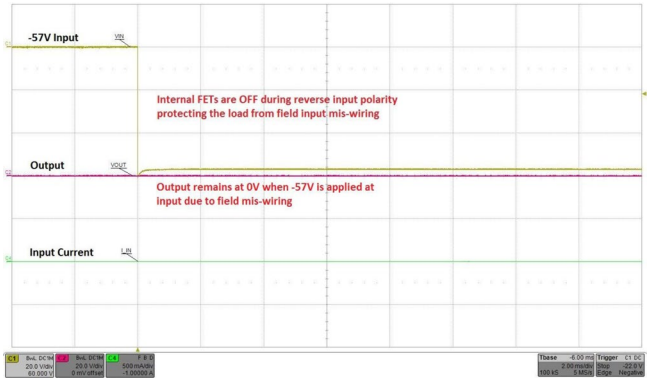

Figure 11: TPS2660 eFuse response to a reverse input supply.

Dynamic Reverse Polarity

Industrial subsystems powered through an external field power supply are subjected to various electrical transients on their power-supply lines and are expected to function without interruption. Positive and negative surges specified by International Electrotechnical Commission (IEC) 61000-4-5 are applied at the input, where a negative transient voltage as low as –500 V is applied for 20 µs.

Figure 12 shows the response of the TPS2660 eFuse to a –500-V negative surge on the input. The output remains protected from the negative surge and the output turns on after the surge event is over.

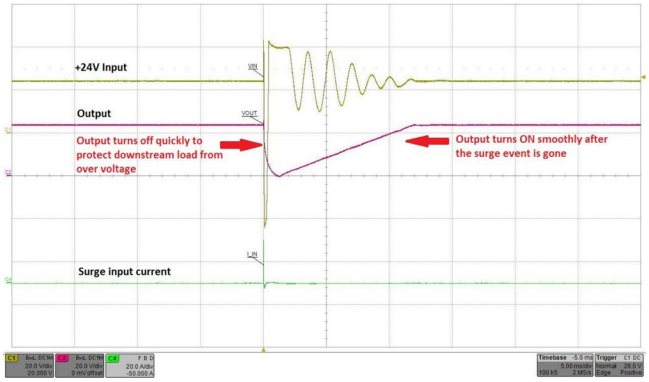

***Figure 12:** TPS2660 eFuse response to –500-V surge at input.*

The benefits of using ideal diode controllers and eFuse devices for reverse polarity protection include:

- **Improved efficiency.** A low forward conduction loss leads to efficiency improvements.
- Integrated reverse polarity and reverse current blocking.
- **Reduced system cost and space.** Thermal management is simpler compared to a Schottky diode.

Conclusion

Using an N-Channel MOSFET-based ideal diode controller and fully integrated eFuse solutions provides robust reverse polarity and reverse current blocking with high power dissipation, thermal performance and space savings against discrete diodes and FET implementations.

Chapter 6 Overvoltage Protection

Authors: Matthew Xiong and Praveen Durga Gorrela

Abstract

Every electronic component is rated for a particular voltage level. The presence of any voltage higher than the rated maximum voltage can cause damage to a component or degrade its performance, which could lead to system failures. Therefore, it's important to prevent components from exposure to these higher voltages by adding overvoltage protection, which will isolate any overvoltage condition from downstream devices or clamp the output to a safe level.

Common Sources of Overvoltages

Overvoltages applied to electronic components can exceed avalanche breakdown voltage of internal device structures, resulting in excessive current flow. This excessive current flow can severely weaken, damage or overheat the entire device.

Increased demand for lower power consumption and cost drove technological improvements, such that many semiconductor components now operate at lower voltage ranges. Lowering the voltage of individual components decreases overall system power consumption, which increases battery life, improves efficiency and indirectly lowers cost (since the system uses less energy). Cost is also directly lowered, since lower-voltage components are cheaper to manufacture than their higher-voltage counterparts. However, with improved power efficiency and lower costs comes a side effect—increased component sensitivity to overvoltage conditions.

There are two types of overvoltages: transient overvoltages and continuous overvoltages. Transient overvoltages are commonly caused by electrostatic discharge (ESD), electrical fast transients, voltage ringing from hot-plug events, or inductive switching surges from nearby power supplies or converters. Transient overvoltages can approach kilovolts but typically have short durations; for example, ESD events may last from 60–100 ns and surge events can last about 20 µs. Telecommunications equipment may be exposed to lightning surges—another form of transient overvoltage that can be orders of magnitude higher than regular surge pulses. These systems require special protection requirements, since both the intensity and duration of lightning transients are more severe.

Continuous overvoltages are overvoltage conditions that remain present for long periods and stress systems indefinitely. Continuous overvoltages can be caused by the failure or miswiring of upstream power supplies, voltage regulators, converters or the insertion of noncompliant adapters into a system.

Common Overvoltage Protection Methods

There are several different overvoltage protection topologies for the two types of overvoltages.

Protection Against Transient Overvoltage Conditions

Since transient overvoltage conditions have short durations, the response time of any solution must be fast enough to minimize the exposure of downstream circuitry. ESD diodes, transient voltage suppressors (TVSs) and Zener diodes can steer transient overcurrents to the ground plane and clamp transient overvoltages so that downstream loads remain within the safe operating area.

Metal-oxide varistors (MOVs) can also clamp transient overvoltages, and absorb significantly more energy than ESD diodes, TVSs and Zener diodes, but with a penalty of slower response time. While TVSs can respond in nanoseconds, MOVs will respond in microseconds. As a result, ESD diodes and TVSs are used for low-voltage circuitry, while MOVs offer better protection for AC mains or high-voltage DC stages.

For more information regarding ESD scenarios, causes and solutions, see the TI training, "ESD Essentials."

Protection Against Continuous Overvoltage Conditions

While transient overvoltage protection solutions can respond quickly to transient overvoltages, they are ill-suited for continuous overvoltage conditions, which characteristically last longer. After all, MOVs, ESD diodes, TVSs and Zeners can only clamp and dissipate a certain amount of energy before failing. To protect against steady-state or continuous overvoltages, there are two common methods: overvoltage lockout and overvoltage clamping.

Overvoltage Lockout

Overvoltage lockout (also known as output voltage cutoff) is the simpler and more common solution of the two. TI power switches with an overvoltage lockout feature generally have an OVLO pin that monitors the input voltage rail through a configurable resistor divider. Once the voltage at the OVLO pin increases beyond a certain threshold, the internal comparator turns the pass field-effect transistors off. **Figure 1** is a typical block diagram of an overvoltage lockout circuit implementation.

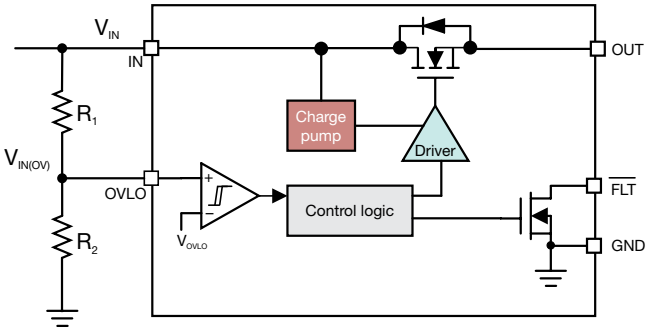

Figure 1: Overvoltage lockout block diagram.

For robust protection, faster response time is critical in continuous overvoltage protection to ensure the downstream loads are not exposed to the overvoltage condition for a long duration. For example, the TPS2595 eFuse responds to overvoltage condition within 3 µsec as shown in **Table 1** below.

Overvoltage lockout power switches will remain off as long as the input voltage exceeds the set overvoltage lockout threshold. Once the input voltage falls below the threshold, the device turns back on.

Figure 2 shows the overvoltage lockout operation of the TPS259573. The overvoltage lockout set point is 17 V. An overvoltage of 18 V is applied on a 12-V nominal operating voltage. Once the input voltage rises above the set threshold, the internal field-effect transistor (FET) turns off and the output starts discharging. To report the overvoltage event, the fault signal ($\overline{FLT}$) is asserted immediately after the pass FET turns off.

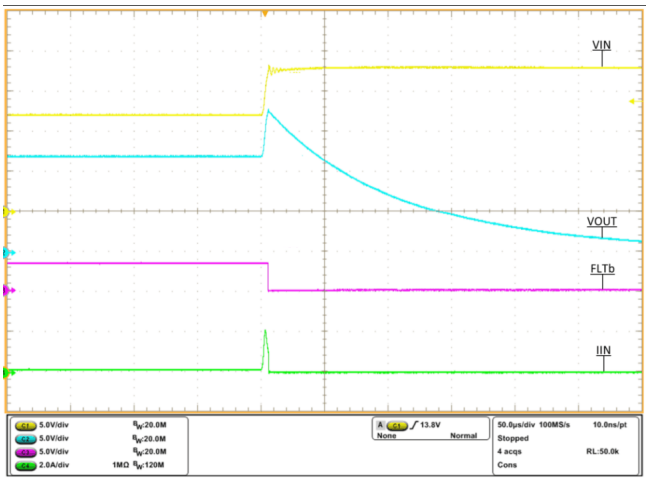

Figure 2: TPS259573 overvoltage lockout response.

Overvoltage Clamp

Overvoltage clamp is another way of responding when temporary overvoltage conditions occur in the system. During temporary overvoltage events, it is best not to interrupt the downstream load and clamp the system voltage at a safe level, rather than cutting off the system voltage. Temporary overvoltage may occur because of events such as:

- The load of an upstream, low-switching-frequency DC/DC converter drops suddenly.
- An upstream DC/DC converter operates in pulse-frequency modulation at light loads.
- The load is being powered by an unregulated, secondary winding of a multioutput flyback converter.

The duration of these events can range from few tens to hundreds of microseconds.

When the input voltage exceeds a certain threshold, the internal clamp activates and limits the output voltage to the programmed level as shown in **Figure 3** on the folowing page. This ensures that the load is not exposed to the high voltages present at the input supply while still powering the load.

As long as an overvoltage condition is present on the input, the output voltage will clamp to the programmed

Parameter	Test condition	Typical value
t_{OVLO} Overvoltage lockout response time	$V_{IN(OV)} > V_{OVLO}$ to $\overline{FLT}$ ↓	3 µs

Table 1: TPS2595 overvoltage lockout response time.

level. When the input drops below the output clamp threshold, the clamp will release and normal operation continues.

During an input overvoltage condition, the gate of the FET is regulated to drop additional voltage across it and clamp the output voltage. The power dissipation in the FET is the product of the voltage drop across it and the current flowing through, as shown in **Equation 1**:

$$PD = (V_{IN} - V_{OUT}) \times I_{LOAD} \quad (1)$$

The higher the input overvoltage and load current, the higher the power dissipated in the FET to keep the output voltage constant. As a result, there could be significant heat dissipation in the internal FET and a rise in the device's junction temperature. If the overvoltage condition persists, thermal shutdown can occur and disconnect the load from the supply, at which point the device will either stay latched off or start an auto-retry cycle.

