

# Application Note for HITFET+

## Smart Low-Side Power Switch

### Application Note

Rev.1.0 2015-01-12

Automotive Power



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**Abstract**

**1 Abstract**

*Note: The following information is given as a hint for the implementation of the device only and should not be regarded as a description or warranty of a certain functionality, condition or quality of the device.*

This Application Note is intended to provide useful information about HITFET+ smart low-side power switches in the automotive environment as well as industrial. Starting from a design perspective, the Application Note describes the application requirements and concludes at the device level.

**Table 1 Terms in Use**

Abbreviation	Meaning
HSS	High Side Switch
LSS	Low Side Switch
RDS	Varying Resistance between Drain and Source during ON state
RDS(ON)	Least Value of $R_{DS}$ as defined in the data sheet for the device.
VBAT	Voltage Measured at the Battery terminals
VIN	Input voltage to the device, measured at the input (IN) pin of the device
VDD	Supply voltage for the device, measured at the $V_{DD}$
AWG	American Wire Gauge
DMOS	Double diffused MOS
ESD	Electrostatic Discharge
EMC	Electro Magnetic Compatibility
ECU	Electronic Control Unit
GND	Ground
IN	Input
OEM	Original Equipment Manufacturer. In this document, car maker.
PWM	Pulse Width Modulation
PLAMP	Lamp Power, expressed in watts
TA	Ambient temperature
TJ	Junction Temperature
TC	Case Temperature or temperature of the solder
VPWM	Root Mean Square Voltage across the load during PWM
Tier1	Supplier of the ECU to the OEM

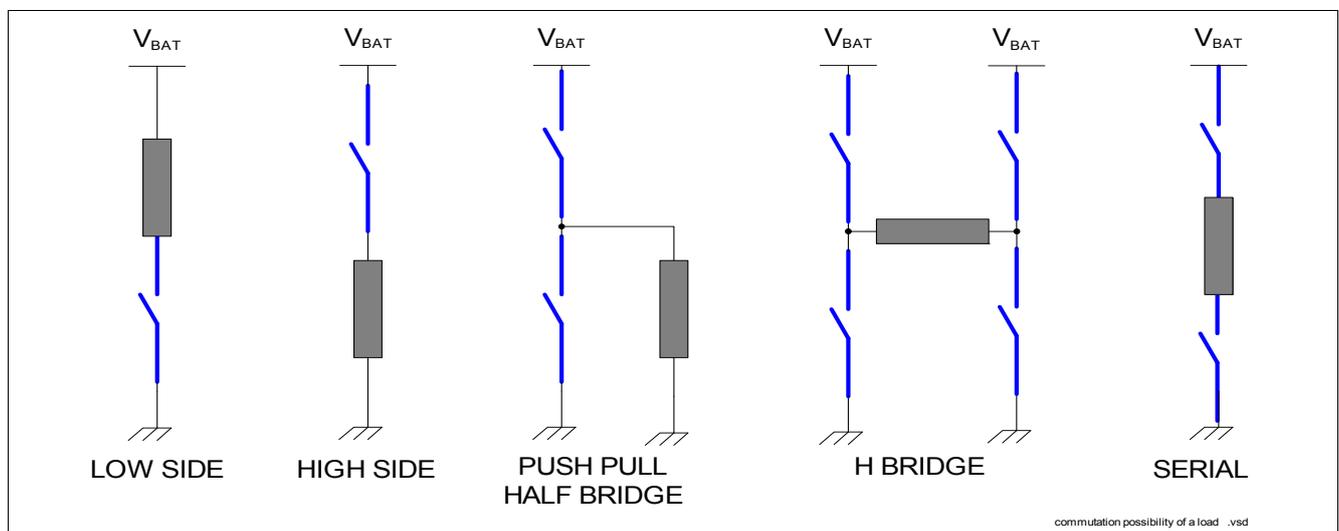
Introduction

## 2 Introduction

The following chapter introduces the general advantages of a low side switch over other switching configurations and gives a short introduction to the Infineon HITFET+ Family.

### 2.1 Why Low Side Switches?

In an automotive system, a single electrical supply  $V_{BAT}$  potential is available. Five possible solutions exist (refer to **Figure 1**) to switch electrical loads ON and OFF. The automotive engineering community defines low side switches as switches that sink/commute the load current.



**Figure 1 Commutation Possibilities of a Load**

Low side switches are found in a wide range of applications worldwide, including automotive applications. The main reasons for their popularity are provided below:

#### 2.1.1 Better Driving Capability

Unlike a high side switch which has to pull up the Output Voltage to around  $V_{BAT}$  with an even higher gate voltage, a low side switch drops down the Output Voltage to almost ground level. For this, the gate voltage of a low side switch requires to increase only up to the  $V_{BAT}$ . This avoids the need of a charge pump, needed by the high side switches. The following figure provides a comparison of the gate voltages required to drive a high side switch vs. a low side switch.

Introduction

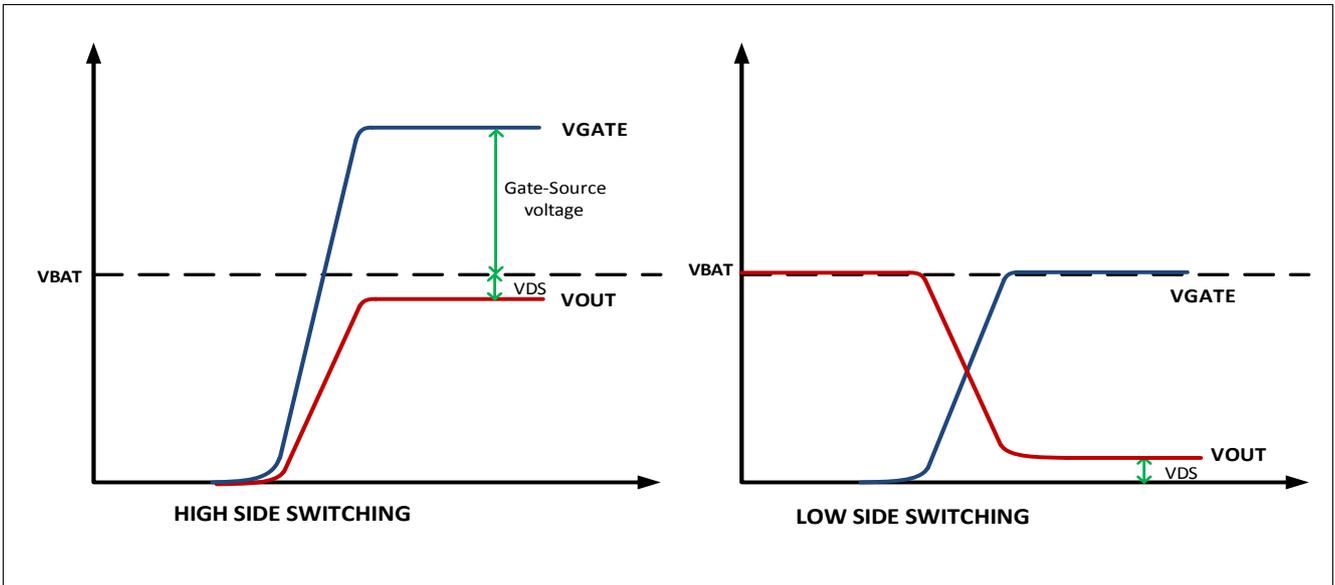


Figure 2 Driving Capability Comparison

A low side switch does not require a specialized driver thus making it less sensitive to noise in comparison to the High Side device.

Introduction

2.1.2 Robust Ground

Because of the way the Low Side is connected, it only has a single ground. This makes it more robust because the risk of stray currents caused by ground shifts is eliminated. See **Figure 3**.

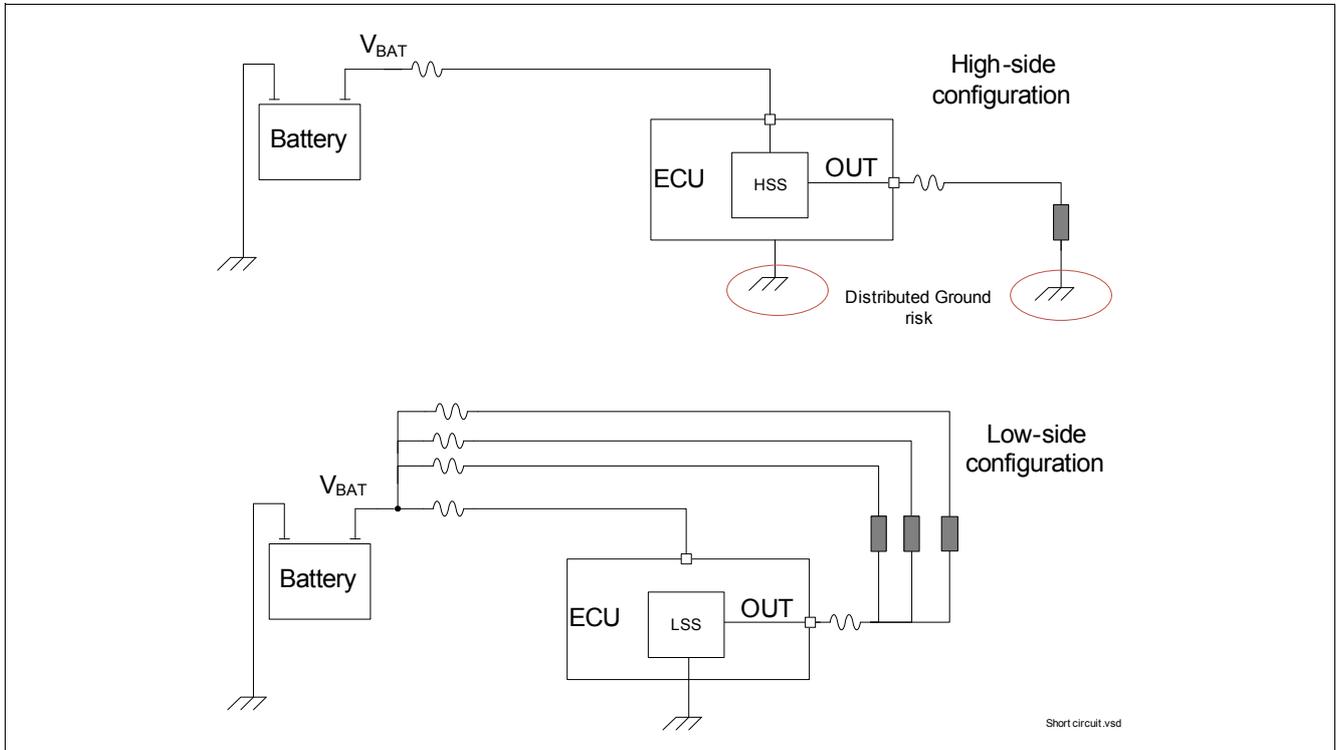
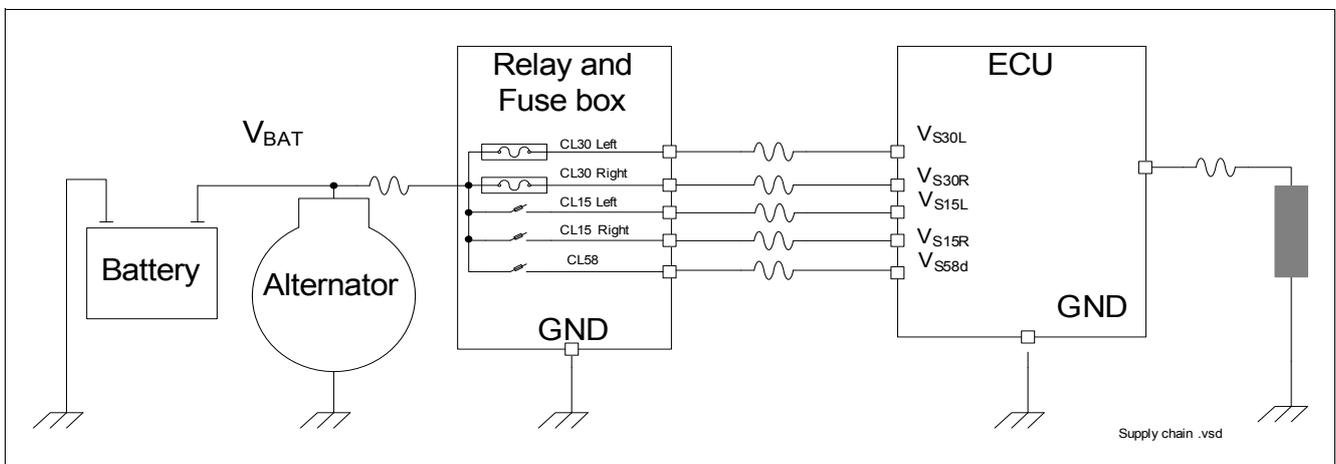


Figure 3 Ground Connection Difference Between HSS and LSS

### 3 Automotive Environment

#### 3.1 Battery Voltage Supply

Only one supply potential, ( $V_{BAT}$ ) is available in the vehicle. This supply comes from the battery when the engine is off and from the alternator when the engine is running. **Figure 4** shows the typical supply topology. The battery voltage is typically 12.6V (engine off) and 13.5V when the engine is running although this figure is different for different OEMs. These values can vary in different phases of the mission profile. For simplicity,  $V_{BAT}$  will be used for both the real battery voltage and  $V_{ALT}$ , the alternator voltage (engine running).



**Figure 4** Typical Supply Chain in a Vehicle

Automotive Environment

3.1.1 Alternator Control Loop

The alternator provides current as soon as the engine reaches idle (typically 800RPM). If there is no diode or battery to limit the voltage, an alternator can provide a transient voltage of greater than 100V in load dump condition. The current the alternator can provide is typically between 55A and 200A. This value is mainly dependent on the engine RPM and engine cooling. The alternator current rating is defined by the total vehicle load. The control voltage is specified as a function of the alternator temperature ( $T_{ALT}$ ). The voltage usually decreases with temperature so that the maximum battery voltage is reached when  $T_{ALT}$  is  $-40^{\circ}C$ . Refer to [Figure 5](#).

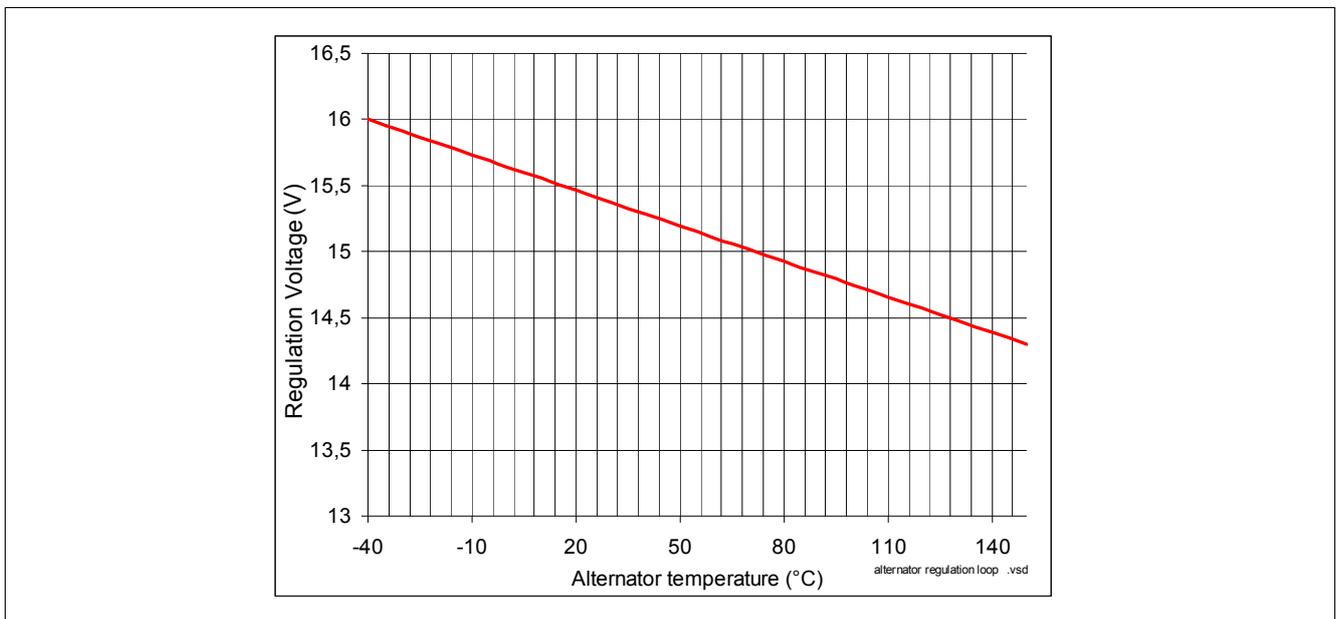


Figure 5 Alternator Control Voltage Function of Temperature

3.1.2 Alternator Ripple

The more the alternator is loaded and providing its maximum possible current, the more ripple on the supply line can be observed.  $V_{BAT}$  looks similar to [Figure 6](#). The frequency  $f_{AR}$  and the voltage swing depends on the engine’s RPM chosen by the OEM. As an umbrella specification, the following figures may be used:  $V_{AR} = 3V$  peak to peak,  $f_{AR} = [1kHz; 20kHz]$ .

Automotive Environment

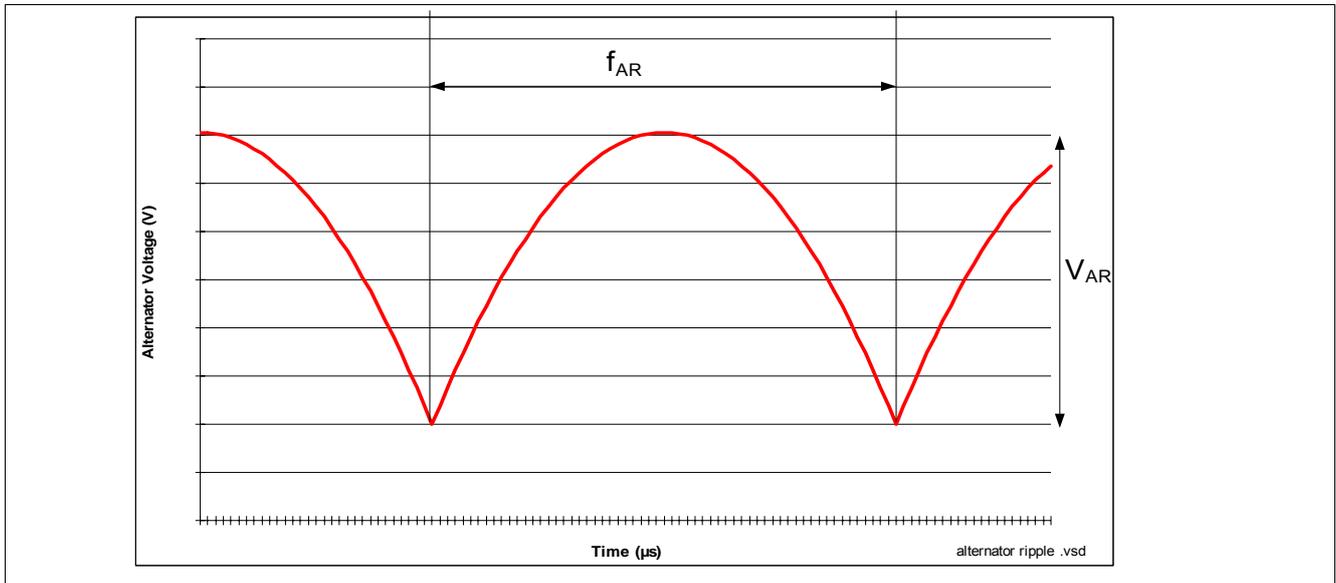


Figure 6 Typical Alternator Ripple Voltage as a Function of Time

### 3.1.3 Start-Stop Application. Regenerative Braking

The alternator can be a starter-alternator and it can also be used for regenerative braking. Each time the car is stationary, the engine is stopped. Engine restart strategies vary between OEMs, however the most common method for automatic cars is when the driver releases the brake pedal and for cars with manual transmission is when the driver shifts into the first gear. This restart will be called in the document “hot ignition”, in contrary to “cold ignition” when the car driver turns the ignition key.

A significant increase in "hot ignition" starts needs to be considered. A typical value is 30 "hot ignition" starts per "cold ignition" start ( $i_{cold}=30*i_{hot}$ ). Since the ignition phase consumes a great deal of power (200A for hot ignition, 1000A for cold ignition), it is necessary to recharge the battery quickly. This can be achieved by increasing  $V_{BAT}$  artificially; typically to 18V. An increase in  $V_{BAT}$  results in an increase in electrical power that also increases the engine's resistive torque and, consequently, the fuel consumption. This consequence is not acceptable except during braking when kinetic energy is converted into electrical energy.

During acceleration, the resistive alternator torque can be too high and the alternator can be turned OFF during strong acceleration. **Figure 7** shows the shape of the battery supply voltage, assuming a starter-alternator with regenerative braking.

As an example, a 14.5V controlled alternator providing 70A DC current corresponds to 1kW electrical power. Assuming 30% efficiency, the mechanical energy required to provide this 1kW of electrical power is 3.2kW or 4 horse power (hp). Taking a standard 100hp engine, the driven alternator can offer up to 5% of power increase.

Automotive Environment

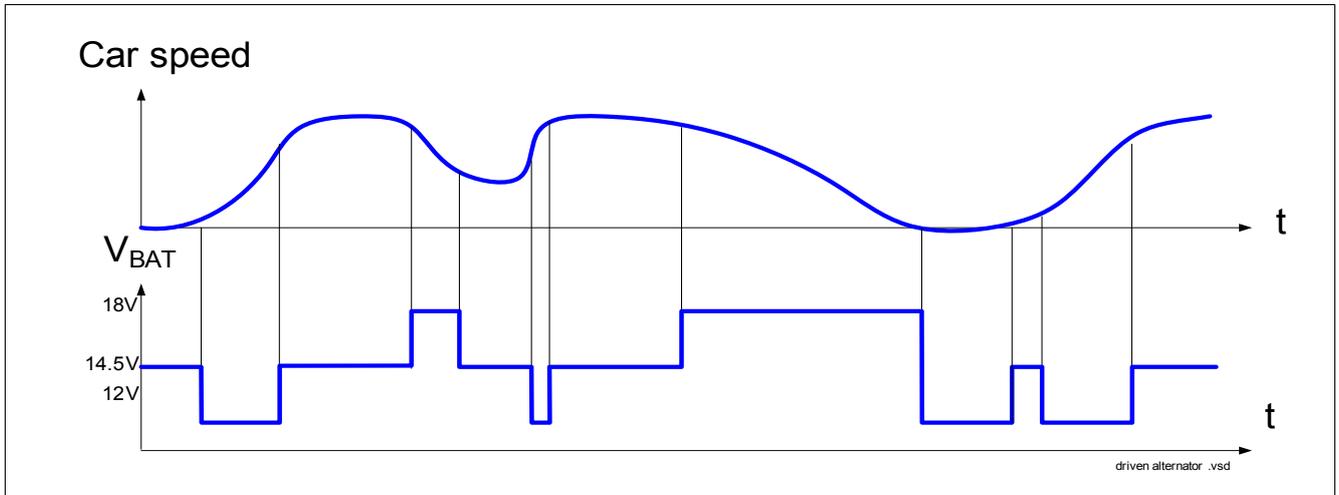


Figure 7 Battery Voltage as a Function of Car Speed

Automotive Environment

3.1.4 Low Battery Voltage Supply

Low voltage supply phases can be either due to a weak battery (discharged) or during engine cranking. The weak battery is a permanent state (from a semiconductor perspective) while cranking is a transient phenomenon.

3.1.4.1 Discharged Battery

A discharged battery is usually due to parasitic leakage current in the vehicle when it has been parked for long periods of time. The minimum battery voltage at which the car can still start is OEM dependent. This voltage is considered as the minimum nominal voltage. Typically, this is 8V.

3.1.4.2 Engine Ignition

The voltage during the ignition phase is complex to describe and the values are very dependent on the vehicle OEM as well as the type of engine. All OEMs specify different ignition voltage pulses  $V_{CRK\_MIN}$  between 3 and 5.5V (refer to **Figure 8**).  $V_{CRK\_OSC}$  is usually 7V and oscillations range from a couple of Hertz to 800Hz (800RPM).  $V_{BAT\_STD}$  is the battery voltage during the engine stand-by phase and is usually 12.6V.  $V_{BAT\_RUN}$  is the battery voltage when the engine is running and is usually 14.5V. For simplicity the red curve is used with  $V_{CRK\_MIN} = 3$  to 5.5V, typically 4.5V.  $t_{CRK} = 65ms$ ,  $t_{LAUNCH} = 10s$  and  $V_{CRK\_LAUNCH} = 5.5$  to 8V.

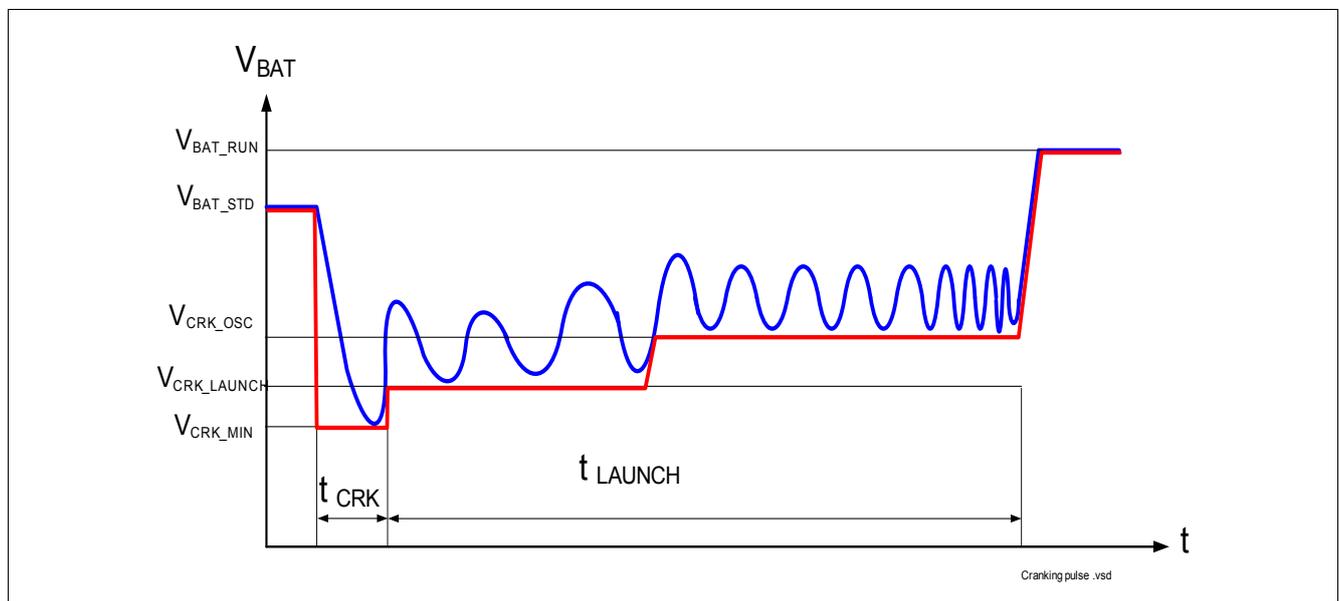


Figure 8 Ignition Pulse

3.1.5 High Battery Voltage Supply

A high battery voltage can occur due to different conditions such as jump start, load dump, faulty alternator control and high alternator ripple capabilities.

Automotive Environment

3.1.5.1 Jump Start

A jump start for a car (12V) is a situation where a truck battery (24V) is bypassing the battery to start the engine. The voltage and the time of the jump start is OEM dependent. A worst case is 28V for 2minutes. For trucks, a jump start occurs when a special electrical device connected to a power outlet supplies 48V to the truck battery for several minutes.

3.1.5.2 Load Dump

Load dump occurs when the battery terminal is suddenly disconnected while the alternator is providing current. The battery is essentially a capacitor and hence stabilizes the system. Load dump can also occur when switching off high current loads. Refer to [Figure 9](#).

