AN1009: Driving MOSFET and IGBT Switches Using the Si828x

The Si828x products integrate isolation, gate drivers, fault detection protection, and operational indicators into one package to drive IGBTs and MOSFETs as well as other gated power switch devices. Most Si828x products (except the Si8286) have three separate output pins to provide independent rise and fall time settings and low impedance clamping to suppress Miller voltage spikes. This application note provides guidance for selecting the external components necessary for operation of the driver.

The figure below depicts an isolated gate drive circuit with three output pins for driving the gate of the IGBT. Input signals from the IN+ pin are RF encoded to transmit across the isolation barrier, and the RF signal is decoded back to its original signal format on the secondary side to connect to the driver input. The driver has three output pins. The VH pin provides positive drive current to turn the IGBT on while the VL pin provides sinking current to turn the IGBT off. The VH and VL pins are connected to the gate of the IGBT through the RH and RL resistors. The CLMP pin is connected to the gate of the IGBT directly to provide a low-impedance clamp between the IGBT Gate and Source terminals during the off time. This low-impedance CLMP connection prevents the high Miller current from generating voltage spikes that might be higher than $V_{th}$ and turn the IGBT on during the off period. The following sections will explain the operation of the gate drive circuits in more detail.

**KEY FEATURES**

- Simple Driver with One Output
- Driver with Independent Rise and Fall
- Driver with Miller Clamp
- Driver Power Dissipation
- Gate Resistor Calculations
- DESAT Detection and Protection
- Half-Bridge Shoot-Through Current
- Example Design

![Gate Drive IC Diagram](image-url)
1. IGBT and MOSFET Gate Drive Models

The figure below shows the simple driver with one output pin driving the gate of a MOSFET through the gate resistor $R_g$. Since both gate charging and discharging current levels are determined by $R_g$, the rise and fall times are the same. This driver model works well for applications where independent rise and fall time is not necessary and the Miller current generated by the $dV/dt$ at the MOSFET Drain terminal through the parasitic $C_{GD}$ is small, hence a Miller clamp is not needed.

![Figure 1.1. Simple Driver Model](image-url)

The figure below illustrates a driver with separated VH and VL (separate output high and output low) pins that allows independent setting of the switch’s turn ON and OFF times. This capability is quite helpful in half-bridge configuration (Figure 2.3 Half-Bridge Circuit with Three-Output-Pins Driver on page 5) where the speed of the turn ON and OFF are aligned to minimize shoot-through current (see section 2.1). The low-impedance Miller clamping pin (CLMP) attenuates the Miller voltage spikes and renders the driver with three output pins more suitable for driving high power MOSFET/IGBT with larger input gate $C_{GS}$ and $C_{DS}$ capacitances.

![Figure 1.2. Driver Model with Three Output Pins](image-url)

The following figure demonstrates an option for the Simple Driver circuit to achieve independent control of the rise and fall times by the use of the steering diode. It is recommended to use 100 V, 1 A fast diode, similar to an ES1B diode, to ensure adequate turn-off speed and meet the operational voltage-current requirements of the driver circuit. Note that since the turn off current flows through both $R_H$ and $R_L$, adjustment to the $R_L$ value is required to ensure the desired turn-off transition time (see 3.1 Calculating $R_H$, $R_L$ Gate Resistors).
Figure 1.3. Simple Driver with Independent Rise and Fall Times
2. Switching Characteristics

Since IGBTs and MOSFETs are gated power switch devices there is no gate current flow during the ON and OFF states, and the voltage level at the gate controls the device between the ON and OFF states. However, there are parasitic capacitors between gate to source (C_{GS}) and gate to drain (C_{GD}) that require rapid charging and discharging during switching. The figure below provides graphical presentation for the gate voltage and current characteristics.

![Diagram of gate current during switching](image)

Figure 2.1. Gate Current during Switching

Turning the power switch ON and OFF requires current flow to charge and discharge the gate capacitors. Power switch datasheets provide the total gate charge (Q_{total}) required to switch a device between the ON and OFF state. The larger the available gate current, the less time it takes to provide the switching gate charge. Equation 1 defines the relationship between the gate current I_{g}, the total gate charge Q_{total}, and the switching transition time t.

\[ Q_{total} = I_{g} \times t \]  
\[ I_{g} = \frac{Q_{total}}{t} \]

Equation 1. Gate Charge, Gate Current over Transition Time

From the above equation, it is easy to see that the switching time can be reduced with higher gate current. The figure below illustrates faster switching time (compared to the figure above) with higher gate current.