The benefit of a quick response is that it is possible to lower or eliminate the dependence on external protection devices such as TVSs and Zener diodes. As an example, the TPS259531's output clamp response time is 5 µSec.

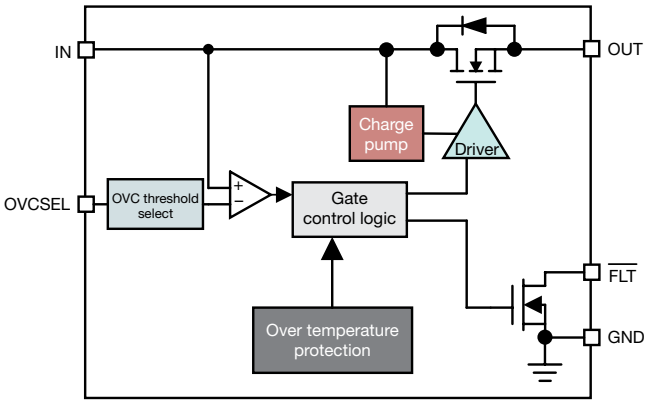

Figure 3: Overvoltage clamp block diagram.

Figure 4 shows the overvoltage clamp operation of the TPS259531, which has an overvoltage clamp set to 5.7 V. An overvoltage of 8 V is applied on a 5-V nominal operating voltage. Once the input voltage rises above the 5.7-V threshold, the gate of the internal FET is regulated to clamp the output voltage at 5.7 V.

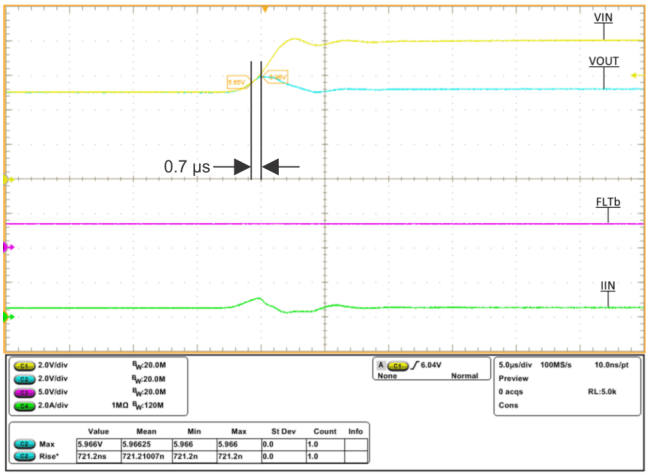

Figure 4: TPS259531 overvoltage clamp response.

Conclusion

To reliably protect power path circuits, overvoltage protection is critical. Different system use cases demand for different responses during overvoltage conditions such as overvoltage lock out or overvoltage clamping. eFuse devices with integrated overvoltage protection provide flexible, fast and compact solutions to achieve robust power path protection in your system.

Chapter 7 Thermal Shutdown

Author: Sreenath Unnikrishnan

Abstract

In power integrated circuits (ICs), thermal shutdown protects the device from failure when its junction temperature exceeds its safe limit. In power switches, fault conditions in load or thermal systems can cause high device junction temperatures.

When the junction temperature exceeds the threshold for thermal shutdown, in order to self-protect, the power switch turns itself off to reduce the power dissipation. This chapter describes the functionality of the thermal shutdown, the behavior of power switches when reaching thermal shutdown conditions, and their recovery to normal operation when the fault condition is no longer present.

Introduction

Many power ICs have a self-protection mechanism to shut down the device in the event of a high junction temperature. Junction temperatures beyond normal operation levels can occur if high power dissipation exists for a sufficient duration, or the heat dissipation from within the device to the rest of the system, or ambient air is not effective. Without a built-in protection mechanism, the junction temperature may exceed the safe limit and permanently damage the device. In most cases the protection mechanism turns off the function in the IC that caused the excessive power dissipation.

A thermal shutdown circuit typically detects that a power IC is overheating by measuring the absolute junction temperature of the hottest areas on the chip. A thermal monitoring circuit translates the junction temperature to a voltage or current level that is then compared to a reference level (the absolute threshold, or T_{ABS}) corresponding to a temperature below the safe limit for the IC. If the junction temperature is above T_{ABS}, then the device function will partially or fully shut down. The implementation of a thermal hysteresis function requires that the junction temperature fall below a specified amount (set as T_{HYS}) before normal circuit function can resume. Thermal hysteresis ensures that the device does not transition into and out of thermal shutdown at a high rate.

Thermal Shutdown in Power Switches

Thermal protection is an integral part of power switches, which themselves are used as input-protection devices or as high-side switches driving offboard loads. Monitors are usually positioned in the power transistor area where most of the power dissipates. Some multichannel switches (TI's TPS2HB08-Q1, for example) may include more than one junction monitor.

In such multichannel power switch, the protection mechanism will only turn off the field-effect transistor (FET) with a high junction temperature; the switches pertaining to the other channels retain their functionality. **Figure 1** shows an example of the thermal shutdown function.

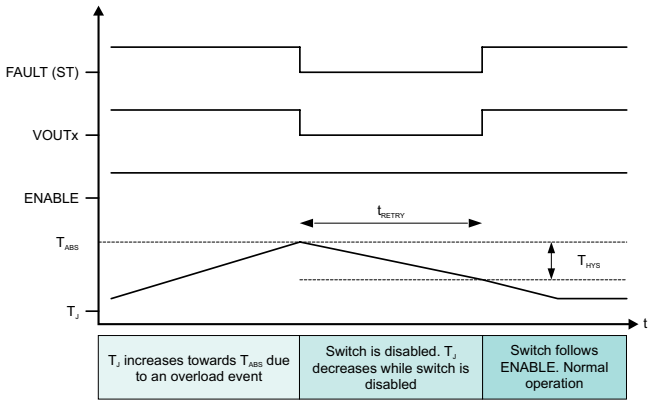

Figure 1: Thermal shutdown function example.

In **Figure 1**, the junction temperature rises due to a temporary overload event. When the junction temperature reaches T_{ABS}, the switch function is disabled. As a result, excess power dissipation ends and the junction temperature starts to fall. When the temperature has decreased by T_{HYS}, the device can retry turning back on. Normal operation will resume, since the temporary overload is gone.

In cases where an overload condition persists, the device will experience thermal shutdown again. Some power switches implement a minimum duration for t_{RETRY} to reduce the average junction temperature.

Higher-than-normal power dissipation can occur in power ICs when the load current exceeds normal operational levels or when a high voltage drop appears

across the FET with high inrush current. The heat generated on the IC must transfer onto the printed circuit board and then to ambient air.

Equation 1 calculates the junction temperature (T_J) resulting from constant power dissipation (P_{diss}) on an IC:

$$T_J = T_A + P_{diss} \times \Theta_{JA} \quad (1)$$

where Θ_{JA} is the thermal resistance and T_A is ambient temperature.

Equation 2 calculates the power dissipation in the FET as:

$$P_{diss} = V_{ds} \times I_{ds} \quad (2)$$

where V_{ds} and I_{ds} are the FET's drain-to-source voltage and current, respectively.

The V_{ds} and the P_{diss} can be high when supporting high inrush currents; for example, during initial turn-on into a capacitive load. When delivering a steady-state load current (I_L), **Equation 3** expresses the power dissipated in the FET with the on-resistance (R_{on}) as:

$$P_{diss} = I_L^2 \times R_{on} \quad (3)$$

Design Considerations

You will need to design the maximum device power dissipation and thermal system parameters to keep the junction temperature below the absolute maximum temperature specification in the data sheet. The thermal shutdown threshold is designed to protect within the safe limit of the IC. If the power dissipation in the IC is higher than the expected system design target—due to a load fault, or if the thermal system is degraded due to a defect—the actual junction temperature could exceed the worst-case maximum design target and thermal shutdown protection will kick in.

Overload Protection

In power switches with current-limiting overload protection, the current is controlled to the current-limit threshold level. While the current is limited to that safe level, the power dissipation can be quite high during current limiting. Thermal shutdown occurs after a duration of time under these conditions depending on the thermal impedance. The current from the supply to the load will reduce.

The thermal shutdown feature in power switches supports an additional function—that of a fuse limiting the duration of an overload fault event. In the case of limiting the duration of an overload event, thermal shutdown may also potentially help avoid damage to upstream and downstream components and connectors. The combination of overload and built-in overtemperature protection ensures safe operating area protection for integrated power switches.

Conclusion

Many TI power switches that are at risk for exposure to fault conditions include a thermal shutdown function as a protection mechanism. You must be aware of the absolute maximum temperature specification in the data sheet and design the thermal system and the worst-case IC power dissipation to keep the junction temperature below this level.

Chapter 8: Current Monitoring

Authors: Abhinay Patil and Paul Kundmueller

Abstract

Many applications require output load current monitoring. Actual needs vary from system to system, ranging from simple parameter data logging and reporting to diagnostics and control. Multiple solutions are available to monitor load current, each having their own advantages and challenges. Choosing the right solution requires an assessment of system requirements and trade-offs between factors that include performance, cost, area and complexity.

This chapter addresses the output load current-monitoring function available in integrated power switches rather than stand-alone current-monitoring devices/solutions.

Current-Monitoring System Applications

The need for current monitoring in a whole system or a subsystem varies. Here are a few examples of applications that benefit from current monitoring:

- **System power reporting.** The most common reason to monitor basic electrical parameters such as voltage and current is for reporting and data-logging purposes. For example, in a data center with thousands of servers, monitoring the power consumption of individual server nodes helps redistribute the workload uniformly, thereby ensuring that the data center is running at its most optimal efficiency.

  System current monitoring is applicable in factory automation as well, especially in remote input/output (I/O) modules that use M16 and M18 cables to distribute power. Having a way to monitor the current of the ports and the system is important in order to not exceed cable current ratings. Otherwise, damage to the system could occur.

- **Dynamic/real-time control.** Some systems use load current measurements to implement real-time control. In such applications, apart from accuracy, it's equally important to have high bandwidth and low latency. For example, in servers, the power monitor can check if the overall system current is rapidly rising beyond a certain threshold and send a signal to the central or graphics processing unit to quickly throttle back its performance, thus preventing the entire system from going into shutdown. Typically, this signal needs to be very fast (<10-µs latency).

- **Diagnostics.** It's possible to identify faults in a system by looking at the current consumption of different sections and comparing this consumption to a threshold based on the expected value during normal operation. Any section consuming distinguishably high or low current could indicate a fault. For example, in factory automation, wiring is a must for power and signal paths because of its distribution across large factory environments. Digital output modules commonly specify the maximum wire length as 100 m. With wires of this length, the wire could accidentally break. Wire breaks are detectable by monitoring the current in the loop, either through an analog current monitor or an open-load detection feature.