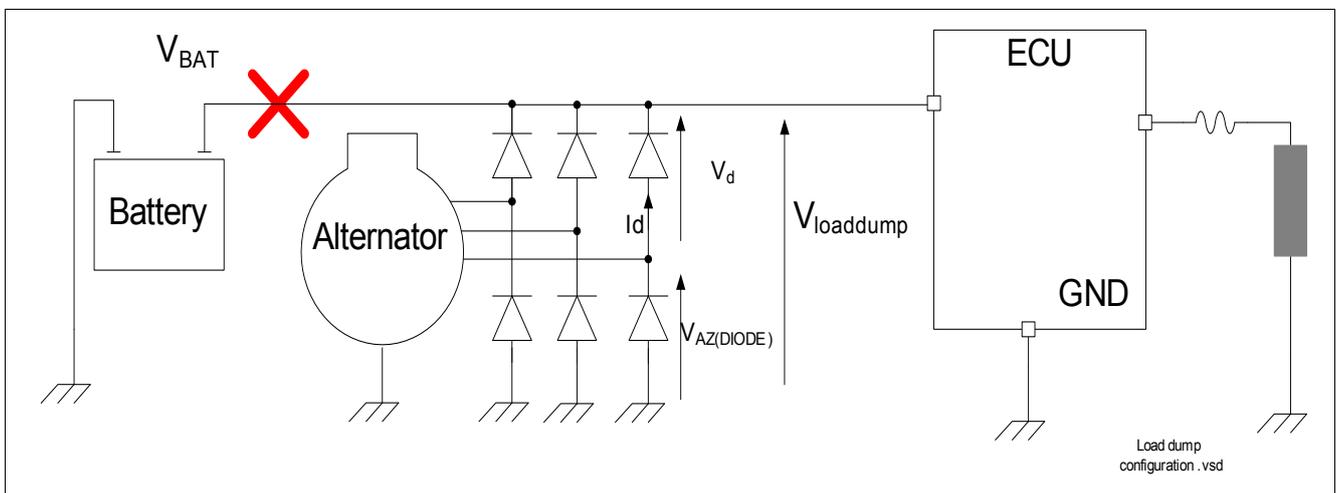


Figure 9 Load Dump Configuration

When the battery is disconnected, the system becomes unstable and the voltage rises until the alternator low side diodes reach avalanche, limiting the voltage to  $V_{loaddump}$ . Some OEMs replace the diodes with Zener diodes. The advantage Zener diodes provide is to reduce the load dump (avalanche) voltage to the Zener voltage.  $V_{loaddump}$  and  $t_{loaddump}$  are specified by the OEM. After a delay, ( $t_{loaddump}$ ), the alternator takes back control and the voltage decreases. Infineon considers  $V_{loaddump} = 40V$  for  $t_{loaddump} = 400ms$  typical. After the load dump event, a high ripple voltage is observed on the battery line while the battery remains disconnected. Infineon considers  $V_{ALT\_MAX} = 18V$  and  $V_{ALT\_MIN} = 12V$  typical. The oscillation frequency is considered to be between 1kHz and 20kHz and can last to 10 hours long. Refer to [Figure 10](#).

Automotive Environment

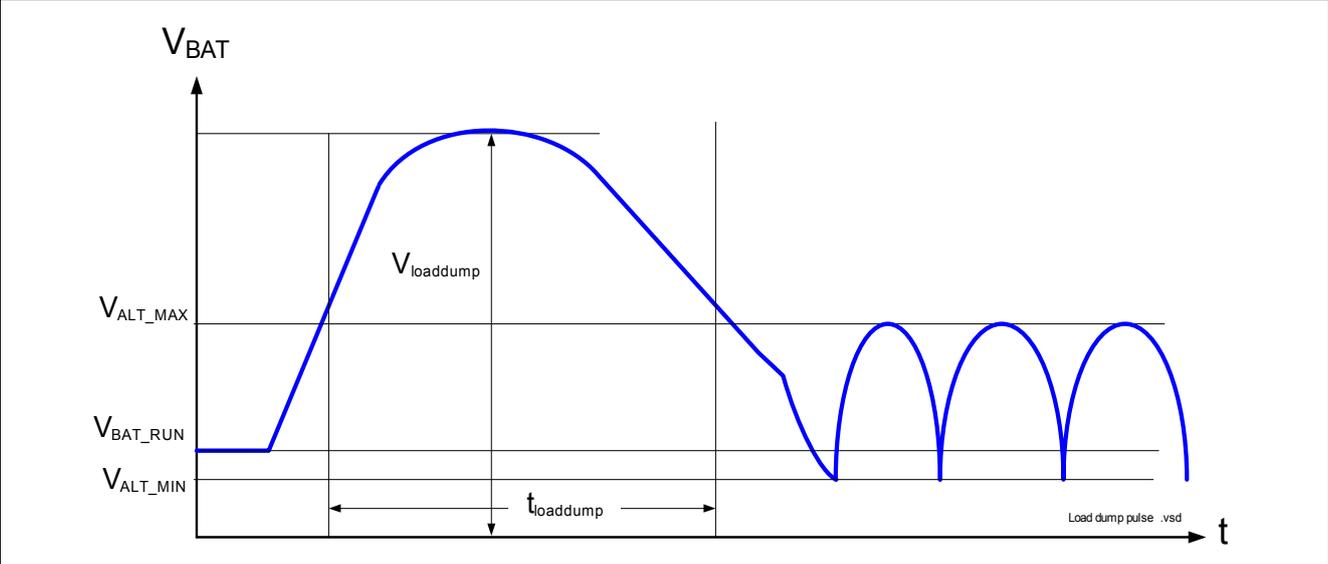


Figure 10 Load Dump Pulse

Automotive Environment

3.1.6 Reverse Polarity

A Reverse polarity condition exists when the battery supply line  $V_{BAT}$  is connected to ground and the ground line GND to the battery supply. Reverse polarity mainly occurs for two reasons. During module handling and installation, where some unnatural movements can be assumed, or when the vehicle has a low battery and the driver connects jumper cables incorrectly from an external battery (Start help), reverse polarity can result. The voltage and time for which the vehicle can withstand this reverse polarity is defined by the OEM. Infineon considers -16V for 2mn at ambient temperature +25°C typical. Some loads such as a lamp or a resistor can tolerate current flowing in the reverse direction whilst others cannot (e.g. motors, polarized capacitors, etc.).

3.1.7 Loss of Battery

In an architecture with a switched supply line like a CL15, a loss of battery is normal. In a shared-fuse architecture, the loss of the supply line should not result in a module failure when the fuse blows due to a short circuit somewhere else.

3.1.8 Specification for Battery Voltage

To sum up the above discussion, refer to [Figure 11](#).

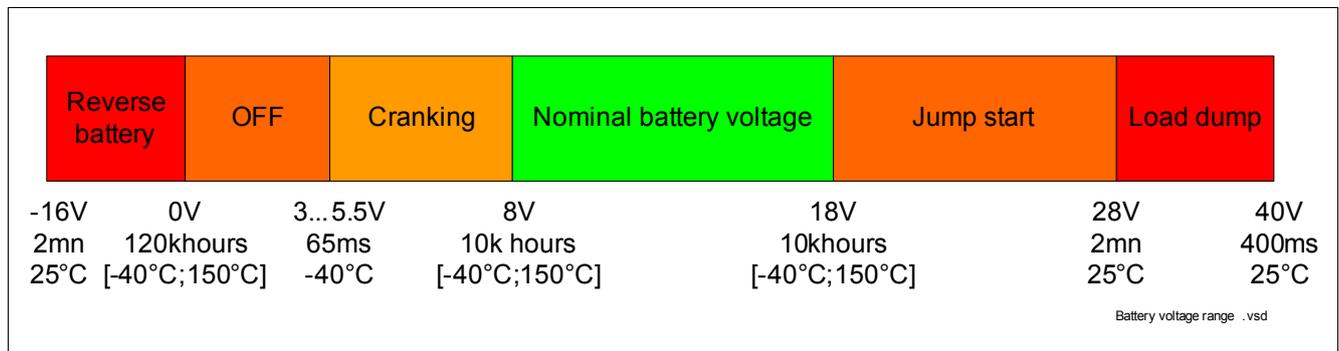


Figure 11 Infineon Specification for Battery Voltage

Automotive Environment

3.2 Temperature

The ambient temperature  $T_A$  range in an automotive application is one of the harshest found in electronics. Only space and aeronautical activities can be more challenging. As the minimum temperature is universally agreed to be  $-40^{\circ}\text{C}$ , the maximum temperature varies according to application, OEM, tier 1, module housing, etc. Infineon considers  $T_{A\_MAX} = +85^{\circ}\text{C}$  for cockpit applications and  $T_{A\_MAX} = +105^{\circ}\text{C}$  for under hood application typical.

3.2.1 Ambient Module Temperature

Ambient module temperature follows the seasons as shown in **Figure 12**. Ambient module temperatures are cold in winter, hot in summer. While  $-40^{\circ}\text{C}$  is considered to be the minimum temperature to start the car in winter, it is not valid for every engine start in winter as the system heats up during driving. The same logic can be applied to the hot season. It is possible to assume  $+85^{\circ}\text{C}$  or  $+105^{\circ}\text{C}$  for example, as the maximum ambient temperature (car parked in summer) at start up, but it is incorrect to assume that  $T_{A\_MAX} = +85^{\circ}\text{C}$  is a permanent condition during summer. In other words,  $-40^{\circ}\text{C}$  and  $+85^{\circ}\text{C}$  are considered as starting points, but not as permanent conditions. Infineon considers an ambient temperature profile shown in **Figure 13** typical.

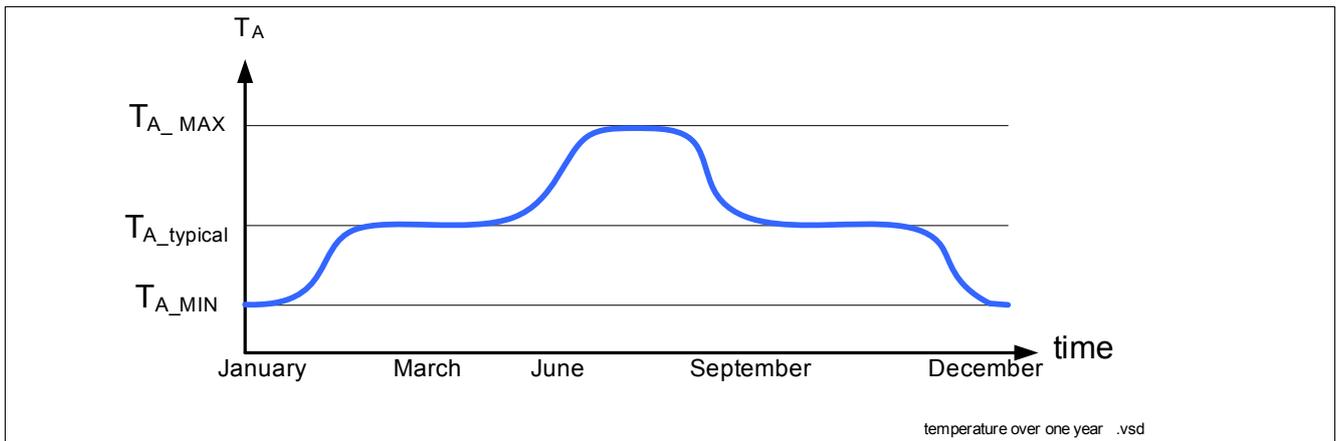


Figure 12 Suggested Ambient Module Temperature Over One Year (North Hemisphere)

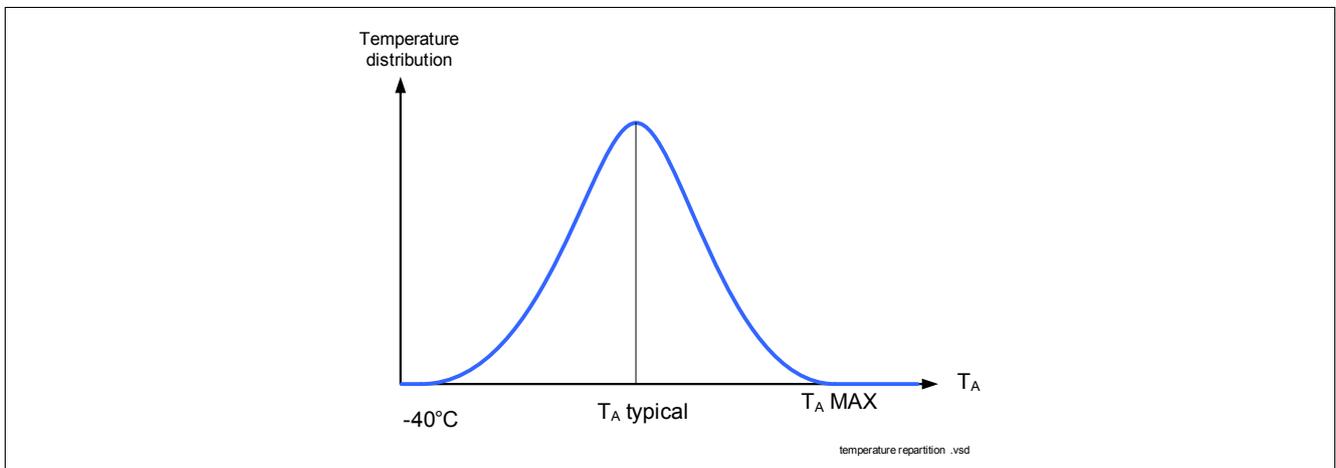


Figure 13 Suggested Temperature Distribution Over Car Lifetime

Automotive Environment

3.2.2 Internal Module Temperature

The devices on the PCB are subject to the heat radiated by neighboring devices. Of course, the heat generated depends on the module’s design. Typically, it is considered the temperature of a module increases by +15°C during operation.

3.3 Ground

As described in Chapter 4.5, a ground shift  $V_{SHIFT}$  can exist between the module ground and device ground. Loss of ground should also be a consideration in module design. There are two possible failures, loss of device ground and loss of module ground. As a device supplier, Infineon assumes any loss of ground to be loss of device ground unless explicitly indicated.

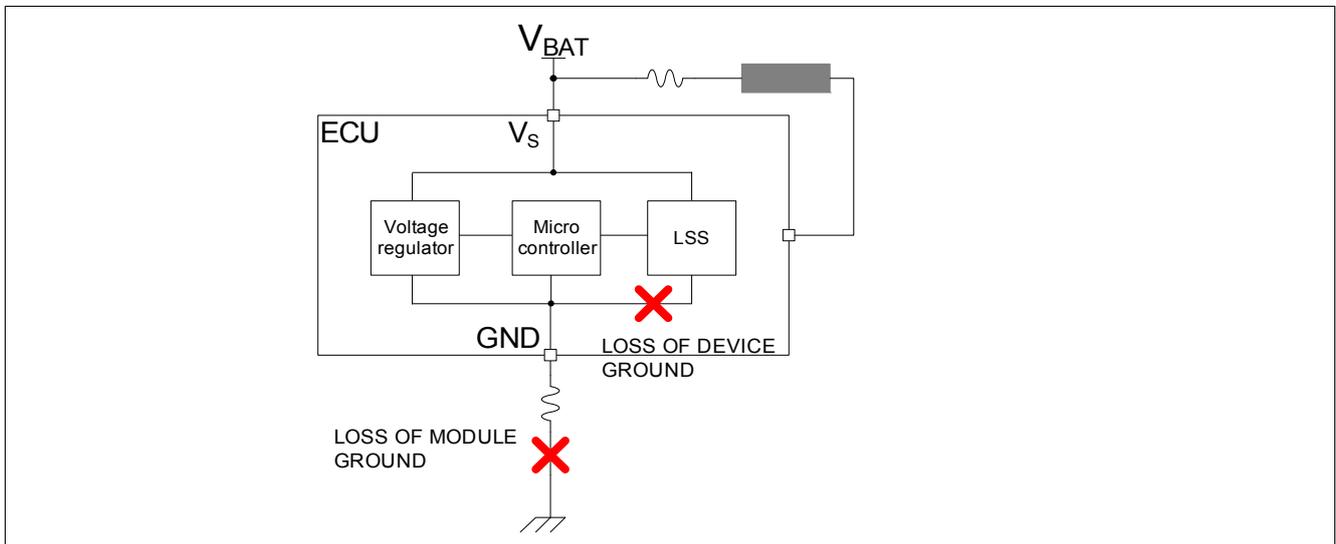


Figure 14 Loss of Module or Device Ground

3.4 Lifetime

The life time of a car/module/device is assumed to be 15 years or 131.400 hours.

3.4.1 Running Time

Running time is an accumulation of time over which the module is in operation (micro controller active, load activated or ready to be activated) is assumed to be 10.000hours. (~2hours per day for 15 years).

3.4.2 Stand-by Time

Stand-by time corresponds to the remaining time over 15 years where the module is not in operation. With the above assumptions, this is 121.400hours.

#### 3.4.3 Number of Ignitions

The number of ignitions cycles is determined by the strategy of the car OEM. As an umbrella specification, Infineon consider 100 000 cold ignitions over the car life time. This leads to almost ~20 (18.6) ignitions per day. This number does not include the additional start and stop cycles due to start-stop systems implemented in some architectures. Several OEMs require the number of start-stop cycles to be at one million.

Type Of Supply

## 4 Type Of Supply

### 4.1 Module Un-powered During Stand-by

Figure 15 shows a typical application where the ECUs are de-powered when the vehicle is parked with the engine off. This type of battery supply is commonly called KL15 (e.g. in Germany).

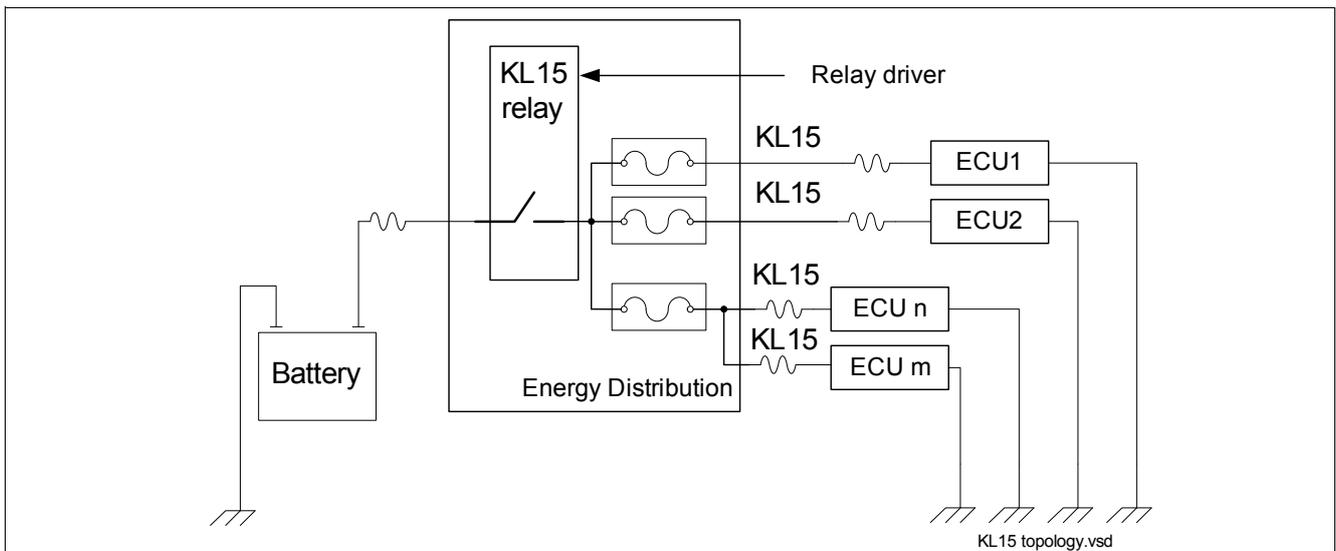


Figure 15 Clamp 15 Application

### 4.2 Module Supplied During Standby

Figure 16 shows a typical application where the ECUs remain powered with the engine off. This type of battery supply is commonly called KL30 (e.g. in Germany).

Type Of Supply

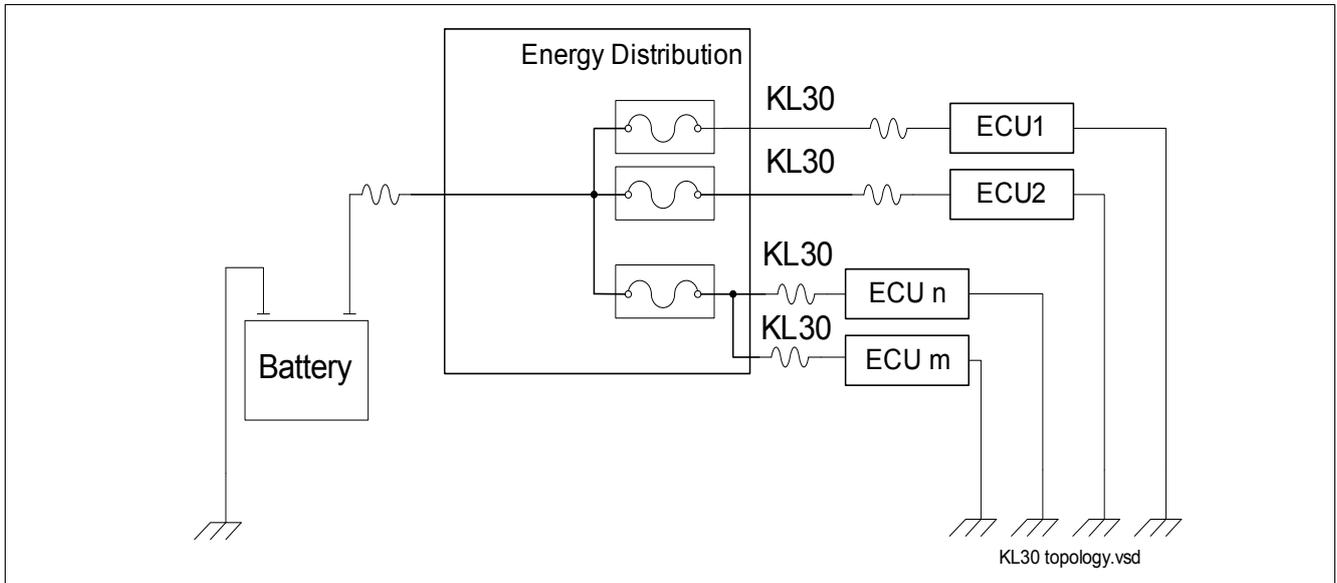


Figure 16 Clamp 30 Application

### 4.3 Left/Right Front/Rear Separation

For safety reasons, supply redundancy is often necessary. Redundancy of the supply is often based on the separation of the left and right side of the vehicle. This is where one battery line supplies all loads on the left side of the vehicle and another line supplies all loads on the right side. The same redundancy can be found with front and rear separation. Adding to this the KL15 and KL30 concepts, a complex ECU can be supplied by up to 8 different supply lines. **Figure 17** shows such a supply architecture.

Type Of Supply

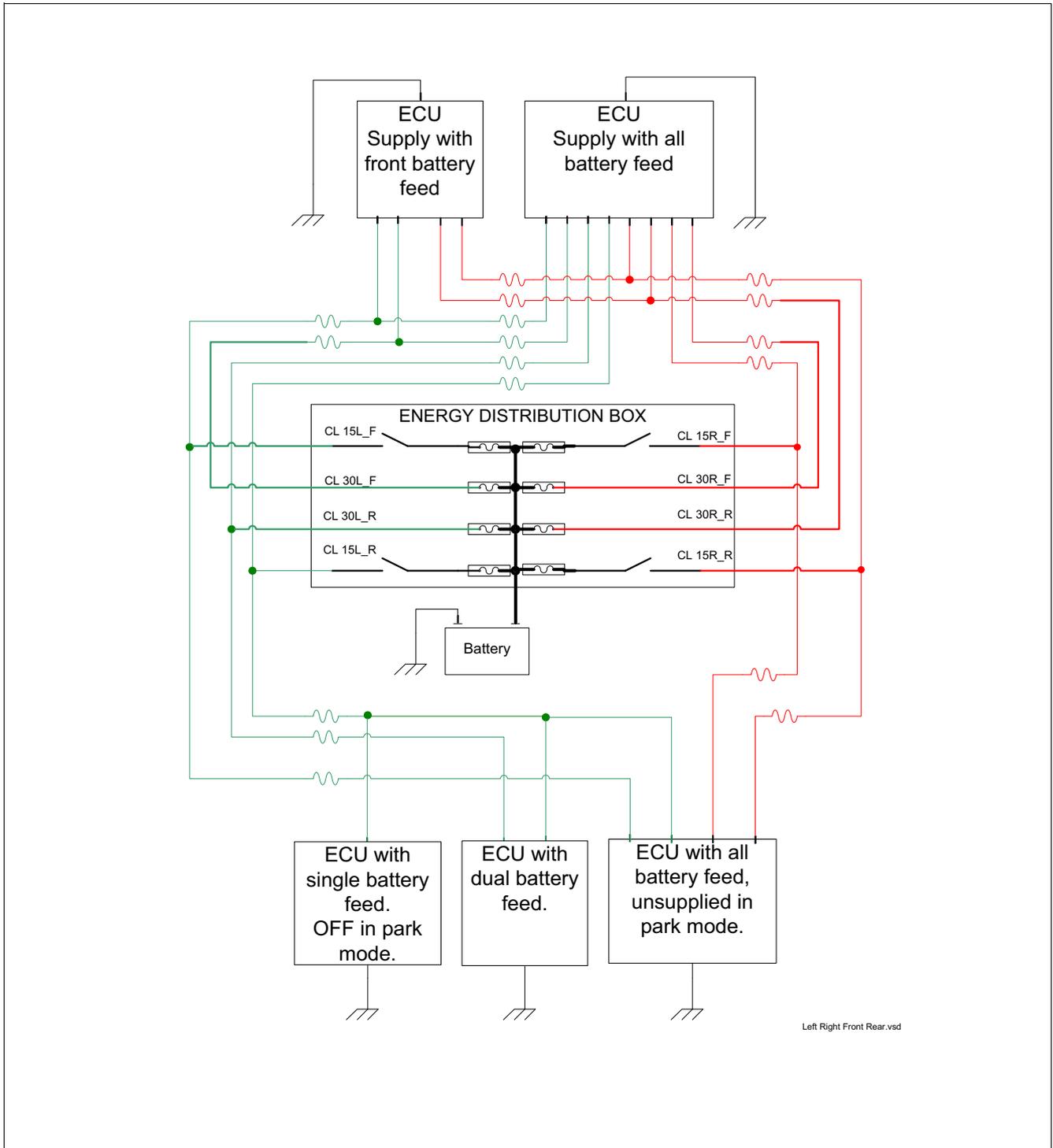


Figure 17 Complex and Mixed Supply Line Architecture

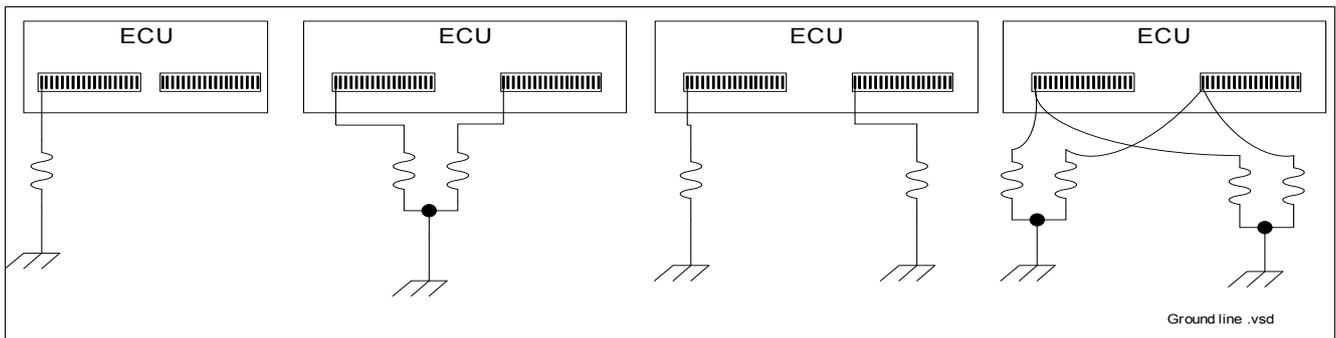
#### 4.4 Secondary Supply

Some modules also provide a secondary supply to sub-systems. This architecture is common in door modules and climate systems which can be supplied by a dedicated battery feed switched from the master door or climate ECU. Typical example is the KL58 supply line used to supply the dashboard.