![Diagram illustrating higher gate current reduces switching time](image)

Figure 2.2. Higher Gate Current Reduces Switching Time
2.1 Coordinating RH and RL Values to Minimize Shoot-Through Current

The following figure illustrates a half-bridge power circuit with three-output-pin drivers. This type of power circuit can suffer shoot-through current (current created when both QH and QL are ON) if driven by complementary (+IN, –IN) input signals and having the same values for RH and RL.

Moreover, power devices are not idealistic switches and the transition between ON and OFF states does not occur cleanly. There is a period, during the switching transition when both switches are in the half-on and half-off state. This condition allows current from the HV-DC rail to flow directly through both QH and QL to ground (see figure above). Shoot-through current dissipates excessive amounts of power in the switching devices since it flows directly across the high voltage dc rail. This is part of switching loss and should be minimized for robust operation.

One method to minimize shoot through current is to turn the switches on slowly and to turn them off quickly. This method ensures that one device is closer to a complete turned-off state before the other one is turned on. This method can be easily implemented with the 3-output-pin driver by having larger RH and smaller RL resistor values. A large RH value reduces the charging current to the gate capacitor and lengthens the turn-on time while the smaller RL value speeds up the turn-off time. The figure below illustrates a smaller shoot-through current using this method where RH >> RL.

Figure 2.3. Half-Bridge Circuit with Three-Output-Pins Driver

Figure 2.4. Shoot-Through Current at Switching, RH = RL

One method to minimize shoot through current is to turn the switches on slowly and to turn them off quickly. This method ensures that one device is closer to a complete turned-off state before the other one is turned on. This method can be easily implemented with the 3-output-pin driver by having larger RH and smaller RL resistor values. A large RH value reduces the charging current to the gate capacitor and lengthens the turn-on time while the smaller RL value speeds up the turn-off time. The figure below illustrates a smaller shoot-through current using this method where RH >> RL.
Notice that this RH >> RL method for minimum shoot-through current can also be implemented on the simple driver with one output pin by adding a steering diode to the gate circuit (see Figure 1.3 Simple Driver with Independent Rise and Fall Times on page 3).

Figure 2.5. Lower Shoot-Through Current, RH >> RL
3. Turn-On Characteristics

In the MOSFET circuit in Figure 1.2 Driver Model with Three Output Pins on page 2, the gate capacitances, \( C_{GS} \) and \( C_{GD} \), need to be charged to a critical voltage level (bringing \( V_{GS} \) toward \( V_{th} \)) to initiate \( I_{DS} \) current flow (t0–t1 in the figure below). The \( I_{DS} \) increases rapidly as \( V_{GS} \) goes above \( V_{th} \) (t1–t2 in the figure below) and it reaches its maximum current level when \( V_{GS} \) reaches a voltage point that generates \( I_{DS-max} = \frac{V_{DD}}{Z_L} \).

After the \( I_{DS} \) reached its maximum level at t2, the \( V_{DS} \) voltage drops rapidly resulted in a large dV/dt at the Drain terminal. This dV/dt sinks large current flow from of the gate of the device through \( C_{GD} \). This is the Miller current flow and the dV/dt effectively “amplifies” the current through \( C_{GD} \). The driver’s current must counter the Miller current and results in a constant voltage (almost constant) level at the gate during the Miller current flow period (t2 to t3). The dV/dt at the Drain terminal ends at t3 when the \( V_{DS} \) reaches \( V_{DS-min} = R_{on} \times I_{DS-max} \). From this point on, any further increase in gate voltage does not affect the drain to source current and the driver current continues to charge \( C_{GS} \) and \( C_{GD} \) toward the driver’s \( V_{DRV} \) voltage level.

The figure below provides the relationship between the gate voltage \( V_{GS} \) and the gate charge \( Q_g \) during the turn on switching transition. Notice that this curve looks similar to the \( V_{GS} \) vs. time curve in the above figure.