  Similarly, automotive driver assistance system sensors and antennas are often connected via long cables due to their location in automobiles. Having the ability to quickly alert and diagnose a wire break is critical to ensure safe operation.

- **Predictive maintenance.** Some systems (servers, line cards and programmable logic controller I/O modules) monitor an electrical parameter's profile and look for changes in the time/frequency signature to predict impending faults. This information helps system operators schedule maintenance before a fault occurs. Predictive maintenance can help avoid unplanned downtime, a loss of productivity and even safety hazards. In these applications, the current monitor needs to have good resolution to capture subtle changes in the current, as well as sufficient bandwidth to capture any transients or instantaneous changes in the current.

Current-Monitoring Solutions

There are different types of current monitoring solutions based on the sense element used (external/internal) and the type of output (analog/digital).

External Sense Elements

Hot-swap controllers use an external resistor (R_{SENSE}) in the power path to sense the load current. The voltage drop across R_{SENSE} is amplified by an internal high-side current-sense amplifier and then either driven as an analog current output to an external analog-to-digital converter (ADC), or fed into an internal ADC. **Figure 1** shows current monitoring with an external sense element.

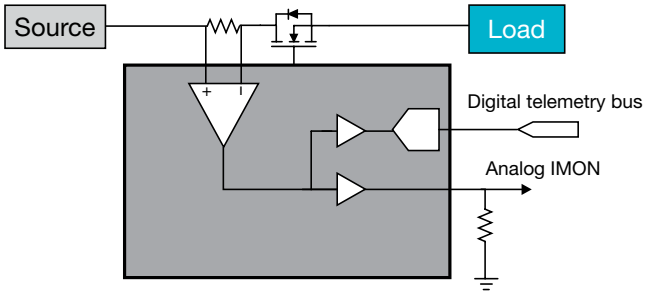

Figure 1: Block diagram of current monitoring with an external sense element.

You'll need to select the R_{SENSE} value based on the maximum system load current (I_{MAX}) and the maximum sense voltage ($V_{SNS,MAX}$) specified in the hot-swap controller data sheet. **Equation 1** calculates R_{SENSE} based on these specifications:

$$R_{SENSE} = V_{SNS,MAX}/I_{MAX} \quad (1)$$

Current-sense accuracy is a function of the R_{SENSE} tolerance and the current-sense amplifier gain/offset error, along with their associated temperature coefficients. It is possible to minimize the error by performing a system-level calibration with an accurate load current reference, but it's not possible to eliminate the temperature drift. Choose sense resistors with a low temperature coefficient, such as thin film or bulk metal.

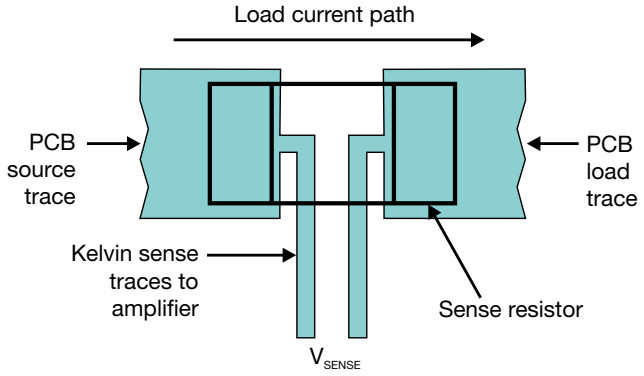

Figure 2: Kelvin sense layout for an external current-sense resistor.

The other sources of error stem mainly from an improper printed circuit board (PCB) layout around the current-sense path. For best results, use a Kelvin sense connection, as shown in **Figure 2**.

The main advantages of an external current-sense resistor-based solution are the ability to achieve high linearity and accuracy when sensing current over a wide range. The disadvantages of an external current sense-resistor are increased cost and board space, along with the additional voltage drop, power dissipation and self-heating in the sense resistor.

Internal Sense Elements

eFuses and smart high-side switches use internal current-sense circuits and don't need external components in the power path for current sensing. Current sensing is usually implemented using some form of current mirror circuit, which drives a fraction of the main power field-effect transistor (FET) current on the current-monitor output pin. **Figure 3** shows a block diagram of the TPS2663 eFuse, which uses an integrated current-sense circuit to provide an analog current-monitor signal output on a dedicated pin (I_{MON}).

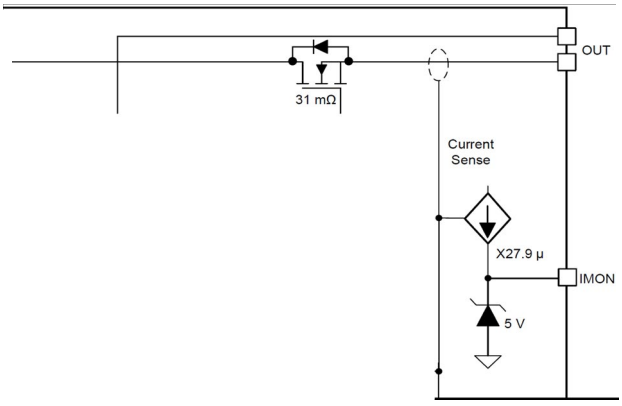

Figure 3: TPS2663 internal block diagram.

Some devices, like the TPS2596 eFuse, provide the analog current-monitor output on a dual-purpose pin (I_{LIM}), which is primarily used to set the adjustable current limit, as shown in **Figure 4** on the following page. The voltage drop across the current-limiting resistor (R_{ILM}) provides a measure of the output load current.

The accuracy of an integrated current-sense solution is governed by the mismatches and other errors

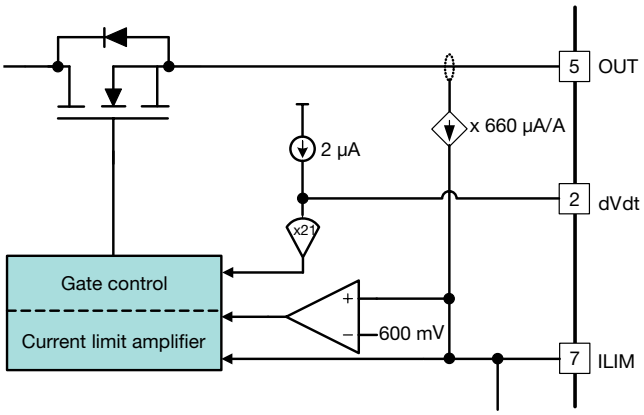

Figure 4: TPS2596 internal block diagram.

associated with the internal circuits, which are factory-trimmed to provide high accuracy across a range of load currents and operating temperatures. Some devices, such as the TPS25982, also employ circuit techniques like a copper-stabilizer amplifier to minimize offset error and drift over time and temperature. The use of factory trimming and other circuit techniques minimizes the need for any additional calibration in the final system to achieve highly accurate current measurements.

An internal current-sensing scheme offers significant advantages over external sense resistor-based solutions by eliminating the added cost, size, voltage drop and heating associated with a sense resistor.

Analog Output Types

The output is an analog current signal proportional to the load current. **Equation 2** and **Equation 3** express the relationship between the output load current (I_{LOAD}) and the current-monitor signal (I_{MON}):

$$I_{MON} = I_{LOAD} \times G_{MON} \quad (2)$$

where G_{MON} is the current-monitoring gain.

Alternately:

$$I_{MON} = I_{LOAD}/K_{MON} \quad (3)$$

where K_{MON} is the current-sense ratio.

Connecting a resistor (R_{MON}) from the current output pin (I_{MON}) to GND converts the sense current to a voltage, as expressed in **Equation 4**:

$$V_{MON} = I_{MON} \times R_{MON} \quad (4)$$

To ensure maximum use of the measurement's dynamic range, select R_{MON} such that V_{MON} at the full-scale load current matches the full-scale voltage input of the ADC or any other downstream circuit it feeds into.

It is possible to improve sensitivity/resolution at low currents by using a larger R_{MON}, as it increases the available signal voltage relative to the noise. However, a larger R_{MON} also increases the voltage swing at higher currents and might lead to clipping/saturation near full scale due to headroom limitations on the I_{MON} pin driver circuit.

High-/Low-Current Digital Indication

Many solutions provide a digital output flag pin that is asserted/deasserted by comparing the load current with a certain threshold.

Smart high-side switches like the TPS27S100A provide an open-drain status output pin to flag the system that an open load has occurred. One important aspect to consider for open or light-load detection is the normal operating range of the output current. To avoid false fault reporting, the nominal operating range needs to be higher than the maximum open-load detection threshold listed in the data sheet. **Table 1** on the following page is a reference to this specification in the TPS27S100A data sheet.

Additionally, the integrated analog current-monitor output (I_{MON}) of the TPS25940, TPS25942, TPS25944, TPS1663 and TPS2663 eFuses can be repurposed together with the integrated comparator (PG, PGTH) to provide an open or light-load detection function, as shown in **Figure 5**.

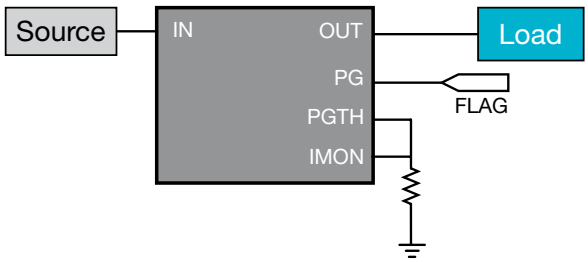

Figure 5: Open or light-load detection using the IMON, PGTH and PG pins.

Electrical characteristics

5 V < V_{IN} < 40 V; –40°C < T_J < 150°C unless otherwise specified

Parameter		Test Conditions	MIN	TYP	MAX	Unit
Diagnostics						
$V_{(ol,off)}$	Open-load detection threshold in off-state	V_{EN} = 0 V, when $V_{IN} - V_{OUT} < V_{(ol,off)}$, duration longer than $t_{d(ol,off)}$. Open load detected.	1.4	1.8	2.6	V
$I_{(ol,off)}$	Off-state output sink current with open load	V_{EN} = 0 V, $V_{IN} = V_{OUT}$ = 24 V, T_J = 125°C.	–150			µA
$t_{d(ol,off)}$	Open-load detection-threshold deglitch time in off state	V_{EN} = 0 V, when $V_{IN} - V_{OUT} < V_{ol,\ off}$, duration longer than $t_{ol,off}$. Open load detected.		600		µs
$I_{(ol,on)}$	Open-load detection threshold in on state	V_{EN} = 5 V, when $I_{OUT} < I_{(ol,on)}$, duration longer than $t_{d(ol,on)}$. Open load detected. Version A only.	2	6	10	mA
$t_{d(ol,on)}$	Open-load detection-threshold deglitch time in on-state	V_{EN} = 5 V, when $I_{OUT} < I_{(ol,on)}$, duration longer than $t_{d(ol,on)}$. Open load detected.		700		µs

Table 1: TPS27S100A electrical characteristics.