Type Of Supply

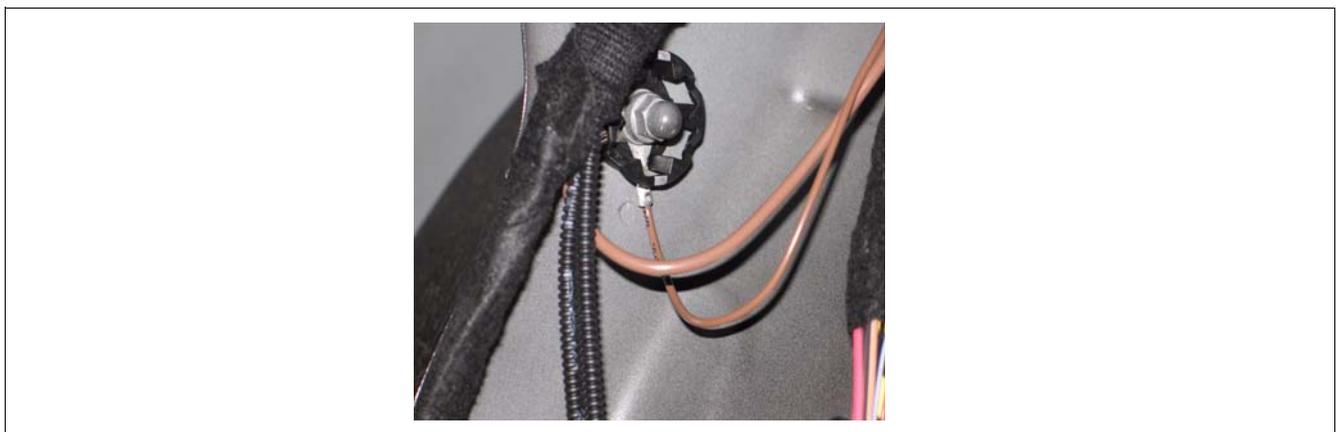
4.5 Ground Line

The ground (GND) in a car is provided by the chassis. Therefore, GND is present everywhere and access to GND is always available. In most cases, there is at least one GND pin per module connector. This GND pin is connected to the chassis by a wire. **Figure 18** shows different ways to implement a GND connection. On the left hand side is the cheapest method. The most expensive but safest and recommended method is shown on the right hand side. **Figure 19** shows a picture of a GND connection realized on vehicle.



**Figure 18** Ground Line Concept

One consequence of this architecture can be that some modules don't have the same 0V (GND) reference. For example, a high current application such as power steering, starter motor or alternator doesn't have the same 0V reference as the rest of the vehicle. This can also be the case for applications where the connecting cable to GND is long or thin, causing a noticeable impedance. This ground shift voltage can be either positive or negative. Infineon recommends ISO11898-3 (Low Speed CAN network ISO norm) as an umbrella specification. This standard specifies a  $\pm 1.5V$  between ECU GND and chassis GND.



**Figure 19** Ground Line Example

As already explained in **Figure 3**, low side switches tend to have a more robust ground and possibilities of ground shift between load and switch do not exist.



Load And Application

## 5 Load And Application

The diversity of loads driven by low side switches is enormous. Clustering these loads is always challenging. Nevertheless, three different categories can be outlined. Lighting/heating or capacitive loads, motors or inductive loads and LEDs or resistive loads.

### 5.1 Heating Loads

HITFET+ is used for various heating loads in body applications such as auxiliary heating, seat heating, steering wheel heating, and also lambda heaters as a major application in power train applications. **Table 2** lists examples of some typical heating loads that can be addressed by HITFET+.

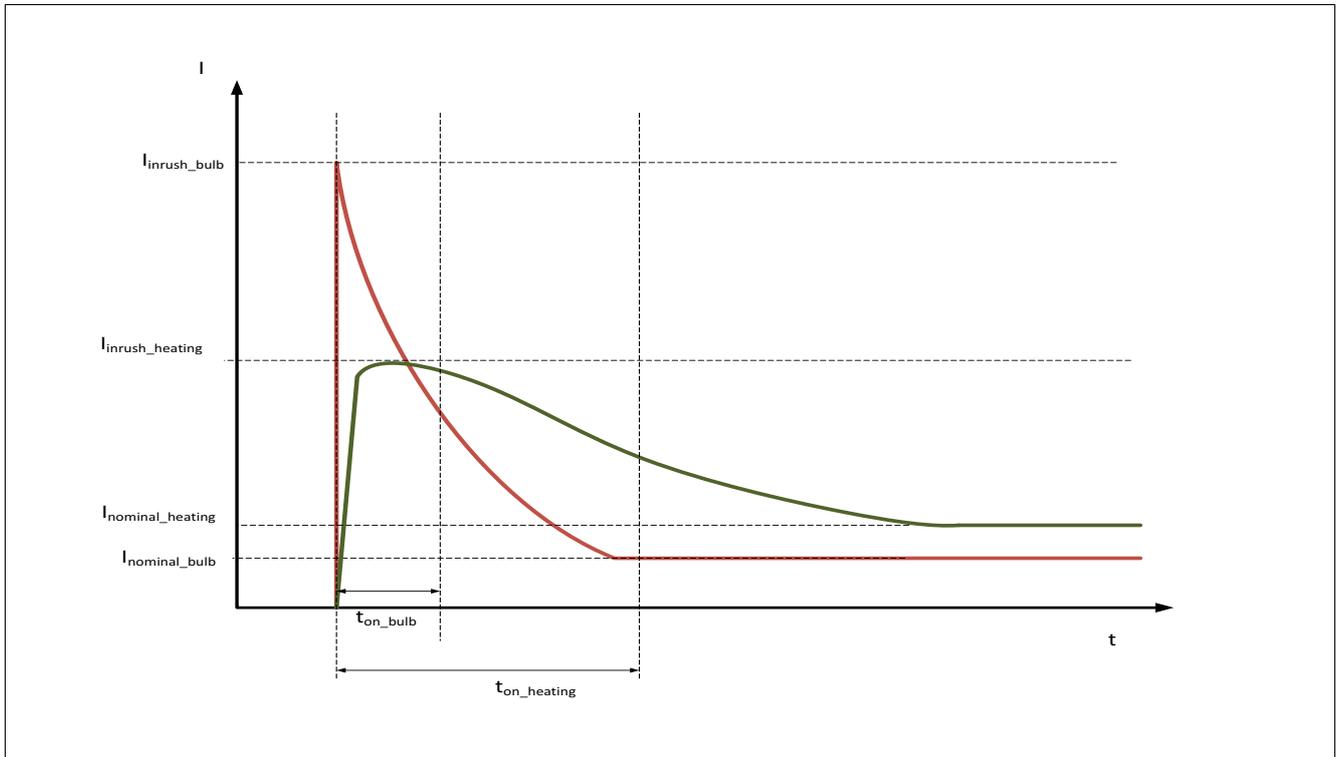
**Table 2 Example Heating Loads**

Type	Example Load	Nominal Current Range (A)
Low power	Rear view mirror heating	2-5
Medium power	Steering wheel heating	5-7
High power	Seat heating	7-9

#### 5.1.1 Inrush Behavior

Heating loads are usually made up of resistive elements that generate heat by blocking the current. And although they are resistive loads, they also have an inrush behavior because the resistance changes with the temperature. **Figure 20** shows a comparison of current profiles for a bulb and a typical heating element.

Load And Application



**Figure 20 Comparison of Typical Inrush Behavior of a Heating Load and a Lighting Load (Not to Scale)**

Depending on the OEM or Tier 1 manufacturer, a certain ratio is applied in inrush which relates to the nominal current of the lamp or heating element. For a lamp, this ratio can be 12 times the nominal current, whereas for heating elements a factor of 2 is considered typical for HITFET+ applications.

For lamps, the inrush current depends on the type of bulb but typical it can be assume about 2ms. The inrush current also defines the time required to switch the lamp on. It can also be seen from **Figure 20**, that heating loads take relatively longer to switch on. Depending strongly on the application the  $t_{on\_heating}$  can take about 1 second or even a couple of minutes. For example, glow plug's  $t_{on\_heating}$  is about ten times faster than a conventional PTC heater.

**5.1.2 Pulse Width Modulation (PWM)**

The life time of a heating load depends strongly on the supply voltage. Heating elements are sensitive to currents. Hence, a constant supply voltage is required. One approach is to use Pulse Width Modulation (PWM) to keep a constant output power. The trick is to use thermal inertia of a heating element to absorb the PWM waveform ensuring that heating is still uniform. The duty cycle can be calculated with the following equation:

$$d = \frac{V_{PWM}^2}{V_{BAT}^2} \tag{5.1}$$

where  $V_{PWM}$  is the optimum voltage needed to maintain a constant power across the heating element, whereas  $V_{BAT}$  is the supply voltage. Refer to **Chapter 10.3** for some details on **Equation (5.1)**

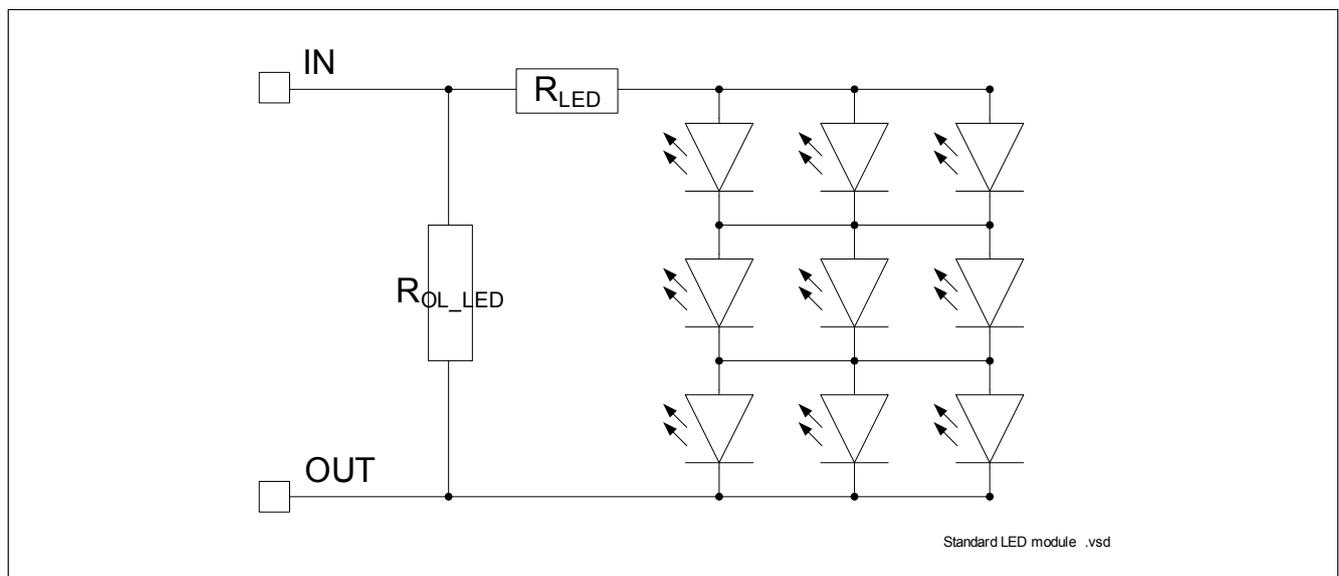
Load And Application

## 5.2 Light Emitting Diode (LED)

Light Emitting Diodes (LED) are increasingly used to replacing standard lamp bulbs. They offer a longer lifetime as well as lower current consumption for an equivalent light intensity output. Two kinds of LED modules are often used, standard and advanced. For practical purposes, the difference between these 2 types of modules is negligible, and they can both be modeled as resistive loads. Unlike lamps, LEDs start emitting light as soon as a voltage is applied that is high enough to overcome the forward bias of the device. This voltage depends mainly on the LED color. A very small current (Infineon considers 10µA typical) is enough to cause a LED to glow. This justifies the use of the  $R_{OL\_LED}$  when open load diagnosis is required.

### 5.2.1 Standard LED Module

In a standard LED module, when one LED is an open circuit, the other LEDs are not affected. This behavior is particularly desirable for rear lighting. The standard LED module shown in **Figure 21** consists of a series resistor  $R_{LED}$  to limit the current and a cluster of LEDs connected in parallel and serial. The advantage of this circuit is its simplicity. The drawback is the continuous power loss in the resistor (at least 500mW) and the susceptibility to transient overvoltages and currents. This kind of LED module is often used for rear light systems. Infineon considers  $R_{LED} = 50\Omega$ ,  $R_{OL\_LED} = 680\Omega$  typical.

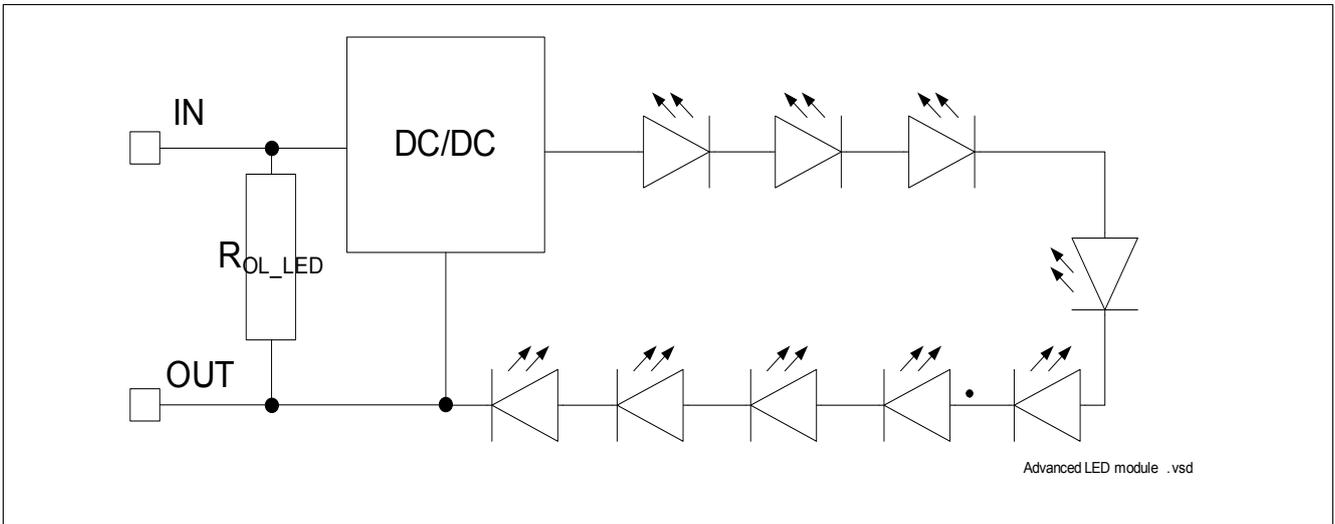


**Figure 21 Standard LED Module**

### 5.2.2 Advanced LED Module

In an advanced LED module, when one LED is an open circuit, the entire module is OFF. This behavior is particularly hazardous for headlights. The advanced LED module shown in **Figure 22** consists of a DC/DC converter driving LEDs in serial. The advantage of this architecture is robustness and immunity to voltage transients. The disadvantage is the relative electronic complexity of the DC/DC converter. Infineon considers the module OFF if  $V_{IN} - V_{OUT} < 7V$  typical. When the LED is broken, the module doesn't consume more than 30mA max, typically 15mA (current needed by the DC/DC supply itself).

**Load And Application**



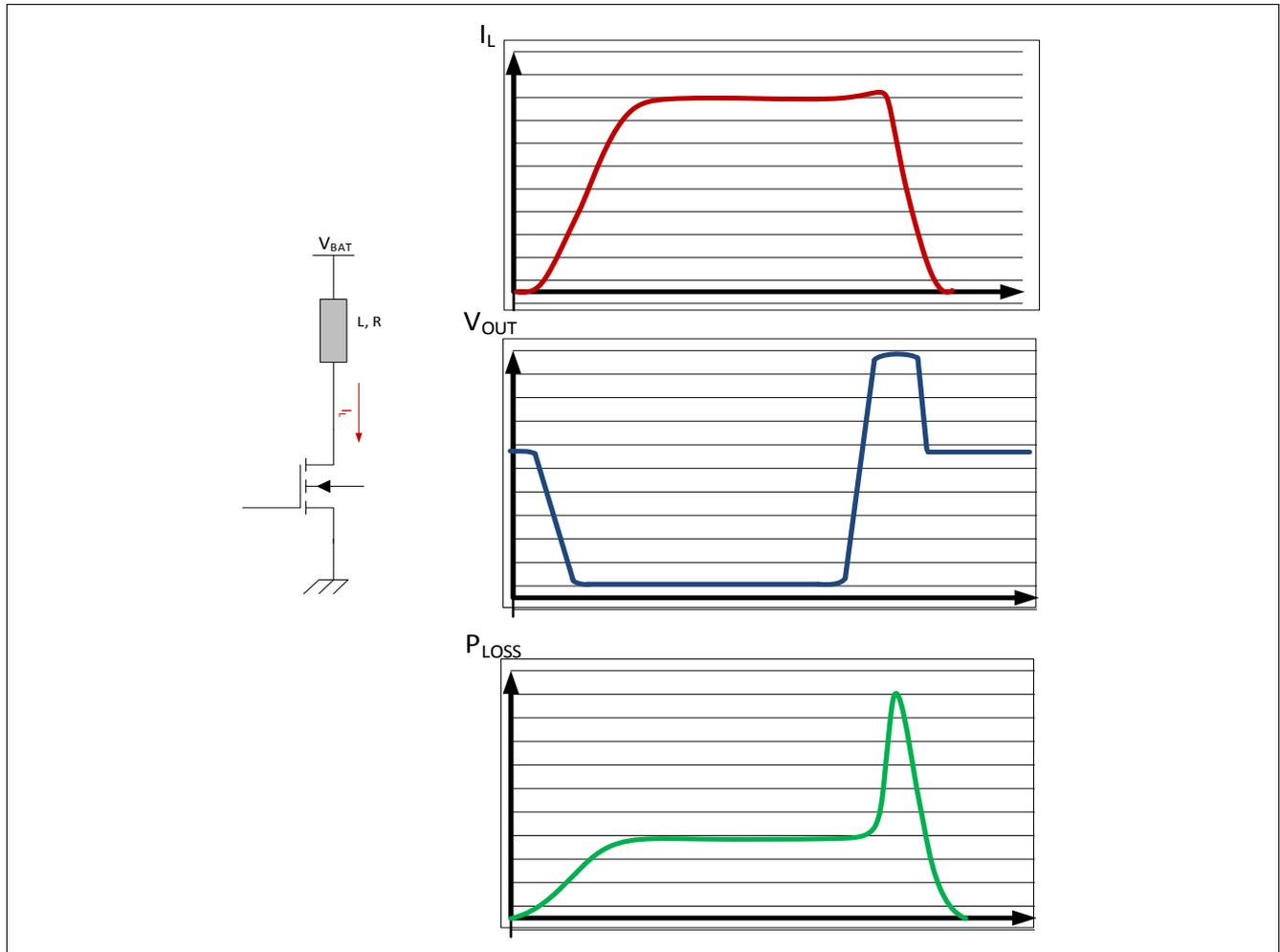
**Figure 22 Advanced LED Module**

**5.3 Inductive Load**

Relay, solenoid and motor driving are major applications of HITFET+. Relay is the oldest switch in the electronic portfolio. Although semiconductors tend to replace them in many applications, mechanical relays are still widely used.

Motors can be clustered depending on their capability to work in one or both directions. HITFET+ family can drive unipolar motors (f.e. safety lock in electric park brake). When the application requirement is to drive a motor in both directions, the drive architecture must be then an H-Bridge with either two HSS or two LSS or one HSS and one LSS.

Load And Application



**Figure 23 Typical Switching for an Inductive Motor Load**

Motors are inductive loads defined usually by inductance L and resistance R. At switch ON, the inductive load causes a slow current ramp-up, based on the time constant  $\tau = L/R$ . At switch OFF due to the inductance, the current attempts to continue to flow in the same direction which causes the load voltage to invert. Refer to **Figure 23** which demonstrates the general voltage and current characteristics of an inductive load at switch ON and OFF. Voltage in blue, current in red, power in green.

Although relay driving is an old technology, it is still challenging to implement. This is mainly due to the wide production spread in manufacturing of mechanical relays which leads to a wide spread of parameters. Also, mechanical relays show dynamic changes during operation. For instance, the inductance of a relay changes as the magnetic resistance changes. When the anchor is lifted, the magnetic flow sees a higher resistance and therefore the inductance changes to a lower value during switch OFF. This causes a current change during switch ON and a voltage change during switch OFF. Therefore, the moment when the load contacts are opening or closing can be seen in the voltage or current profiles. Refer to **Figure 24**.

Load And Application

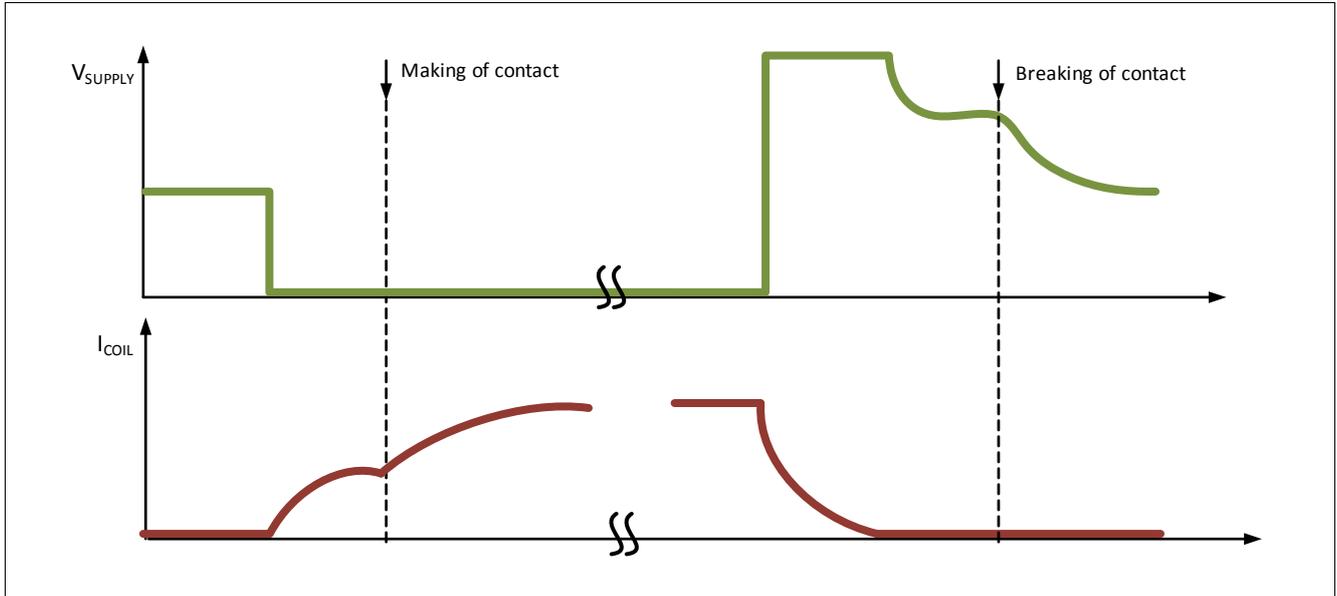


Figure 24 Switching of Relay with HITFET+

Figure 25 shows the electrical schematic of a relay. The circuitry consists of an inductance and a serial resistor with an optional parallel resistor. The serial resistor represents the copper wire resistance of the coil. Typical resistor values at room temperature are 60-9 Ohm while inductance ranges from 400mH to 600mH for an automotive 12V relay.

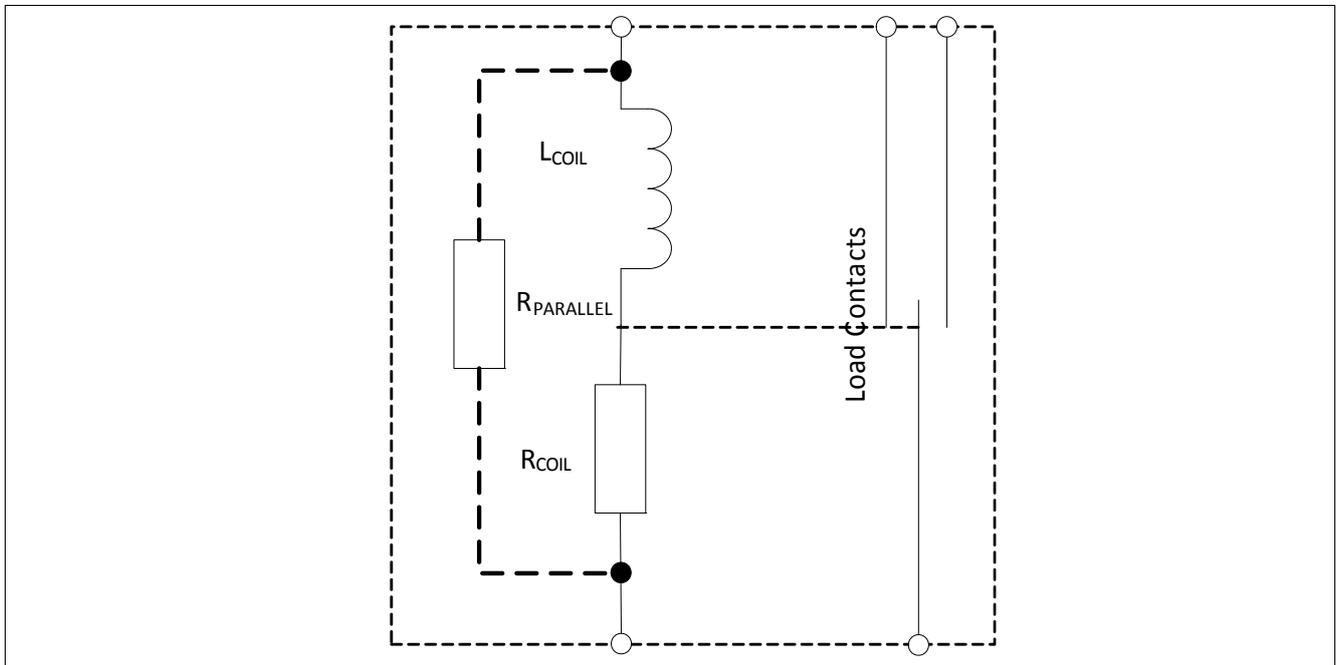


Figure 25 Electrical Schematic of a Mechanical Relay

Load And Application

5.3.1 Demagnetization Energy

As can be seen from **Figure 23**, each time an inductive load is switched OFF, a demagnetization energy has to be considered. If the over-voltage protection limit is known, this demagnetization energy can be calculated according to **Equation (5.2)**. Refer to **Chapter 10.2** for a detailed calculation.

$$E_{AS} = V_{OUT(CLAMP)} \times \frac{L}{R} \times \left[ \frac{V_{BAT} - V_{OUT(CLAMP)}}{R} \times \ln\left(1 - \frac{I_L \times R}{V_{BAT} - V_{OUT(CLAMP)}}\right) + I_L \right] \quad (5.2)$$

5.3.2 Freewheeling Diode

To keep the current flowing after switching OFF an inductive load, and to access the energy stored in the coil, a freewheeling diode can be used. **Figure 26** shows the different configurations in which relays can be used.