- Between t0–t1–t2: The gate current charges \( C_{GS} \), \( C_{GD} \) and the gate voltage \( V_{GS} \) increases linearly.
- Between t2–t3: The gate current charges \( C_{GS} \) and mainly the larger \( C_{GD-M} \) Miller cap (\( C_{GD} \) with the presence of dV/dt at the Drain terminal). The driver current is sourcing the Miller current and results in no increase in the \( V_{GS} \) voltage.
- Between t3–t4: The gate current charges \( C_{GS} \), \( C_{GD} \), and \( V_{GS} \) increases linearly.

This \( V_{GS} \) vs. \( Q_g \) curve is unique for each power switch device and is available in the power switching device’s data sheet. It is an important parameter as it is used to calculate the values of the RH, and RL gate resistors as will be illustrated in an example later in this application note.
Figure 3.2. $V_{GS}$ vs. Gate Charge
3.1 Calculating $R_H$, $R_L$ Gate Resistors

The switching transition time depends on how quickly the total gate charge $Q_g$ is delivered, which in turn, depends on the amplitude of the gate current. Therefore, transition times ($t_{\text{rise}}$ and $t_{\text{fall}}$) can be controlled by selecting the values of the gate resistors to a desired gate current levels. From Equation 1 and the information in the above figure, the gate current required to switch the power device between the ON and OFF states can be approximated:

$$I_{g_{\text{on}}}=\frac{Q_g}{t_{\text{rise}}}, I_{g_{\text{off}}}=\frac{Q_g}{t_{\text{fall}}}$$

Equation 2. Gate Current

Notice that for the driver with three output pins, the gate turn-on ($I_{g_{\text{on}}}$) and turn-off current ($I_{g_{\text{off}}}$) can be different for applications with different rise, and fall time requirements. For the driver with one output pin (see front page figure), the rise and fall times are the same since both charging and discharging currents flow through the same $R_g$ resistor.

The $R_H$ can be calculated using Ohm's law and the driver's operating voltage VDDB (VDRV)

$$R_H = \frac{V_{\text{DRV}}}{I_{g_{\text{on}}}} - R_{OH}$$

Equation 3. $R_H$ Resistor

The power switch turn-off process is the reverse of the turn-on phenomenon. The gate charge built up during the turn-on transition must be removed to bring the power device to a fully off state. The gate voltage from the ON state is equal to VDDB (sustained by the gate capacitors). This voltage level needs to “discharge” to 0V to turn off the power device.

$$R_L = \frac{V_{\text{DRV}}}{I_{g_{\text{off}}}} - R_{OL}$$

Equation 4. $R_L$ Resistor

The calculation for $R_L$ in the simple driver circuit with steering diode in Figure 3 requires additional adjustment to reflect the fact that the turn off current flows through both $R_H$ and $R_L$. Using the calculated $R_L$ value from Equation 4, the actual $R_L$ value can be derived from the parallel equation for $R_H$ and $R_L$.

$$R_{L_{\text{adjusted}}} = \frac{R_L \times R_H}{R_H - R_L}$$

Equation 5. Adjusted for $R_L$ Circuit in Figure 1.3 Simple Driver with Independent Rise and Fall Times on page 3
3.2 RH, RL Power Dissipation

In Figure 1.2 Driver Model with Three Output Pins on page 2, the current flows through resistors RH and RL to charge and discharge the gate capacitors. Depending on the switching frequency, VDDB voltage level, and the total gate charge of the power switch device, the power dissipation on these gate resistors must be considered. Therefore, proper package selections are required to ensure that the resistors can handle the power dissipation appropriately. Power dissipation on the gate resistors occurs only during the turn-on and turn-off transitions.

Upon turn-on, current flows from VDDB through the $R_{OH}$ (Si828x internal source impedance) and RH resistors (see figure below) to charge the gate capacitor, $C_g$. At the end of the turn-on transition, the voltage at the gate capacitor reached the VDDB level. Half of the total charging energy from VDDB is stored in the gate capacitor and the other half is dissipated on the gate resistors RH and $R_{OH}$. The power dissipation on the resistor is split between $R_{OH}$ and RH proportionally to their resistance.