It's also possible to achieve open or light-load detection with other devices that have an analog current-monitor output by adding an external comparator, as shown in **Figure 6**.

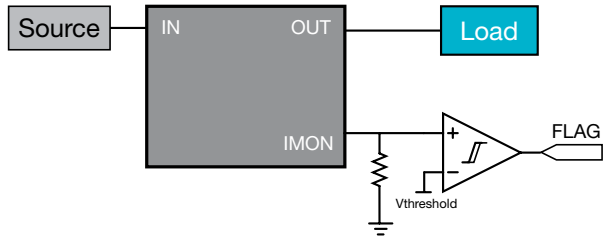

Figure 6: Open or light-load detection using an external comparator.

Digital Telemetry Output Types

Another approach to digital current monitoring is telemetry. This digital approach to current monitoring involves sensing an analog current and converting it to a digital signal through an integrated ADC, which is then stored in internal digital registers. A digital bus such as an I^2C or Serial Peripheral Interface bus then reports the data back to a central processing unit. The highlighted portions in **Figure 7**'s block diagram demonstrate both the sensing and analog-to-digital conversion of the current.

The data format of the stored analog current information can be defined by industry standards such as PMBus® or customized depending on the device. PMBus is a popular specification that helps standardize the reading and writing of telemetry data through I^2C. PMBus also defines the data format for storing power and voltage data.

Figure 7: LM5066I internal block diagram.

TI offers many devices that conform to the PMBus standard. For more information on how PMBus is used in a device, see the LM5066I data sheet. Sections 8.5.4 through 8.5.8 in the data sheet thoroughly explain the data format and equations required to have the system digitally interpret analog current information through I^2C.

Key Performance Specifications Relevant to Current Monitoring

Here are some of the key specifications you should consider while choosing a current-monitoring solution for your application:

- Dynamic range refers to the range of load currents that the current-monitor circuit can sense and report accurately (see **Figure 8**). Signals outside the supported dynamic range would have a significantly higher error in measurement, or may not be sensed at all. The dynamic range is usually limited at the upper end by the headroom of the current-monitor amplifier circuit and on the lower end by noise and offset errors. There's usually a trade-off between accuracy at the lower or upper ends of the range. Using a higher gain or larger sense resistor in the circuit improves the performance near zero scale by increasing the signal relative to the offset and noise, but leads to clipping and/or saturation of the output near full scale. Conversely, using a lower gain or smaller sense resistor can ensure better linearity and accuracy near full scale, but leads to poor measurement accuracy near zero scale because the measurement is dominated by offset and noise. Some current-monitoring devices use techniques such as chopper-stabilized amplifiers, auto-zero amplifiers and multipoint trim to minimize offset errors and improve performance at the lower end of the range.

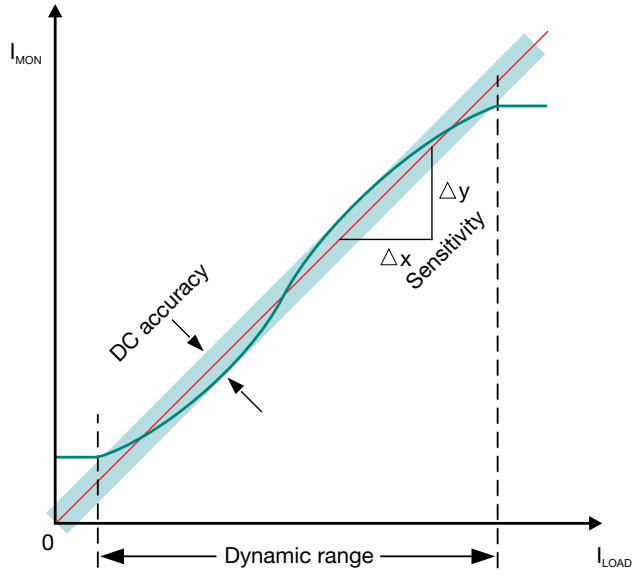

Figure 8. Current-monitoring performance specifications.

- DC accuracy is a measure of the absolute error/deviation of the current monitor's output from the expected value, expressed in terms of %FS or %Reading. The error is a combination of components such as gain error, offset error and nonlinearity. While gain error dominates at higher currents, offset error dominates at lower currents. For devices with an external current-sense resistor, the tolerance of the sense resistor also contributes to the overall current-monitoring error. You will need to account for the temperature coefficient of the error components to determine the accuracy of the measurement across temperatures.

- Resolution/sensitivity is a measure of the smallest change in current that the monitoring circuit can sense. There is usually a trade-off between resolution and dynamic range. For analog monitoring solutions, higher sensitivity means higher gain in the sense circuit. For digital output solutions, higher sensitivity also depends on the resolution of the ADC.

- Bandwidth is a measure of the fastest change in current that the monitoring circuit can sense. Any signals within the circuit bandwidth will be reproduced faithfully, while signals outside the bandwidth will be attenuated significantly, leading to erroneous measurements. The bandwidth also determines the circuit's ability to detect instantaneous changes in the current such as load transients, as shown in **Figure 9**.

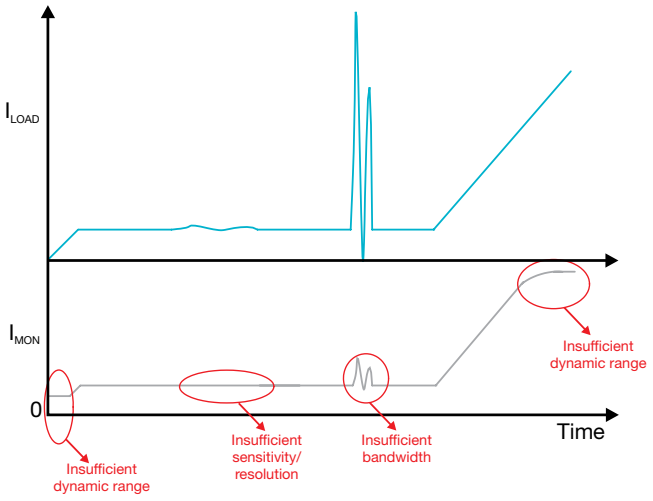

Figure 9. System impact of current-monitoring performance specifications.

- Latency is a measure of the delay between any change in the load current and the time at which that change is reflected in the measured output. Latency is usually a factor in digital current-

monitoring solutions because of the analog-to-digital conversion time and propagation delay through the digital circuits.

Conclusion

Table 2 lists the various power-switch solutions offered by Texas Instruments with current-monitoring capabilities.

In addition to providing power-path control and protection, power switches like hot-swap controllers, eFuses and smart high-side switches can provide added value by performing current monitoring as well. While many power switches are available on the market, in some cases it's possible to narrow down the choice based on its integrated current-monitoring capabilities.

	eFuse	Hot-swap controller	Smart high-side switches
Bus voltage	2.5 V–60 V	±80 V	3 V–65 V
Minimum sense current	20 mA	Depends on external sense resistor	5 mA
Maximum sense current	15 A	Depends on external sense resistor	80 A
Internal current-sense element	Yes	No	Yes
Analog current monitor	Yes	Yes	Yes
Digital current monitor	No	Yes	No
Open/light-load detection	Yes	No	Yes

Table 2: Power switches with current-monitoring capabilities.

Chapter 9: Power-Good Signal

Author: Arthur Huang

Abstract

Sensitive loads such as microprocessors or system-on-chips (SoCs) require stable input voltages to remain in regulation. If the voltage rail is experiencing large amounts of inrush current or the application of a high-capacitive load, the voltage could dip, causing the device to enter an undervoltage lockout (UVLO)/reset state or damage downstream circuitry. A power-good signal (PG) can help manage inrush current or control rail sequencing by connecting the signal to a downstream load.

A power-good signal only asserts when the voltage rail reaches a value sufficient for normal operation. Connecting a downstream enable pin or reset supervisor to the power-good signal ensures that the downstream load only operates with a stable, regulated voltage input. Power-good functionality is also a good fit in applications such as power sequencing and power multiplexing.

Common Power-Good Applications

The following sections will include implementations of a power-good signal.

Connecting to Downstream Loads

Many systems require stable, regulated power inputs for proper operation of downstream loads. If the power rail were to fall out of regulation, the load could malfunction and cause the system to fail. Some downstream loads that depend on proper voltage rails include microprocessors, field-programmable gate arrays (FPGAs), SoCs and other sensitive components.

As shown in **Figure 1**, the power-good signal can provide feedback to a microcontroller (MCU) when the voltage rail is on and stable. If the output of the power switch contains a high-capacitive load, turning the MCU on when the output voltage is stable will manage the inrush current. In other words, when the power-good signal deasserts, the MCU will remain in an off state. Since the power-good output is usually open drain, you will need a pullup resistor to bring the pin high during regulation.

If multiple supply rails need regulating, connecting the power-good signals together to a reset supervisor or

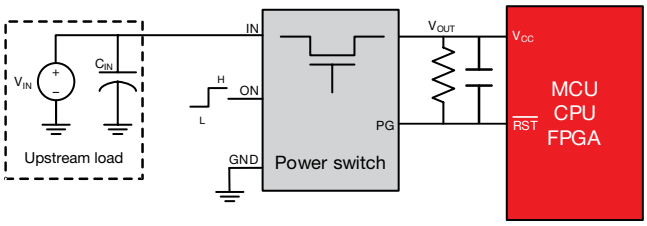

Figure 1. Power-good reset configuration.

logic integrated circuit (IC) ensures that the reset signal will assert if any voltage rail falls outside of regulation. **Figure 2** shows this configuration by connecting PG from a DC-DC converter and a power switch together.

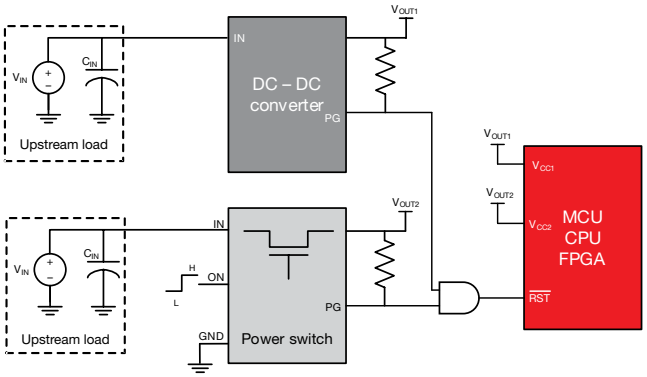

Figure 2. Multiple power-good reset configuration.