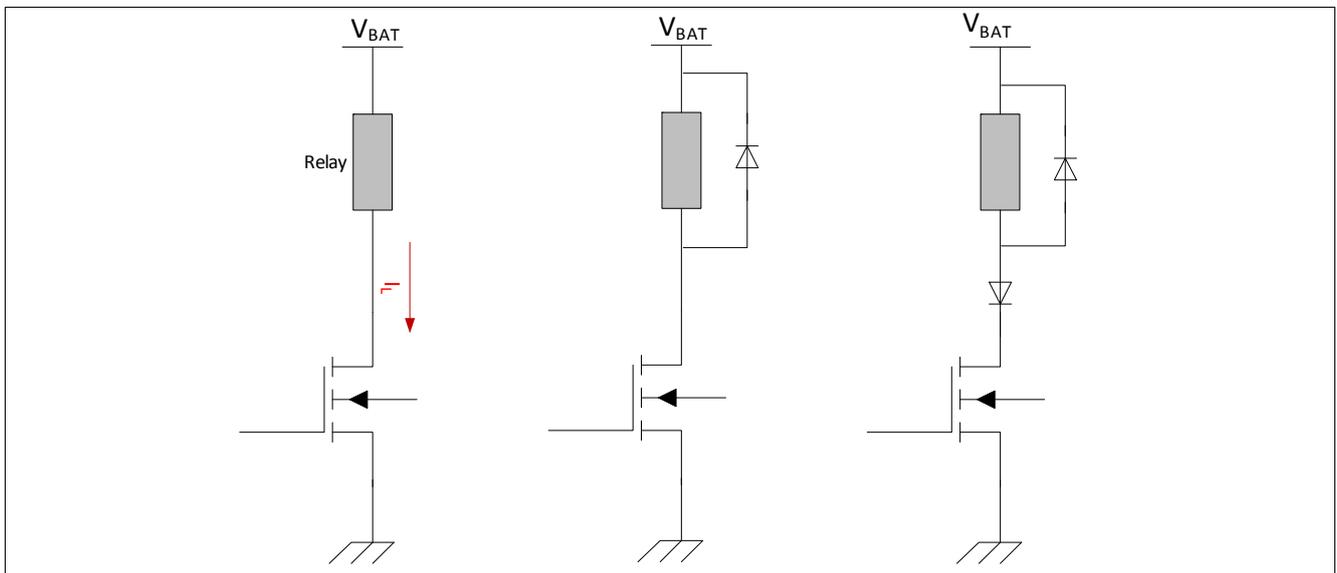


Figure 26 Different Relay Configurations with HITFET+

The configuration in the center of **Figure 26** shows the use of a freewheeling diode as compared to the one on the left without any freewheeling diode. The configuration on the right has an extra diode to prevent inverse current.

**Load And Application**

**5.4 Number of Activations**

The total number of activations (brake pedal depressed, low beam activation, compressor activation, etc.) depends largely on the habits of the vehicle driver. This does not including extra switching done by the ECU e.g. PWM, software retry strategies etc. The exact mission profile is usually provided by the OEM, but nevertheless loads can generally be placed in one of three categories as defined in the table below.

**Table 3 Load Activations per Engine Ignition**

No. of Activations	No. of Activations/Ignition	Average Activation Time	No. of Activation/Year
High	30	<1min or>1min	220 000
Mid	1 or 2	<1min or>1min	15 000
Low	1/3	>1 min	2500

**5.5 Wiring**

Four parameters are needed to define a wire: diameter, length and core and insulator materials. The diameter and the length determine the electrical characteristics ( $\Omega/\text{km}$  and  $L_{\text{cable}}/\text{km}$ ). The material and the environment determine the maximum current.

**5.5.1 The Wire as a Parasitic Electrical Load**

Although the wire is not a load, it has to be considered in automotive applications during the design phase. Wires offer a benefit to the system by limiting surge currents such as bulb lamp inrush current thanks to parasitic inductance ( $L_{\text{wire}}$ ), as well as resistive ( $R_{\text{wire}}$ ). The wire limits the current. On the other hand, the inductive energy stored in the cable is sometimes not negligible, especially for long wire harness found in truck or trailer application.

**5.5.2 Maximum Current in a Wire**

Wires require protection from high temperature induced by excessive current. The maximum current which can flow in the wire is time dependent and defined by a square law function  $I^2t = \text{constant}$ . The maximum current the wire can handle is limited by the insulation material. The OEM defines the wires to be used in a vehicle and this information is usually kept confidential. **Figure 27** shows an example of the current time coupling limitation of a wire as a function of the time.

The maximum current in a wire is defined by a thermal law. This is constant, as previously stated on the insulation material and also neighboring cables. For example, a wire within a group of 20 wires in a wire harness will have a lower maximum current rating than the same wire when it is not in a group. Infineon considers a reduction of 40% of the nominal current typical.

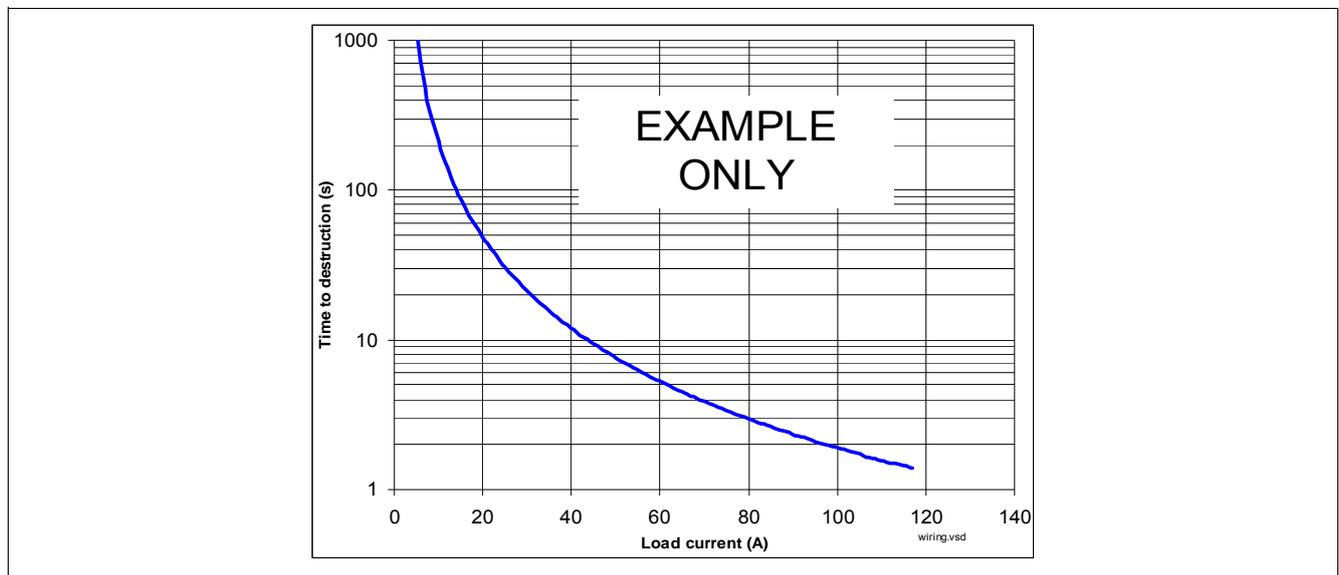
**Load And Application**

**Table 4** sums up the types of wire often use in an automotive environment. Note that these values are indicative and must be cross-checked with the application and the OEM.

**Table 4 Wire Characteristics as a Function of Diameter**

Cross section (mm <sup>2</sup> )	Gauge (AWG) <sup>1)</sup>	Impedance (Ω/km)	Inductance (mH/km)	Max DC current (A) <sup>2)</sup>
50	0	0.4	1.1	228
25	3	0.8	1.16	150
10	7	1.9	1.20	85
6.0	9	3.1	1.25	60
4.0	11	5	1.30	45
2.5	13	7.6	1.36	34
1.5	15	12.7	1.4	24
1.0	17	18.5	1.45	19
0.75	19	24.7	1.49	16
0.50	20	37	1.55	12
0.30	21	56	1.65	9

- 1) Approximation only.
- 2) Assuming  $T_{ambient} = 85^{\circ}C$  and wire alone in free air. Approximation only.



**Figure 27 Example of Current Limitation by Wire Harness**

The maximum current in the wire is a thermal law. This constant depends, as previously stated on the insulation material and also neighboring cables. For example, a wire within a group of 20 wires in a wire harness will have a lower maximum current rating than the same wire when it is not in a group. Infineon considers a reduction of 40% of the nominal current typical.

Typical wire characteristics are given as a function of diameter/cross-sectional area and can be provided by the wire manufacturer.

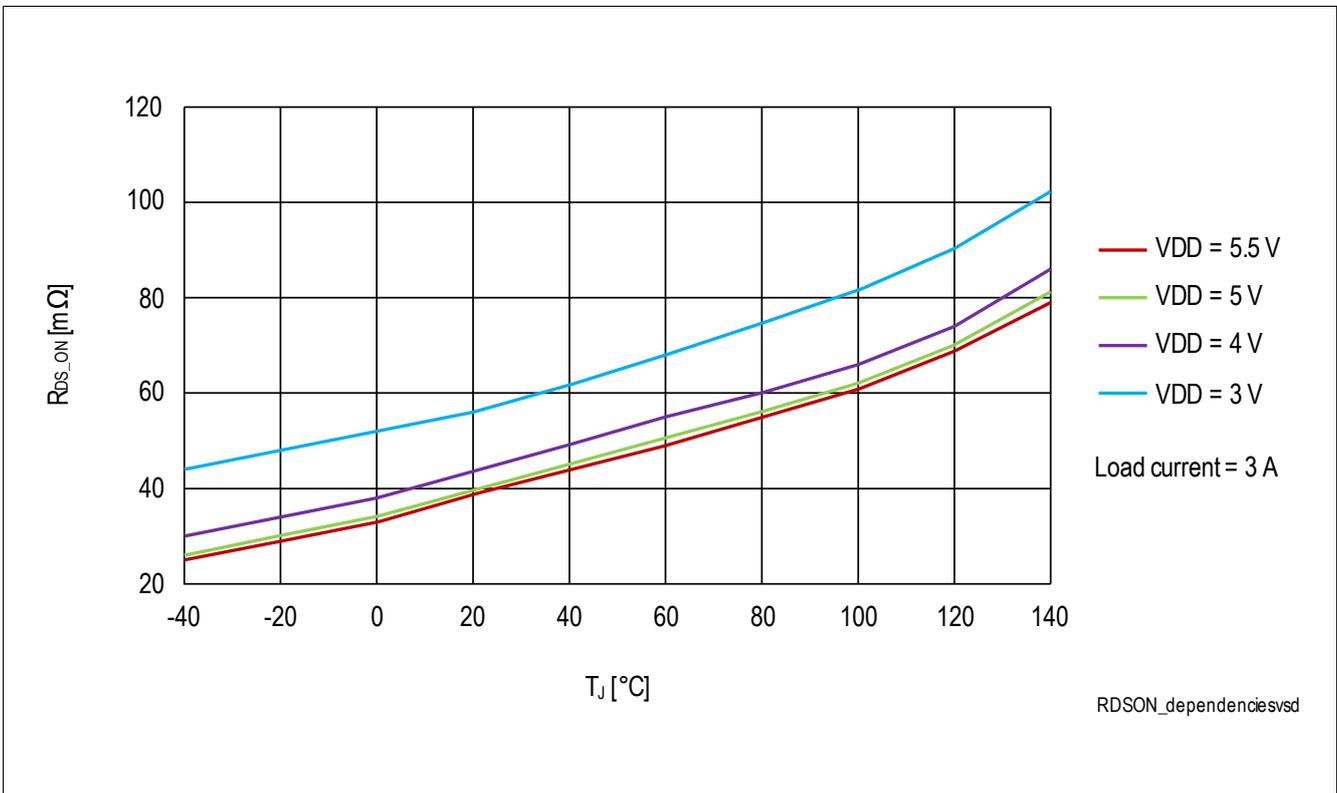
Power Stage

## 6 Power Stage

The power stage of HITFET+ is a low side switch consisting of a N-channel vertical Power MOSFET. The capability of this power element to pass current can be expressed in terms of its drain-source resistance in on-state,  $R_{DS(ON)}$ . The smaller the  $R_{DS(ON)}$ , the higher the current capability.

### 6.1 Power Element

As mentioned before, the capability of the power element is defined by its  $R_{DS(ON)}$ . For HITFET+,  $R_{DS(ON)}$  depends on both the supply voltage and the junction temperature  $T_J$ . **Figure 28** shows these dependencies for BTF3050TE. Hence,  $R_{DS(ON)}$  indication in the device naming is defined as the maximum  $R_{DS(ON)}$  measured at  $T_J=150^\circ\text{C}$



**Figure 28** Typical  $R_{DS(ON)}$  of BTF3050TE as a Function of Temperature and Supply Voltage

Power Stage

6.2 Switching

To understand how the slew rate control works, first let us have a look at the different stages of switching with a HITFET+. **Figure 29** shows the switching characteristics using a resistive load. Since this is a low side switch,  $V_{OUT}$  drops to almost zero volts during switch on and rises back to  $V_{BAT}$  during switch off. As can be seen from **Figure 29**, there are three stages of switching. The first stage is the delay after the  $V_{IN}$  goes up. The second stage is a fast drop in voltage and the third stage is a slower drop. The second part is where the slew rate (SR) is define as a voltage change where the voltage drops from 90% of  $V_{BAT}$  to 50% of  $V_{BAT}$ .

$$\left(\frac{dV}{dt}\right) = \frac{(V_{OUT(90\%)} - V_{OUT(50\%)})}{|(t_{OUT(90\%)} - t_{OUT(50\%)})|} \tag{6.1}$$

The third part is a slower voltage drop after the voltage has dropped by more than 50%. At switch off, the opposite happens. Initially, there is a voltage rise delay followed by a slow rise in voltage up to around 50%, followed by the slew rate voltage rise.

Besides the slew rate, the data sheet also defines the delay time ( $t_{dON}/t_{dOFF}$ ) as the time it takes for the voltage to drop to 90%-/rise to 10% of  $V_{BAT}$  depending on if it is switch on- or switch off operation. And fall/rise time ( $t_f/t_r$ ), i.e. the time it takes for the voltage to fall/rise from 90%/10% of  $V_{BAT}$  to 10%/90% of  $V_{BAT}$ .

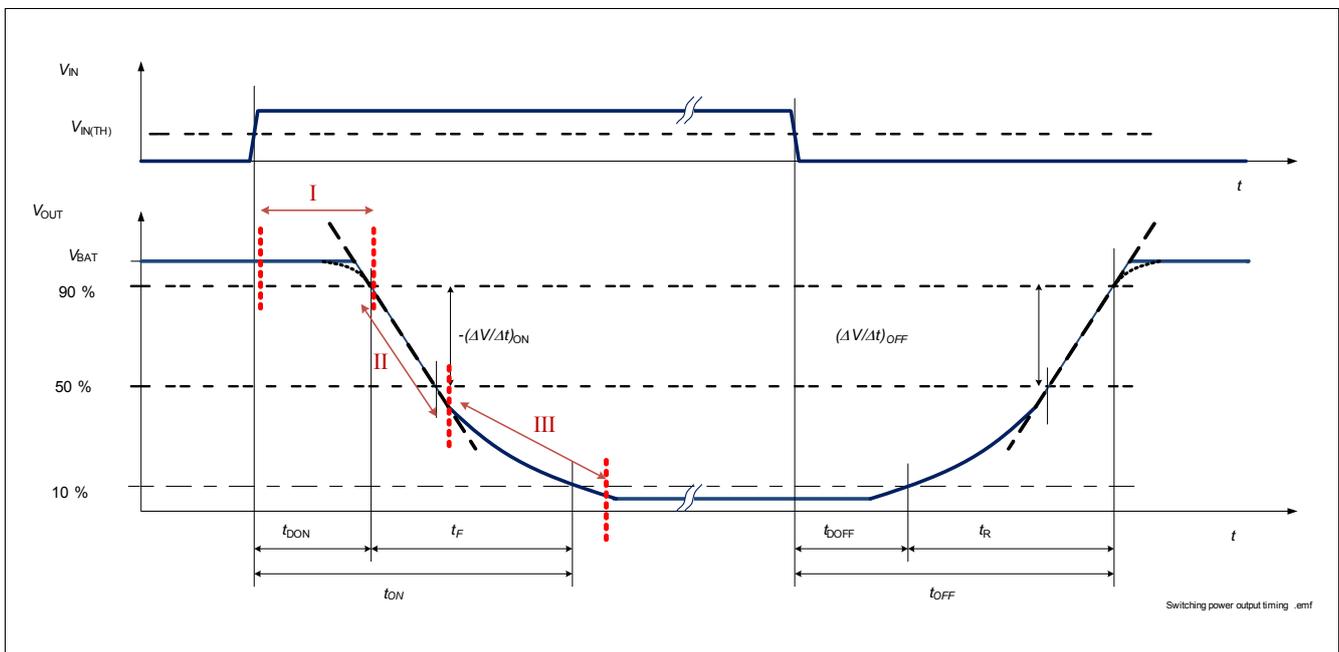


Figure 29 Definition of Power Output Timing for a Resistive Load

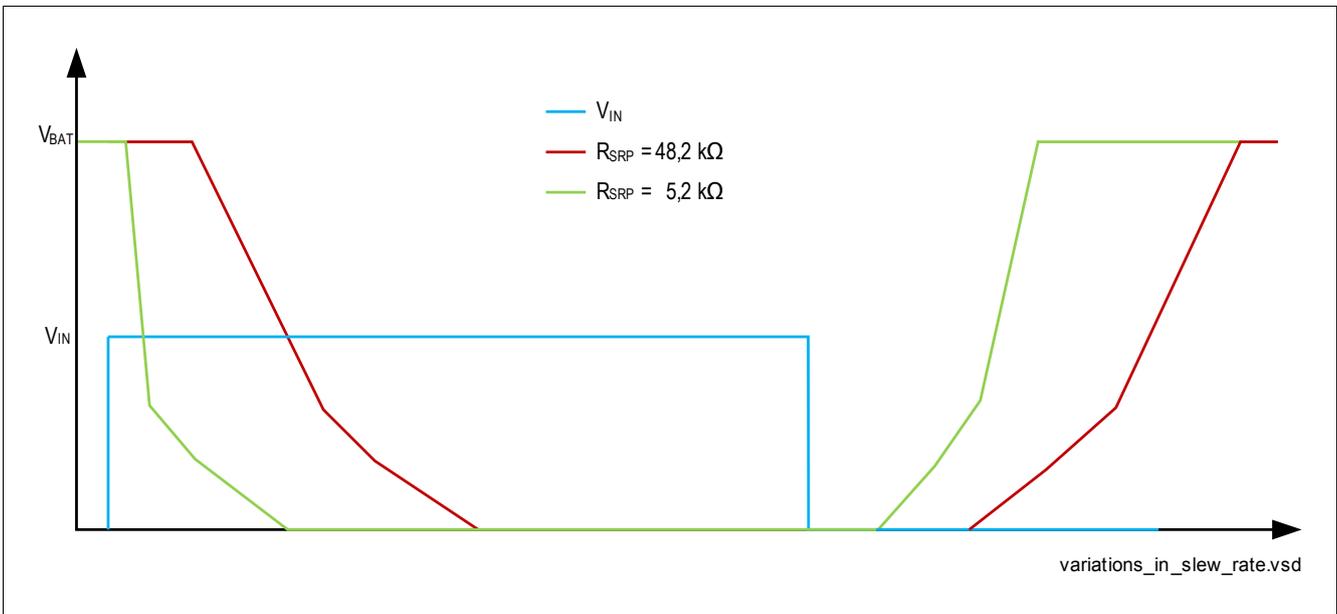
Power Stage

### 6.3 Slew Rate Control

Within the HITFET+ family, some devices are able to control the Slew Rate. This can be identified with their naming as BTF. Devices without this function are identified with the naming BTS.

In order to optimize electromagnetic emission, fast HITFET+ devices provide a SRP pin to control the switching speed of the MOSFET. By connecting an external resistor between SRP pin and GND, the switching speed can be adjusted. This allows for balancing between electromagnetic emissions and power dissipation especially when using them in PWM operation.

Shorting the SRP pin to GND represents the fastest switching which keeps decreasing as resistance between SRP pin and GND is increased. Open condition represents the slowest switching speed. **Figure 30** shows the variation in Slew Rates offered by the HITFET+.



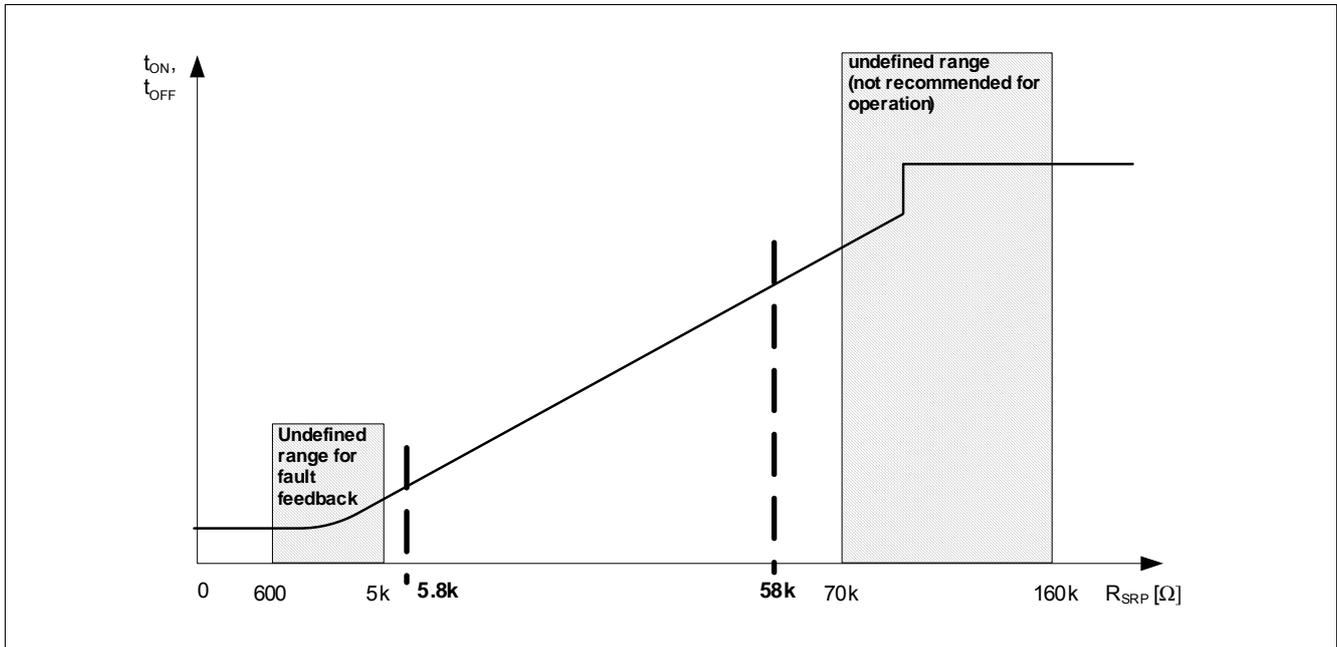
**Figure 30** Variation in Slew Rates

The accuracy of the switching speed depends on the precision of the external resistor used. Hence, it is recommended to use accurate resistors. Also, it is not recommended to change the slew rate resistance during switching ( $V_{DD} > V_{DD(UV\_ON)}$ ) as the resultant switching times will be undefined.

**Figure 31** shows the switching timing range in dependency of the  $R_{SRP}$

*Note: It is recommended for the BTF3050TE a maximum  $R_{SRP}$  to be 70kΩ due to its diagnosis features through the same pin mentioned in **Chapter 8**.*

Power Stage



**Figure 31 Typical Simplified Relation Between Switching Time and  $R_{SRP}$  Resistor Value Used on SRP Pin ( $V_{BAT} = 13.5V$ )**

**Slew Rate in Fault Mode**

The SRP pin has a hybrid function as input and output pin. In case of a latched fault caused by overtemperature, the SRP pin is internally pulled to  $V_{DD}$ . In this operation mode, the slew rate control with the  $R_{SRP}$  is ignored and a fault mode default slew rate, equivalent to a slew rate with  $R_{SRP} = 5.8k\Omega$ , is set. If the SRP pin is externally pulled above the normal SRP pin voltage  $V_{SRP(NOR)}$ , again the slowest slew rate (equivalent to  $R_{SRP} = 5.8k\Omega$ ) is set.

The fault mode can be reset by externally pulling down the  $V_{SRP}$  to 0V for a time greater than  $t_{RESET}$  as defined in the data sheets.

For more information on the Diagnostic function of the SRP pin through fault feedback and on using  $R_{SRP}$  above 70k $\Omega$ , refer to [Chapter 8](#).

Power Stage

6.4 Power Losses Calculation

Switching of a MOSFET can be represented by a load line on  $V_{DS}$  vs  $I_L$  curve as shown in **Figure 32**. While switching ON, MOSFET parameters ( $I_L$ ,  $V_{DS}$ ) move along the load line from B to A and from A to B while switching OFF. Moving from B to A in **Figure 32** implies decreasing  $R_{DS}$  while  $R_{DS}$  increases moving from A to B. Point A is the optimal operating point at which the device’s resistance is lowest and which is also rated as the  $R_{DS(ON)}$  in the data sheet.

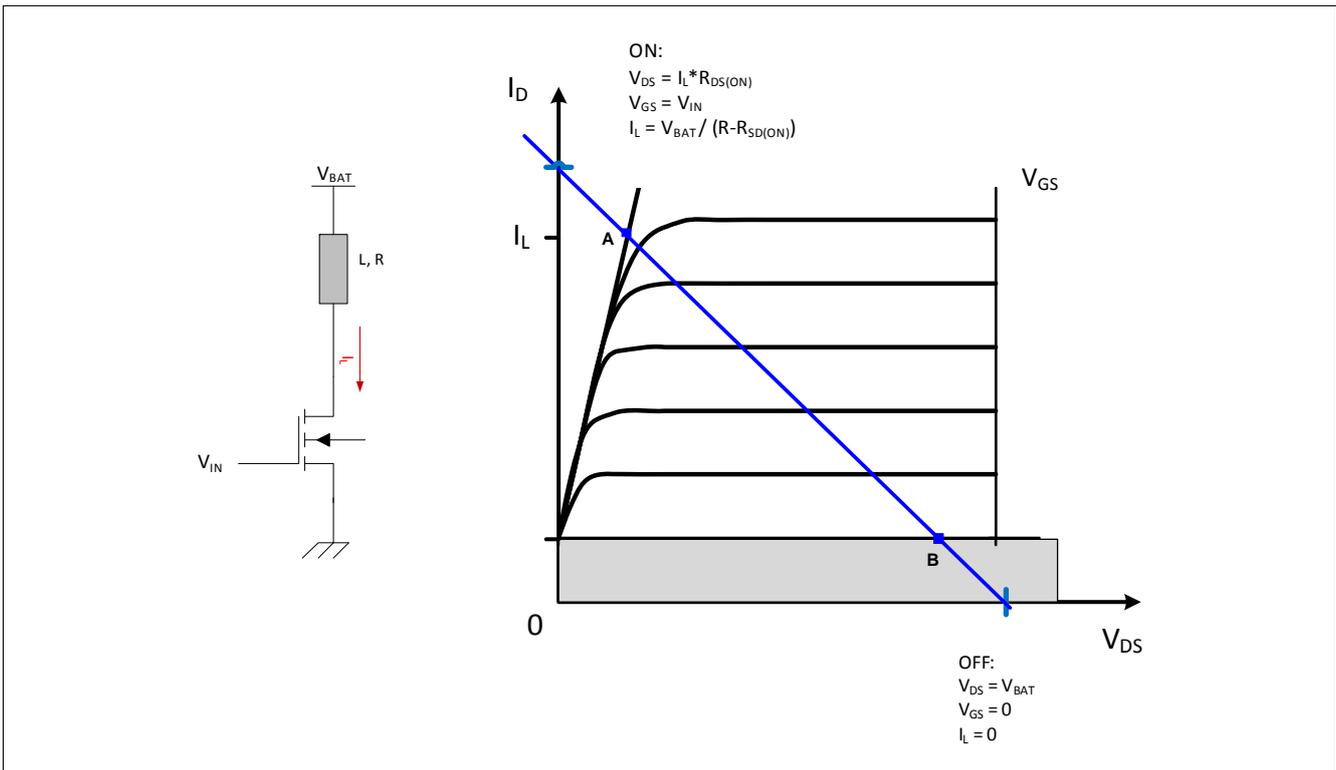


Figure 32 Switching Characteristics of a MOSFET with a Resistive Load

6.4.1 General Calculation

The power loss P in the device can be calculated as follows (assuming a resistive load  $R_L$ ). This works because any load will have a resistive component which will be responsible for major losses during operation. Capacitive and inductive components are not lossy during steady state but do affect losses during switching. However, they are dependent on the type of load and application and have to be calculated accordingly. A general loss calculation during usage of the switch can be performed using a resistive load  $R_L$ .