Systems that require robust input protection can use the power-good signal to reduce system stress during startup. Some devices such as hot-swap controllers monitor the stress across an external field-effect transistor (FET) so that it remains within the safe operating area (SOA). Having the downstream load turn on during startup while the FET is not fully enhanced could add additional stress on the external FET and system. Connecting the power-good signal to the enable signal of the downstream load, as shown in **Figure 3**,

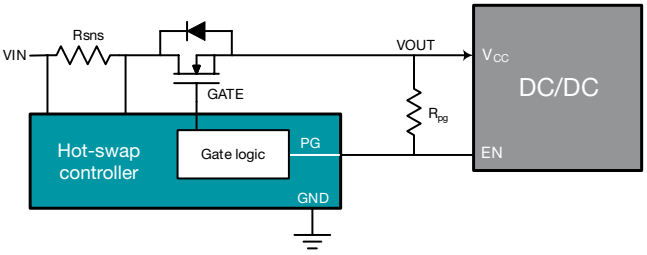

Figure 3. Hot-swap power-good configuration.

ensures that the downstream load will only turn on when the hot-swap FET is fully enhanced.

Power Sequencing

Many applications require controlled power-up and power-down sequences to properly operate subsystems and downstream components. Sequencing is critical on voltage rails that need to turn on in a specific order to ensure operational safety and reliability. Sequencing rails also helps stagger the inrush current during power-up, which reduces system stress and input voltage dip while also helping prevent potential reverse-bias conditions.

It is possible to use the power-good signal to sequence voltage rails. For example, by connecting the power-good signal of a load switch to the enable pin of a second load switch, the second load switch will only turn on after the first load switch is stable. This condition delays the turn on of the second load switch, effectively staggering the voltage rails and reducing the inrush current stress on the system. When the first switch is disabled, the power-good signal pulls the EN/ON pin low and prevents the second switch from turning on. For more information on inrush current, see Chapter 1.

Figure 4 demonstrates a power-sequencing application using power good. You can also add delay to the power-good signal to further delay the turn on/off sequencing of the rails, as shown in **Figure 5**.

Power Multiplexing

Power multiplexing is a set of electronic switches used to select and transition multiple input power rails to a single output. Systems that include backup power rails or backup batteries may require a fast switchover between supplies to minimize the output voltage dip and

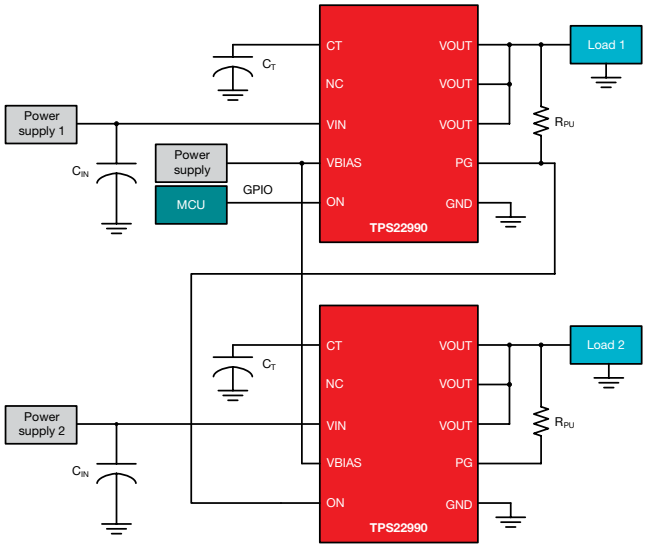

Figure 4. Power sequencing.

prevent system reset. In order to prevent reverse current conditions, you can use the power-good signal to toggle the switchover between inputs. For more information on reverse current blocking, see Chapter 4.

By connecting the power-good signal of a device like a load switch to the gate of an external P-channel FET (PFET), the power-good signal enables seamless switchover between two power rails and eliminates the need for a discrete timing circuit.

As **Figure 6** on the following page illustrates, when the TPS22990 power-good pin pulls high, the gate of the PFET also pulls high, and this turns off the FET. When the power-good pin asserts low, the gate of the PFET pulls low and turns on the PFET, allowing current to flow from IN2.

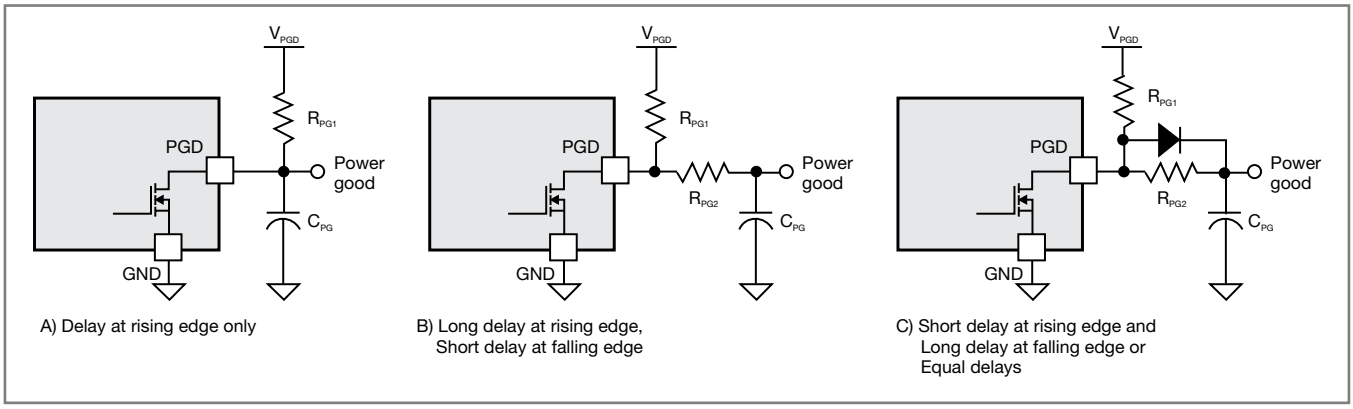

Figure 5. Adding delay to the power-good output pin.

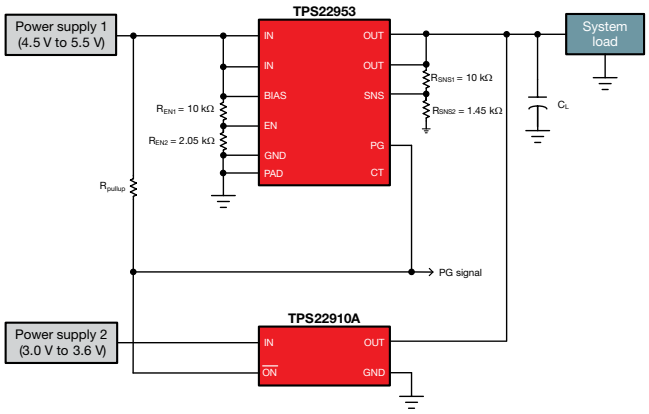

Figure 6. Make-before-break power multiplexer schematic.

Power-Good Signal Architecture

Power Good Signals can be implemented either discretely or with an integrated implementation. The following section will showcase both solutions

Discrete Implementation

Figure 7 shows a discrete implementation of a power-good signal. In this implementation, the assertion of the power-good signal depends on the output voltage. Power good asserts once the output voltage reaches a certain threshold to bias the discrete FETs. In this example, when V_{OUT} is low, the voltage at the gate of FET Q2 pulls low, turning it off. This leaves the gate of Q3 pulled up to V+, turning it on and pulling the power-good signal to ground. When V_{OUT} reaches a certain level specified by the resistor divider, the gate on Q2 asserts high, which pulls the gate of Q3 low, turning Q3 off. This pulls the power-good output high to V_{OUT}, asserting the power-good signal.

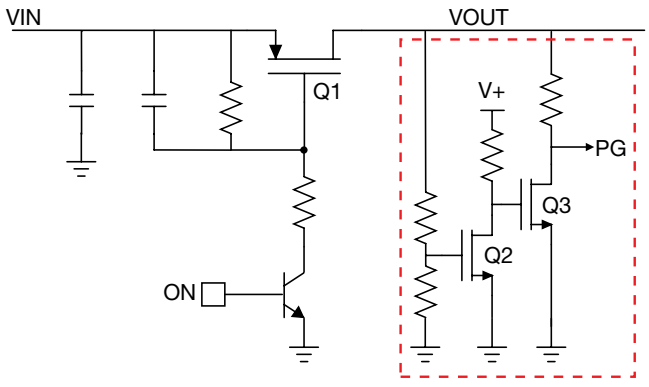

Figure 7. Discrete power-good feature.

One limitation of this circuitry is that the power-good signal is only influenced by V_{OUT}. When V_{OUT} is present, the power-good signal will assert even if the V_{OUT} rail fluctuates. For example, if V_{OUT} is 12-V nominal but a high-capacitance load connects to the output, the input voltage could dip down to 10 V. The power-good signal will remain asserted as long as the left FET is biased.

Integrated Implementations

Depending on the device, there are a few methods for implementing power good using an integrated IC. The first method, V_{OUT}/V_{IN} power good is similar to the discrete power-good solution. As **Figure 8** illustrates, the power-good signal asserts when the output voltage reaches a certain percentage of V_{IN}. For this example, the power-good signal will assert once V_{OUT} reaches 90% of V_{IN}. This offers the simple functionality of power good using a comparison between V_{IN} and V_{OUT}.

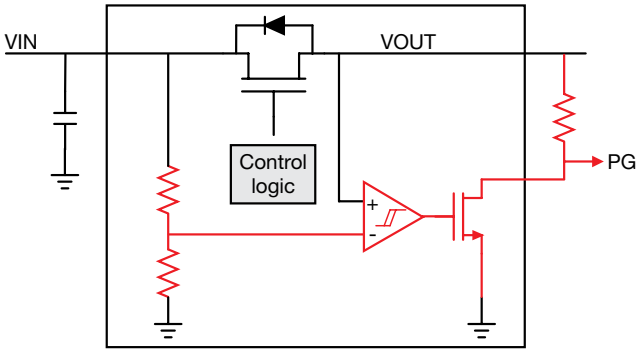

Figure 8. Integrated V_{OUT}/V_{IN} power good.

However, V_{OUT}/V_{IN} power good comes with a limitation when turning on into a high-capacitive load. Since the power-good signal only depends on the V_{IN}-to-V_{OUT} ratio, power good will assert even if the voltage fluctuates. If the downstream load is pulling a lot of current due to a high-capacitive load, there could potentially be a large inrush current event during startup. As mentioned in Chapter 1, charging the load capacitance could cause the input voltage to dip. Since power good ultimately depends on the V_{IN}/V_{OUT} ratio, the power-good signal will still assert if V_{IN} and V_{OUT} dip.

V_{OUT}/V_{IN} power good also has another limitation at lower voltage applications. At lower V_{IN}, the power-good signal could potentially not assert due to the on-state resistance.