The instantaneous power in the switch is the result of the load current  $I_L$  multiplied by the drain to source voltage, which for a low side is  $V_{DS} = V_{OUT}$ . The resulting curve is shown in **Figure 33**. A good approximation is provided by the orange triangles and rectangle.

**Power Stage**

As the  $R_{DS}$  decreases during switching ON, the point where it corresponds exactly to  $R_L$  is where the triangle has its vertex at  $P_{MATCH}$ . In general, P is given by [Equation \(6.2\)](#). Refer to [Chapter 10.1](#) for a detailed calculation.

$$P = \frac{V_{BAT}^2 \times R_{DS}}{(R_L + R_{DS})^2} \tag{6.2}$$

Replacing  $R_{DS}$  by  $R_L$  to calculate  $P_{MATCH}$  gives us [Equation \(6.3\)](#)

$$P_{MATCH} = \frac{V_{BAT}^2}{4 \times R_L} \tag{6.3}$$

Energy dissipated during switch ON,  $E_{SON}$ , can thus be calculated as the area of the triangle given by [Equation \(6.6\)](#)

$$E_{SON} = \frac{1}{2} \times P_{MATCH} \times (t_{ON} - t_{DON}) = \frac{1}{2} \times P_{MATCH} \times t_F \tag{6.4}$$

The orange rectangle represents the energy  $E_{RON}$  lost during the ON state of the DMOS power element and is easily calculated by [Equation \(6.5\)](#)

$$E_{RON} = R_{DS(ON)} \times I_L^2 \times t_{RON} \tag{6.5}$$

And since the typical values of the rise time ( $t_r$ ) and the fall time ( $t_f$ ) are similar (see above),  $E_{SOFF}$  can be calculated similarly by [Equation \(6.5\)](#) replacing  $t_f$  with  $t_r$ . In conclusion, the power losses in the DMOS power element can be calculated by using:

$$P = \frac{(2 \times E_{SON} + E_{RON})}{t_{CYCLE}} \tag{6.6}$$

Power Stage

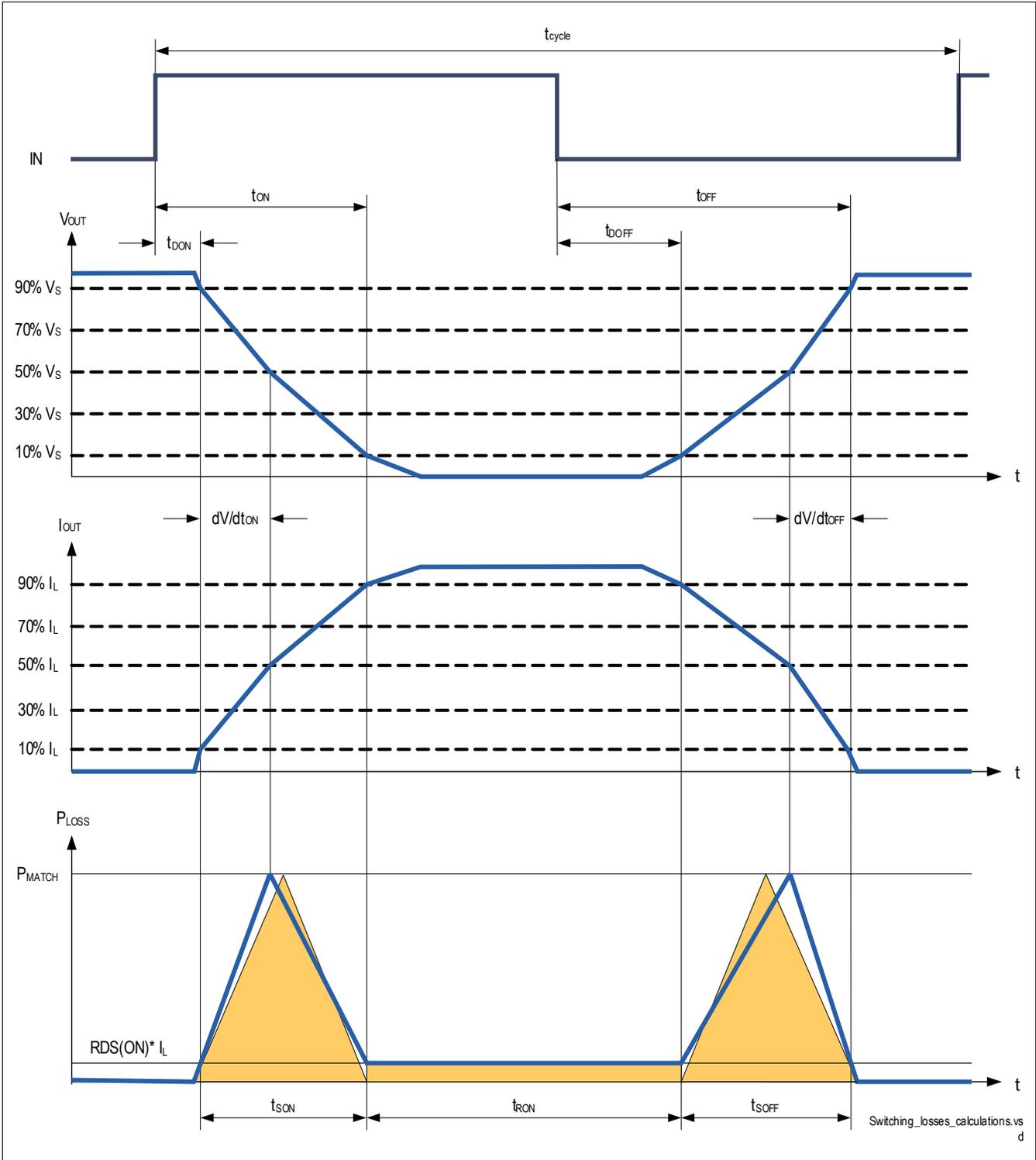


Figure 33 Power Losses Calculation

Power Stage

6.5 Switch Behavior with PWM Input

Pulsed Width Modulation is a special case where the cycle time,  $t_{CYCLE}$  is the inverse of the PWM frequency  $f_{PWM}$ . There are certain factors to be considered when using a PWM waveform. These are power loss, switching time and diagnostic limitations. Power loss and switching time are described below. Diagnostic limitations are described in [Chapter 8.1](#).

6.5.1 PWM Limitations due to Switching Time

As has been discussed before, switching ON of a HITFET+ device is defined when the  $V_{OUT}$  drops to 10% of  $V_{BAT}$ . Now,  $t_{DOFF}$  can be considered as the shortest ON time that the switch can reach ( $t_{RON}$ ). [Figure 34](#) shows what happens when the time to switch on the device ( $t_{ON}$ ) is given exactly as the input pulse to the device.

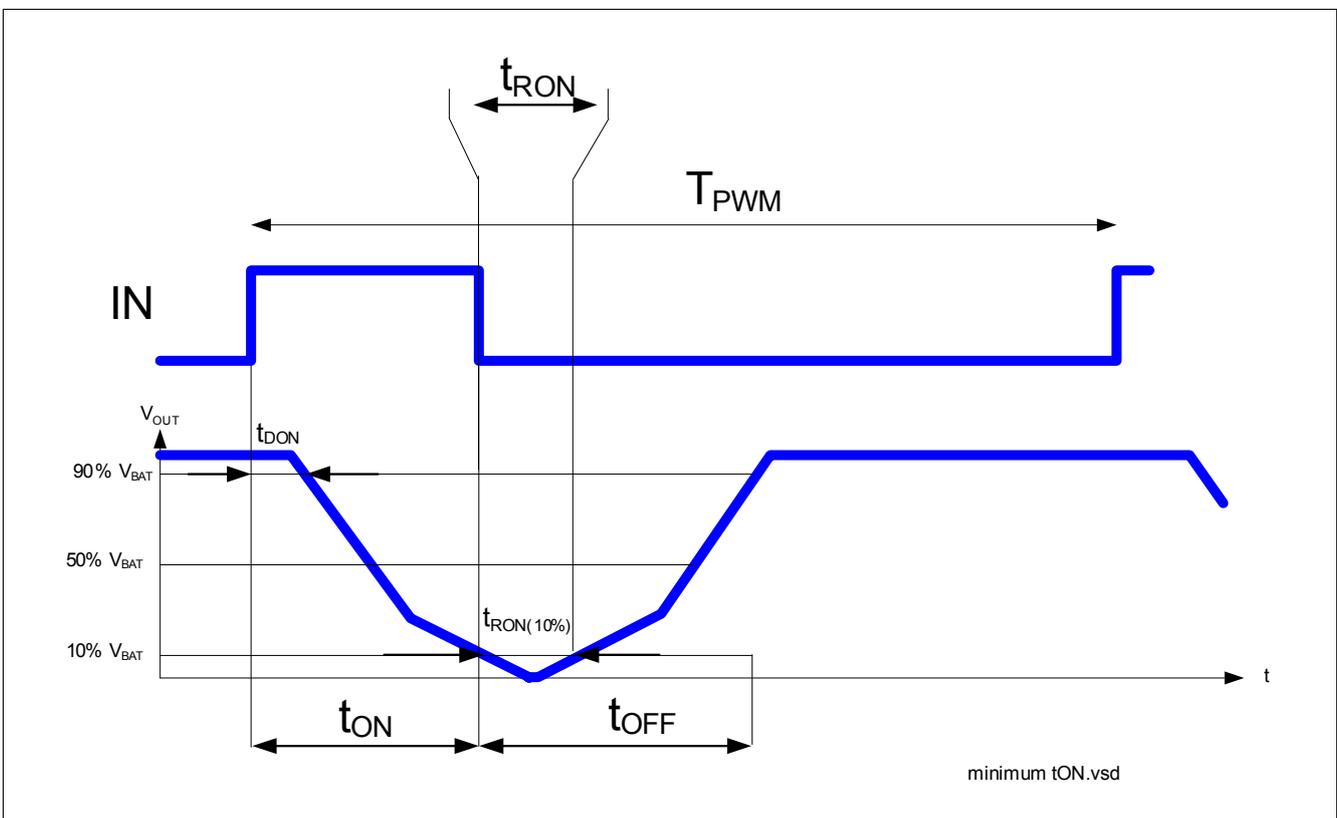


Figure 34 Minimum  $t_{RON}$

Since  $t_{ON}$  is defined until  $V_{OUT}$  drops to 10% of  $V_{BAT}$ , the resulting delay  $t_{RON(10\%)}$  in switching immediately at the point when  $V_{OUT}$  becomes 10% will be much less than the typical  $t_{DOFF}$ . This ON time,  $t_{RON(10\%)}$  is the shortest ON time achievable on the HITFET+ as the voltage does not exactly reach GND but keeps decreasing due to electrical inertia and then starts rising again for a short time (see [Figure 34](#)).

A similar example can be used to show that the minimum length of a Switch OFF Pulse through PWM, which will be limited by  $t_{DON}$ . And there will be a minimum OFF time that the device can reach,  $t_{ROFF(90\%)}$ .

**Power Stage**

To define the maximum and minimum duty cycle, refer again to **Figure 33**. The minimum duty cycle is determined by the shortest ON time of the IN pulse. The shortest ON time of the IN pulse will be the smallest time that switches on the HITFET+. Hence the shortest duty cycle can be defined by the  $t_{ON}$ . Similarly, the highest duty cycle will be decided by the shortest OFF time of the IN pulse. And the shortest OFF time of the IN pulse will be defined by the minimum time the device takes to switch off,  $t_{OFF}$ . Refer to **Table 5**.

**Table 5 BTF3050TE typical PWM Timing Limitation at  $R_{SRP} = 0k\Omega$**

Parameter	Symbol	Formula	$f_{PWM} = 100Hz$	$f_{PWM} = 1kHz$	$f_{PWM} = 10kHz$	$f_{PWM} = 20kHz$	Unit
Period	tCYCLE	$1/f_{PWM}$	10	1	0.1	0.05	ms
Min. duty cycle	dMIN	$t_{ON}/t_{CYCLE}$	0.05	0.5	5.3	10.6	%
Max. duty cycle	dMAX	$1-t_{OFF}/t_{CYCLE}$	99.95	99.5	94.7	89.4	%

**Attention:** *The limitations in **Table 5** are theoretical limits considering typical switching times. Working with a PWM duty cycle near these limits might lead to different results due to differences in test setup and production spread. It is recommended to consider a safety margin above the limits. To adequate even further the duty cycle and PWM to a specific application, interleaving method can be applied to the maximum values given.*

**6.5.2 PWM Limitations due to Power Losses**

The formula derived at the end of **Chapter 6.4.1** can be put to use here to calculate the power losses during PWM. At the end of **Chapter 6.4.1**, we got **Equation (6.6)**

$$P = \frac{(2 \times E_{SON} + E_{RON})}{t_{CYCLE}} \tag{6.7}$$

where  $t_{CYCLE} = 1/f_{PWM}$ . By replacing this, we obtain

$$P = 2 \times E_{SON} \times f_{PWM} + E_{RON} \times f_{PWM} \tag{6.8}$$

and expanding **Equation (6.8)** using **Equation (6.4)** and **Equation (6.5)**, we obtain

$$P_{PWM} = \frac{V_{BAT}^2}{4 \times R_L} \times t_F \times f_{PWM} + R_{DS(ON)} \times I_L \times t_{RON} \times f_{PWM} \tag{6.9}$$

**Equation (6.9)** can be used as the general formula for calculating power losses during PWM. The exact losses during usage will of course depend on the application and other lossy components present in the module. The application also determines how PWM is used. PWM may be used to control either the load current ( $I_L$ ), (for e.g. in LED and Relays) or to control the power (in lighting). Both cases are considered below:

Power Stage

6.5.2.1 PWM to Control Load Current

To use Equation (6.9) to calculate Power Losses, it is needed to modify it slightly.

Defining the duty cycle

$$d = t_{RON} * f_{PWM} \text{ and } I_L = V_{BAT} / R_L,$$

we can write:

$$P_{PWM} = \frac{I_L \times V_{BAT} \times t_F \times f_{PWM}}{4} + R_{DS(ON)} \times I_L \times d \tag{6.10}$$

Varying  $f_{PWM}$  and  $d$  shows how the  $P_{PWM}$  is affected by it. Figure 35 shows that PWM might now always be beneficial in terms of saving power. As can be seen, a combination of PWM frequency and duty cycle ensures power efficient current control.

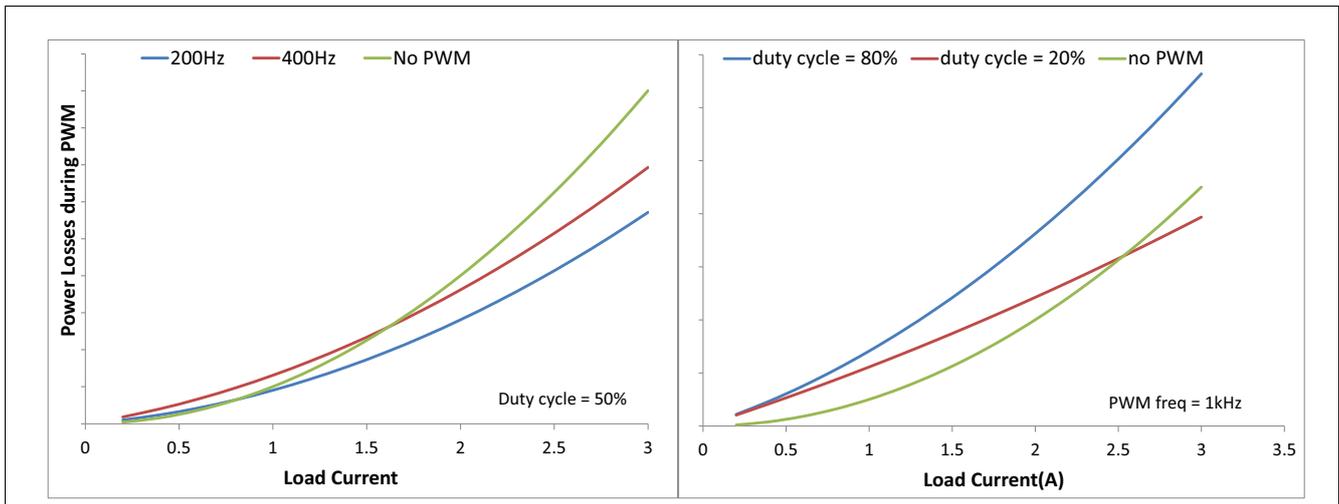


Figure 35 Power Losses due to PWM Control of Current with BTF3050TE

Power Stage

6.5.2.2 PWM to Control Power

This is typically deployed for light bulbs which are usually rated by max. power and take up a lot of it during starting. A  $V_{P_{PWM}}$  is defined as the optimum voltage at which PWM should be run so as to control the power. Refer to Load and Application chapter to understand how the next equation is derived. The relation of  $d$  and  $V_{P_{PWM}}$  is given by:

$$d = \frac{V_{P_{PWM}}^2}{V_{BAT}^2} \tag{6.11}$$

Using this in Equation (6.9) gives us the following equation:

$$P_{P_{PWM}} = \frac{V_{BAT}^2 \times t_F \times f_{P_{PWM}}}{4 \times R_L} + \frac{R_{DS(ON)} \times V_{P_{PWM}}^2}{R_L^2} \tag{6.12}$$

Using this equation with BTF3050TE and a 21W light bulb gives us Figure 36. It shows PWM control voltage of 13V with PWM frequencies of zero, 200Hz and 400Hz to give an idea of how PWM control will actually work. It is clear from these graphs that BTF3050TE has higher losses at higher frequencies and again, a correct combination of supply voltage and frequency is required to achieve efficient control.

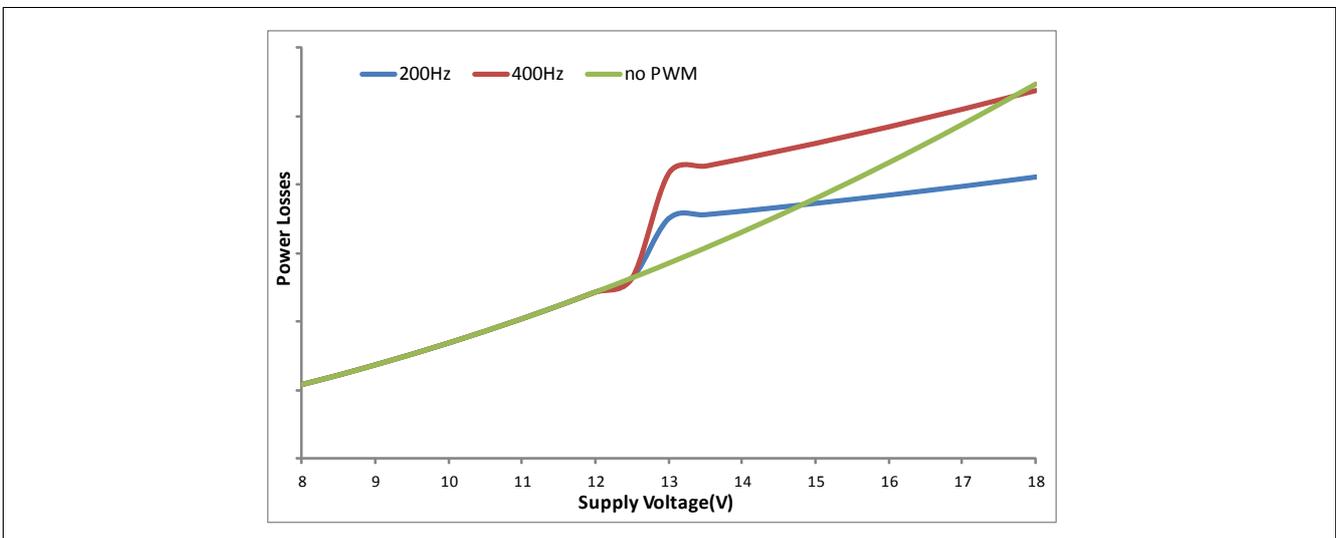


Figure 36 BTF3050TE Power Losses in PWM with a 21W Bulb Load.  $V_{P_{PWM}} = 13\text{ V}$

Power Stage

## 6.6 Thermal Considerations

Thermal considerations are important as they define the maximum functional power that can be dissipated by the device. Reaching temperature limits of the device triggers protection which will shut down the device. Hence it is important to consider thermal limitations of the device along with the application to ensure the application runs smoothly.

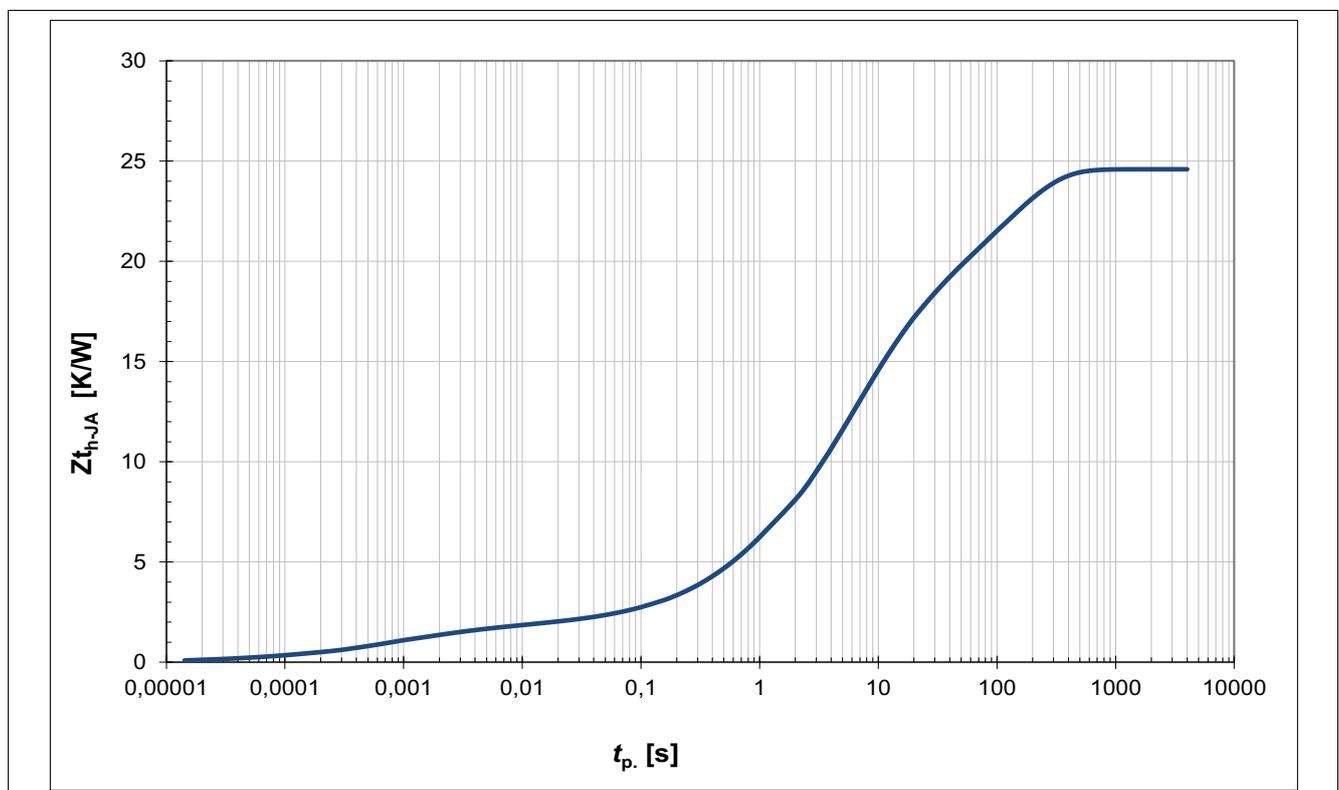
### 6.6.1 Maximum Junction Temperature

HITFET+ devices are embedded in exposed pad packages which offer excellent thermal resistance ( $Z_{thJC}$  characteristics) between junction and the case compared to non-exposed packages.

HITFET+ should be kept below a maximum junction temperature  $T_{J(max)} = 150^{\circ}C$ . The formula below expresses this constraint mathematically.

$$P_{MAXTJ} = \frac{T_{J(max)} - T_{AMB}}{R_{thJA}} \tag{6.13}$$

Here,  $T_{AMB}$  is the ambient temperature at which the application is running.  $R_{thJA}$  for HITFET+ depends on the module design (cooling, type of PCB etc.). For a 1s0p board and a  $T_{AMB} = 85^{\circ}C$ , we can calculate a  $P_{MAXTJ}$  of 1.7W. This is a rough calculation guideline. In actual application  $R_{thJA}$  will also change depending on the pulse length of the current as shown in [Figure 37](#).



**Figure 37 Typical Transient Thermal Impedance  $R_{thJA}$  at  $T_{AMB} = 85^{\circ}C$  for BTF3050TE. Graph Drawn According to Jedec JESD51-3 at Natural Convection on FR4 2s2p Board. The Device is Dissipating 1W of Power.**

Power Stage

Figure 38 and Figure 39 show the cross-section and layout of a 2s2p board.

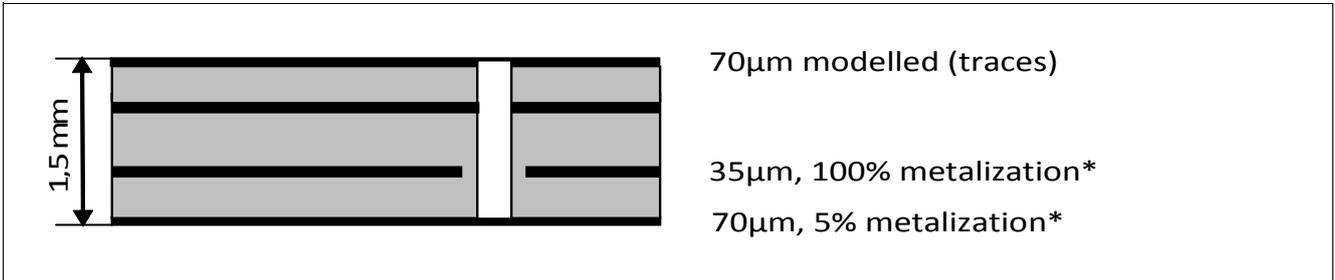


Figure 38 Cross Section of JEDEC2s2p

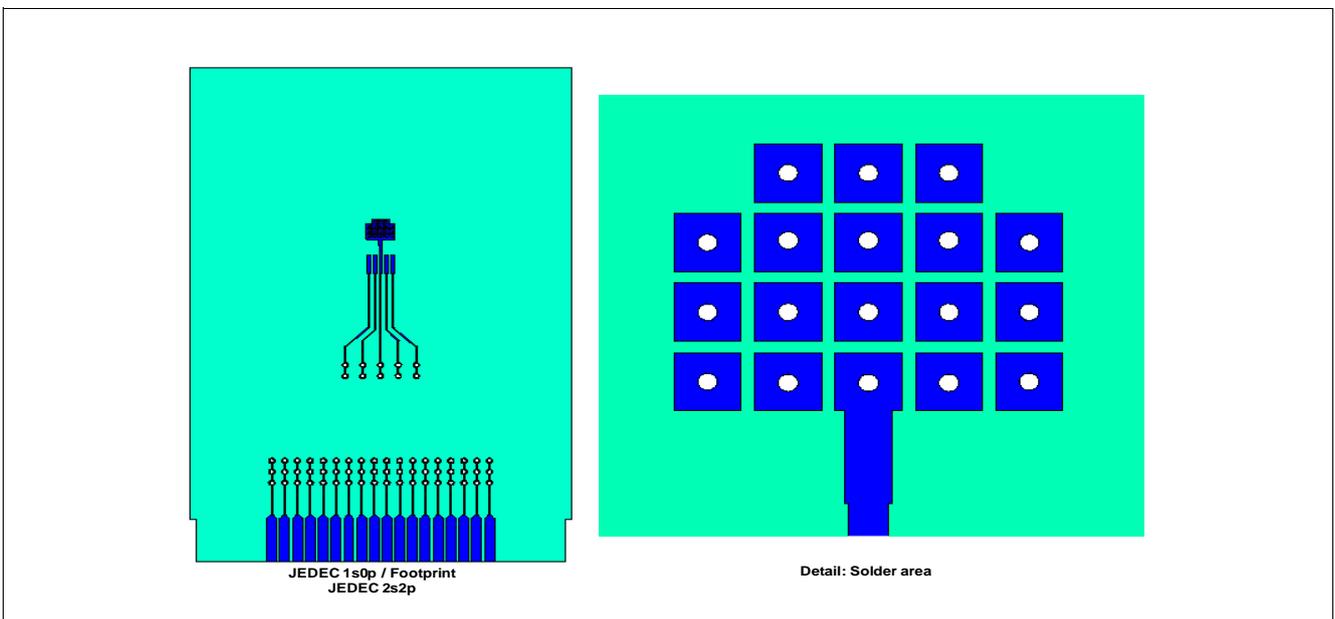


Figure 39 PCB Layout

Power Stage

6.7 Reverse/Inverse Current

A reverse battery situation means the OUT pin is pulled below GND potential to  $-V_{BAT}$  via the load  $Z_L$ .