For example, looking at **Equation 1**, if the device is operating at 0.5 V at 1 A with an on-state resistance of 100 mΩ, the output voltage will be 0.4 V, which is outside the range of 90% V_{IN}. Therefore, the power-good signal might not assert.

$$V_{OUT} = V_{IN} - I_{LOAD} \times R_{ON} \quad (1)$$

An effective way to implement power good is by measuring the on-resistance of the pass FET. Sometimes called "V_{GS} good," this method asserts power good when the V_{GS} voltage rises to the point where the FET on-resistance approaches its nominal value. A delay is also introduced into the circuitry to prevent any transient or inrush event from accidentally asserting V_{GS} good. After the voltage on the EN/ON pin exceeds the VIH level of the device, the device will measure the on-resistance and wait a certain time (t_{BLANK}) before asserting high. If the voltage on the EN/ON pin falls below V_{IL} or the V_{GS} of the FET falls below the threshold, power good deasserts. **Figure 9** shows a common implementation of V_{GS} power good.

V_{GS} power good offers the advantage of working at lower V_{IN} values since the on-state resistance fluctuations are small, and prevents transient or inrush conditions from causing an accidental assertion of the power-good pin. One limitation with V_{GS} good involves V_{BIAS} and V_{ON} during startup. If the V_{BIAS} and V_{ON} pins are high, and V_{IN} starts to ramp, power good will assert since the V_{GS} threshold remains high. Since V_{BIAS} powers the voltage supplied to the gate, the V_{GS} threshold will remain high.

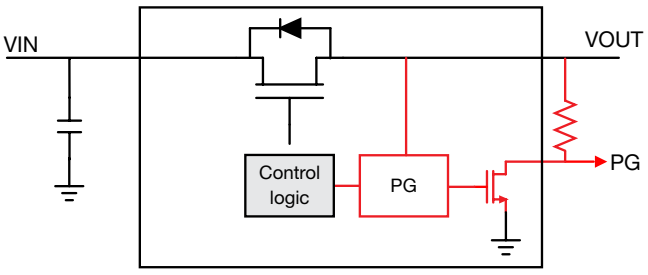

Figure 9. Integrated V_{GS} power good.

Conclusion

Sensitive applications that require stable input voltages can use a power good signal to remain within regulation. A power good signal can help manage inrush current or control rail sequencing, which is useful in applications such as power sequencing or power muxing.

Chapter 10: Safely Driving an Inductive Load

Author: Mahmoud Harmouch

Abstract

Inductive loads are relays, solenoids, electric motors and even loads connected through a long cable. Their impedance consists of both a resistance (R) and an inductance (L) in series. The R value determines the steady-state current, and the L value determines the stored magnetic energy. This stored magnetic energy in the inductor can cause system- or component-level damage if not properly dissipated.

The opening or closure of a magnetic contact in a relay or solenoid requires the storage or dissipation of magnetic energy. In the case of an electric motor, this stored energy is necessary for mechanical rotation. Inductive loads are continuously energized and de-energized for opening or closing contacts (relays and solenoids) and rotation or idle (electric motors). Disconnecting an inductive load from an energized state creates a high-voltage spike that can lead to system damage. Safely de-energizing an inductive load requires the implementation of an appropriate clamp.

Introduction

When encountering an inductive load, you must take into account the amount of energy stored, as this energy can damage components in the system if not properly managed.

There are two states to consider when driving an inductive load:

- Energizing or connecting the inductive load to a voltage source such as a battery.
- De-energizing or disconnecting the inductive load from the voltage source.

You can reach either state by using switches such as a bipolar junction transistor, a power field-effect transistor, or an integrated switch that connects one side of the switch to the voltage source (the high-side switch) or to ground (the low-side switch). In some instances, the integrated switch option may integrate a voltage clamp.

In the energizing state, the switch drives the steady-state load current and the inductance stores magnetic energy equal to half the L value and the square of the load current.

In the de-energizing state, the current decays from the steady-state value to zero; a voltage spike proportional to the current slope appears across the switch. The voltage spike must be limited and safely dissipate the stored energy, or it can damage the system.

This chapter focuses on calculating inductive load parameters for optimizing drive circuit capability. For both high- and low-side switches, the energizing and de-energizing mechanism is the same.

Challenges of Inductive Loads

The voltage across an inductive load is time-dependent. **Equation 1** calculates the inductive load parameters necessary for selecting a reliable drive circuit:

$$V(t) = R\,I(t) + L\,\frac{dI(t)}{dt} \qquad (1)$$

Equation 1 shows that an overvoltage spike can occur when disconnecting an inductive load due to a sudden change in current—the dI(t)/dt component. The magnitude of this overvoltage spike is the element that can cause component damage, and, if necessary, should be clamped.

Energizing an Inductive Load

An inductive load is energized when connected to a voltage source. The current ramps up exponentially to a steady-state value and magnetic energy is stored in the coil. **Figure 1** illustrates this behavior, using a high-side switch to drive the load.

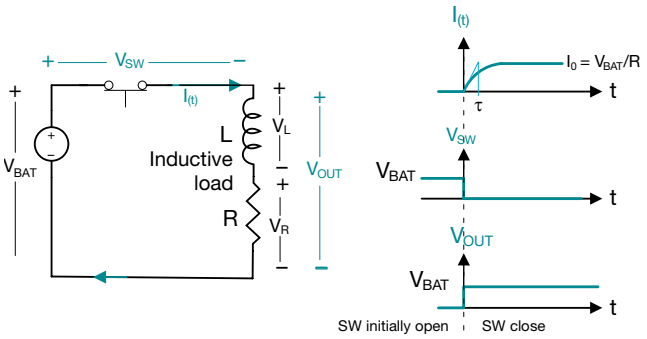

Figure 1. Energizing an inductive load waveform.

When energizing an inductive load in a resistor-inductor (RL) circuit, the voltage loop is a first-order differential equation (**Equation 2** on the following page):

$$V_{BAT} = R\, I(t) + L\frac{dI(t)}{dt} \quad (2)$$

Solving **Equation 2** gives you the instantaneous inductive load current. You can calculate the important circuit parameters using **Equation 3**:

$$I(t) = \frac{V_{BAT}}{R}\left(1 - e^{\frac{-t}{\tau}}\right) \quad (3)$$

The challenge at this stage is to select a switch that can carry the steady-state load current and keep the junction temperature below the thermal shutdown temperature. The best switch options include a low dropout voltage, like in a low on-resistance switch.

Table 1 shows the parameters introduced by **Equation 1**, and their impact on the system.

Parameter	Time constant	Steady-state current	Stored magnetic energy
Equation	$\tau = L/R$	$I_0 = V_{BAT}/R$	$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$
Impact	Predicts the energizing time	Helps in R_{ON} selection	Helps in clamp circuit selection

Table 1: Calculated parameters of an energizing inductive load.

Disconnecting an Inductive Load from a Voltage Source

The voltage across an inductive load is expressed by **Equation 4**:

$$V(t) = R\, I(t) + L\frac{dI(t)}{dt} \quad (4)$$

A sudden change in the current (dI(t)/dt) induces a very high voltage as the current changes from V_{BAT}/R to zero.

Disconnecting an inductive load from a voltage source generates a huge negative voltage spike across the inductor, as shown in **Figure 2**. The voltage spike will appear across the switch and be much higher than the switch's rated breakdown voltage. Adding an appropriate voltage clamp will avoid having a voltage spike cause severe damage.

Figure 2 shows the expected waveforms in the case of a disconnected energized inductive load through a high-side switch, both with and without a clamp present in the system.

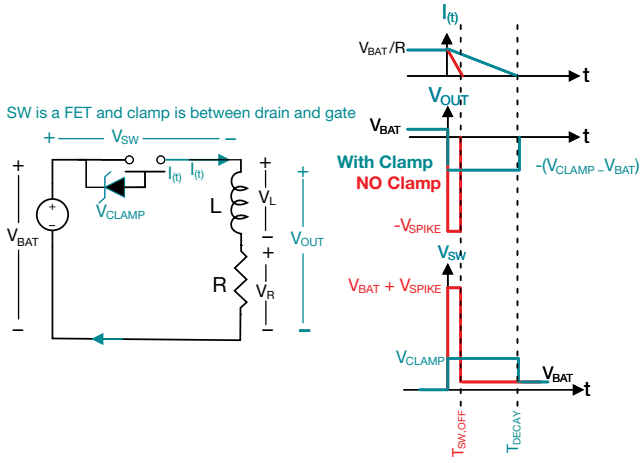

Figure 2. Waveform of an inductive load disconnected from a voltage source with a clamp (teal) and without a clamp (red).

Table 2 shows the effectiveness of implementing a clamp voltage across the switch. The clamp slows down the current slope from 250 mA/10 µS to 250 mA/419 µS and safely dissipates 3.125 mJ of stored energy. Without the clamp, the 2512-V voltage spike can easily damage the switch.

Clamp	DC resistance (R)	Inductance (L)	V_{BAT}	On-state: current switch on V_{BAT}/R	Off-state: current switch off	Stored magnetic energy $E = \frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$	$T_{SW,OFF}$	T_{DECAY}	Voltage spike or clamp $L\frac{dI(t)}{dt}$	Voltage across the switch (V_{SW})
Without clamp	48 Ω	100 mH	12 V	250 mA	0 mA	3.125 mJ	10 µS		2500 V	2512 V (high voltage spike)
With clamp	48 Ω	100 mH	12 V	250 mA	0 mA	3.125 mJ		419 µS	48 V	60 V

Table 2: Voltage levels caused by switching off an inductive load, with and without a clamp.

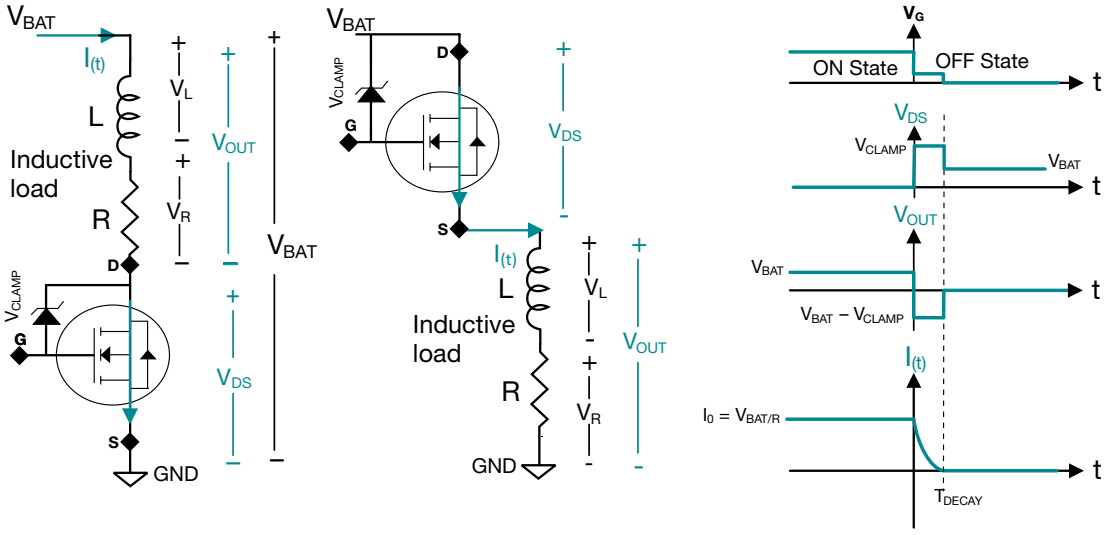

Figure 3. De-energizing an inductive load using a transient voltage suppression (TVS) clamp across the switch.