An inverse current situation means the OUT pin is pulled below the GND potential by the current flowing from GND to OUT.

In both situations, the load is driven by a current through the intrinsic body diode of the MOSFET and all protection, such as current limitation, overtemperature or over voltage clamping, are inactive.

Figure 40 shows how a reverse diode can be used (B) to prevent inverse/current operation (A).

In both situations, power loss is defined by the current driven and the voltage drop on the body diode  $-V_{DS}$ .

During Inverse Current, an increased supply current  $I_{DD}$  flowing into  $V_{DD}$  needs to be considered. The device could be reset by inverse current too.

During inverse/reverse current situations, it is important to note that the parameters do not cross the absolute maximum ratings as given in the respective data sheets.

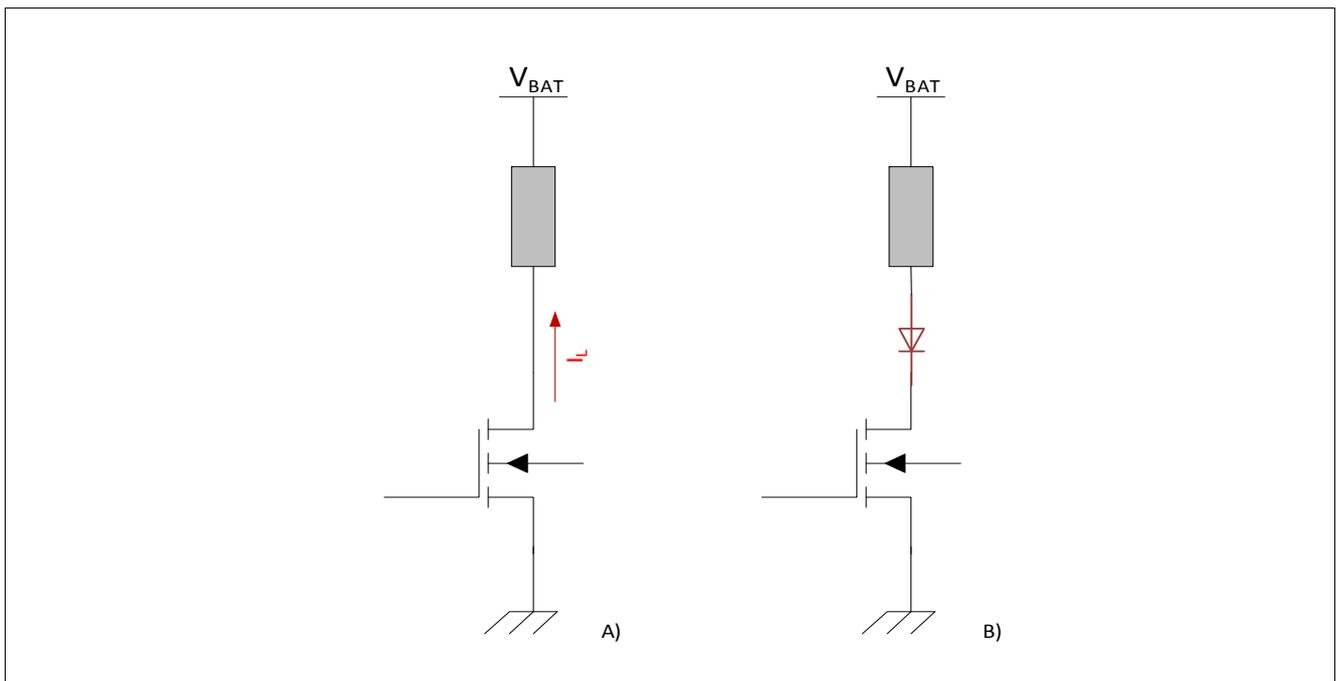


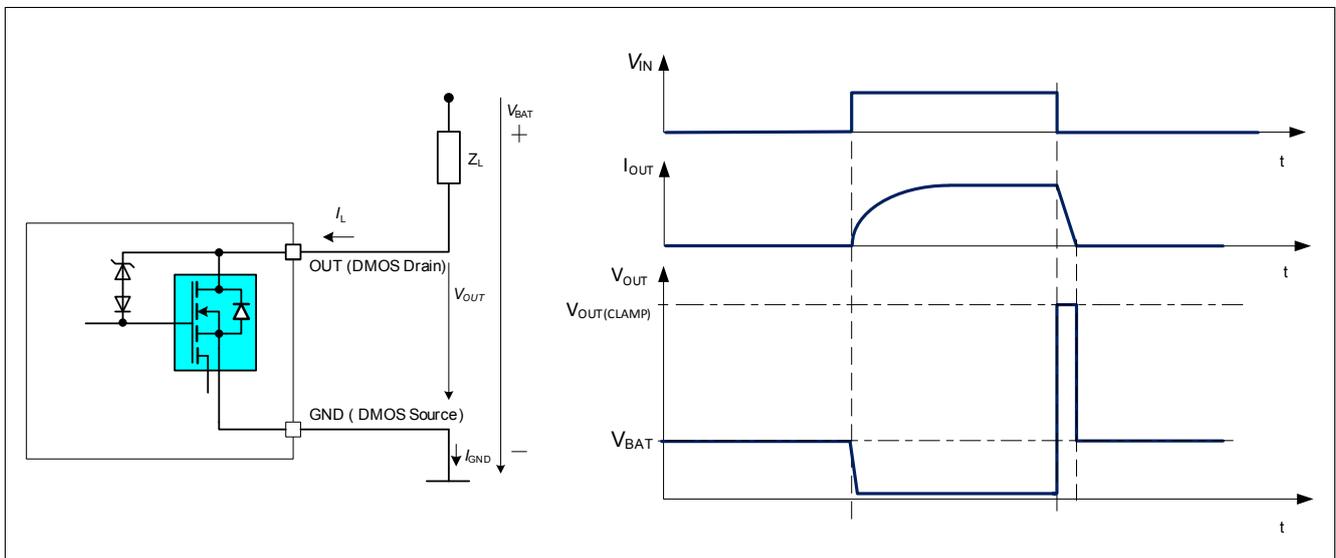
Figure 40 Inverse Diode

Power Stage

6.8 Output Clamping

When switching off inductive loads with low side switches, the drain-source voltage  $V_{OUT}$  rises above battery potential, because the inductance tends to continue driving the current (Refer to **Figure 41**). To prevent unwanted high voltages the device has a voltage clamping mechanism to keep the voltage at  $V_{OUT(CLAMP)}$ . During this clamping operation mode the device heats up as it dissipates the energy from the inductance. Therefore, the permissible inductance is limited.

**Figure 41** shows the output clamp circuitry for HITFET+ devices. The clamp circuitry is only responsible for clamping the  $V_{OUT}$  and not for providing an additional path to let out the demagnetization energy. The demagnetization energy is spent by the inductor in pushing up  $V_{OUT}$ .



**Figure 41 Output Clamp Circuitry and Switching**

Device data sheet mentions the minimum value of  $V_{OUT(CLAMP)}$  at which clamping becomes active and it has been designed such that it never reaches the technology breakdown limit.  $V_{OUT(CLAMP)}$  is also very stable with temperature as can be seen in **Figure 42**.

Power Stage

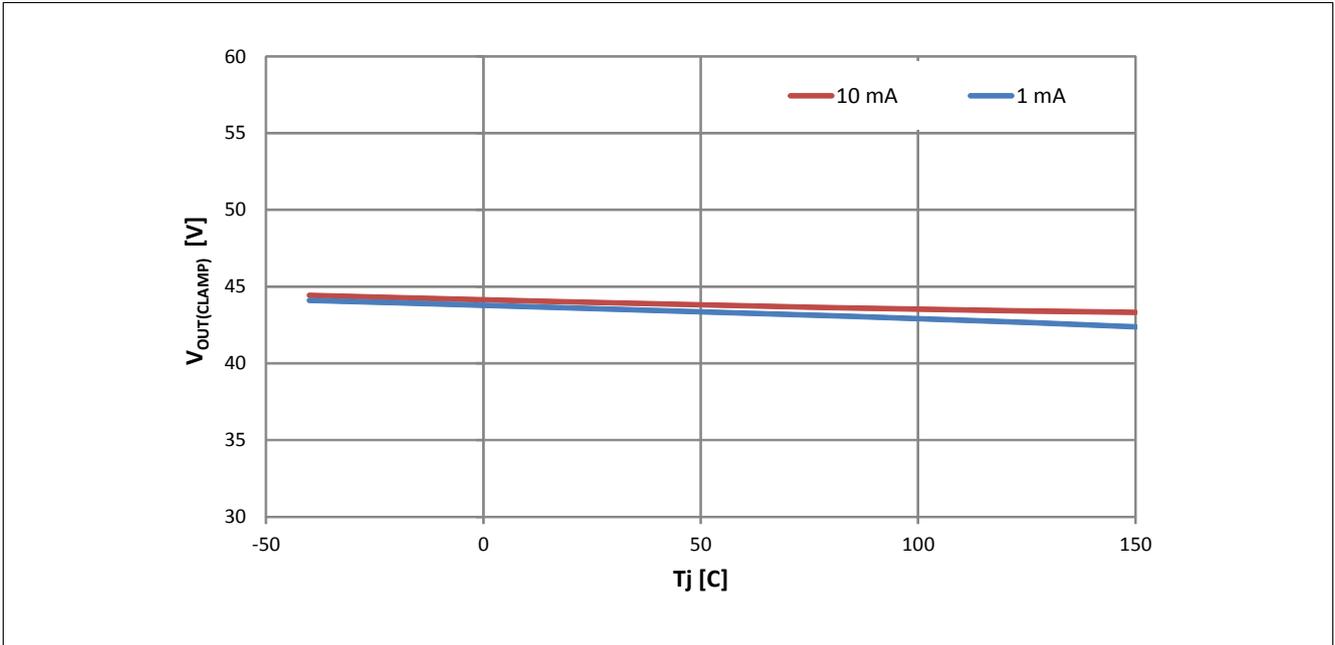


Figure 42 Typical  $V_{OUT(CLAMP)}$  vs.  $T_j$  for BTF3050TE

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**Protection**

## 7 Protection

The comprehensive set of protection functions is one of the most important features offered by HITFET+ switches. They are integrated and are designed to prevent IC destruction under fault conditions. Fault conditions are defined as conditions outside normal operation of the device. Protection functions are not designed for continuous repetitive operation. Also, protection functions are not available during a reverse/inverse current condition.

### 7.1 Over Voltage Clamping on OUT pin

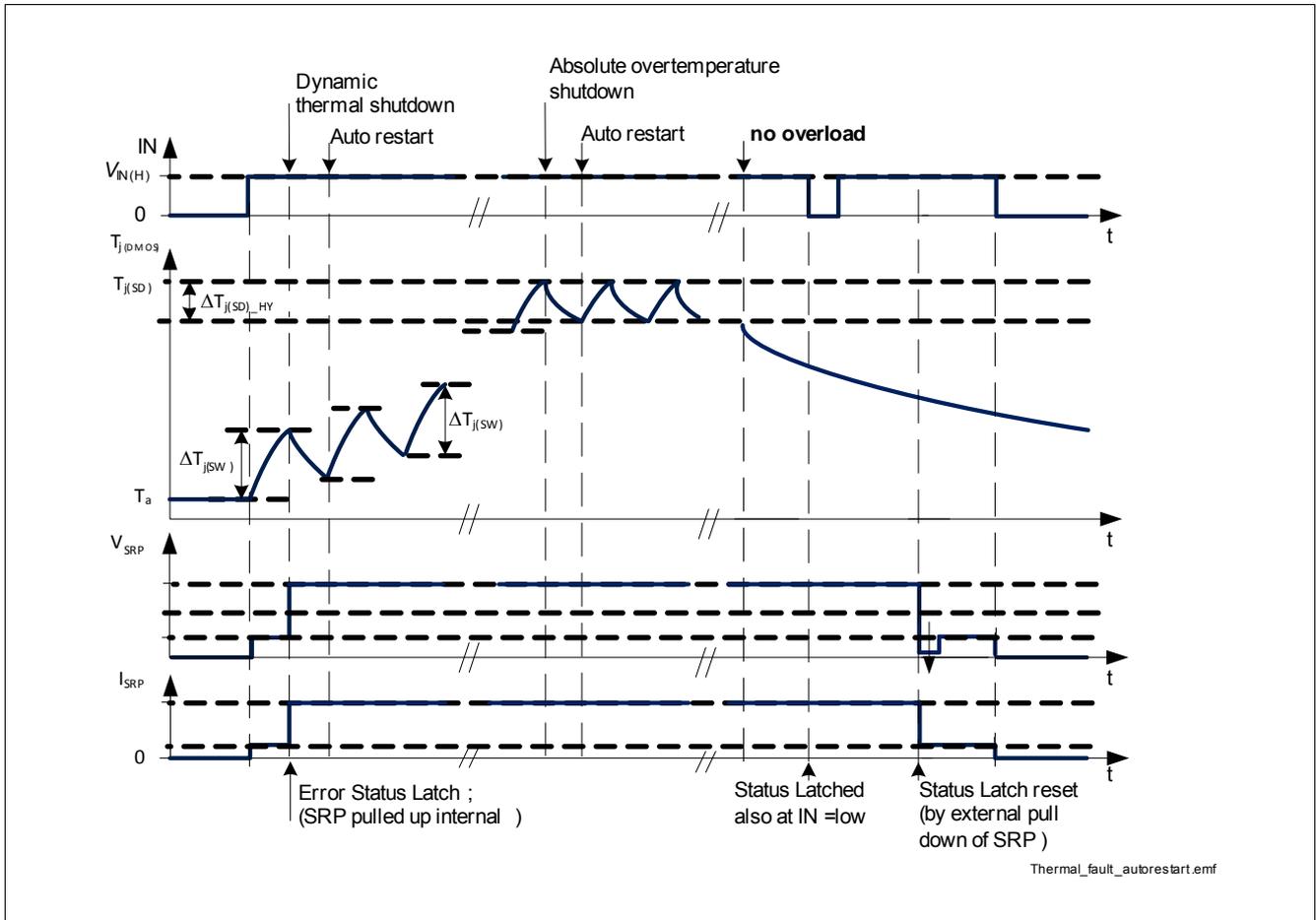
HITFET+ is equipped with a voltage clamp circuitry that keeps the drain-source voltage, specifically  $V_{OUT}$  at a certain level  $V_{OUT(CLAMP)}$ . Functioning of the clamping is defined in [Chapter 6.8](#) and energy considerations are included in [Chapter 5.3.1](#).

It is important to note that the overvoltage clamping overrules all other protection functions and power dissipations must be limited to stay below the maximum allowed junction temperature as discussed in [Chapter 6.6.1](#).

### 7.2 Thermal Protection

Thermal protection against overtemperature due to overload and/or bad cooling conditions. Two temperature sensors are integrated into the device to implement two kinds of thermal protection, absolute ( $T_{J(SD)}$ ) and dynamic ( $D T_{J(SW)}$ ) temperature limitation. Triggering either of these will cause the output to switch off. However, thermal protection has an automatic restart and the device will switch ON again after the drop in temperature is more than the thermal hysteresis ( $D T_{J(SD)_HYS}$ ).

Protection



**Figure 43 Thermal Protective Switch OFF Scenario in Case of Overload**

**Figure 43** shows how thermal protection switch OFF works showing how different parameters react. The moment dynamic protection ( $dT_{J(SW)}$ ) is triggered, the device shuts down with the fastest slew rate and  $V_{SRP}$  is pulled up internally to  $V_{SRP(FAULT)}$ . During this process, a fault current  $I_{SRP}$  has to be considered. The latched state is independent of the IN signal, providing a stable fault signal to be read out by a microcontroller. The latched fault signal needs to be reset externally by low signal ( $V_{SRP} < V_{SRP(RESET\_MIN)}$ ) at the SRP pin, provided that the junction temperature has decreased at least below the thermal hysteresis in the meantime. To reliably reset the latch the SRP pin needs to be pulled down with a minimum length of  $t_{RESET}$ .

As long as the fault signal is set and the SRP pin is not shorted to GND a fast default slew rate adjustment (like for  $R_{SRP} = 5.8k\Omega$ ) will be applied to the device. If the latched fault signal is not reset, the device logic stays active (also if IN = low), not entering the quiescent current mode and therefore reaching the upper limits of the normal supply current  $I_{DD}$ .

Also, important to consider while using BTF3050TE is the variation of  $T_{J(SD)}$  and  $dT_{J(SD)\_HYS}$  with the supply voltage  $V_{DD}$ . Although it is quite stable over  $V_{DD}$ , it is still important to understand the variation in case of poorly controlled supply voltage or for high power applications.

Protection

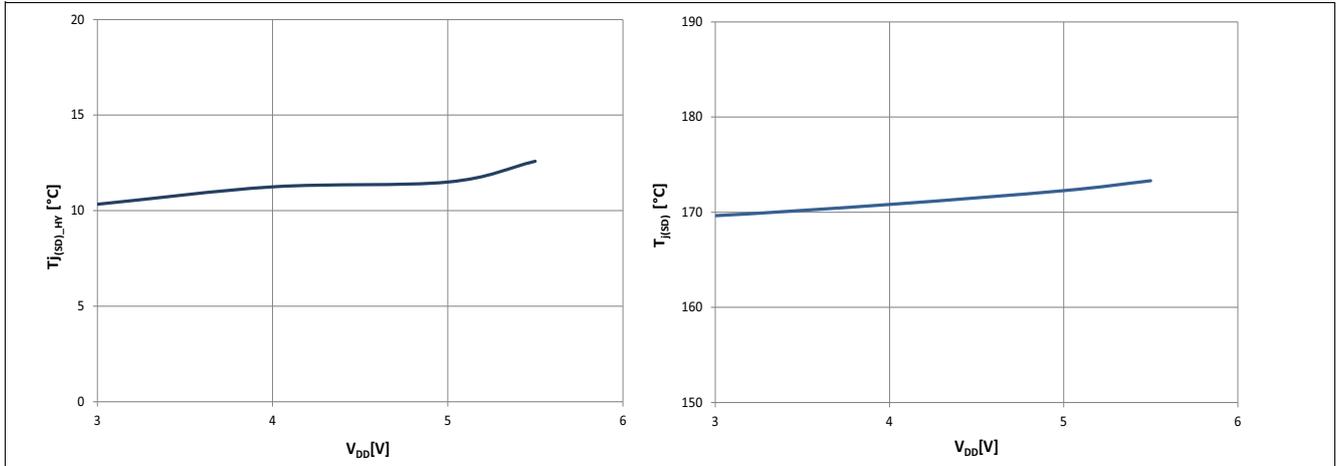


Figure 44 Typical  $T_{j(SD)}$  and  $T_{j(SD)\_HY}$  vs.  $V_{DD}$ .  $I_L = 10mA$  of the BTF3050TE

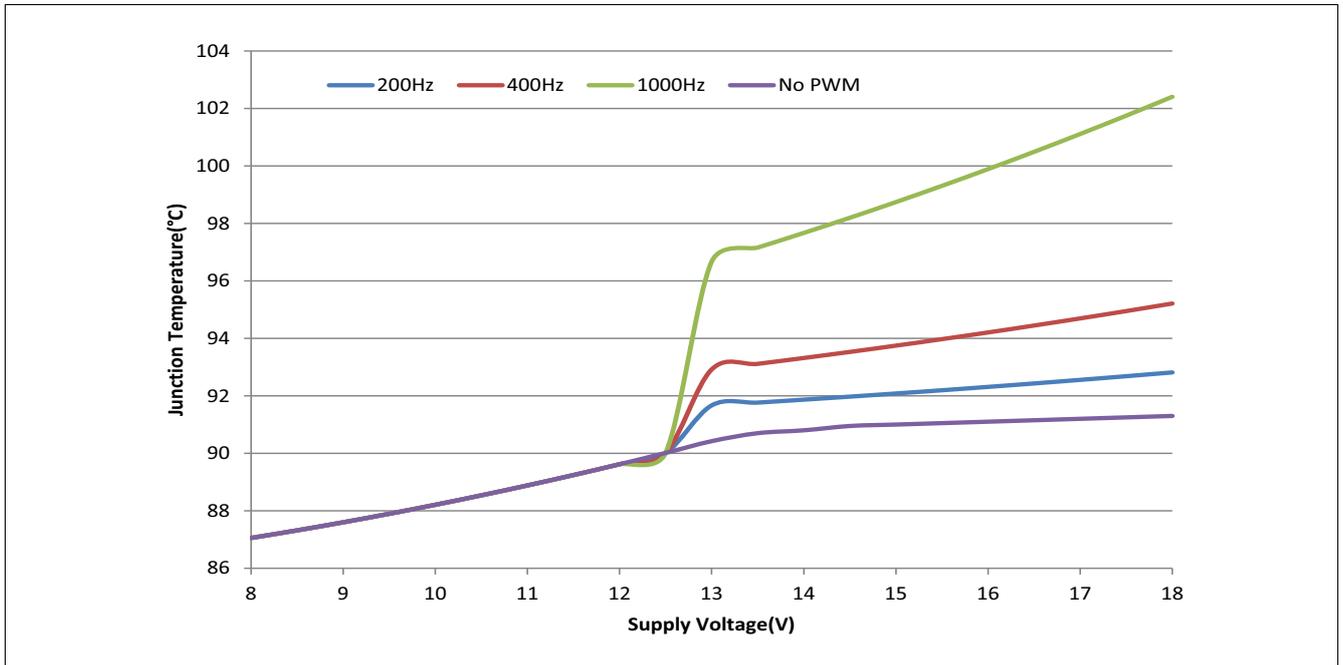
As a rough calculation, Equation (6.13) can be used to calculate the temperature that can be reached in an application. The trends of temperature will follow the same pattern as for the power losses because the relationship is linear. As an example, using Equation (6.12) for Power Losses due to PWM while regulating light bulb with Equation (6.13) will give us:

$$\left( \frac{V_{BAT}^2 \times t_F \times f_{PWM}}{4 \times R_L} + \frac{R_{DS(ON)} \times V_{PWM}^2}{R_L^2} \right) \times R_{thJA} + T_{AMB} = T_{J(SD)} \tag{7.1}$$

Figure 45 depicts Junction Temperature ( $T_j$ ) with the supply voltage and it can be seen clearly that it has the same characteristic as Figure 36.

As can be seen from Figure 43, that restart function will lead to thermal and voltage cycles on the device. The frequency of the thermal cycle will depend strongly on the cooling conditions of the application.

Protection

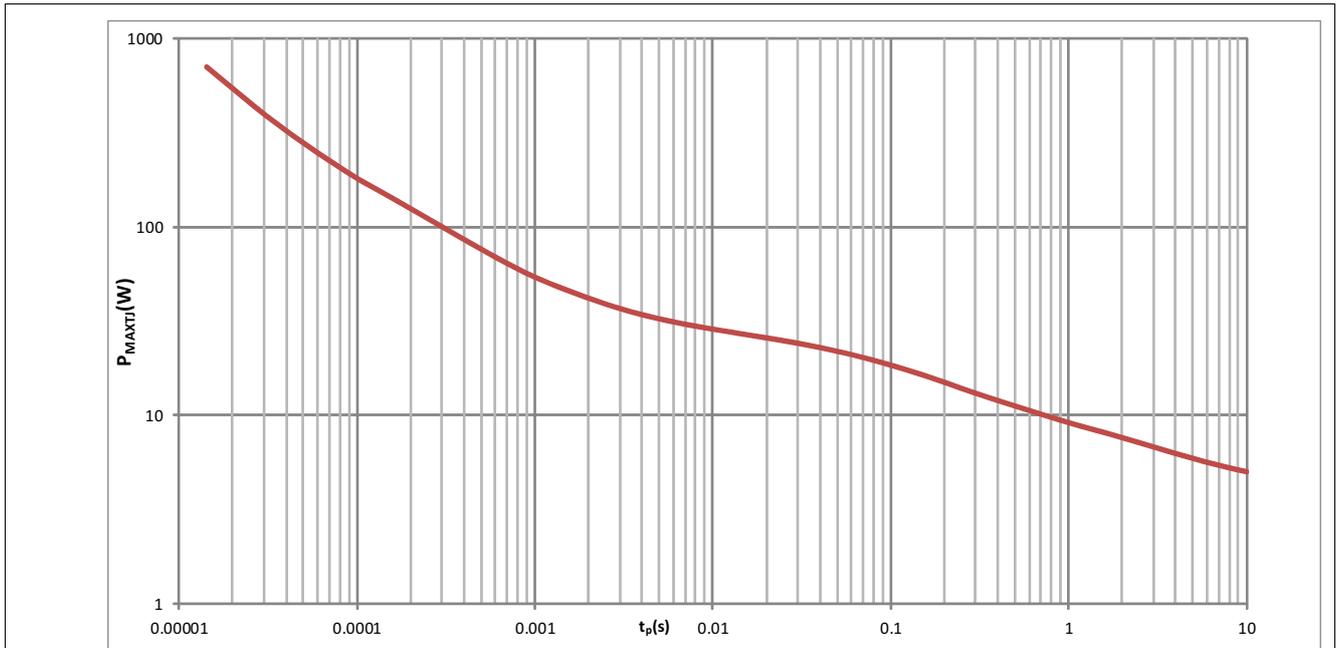


**Figure 45 Junction Temperature Rise in BT3050TE with PWM with 21W Bulb Load.**  
 $V_{PWM} = 13V, T_{AMB} = 85^{\circ}C$

Protection

7.2.1 Maximum Temperature Limitation

The device is qualified for junction temperature up to  $T_J = 150^\circ\text{C}$  continuous. This is the minimum temperature for the activation of the Temperature Protection Sensor. The thermal design must ensure that the device operates below this temperature based on [Equation \(6.13\)](#) or [Equation \(7.1\)](#). However, as discussed above,  $R_{thJA}$  also changes with the length of the pulse.



**Figure 46** Maximum Power Loss Allowed  $P_{MAXJT} = f(t_p)$ ,  $T_{AMB} = 85^\circ\text{C}$  for Typical Thermal Impedance  $Z_{thJA}$  for BTF3050TE

**Figure 46** shows max allowed power loss in the device with for the  $T_J$  to remain below  $150^\circ\text{C}$  so as to not trigger thermal protection. It has been calculated using [Equation \(6.13\)](#) and [Figure 37](#). Any pulse below the red curve can be safely assumed to not trigger overtemperature protection under the given conditions.