Safely De-Energizing an Inductive Load

There are three ways to safely de-energize an inductive load when disconnecting it from the voltage source:

- Use a clamp across the switch.
- Use a clamp and blocking diode across the inductive load.
- Use a freewheeling diode across the inductive load.

In all three methods, the main objective is to keep the voltage across the switch below the breakdown level and safely dissipate the stored energy in the inductive load.

Equation 5, for the clamping circuit, is based on a first-order differential equation:

$$RI(t) + L\frac{dI(t)}{dt} = K \quad (5)$$

where K is a constant depending on the clamp method in each circuit. The application report, "How to Drive Resistive, Inductive, Capacitive and Lighting Loads with Smart High-Side Switches," offers more detailed calculations for driving an inductive load.

Let's review the three methods.

Using a Clamp Across the Switch

Using a clamp across the switch is the most effective method and covers most inductive loads. The clamp is integrated in most smart high-side switch integrated circuits used in automotive and industrial systems.

Figure 3 shows the de-energizing of an inductive load when using a clamp across the switch.

Solving for **Equation 5**, where K represents the clamping method for the circuit shown in **Figure 3**, results in **Equation 6**:

$$RI(t) + L\frac{dI(t)}{dt} = -(V_{CLAMP} - V_{BAT}) \quad (6)$$

Table 3 lists the calculated parameters from **Equation 6**.

Inductance	Resistance	Initial current	Stored energy	Demagnetization time	TVS dissipated energy
L	R	$I_0 = V_{BAT}/R$	$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$	$t_{DEMAG} = \frac{L}{R}\text{Ln}(1 + \frac{R \times I_0}{V_{CLAMP} - V_{BAT}})$	$E_D = \frac{L}{R}V_{CLAMP}\left(I_0 - \frac{V_{CLAMP} - V_{BAT}}{R}\text{Ln}(1 + \frac{R \times I_0}{V_{CLAMP} - V_{BAT}})\right)$
Simplified formula when $V_{CLAMP} > 3 \times V_{BAT}$					$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2 \frac{V_{BAT}}{V_{CLAMP} - V_{BAT}}$

Table 3: Parameters necessary for selecting a switch and clamp (clamp across switch).

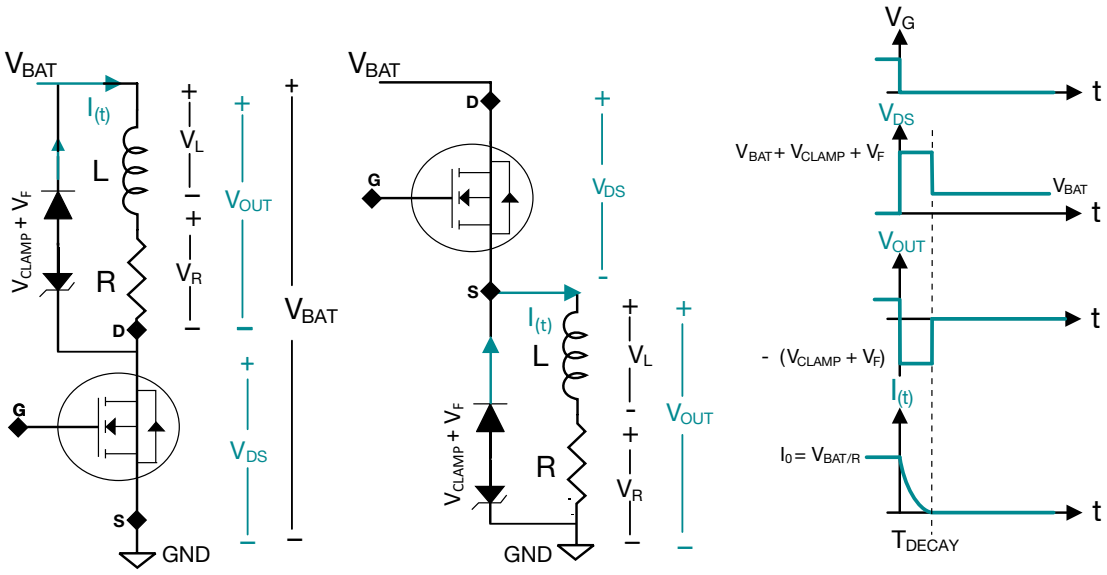

***Figure 4.** De-energizing an inductive load using a TVS clamp across the load.*

Using a Clamp and Blocking Diode Across the Load

Figure 4 shows the configuration in which the clamp and a blocking diode are placed across the load.

Again, substituting the clamping voltage and forward voltage for the blocking diode into **Equation 5** generates **Equation 7**:

$$RI(t) + L\,\frac{dI(t)}{dt} = -(V_{CLAMP} + V_F) \qquad (7)$$

Table 4 lists the calculated parameters from **Equation 7**.

The series diode is a must for preventing current from flowing in the external TVS when in the on state.

Inductance	Resistance	Initial current	Stored energy	Demagnetization time	TVS dissipated energy
L	R	$I_0 = V_{BAT}/R$	$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$	$t_{DEMAG} = \frac{L}{R}\,Ln\left(1 + \frac{R \times I_0}{V_{CLAMP} + V_F}\right)$	$E_D = \frac{L}{R}(V_{CLAMP} + V_F)\left(I_0 - \frac{V_{CLAMP} + V_F}{R}\,Ln\left(1 + \frac{R \times I_0}{V_{CLAMP} + V_F}\right)\right)$
Simplified formula when $V_{CLAMP} > 2 \times V_{BAT}$					$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$

***Table 4:** Parameters necessary for selecting a switch and clamp (clamp and blocking diode across load).*

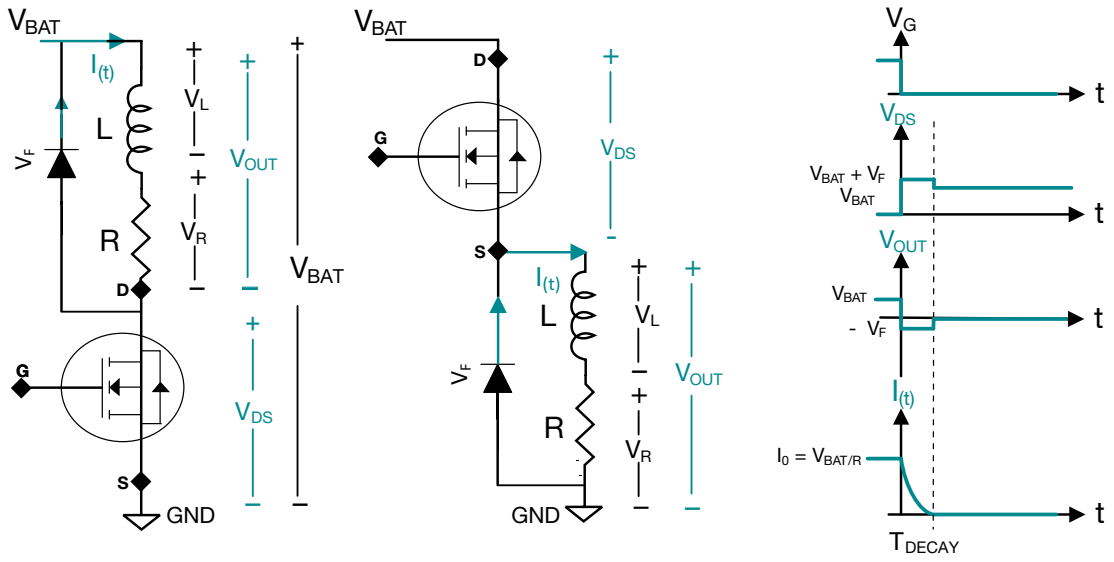

***Figure 5.** De-energizing an inductive load using a freewheeling diode across the load.*

Using a Freewheeling Diode Across the Load

Figure 5 shows the configuration using a freewheeling diode across the load.

Using **Equation 5** and substituting for K, **Equation 8** generates the expression for this de-energizing method:

$$RI(t) + L\frac{dI(t)}{dt} = -V_F \qquad (8)$$

Table 5 lists the calculated parameters from **Equation 8**.

Inductance	Resistance	Initial current	Stored energy	Demagnetization time	Diode dissipated energy
L	R	$I_0 = V_{BAT}/R$	$\frac{1}{2}L\left(\frac{V_{BAT}}{R}\right)^2$	$t_{DEMAG} = \frac{L}{R}Ln(1 + \frac{R \times I_0}{V_F})$	$E_D = \frac{L}{R}V_F\left(I_0 - \frac{V_F}{R}Ln(1 + \frac{R \times I_0}{V_F})\right)$
Simplified formula					$0.8\ L\frac{V_F}{R}\frac{V_{BAT}}{R}$

***Table 5:** Parameters necessary for selecting a switch and freewheeling diode.*

Parameter	Clamp across switch (could be integrated)	Clamp across load	Freewheeling across load
Voltage stress across switch	V_{CLAMP}	$V_{BAT} + V_{CLAMP} + V_F$	$V_{BAT} + V_F$ Lowest stress
Peak power	$V_{CLAMP} \times \frac{V_{BAT}}{R}$	$V_{CLAMP} \times \frac{V_{BAT}}{R}$	$V_F \times \frac{V_{BAT}}{R}$
Demagnetization time	$\frac{L}{R} Ln(1 + \frac{V_{BAT}}{V_{CLAMP} - V_{BAT}})$	$\frac{L}{R} Ln(1 + \frac{V_{BAT}}{V_{CLAMP} + V_F})$	$\frac{L}{R} Ln(1 + \frac{V_{BAT}}{V_F})$ Longest demagnetization time
Dissipated energy	$\frac{1}{2} L \left(\frac{V_{BAT}}{R}\right)^2 \frac{V_{CLAMP}}{V_{CLAMP} - V_{BAT}}$	$\frac{1}{2} L \left(\frac{V_{BAT}}{R}\right)^2$	$0.8\ L \frac{V_{BAT}}{R} \frac{V_F}{R}$ Lowest dissipated energy
Comment	Most suitable for board space and cost because the clamp is small in size and energy dissipates in the switch	Used if the switch is not capable of dissipating the energy; an additional clamp would increase board space and cost	Used if the demagnetization energy is relatively high in cases of high load currents and high inductance; long demagnetization time T_{DECAY} could be an issue; increases board space and cost

Table 6: Pros and cons of the three clamping methods.

Selection Method Pros and Cons

Table 6 summarizes all three ways to safely de-energize an inductive load.

As previously stated, the clamp across the switch is integrated in smart high-side switches. Smart high-side switches have the ability to dissipate the magnetic energy in most cases. In some applications, however, the integrated clamp is not capable of dissipating the magnetic energy, so an external clamp such as a TVS diode is required. When using an external TVS diode, its clamp voltage should be lower than the integrated clamp's lower limit minus the battery voltage, as expressed by **Equation 9**:

$$V_{CLAMP(EXT)} < V_{CLAMP(INT)} - (V_{BAT} + V_F) \quad (9)$$

Table 7 summarizes the important parameters for smart high- and low-side switches.

Conclusion

An inductive load can be driven by a high- or low-side switch. The advantage of a high-side switch is the capability of protection against short to ground and the flexibility of load diagnostics. TI smart high-side switches come in a variety of FET sizes and feature an integrated clamp across the FET.

Using smart high-side switches, you must calculate the inductive load parameters and compare them against the device capability in terms of the on-resistance and demagnetization energy capability. You will not need an external clamp if the switch's integrated clamp is capable of dissipating the demagnetization energy. If the clamp cannot dissipate that energy, you will need an external clamp, TVS diode or a freewheeling diode.

Device family	Low on-resistance	Short-to-ground protection	Integrated clamp	Inductive energy capability	Junction-to-ambient thermal resistance
Smart high-side switches	✔	✔	✔	✔	✔
Low-side switches			✔	✔	✔

Table 7: Important parameters for driving an inductive load.

Chapter 11: Automotive Load Dump

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Abstract

In automotive applications, a load dump is one of the largest transient pulses in modern electronics. This low-impedance pulse, defined by International Organization for Standardization (ISO) 16750 pulse A and B, can be devastating to integrated circuits (ICs), as voltages can rise as high as 120 V for almost half a second.

These transient pulses occur when a generator delivering current is abruptly disconnected from the system. Most electronics qualified to withstand load dump will survive the suppressed version of a pulse while other circuitry clamps the unsuppressed waveform. This chapter will explore the details of load-dump transient pulses, their causes and how devices are able to survive load dump.

Causes of Load Dump

In modern automobiles, the alternator charges the battery and powers the vehicle's electronics when the engine is running. An alternator mirrors the basic properties of an inductor, which when spun by a belt produces a magnetic field that converts to an electric current. **Figure 1** illustrates the connection of a battery in parallel with an alternator and other loads of a vehicle.

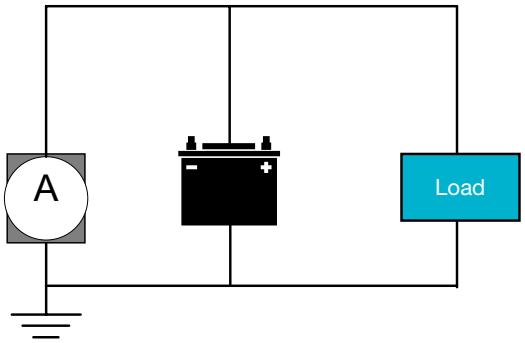

Figure 1. An alternator connected with a battery and load.

In **Figure 1**, the battery acts as a large capacitor, keeping the voltage at a steady 13.5 V. However, if the car battery is disconnected from the line, then the inductive nature of the alternator causes a very large voltage spike on the power line of the vehicle that could take almost half a second to dissipate; see **Figure 2**. ISO 16750 defines this transient pulse as a load-dump pulse.

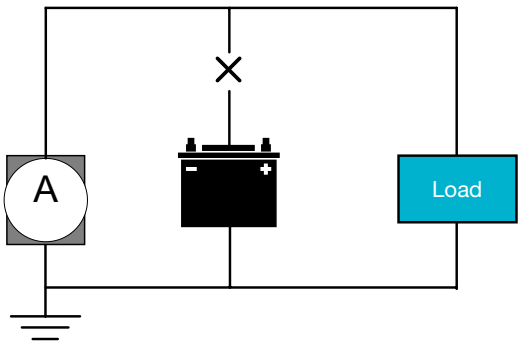

Figure 2. Battery disconnected from car system.

Since the alternator is always on when the engine is on, there is a high probability that a vehicle could experience this battery disconnection scenario. Let's look at the voltage profile of a load-dump transient and show the danger that it poses to systems.

ISO 16750-2

The ISO 16750-2 (2012) automotive standard highlights several different battery discontinuities, such as starting profile (more commonly known as cold crank) and load dump. The standard also covers momentary drops in supply voltage, ground reference and supply offset, as well as open- and short-circuit load conditions.

Pulse A (Unsuppressed)

Figure 3 shows the waveform of a load-dump pulse, while defining the voltage levels that the alternator can reach and how long it will take the voltage to dissipate. **Table 1** on the following page lists the values of **Figure 3**.

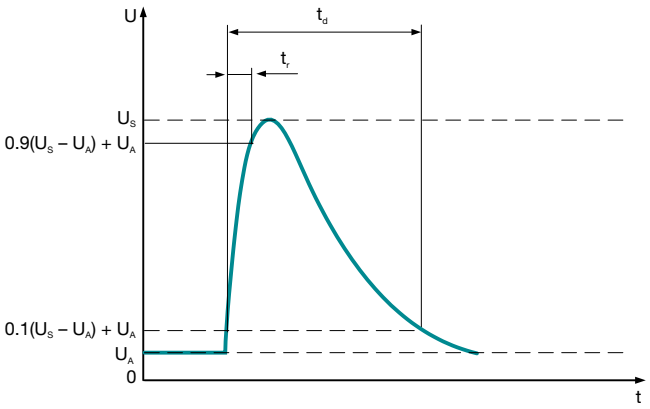

Figure 3. Load-dump test A unsuppressed voltage waveform.

Parameter	Type of system		Test requirements
	12 V	24 V	
U_s	79 V–101 V	151 V–202 V	10 pulses at 1-minute intervals
R_i	0.5 Ω–4 Ω	1 Ω–8 Ω	
t_d	40 ms–400 ms	100 ms–350 ms	
t_r	10 ms	10 ms	

Table 1: Load-dump test A unsuppressed parameters.

There is a large difference in the requirements between 12-V and 24-V automotive systems. A load-dump pulse is a significant challenge for electronic designers because most devices' internal structure will break down before the voltage achieves the maximum level specified by ISO 16750-2.

Equation 1 calculates the maximum current that the IC can withstand from an alternator:

$$(U_S–V_{breakdown})/R_i = (79\ V–40\ V)/0.5\ \Omega = 78\ A \qquad (1)$$

There is a usually small clamp or electrostatic discharge (ESD) structure within the IC that cannot withstand this level of current for 400 ms. With the pulse being such high energy, vehicle manufacturers usually implement a centralized circuit that will clamp the pulse to a more manageable voltage.

Pulse B (Suppressed)

Since the test A pulse is extremely high energy and difficult to dissipate, most vehicles have a suppression circuit in parallel with the alternator and battery that clamps the voltage at a less-stressful level. This means that the internal circuit changes from **Figure 2** to **Figure 4**.

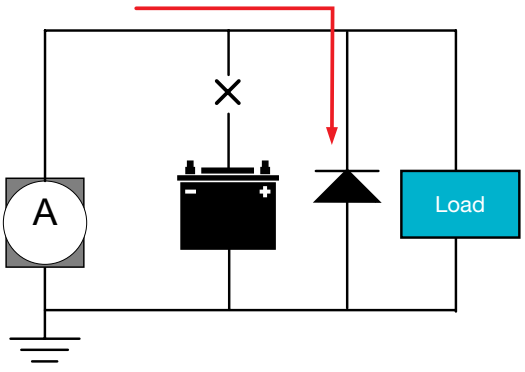

Figure 4. An alternator connected to a battery, load and central clamp.

This clamp will dissipate the excess energy of the alternator and keep the voltage set at a level defined in the pulse in **Figure 5**; **Table 2** shows the subsequent values.

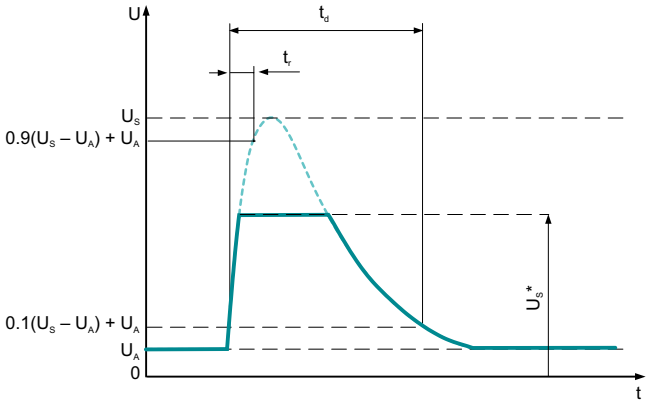

Figure 5. Load-dump test B suppressed voltage waveform.

Parameter	Type of system		Test requirements
	12 V	24 V	
U_s*	35 V	58 V	5 pulses at 1-minute intervals
R_i	0.5 Ω–4 Ω	1 Ω–8 Ω	
t_d	40 ms–400 ms	100 ms–350 ms	
t_r	10 ms	10 ms	

Table 2: Load-dump test B suppressed parameters.

Most vehicles require all battery-connected ICs to survive a suppressed load dump.

Surviving Load Dump

The load-dump pulse is high energy and requires a high-power dissipation clamp to suppress the pulse. To avoid this high dissipation, TI's portfolio of power switches includes a higher voltage tolerance instead of trying to dissipate the entire pulse. This means that the IC connected to the power line will simply allow the pulse to occur without being damaged.

In TI smart high-side switches, the maximum accepted voltage is 40 V to 45V depending on the switch. During a suppressed load-dump event, the transient will not break the switch, as it is lower than the maximum rating for the device.

Similarly, for TI's ideal diode controller family, even though the field-effect transistor (FET) is not integrated, it can

still withstand a high transient during a suppressed load dump. As long as the main FET is rated with a high-enough voltage to survive the suppressed load dump, the system should see no damage.

Conclusion

A car alternator helps charge the battery during normal operation, but due to its inductive nature, if the battery becomes disconnected, the voltage on the power line can spike up very high. This transient is represented in ISO 16750-2 and is characterized for a suppressed pulse and unsuppressed pulse. The two pulses are nearly identical, with one having a central clamp to dissipate a high portion of the energy and lowering the voltage that the IC has to tolerate. TI's portfolio of smart high-side switches are designed to survive these transients and can be used in automotive applications connected directly to the battery.

TI's integrated power switch portfolio arms engineers with reliable solutions. From managing inrush current to surviving stressful transients in an automotive environment, our diverse portfolio of devices can distribute, protect or diagnose power path designs.

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