Protection

7.3 Overcurrent Protection (BTF devices only)

BTF3050TE and all BTF devices, provide a smart overcurrent limitation providing protection against short circuit conditions and any other increased current conditions, while also allowing load inrush currents higher than the current limitation level. To achieve this, the device has a higher current trigger level  $I_{L(LIM)\_TRIGGER}$  which triggers a lower current limitation level  $I_{L(LIM)}$ .

This enables the device to take currents higher than  $I_{L(LIM)}$  (overload condition) provided the device is not heated up so much that the overtemperature protection (OT) is triggered. In case of a short circuit,  $I_{L(LIM)\_TRIGGER}$  will be triggered, which will limit the current to  $I_{L(LIM)}$ . The reason for limiting instead of tripping is to enable HITFET+ devices to drive loads with high inrush currents.

Figure 47 depicts the functioning of the overcurrent protection for both overload (in green) and short circuit (in red) behavior.

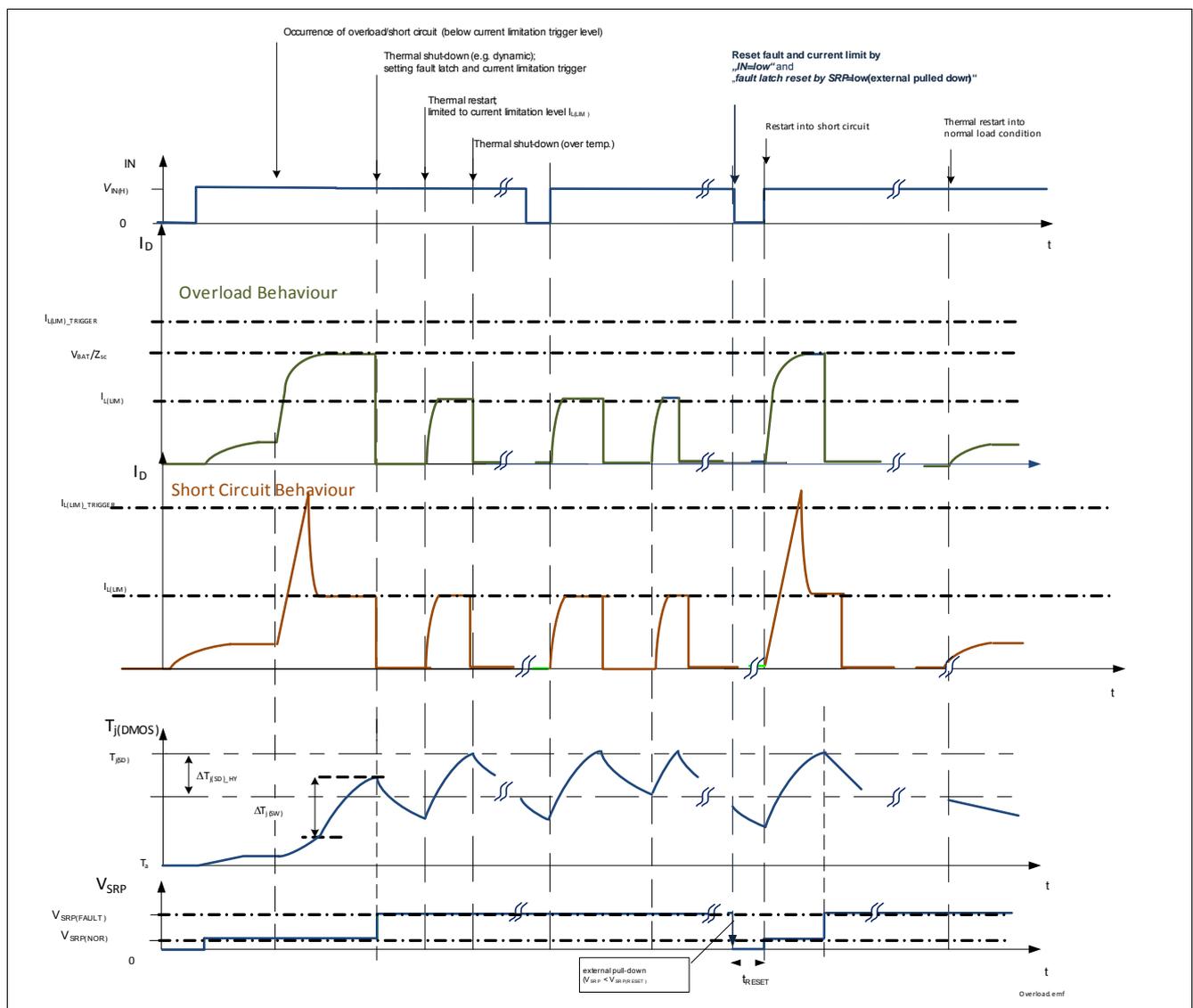


Figure 47 Example of Short Circuit Protection and Overload Behavior with Thermal Protection with Latched Fault Signal on SRP

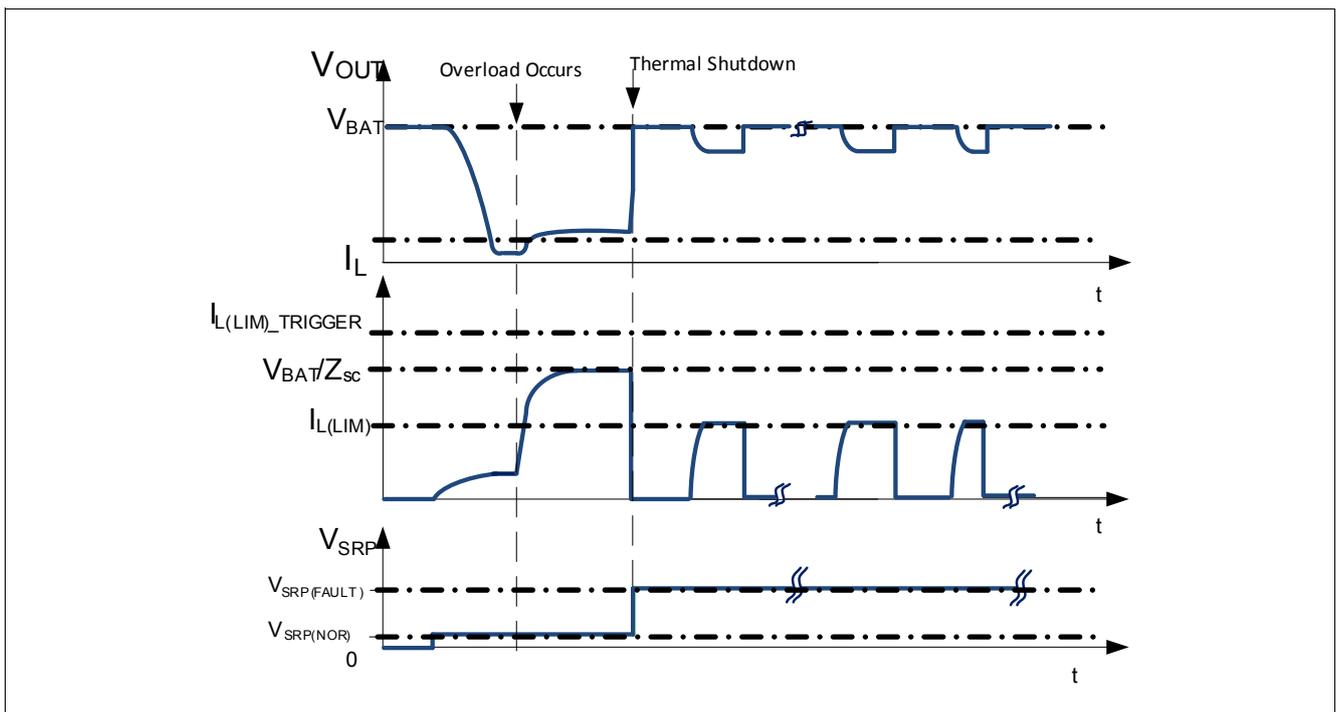
**Protection**

*Note: The time scale is not linear and not similar for short circuit and overload behavior. Real timing for both conditions will be application dependent.*

The current limitation trigger is a latched signal. It will only be reset by input pin (IN) low. The fault latch feedback has to be reset by pulling down the SRP pin (SRP-pin = low (below reset threshold) for  $t > t_{RESET}$ ). This means if the input stays high all the time during a short circuit the current will be limited to  $I_{L(LIM)}$  in the following pulses (during normal restart). It also means that the output current is limited to the current limitation level  $I_{L(LIM)}$  until the current limitation trigger is reset. For more details on how latching works, refer to [Chapter 8.1](#).

**7.3.1 Overload Condition**

Under an overload condition, a current higher than  $I_{L(LIM)}$  but lower than  $I_{L(LIM\_TRIGGER)}$  is flowing through the device. Overload can be both a fault condition or a required temporary phase when switching an output, e.g. inrush of a light bulb or a heating element. [Figure 48](#) shows the typical output voltage characteristics during overload with a resistive load.



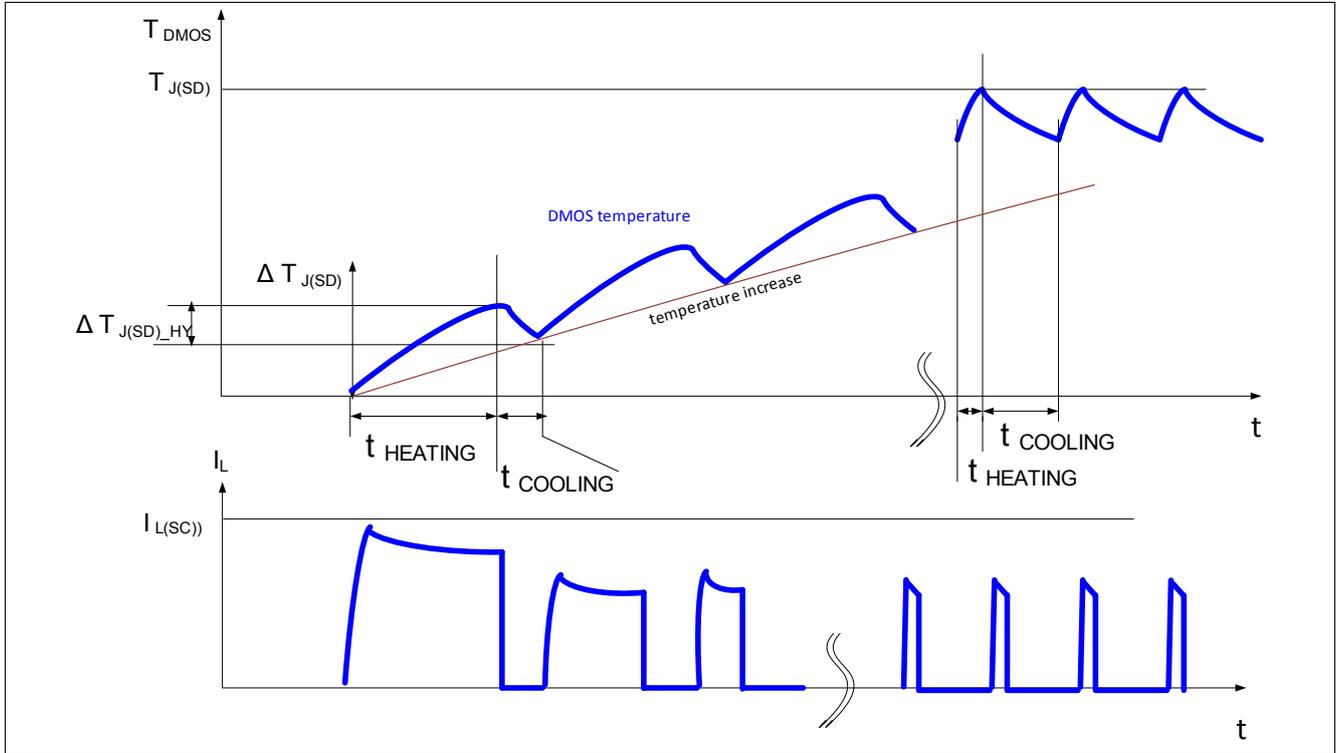
**Figure 48 Overload Behavior of  $V_{OUT}$  and  $I_L$  in BTF3050TE for Resistive Loads**

For overload behavior, the device will allow for current higher than  $I_{L(LIM)}$  as long as it does not heat up to trigger thermal shutdown. Once the device shuts down, it will be restarted after the temperature drops below the temperature hysteresis  $T_{J(SD\_HY)}$  but with the current now limited to  $I_{L(LIM)}$ .

[Chapter 7.2.1](#) provides information on how to calculate whether the overtemperature protection will be triggered or not during overtemperature condition. Repeated overtemperature condition will increase chip temperature and lead to faster heating times thus further stressing the device. Refer to [Figure 49](#).

**Protection**

For multiple restarts during overload, device temperature will slowly increase leading to faster heating time,  $t_{HEATING}$  and slower cooling time,  $t_{COOLING}$ . Both  $t_{HEATING}$  and  $t_{COOLING}$  strongly depend on module cooling conditions.



**Figure 49 Example of Overload Current Behavior during Thermal Shutdown for BTF Devices**

Protection

7.3.2 Short Circuit to Battery

For the short circuit, after the current is limited to  $I_{L(LIM)}$  the device starts heating up. When the thermal shutdown temperature  $T_{J(SD)}$  is reached, the device turns off. The time from the beginning of current limitation until the overtemperature switch off strongly depends on the cooling conditions.

A short circuit event is a very stressful event for the device. Toggling of short circuit current can reach very high frequencies. It can be beneficial to keep the number of restarts to a minimum so that the device's capability to handle short circuit event does not deteriorate. **Figure 50** shows the current and  $V_{OUT}$  profile during a short circuit.

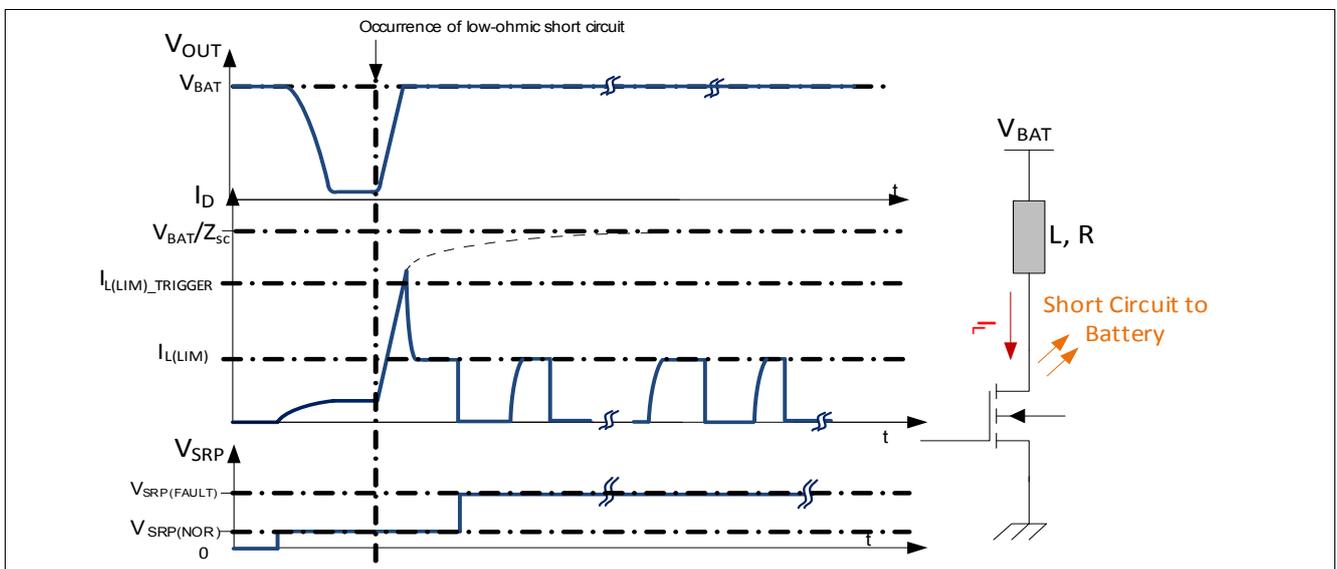


Figure 50  $V_{OUT}$  Profile during Short Circuit for BTF Devices

Protection

7.4 Undervoltage Shutdown

In order to ensure a stable and defined device behavior under all allowed conditions the supply voltage  $V_{DD}$  is monitored. The output switches off when the supply voltage  $V_{DD}$  drops below the switch-off threshold  $V_{DD(TH)}$ , causing all latches to be reset.

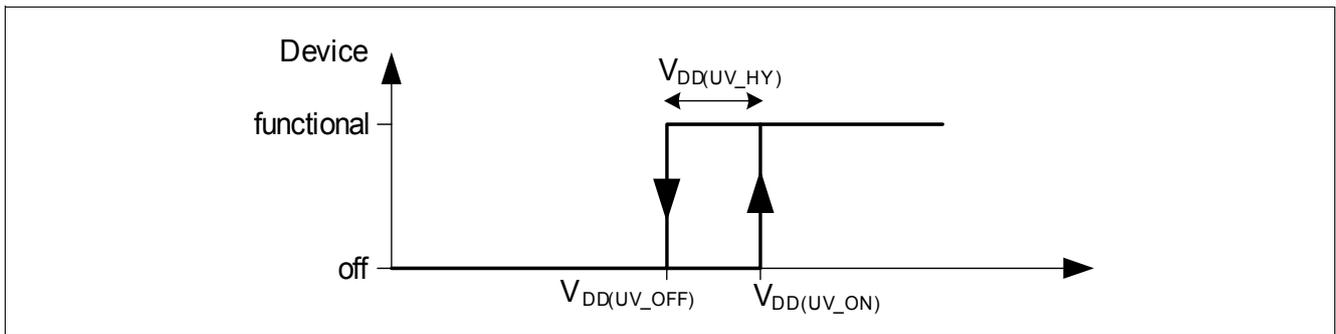


Figure 51 Undervoltage Threshold

Figure 52 shows that  $V_{DD(TH)}$  decreases with rise in temperature. Hence, at higher temperatures, the device can handle higher tolerances on the supply line without affecting the output.

Operation of the device around the  $V_{DD(TH)}$  is not recommended as the device might not be fully ON or OFF near the threshold. The same has to be kept in mind during slow turn ON and OFF of the device.

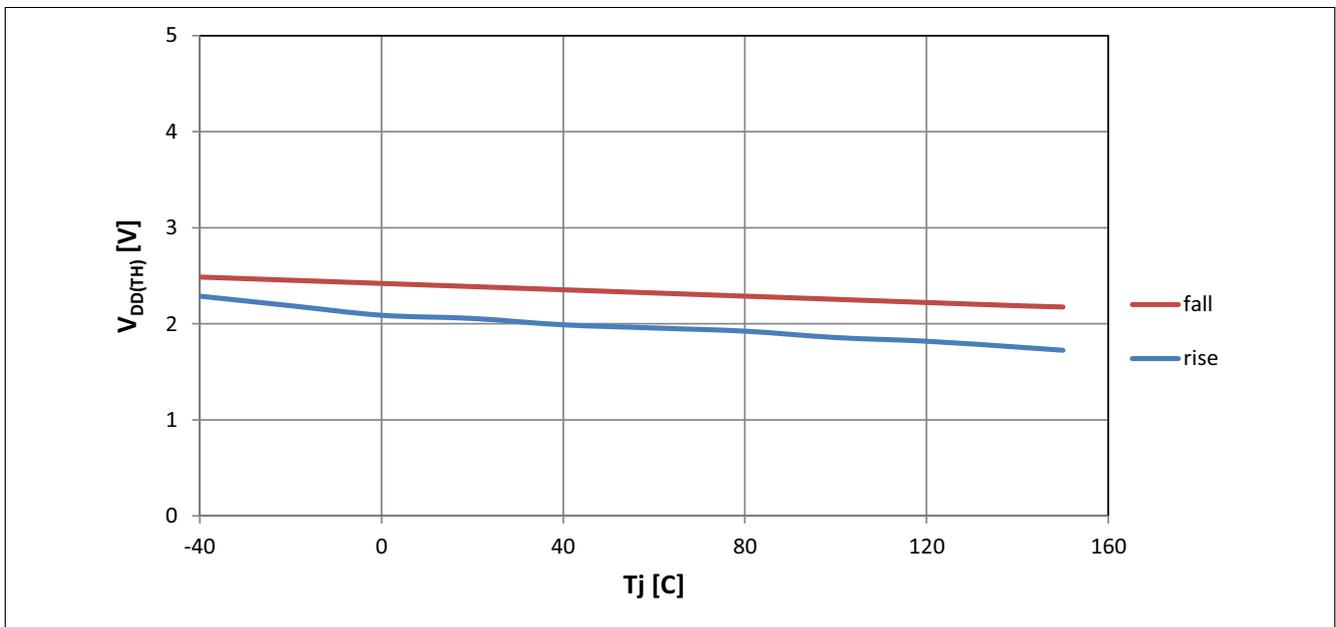


Figure 52  $V_{DD(TH)}$  vs Junction Temperature,  $T_j$  at  $R_L = 4.5\Omega$  for BTF3050TE

Protection

7.5 Load Dump

Load Dump is an extreme application scenario for a HITFET+ and can be destructive due to thermal overstress. Because of the architecture of the Low Side Switching Systems, Load dump ripple is faced by the OUT pin. Refer to [Figure 53](#).

As a result, the same mechanism as for Voltage Clamping is utilized by the device to handle Load Dump in HITFET+. The minimum value of  $V_{OUT(CLAMP)}$  as given in the HITFET+ data sheet is 40V which is above the usual value of  $V_{LOADDUMP}$  limited by OEM with diodes (as shown in [Figure 53](#) with  $V_{AZ(DIODE)}$ ). Also refer to [Chapter 3.1.5.2](#).

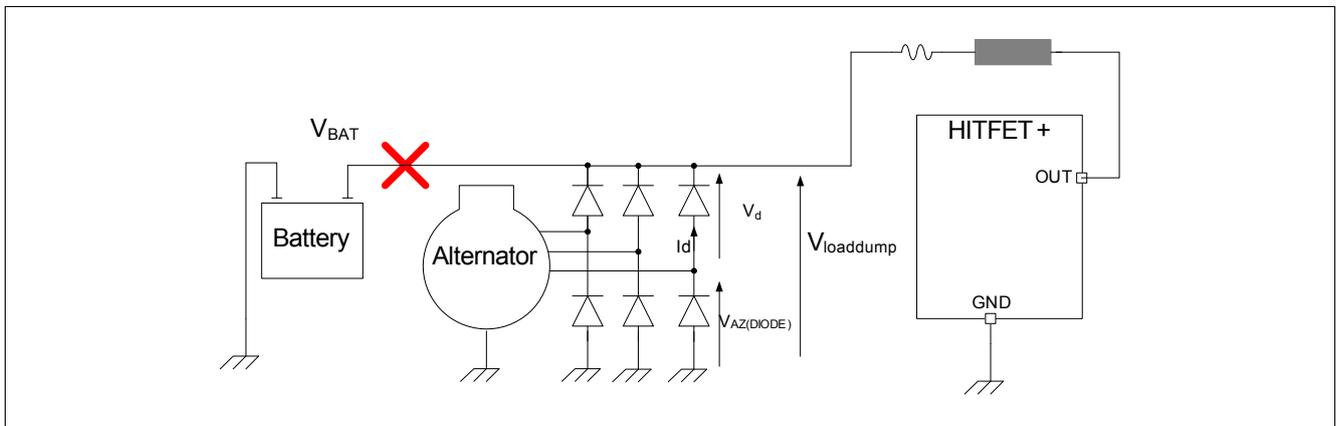
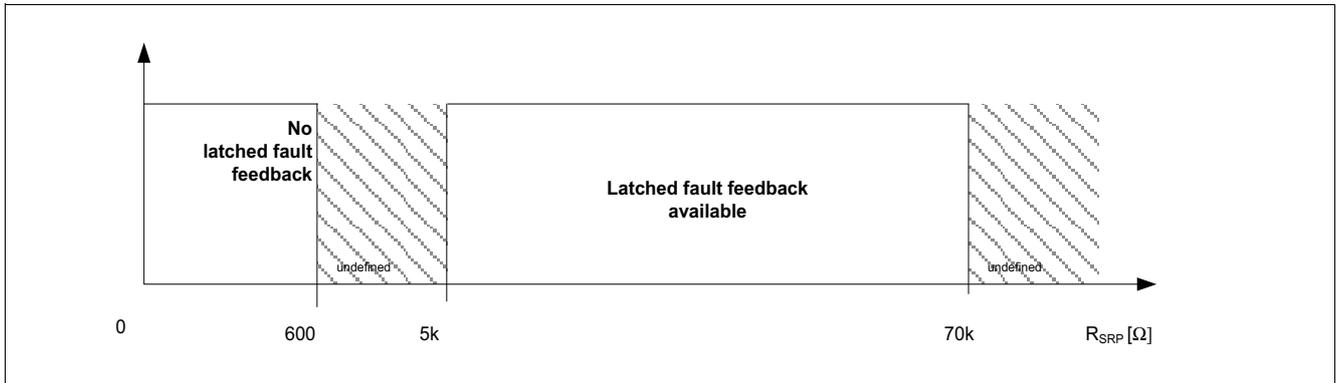


Figure 53 Load Dump Configuration for HITFET+



Diagnostics



**Figure 55 Availability of Latched Fault/Feedback Mode in Dependency of Slew Rate Resistor  $R_{SRP}$  for the BTF3050TE**

Reset Latch

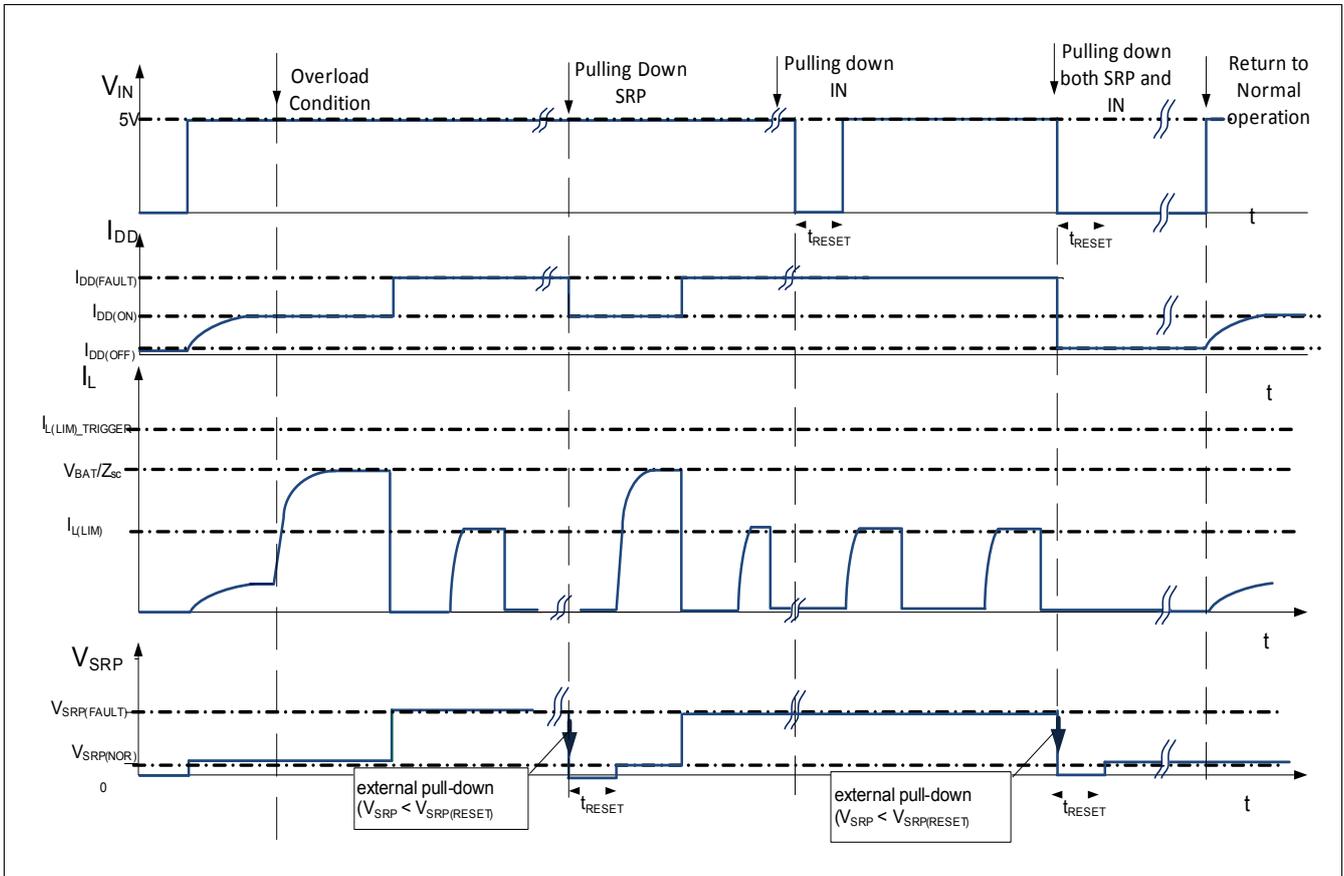
The pin is used as an input pin to set the device back to normal mode and reset the fault latch.

To reset the device, the voltage on the SRP pin needs to be forced below the reset threshold  $V_{SRP(RESET)}$  by an external pull down (e.g. using the microcontroller I/O as a pull-down).

When the SRP pin is pulled down below  $V_{SRP(RESET)}$  for a minimum time of  $t_{RESET}$  the logic resets the feedback latch, provided that its temperature has decreased at least by the thermal hysteresis  $\Delta T_{j(SW)\_HYS}$  in the meantime.

If the input is pulled down as well, the current limitation trigger level is also reset (if the  $I_{L(LIM)\_TRIGGER}$  was reached). As long as the latched fault signal is not reset, the device logic stays active (also when IN = low), not entering the quiescent current mode and therefore reaching the upper supply current limits,  $I_{DD}$  (through the internal pull-up source in [Figure 54](#)).

Diagnostics



**Figure 56 Description of Resetting Fault and SRP Pin in BTF3050TE in Overload Condition**

**Figure 56** shows that to return to normal operation and to detect further faults accurately, it is necessary to pull down both the IN and the SRP pin for a time greater than  $t_{RESET}$  (though not necessarily simultaneously, but possible before the occurrence of another fault). The input pin resets the the current limitation trigger (as shown in **Figure 47**) while the SRP pin resets the the fault signal latch and current limitation. Resetting only the IN pin will not change the latch signal and hence, if the current limitation is reached again leading to shutdown, the fault will not be detected as the fault signal will already be high. Similarly, resetting only the SRP pin will reset only the fault latch signal and current limitation (and pull back  $I_{DD}$  to  $I_{DD}(ON)$ ).  $I_L$  will not be able to reach above  $I_{L(LIM)}$  and the SRP pin will be pulled high again, along with the  $I_{DD}$  the next time the device shuts down because of over heating due to current limitation.

Since the minimum value of  $t_{RESET}$  is 100µs and PWM is determined by the input pin, it is important to consider the effect of  $t_{RESET}$  on the application and on the frequency and duty cycle of the PWM. **Figure 56** shows that if the off time in PWM is less than 100µs, then the current limitation trigger (once set) will not be reset again and the current will be limited to  $I_{L(LIM)}$  for subsequent PWM cycles. Depending on the application’s power loss and current requirements, it has to be decided whether the inrush has to be allowed for each PWM cycle.

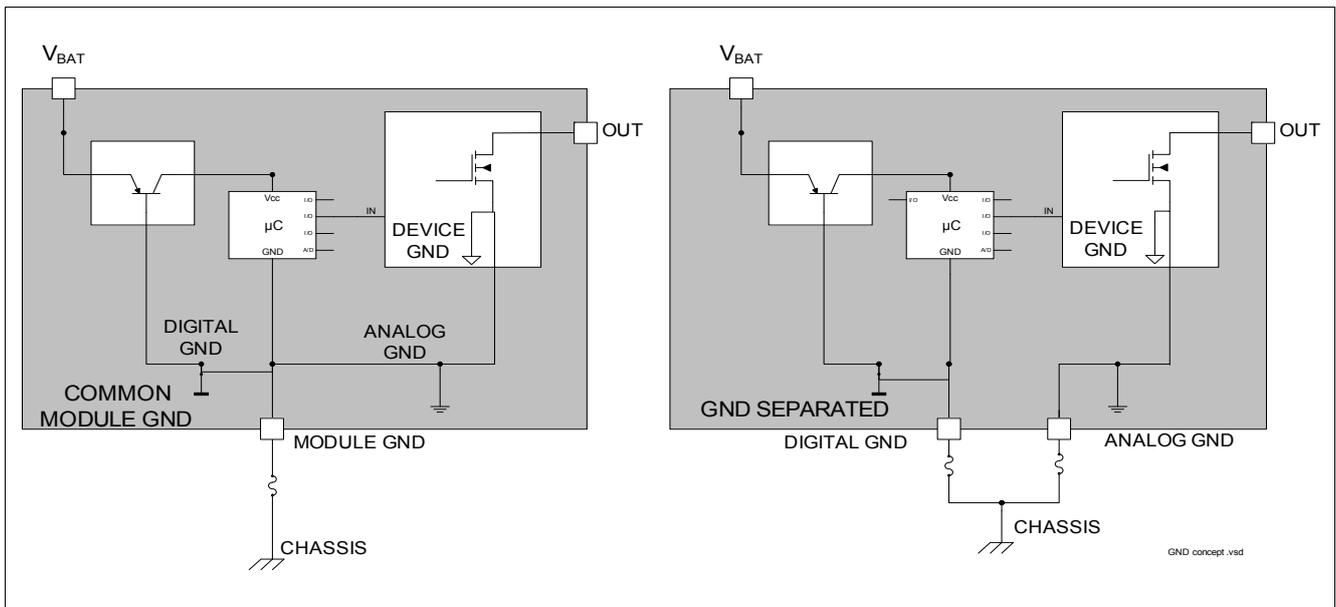
Device Information

## 9 Device Information

Note: The following information is given as a hint for the implementation of the device only and should not be regarded as a description or warranty of a specific functionality, condition or quality of the device.

### 9.1 GND Pin

At least two different grounds are defined at a system level and usually three are necessary for optimum design. The chassis GND is the system 0V reference. The module GND is the 0V module reference. The module GND is sometimes split into digital GND (reference voltage for the digital sections such as the voltage regulator, microcontroller, A/D converter, CAN transceivers etc.) and power GND (reference voltage for the power elements such as LSS, HSS, H bridges etc.). A fourth GND can also be defined which corresponds to the device GND. These different GNDs are shown in **Figure 57**. For simplification it does not describe the redundancy of GND wiring connection as shown in **Figure 18**. In a real system, the GND schematic can be even more complicated!



**Figure 57 GND Definition**

The BTF3050TE has no separate pin for power and logic ground. GND pin acts as both the power ground sourcing all the load current from the device as well as the ground for the supply and input pins. It is therefore important that the ground shift between the ground connection of the  $V_{DD}$ ,  $V_{IN}$  and SRP pins and the OUT pins is referenced to the same GND point.

Device Information

9.2 SRP Pin

To minimize the offset between the ground connection of the slew rate resistor and the ground pin of the device, it is recommended to place the resistor  $R_{SRP}$  as close as possible between the SRP pin and GND pin to avoid any influence of GND shift on the functionality of the SRP pin.

Parasitic capacitance between SRP pin and OUT pin ( $C_{SRP\_OUT}$ ) should be also be minimized as  $V_{OUT}$  changes while switching which might affect the slew rate.  $C_{SRP\_OUT}$  values as low as a few pF can affect the slew rate.

Also, the maximum capacitance between the SRP line and GND ( $C_{SRP\_GND}$ ) has to be less than 100pF. This includes any capacitance between SRP line and GND, be it parasitic or otherwise. An alternate plausible method is to maintain a maximum SRP settling time of 2.5 $\mu$ s. This has to be considered by a proper layout also taking into account of parasitic capacitors. It is recommended to not let the SRP pin floating. A maximum of 200k $\Omega$  to GND is recommended.

9.3 Input Pin

Figure 58 shows the input circuitry of the BTF3050TE. The internal pull down ensures that the device switches off when the input pin is open. A Zener structure protects the input circuit against ESD pulses. As the BTF3050TE has a supply pin, the  $R_{DS(ON)}$  of the power MOS is independent of the voltage on the IN pin (assuming  $V_{DD}$  is sufficient).

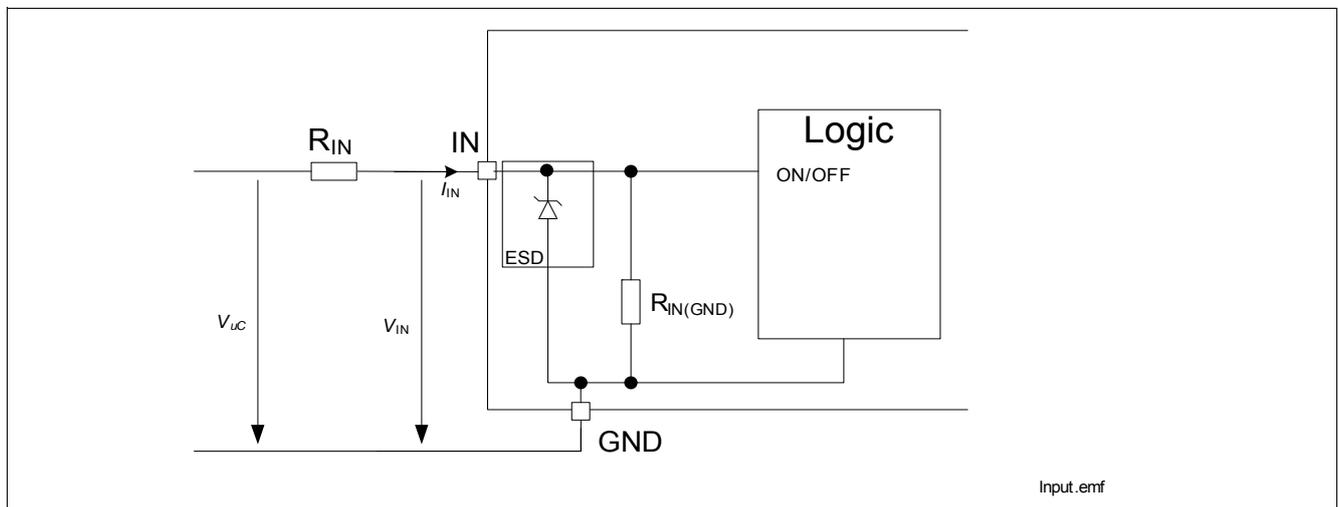


Figure 58 Simplified Input Circuitry

Also, to be noted from Figure 58 is the point where  $V_{IN}$  is specified - at the input pin of the device. If a  $R_{IN}$  is used then care has to be taken of the voltage drop across it. Although this is not specifically required by BTF3050TE, it may be used to limit currents to and from the microcontroller.

The input pin is ESD protected and also stable with respect to transients as long as they are not comparable to  $t_{ON}$  and  $t_{OFF}$ . Since delays,  $t_{DON}$  and  $t_{DOFF}$ , are a big portion of the total  $t_{ON}$  and  $t_{OFF}$ , transients as long as  $0.5 * t_{OFF}$  do not disturb the output.

Device Information

9.4 Supply Pin

Figure 59 shows the circuitry for the supply pin. It is also protected against ESD pulses through a Zener diode. The device supply is not internally controlled but directly taken from an external supply. Therefore, a reverse polarity protected and buffered 5V (or 3.3V) supply is required. To achieve a reasonable  $R_{DS(ON)}$  and the specified switching speed, a 5V (or 3.3V) supply is required.

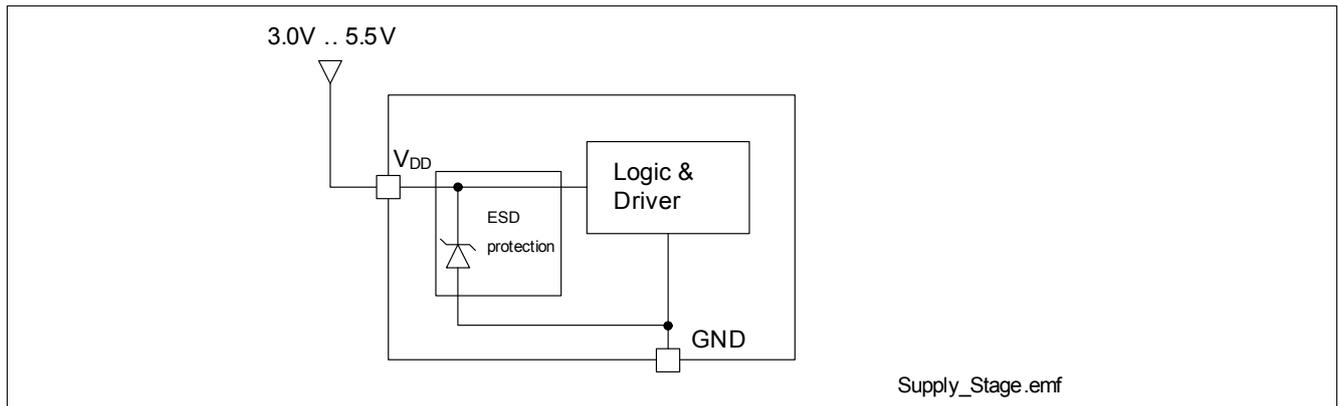


Figure 59 Supply Circuit

9.5 Threshold Region

The undefined region contains the switching thresholds for ON and OFF. The exact value  $V_{TH}$  where this switching takes place is unknown and depends on the device manufacturing process and temperature. To avoid cross-talk and parasitic switching, hysteresis is implemented. This ensures a certain immunity to noise.

This noise immunity can be defined, assuming that the exact turn ON and turn OFF thresholds are known. As an example, a rising or falling signal with parasitic noise will see several ON / OFF states before going to a stable state. Figure 60 gives an example of this situation. At turn ON, the parasitic noise is sufficiently intrusive to turn the device ON and OFF. At turn OFF, the parasitic noise is filtered by the hysteresis circuitry. The bigger the hysteresis, the higher the immunity to noise, but the difference between  $V_{IN(H)_MIN}$  and  $V_{IN(L)_MAX}$  also increases, limiting the application's range. BTF3050TE has a hysteresis voltage of 200mV.

Device Information

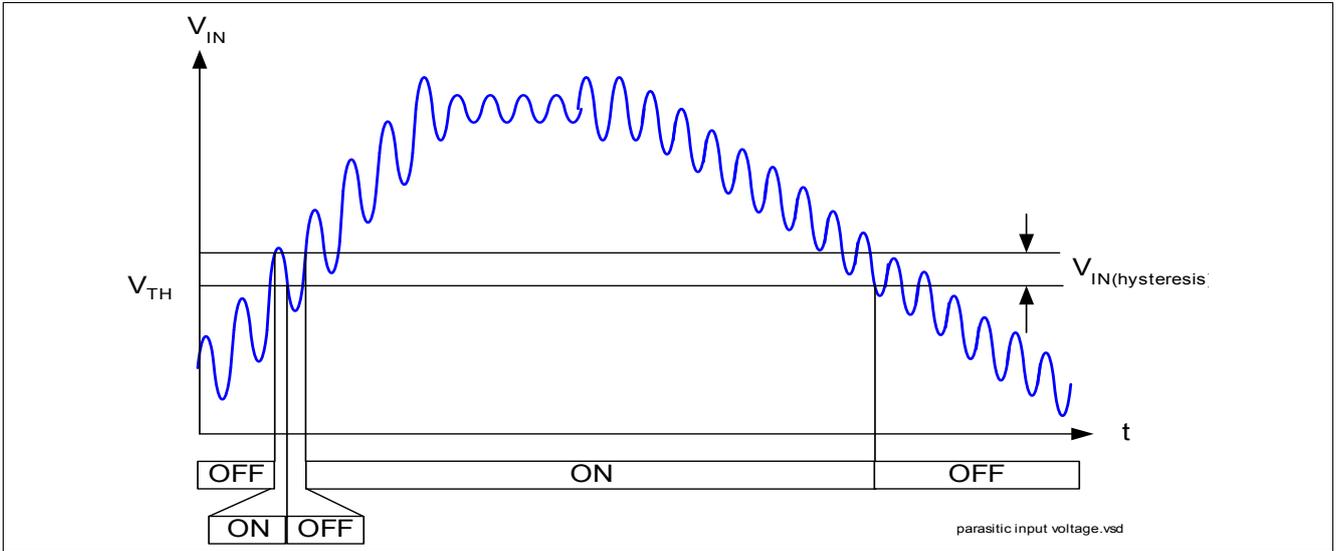


Figure 60 Benefit of the Hysteresis for Immunity to Noise

Device Information

9.6 Thermal Performance of Package

The overall thermal performance of a PG-TO252 (DPAK) package is characterized by a junction to ambient thermal resistance  $R_{thJA}$ . The  $R_{thJA}$  can be calculated using Equation (9.1)

$$R_{thJA} = R_{thJC} + R_{thCS} + R_{thS} + R_{thSA} \tag{9.1}$$

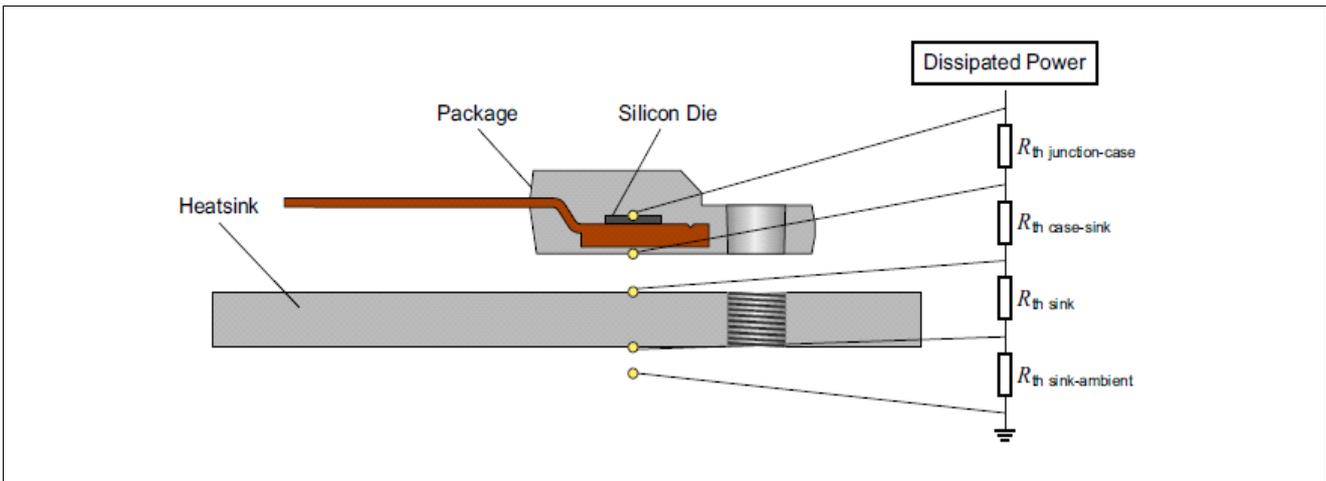


Figure 61 Thermal Model

When mounting the package on a heatsink, it is important to consider the interface resistance  $R_{thCS}$ . In an ideal case, this is zero. In real applications, however, there will be a small air gap because of these three factors:

- Package and heatsink are never perfectly smooth.
- Package and heatsink are never perfectly flat.
- Misalignment of package due to imperfect mounting

This means that  $R_{thCS}$  will always exceed zero.

In many applications, the package must be electrically insulated from its mounting surface. The insulation has a comparatively high thermal resistance, which raises junction operating temperatures.

## Appendix

## 10 Appendix

## 10.1 Power Loss Calculation

Referring back to [Chapter 6.4.1](#) and [Figure 33](#), the power losses P in the device were defined as:

$$P = V_{DS} \times I_L = V_{OUT} \times I_L \quad (10.1)$$

Now to express  $V_{OUT}$  and  $I_L$  in terms of known parameters, we can express  $I_L$  in terms of [Equation \(10.2\)](#)

$$I_L = \frac{V_{BAT}}{R_L + R_{DS}} \quad (10.2)$$

and  $V_{OUT}$  in terms of [Equation \(10.3\)](#)

$$V_{OUT} = I_L \times \frac{R_{DS}}{R_L + R_{DS}} \quad (10.3)$$

Substituting [Equation \(10.2\)](#) and [Equation \(10.3\)](#) in [Equation \(10.1\)](#) gives us:

$$P = \frac{V_{BAT}^2 \times R_{DS}}{(R_L + R_{DS})^2} \quad (10.4)$$

which is the same as [Equation \(6.4\)](#). Now to calculate the  $P_{MATCH}$  we note that in [Figure 33](#),  $P_{MATCH}$  is the peak of the power curve as in the mathematical maxima of the function P as defined in [Equation \(10.4\)](#) above. Hence to calculate the maxima with respect to  $R_{DS}$  we differentiate the function and equate to zero.

$$\frac{dP}{dR_{DS}} = \frac{d}{dR_{DS}} \left( \frac{R_{DS} \times V_{BAT}^2}{(R_L + R_{DS})^2} \right) = 0 \quad (10.5)$$

which gives us:

$$(R_L + R_{DS})^2 \times \frac{dR_{DS}}{dR_{DS}} - R_{DS} \times \frac{d(R_L + R_{DS})^2}{dR_{DS}} = 0 \quad (10.6)$$

which on solving leads to:

$$R_{DS} = R_L \quad (10.7)$$

Thus at the point of maximum power dissipation by the device,  $P_{MATCH}$  the  $R_{DS}$ , which has been decreasing during

Appendix

switching ON, is equal to  $R_L$ . The opposite is true for switching OFF when the increasing value of  $R_{DS}$  matches the  $R_L$  and gives the same maximum power peak  $P_{MATCH}$  during switch OFF. Replacing  $R_{DS} = R_L$  in [Equation \(10.4\)](#)/[Equation \(6.4\)](#) gives us back [Equation \(6.5\)](#).

$$P = \frac{V_{BAT}^2}{4 \times R_L} \tag{10.8}$$

### 10.2 Demagnetization Energy

To calculate the demagnetization energy in a HITFET+, refer to

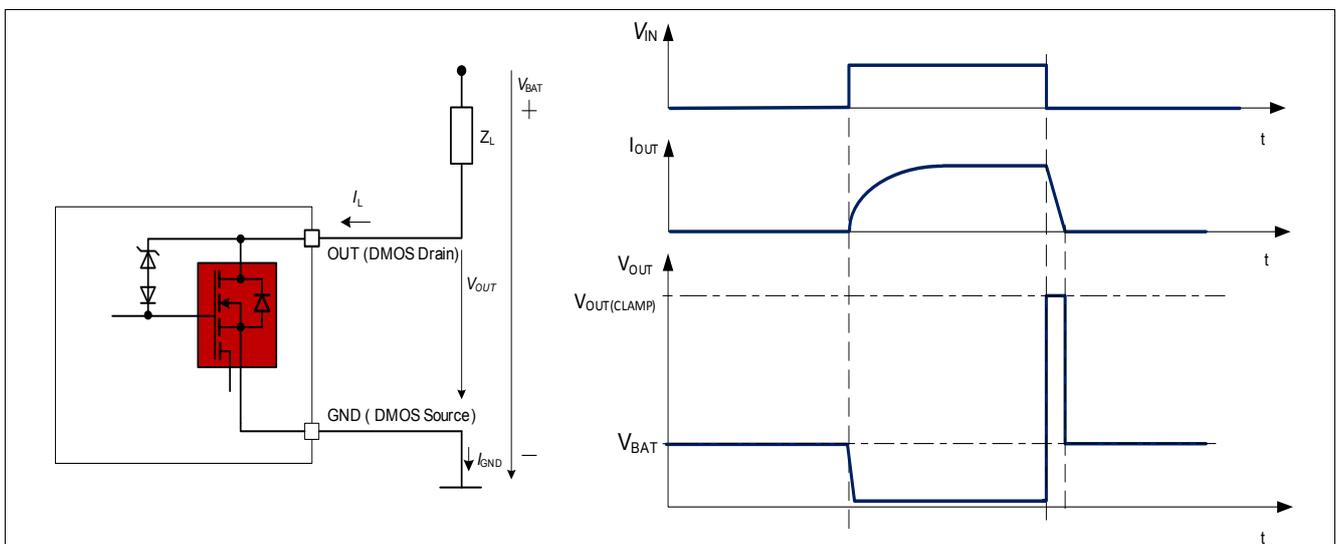


Figure 62 Output Clamp Circuitry

The Equation for the clamp circuit can be written as:

$$-L \times \frac{dI}{dt} - I \times R + V_{BAT} - V_{OUT} = 0 \tag{10.9}$$

which can be simplified to be written as:

$$V_{BAT} - V_{OUT} = L \times \frac{dI}{dt} + I \times R \tag{10.10}$$

Solving this differential equation gives us:

$$I = \frac{V_{BAT} - V_{OUT}}{R} - \frac{V_{OUT}}{R} \times e^{-R \times t/L} \tag{10.11}$$

Using this, we can calculate the time  $t_1$  it takes for the current to go down to zero.

$$0 = \frac{V_{BAT} - V_{OUT}}{R} - \frac{V_{OUT}}{R} \times e^{-R \times t_1/L} \tag{10.12}$$

**Appendix**

which gives  $t_1$  as:

$$t_1 = \frac{L}{R} \times \ln\left(\frac{V_{OUT}}{V_{BAT}}\right) \tag{10.13}$$

Now  $E_{AS}$  can be written as

$$E_{AS} = \int_0^t (V(t) \times I(t))dt \tag{10.14}$$

From **Figure 62**, it can be seen that  $V_{(t)}$  during clamping switch off can be assumed to be  $V_{OUT(CLAMP)}$  whereas  $I_{(t)}$  can be taken to be **Equation (10.11)**. Thus **Equation (10.14)** becomes:

$$V_{OUT(CLAMP)} \times \int_0^{t_1} \left[ \frac{V_{BAT} - V_{OUT(CLAMP)}}{R} - \frac{V_{OUT(CLAMP)}}{R} \times e^{-R \times t/L} \right] dt \tag{10.15}$$

Solving this gives us:

$$E_{AS} = V_{OUT(CLAMP)} \times \frac{L}{R} \times \left[ \frac{V_{BAT} - V_{OUT(CLAMP)}}{R} \times \ln\left(\frac{V_{OUT(CLAMP)}}{V_{BAT} - V_{OUT(CLAMP)}}\right) + \frac{V_{BAT}}{R} \right] \tag{10.16}$$

Using  $V_{BAT}/R = I_L$ , load current in the normal operation, in the **Equation (10.16)**:

$$E_{AS} = V_{OUT(CLAMP)} \times \frac{L}{R} \times \left[ \frac{V_{BAT} - V_{OUT(CLAMP)}}{R} \times \ln\left(1 - \frac{I_L \times R}{V_{BAT} - V_{OUT(CLAMP)}}\right) + I_L \right] \tag{10.17}$$

which is the same as **Equation (5.2)**.

**10.3 PWM Duty Cycle Calculation**

To maintain constant power across a heating load during PWM, the following can be written:

$$\frac{V_{BAT}^2}{R_L} \times d = \frac{V_{PWM}^2}{R_L} \tag{10.18}$$

**Equation (10.18)** simply equates the power dissipated across the load in 2 different ways. Here,  $V_{PWM}$  is the root mean square voltage across the heating load during the PWM. Cancelling out  $R_L$  gives us **Equation (5.1)/Equation (10.19)**.

$$d = \frac{V_{PWM}^2}{V_{BAT}^2} \tag{10.19}$$



Revision History

11 Revision History

Revision	Date	Changes
Rev 1.0	2015-01-12	Application Note released

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