Upon turn-off, current flows from $C_g$ through RL and $R_{OL}$ to ground (see figure below) to discharge the $C_g$ stored energy (from turn on). The voltage on $C_g$ begins at VDDB voltage level and ends at ground voltage level, and all stored capacitive energy is dissipated on resistors RL and $R_{OL}$.

\[ P_{RH} = \frac{1}{2} f C_g VDDB^2 \left( \frac{RH}{R_{OH} + RH} \right) = \frac{1}{2} f Q_g VDDB \left( \frac{RH}{R_{OH} + RH} \right) \]

Equation 6. RH Power Dissipation

\[ P_{RL} = \frac{1}{2} f Q_g VDDB \left( \frac{RL}{R_{OL} + RL} \right) \]

Equation 7. RL Power Dissipation
Dissipate $\frac{1}{2}CV^2$ as heat (from $C_g$)

**Figure 3.4. Rg Turn-Off Power Dissipation**

For the driver circuit with only one Rg gate resistor (see front-page figure), both turn-on and turn-off current flow through the same Rg resistor. The power dissipation on Rg is the sum of the turn-on and turn off power.

$$P_{Rg} = \frac{1}{2} f C_g V_{DDB}^2 \left( \frac{R_g}{R_{OH} + R_g} \right) + \frac{1}{2} f C_g V_{DDB}^2 \left( \frac{R_g}{R_{OL} + R_g} \right)$$

The above equation can be simplified as follows:

$$P_{Rg} = \frac{1}{2} f C_g V_{DDB}^2 \left( \frac{R_g}{R_{OH} + R_{g}} + \frac{R_g}{R_{OL} + R_{g}} \right)$$

where $C_g = \frac{Q_{total}}{V_{DD}}$

**Equation 8. Rg Power Dissipation**
4. Driver’s Power Dissipation Consideration

When all the driver’s external gate component values are calculated, it is important to review the total power dissipation of the driver’s package to make sure that the device’s maximum junction temperature is not exceeded while operating within the desired temperature range. The power equations for the drivers are provided in the data sheet and repeated here for convenience.

Si8285, Si8286, Si8287:

\[
P_D = (V_{DDA})(I_{DDA}) + (V_{DDB})(I_{DDB}) + f \times Q_{int} \times V_{DDB} + \frac{1}{2} \left( f \left( \frac{R_{OH}}{R_{OH} + R_{HL}} + \frac{R_{OL}}{R_{OL} + R_L} \right) \right) \]

Si8281–Si8284:

\[
P_D = (V_{DDA})(I_{DDA}) + 1.05(V_{DDB})(I_{DDB}) + 1.05 \times f \times Q_{int} \times V_{DDB} + \frac{1.05}{2} \left( f \left( \frac{R_{OH}}{R_{OH} + R_{HL}} + \frac{R_{OL}}{R_{OL} + R_L} \right) \right)
\]

Note: The dc-dc circuit adds 5% higher power dissipation to the Si8281-84 packages.

Note: For system with -VSSB use the same equations but substitute VDDB=(VDDB + VSSB) to account for VSSB power source.

Where (Si8285 bias current data):

- PD is the total Si828x device power dissipation (W).
- IDDA is the input-side maximum bias current (6.5 mA).
- IDDB is the driver die maximum bias current (4.5 mA).
- Qint is the internal parasitic charge (3 nC).
- VDDA is the input-side VDD supply voltage (2.7 to 5.5 V).
- VDDB is the total driver-side supply voltage (VDDB + VSSB: 12.5 to 30 V).
- f is the IGBT/MOSFET switching frequency (Hz).
- RH is the VH external gate resistor, RL is the VL external gate resistor (RH = RL = Rg if using Rg only).
- R_{OH} is the RDS(ON) of the driver pull-up switch: (2.48 Ω).
- R_{OL} is the RDS(ON) of the driver pull-down switch: (0.86 Ω).

**Equation 9. Driver’s Power Equations**

The maximum driver’s junction temperature can be calculated given the maximum operating ambient temperature and the driver thermal resistance in the data sheet.

\[
T_j = PD \times \theta_{ja} + TA
\]

Where:

- \(T_j\) < 150 °C
- \(\theta_{ja} = 60 \text{ C/W (Si8285, Si8286)}\)
- \(\theta_{ja} = 60 \text{ C/W (Si8281–Si8284)}\)
- TA = 125°C (maximum temperature range: -40 to 125°C)

**Equation 10. Junction Temperature**
5. DESAT Detection and Soft Shutdown Control

The Si828x has a comparator with its input connected to the DSAT pin to monitor the power switch VCE voltage during on time (when VH is in the high state) to ensure that the power switch is fully on (VDS < 7 V). The DSAT pin is connected to the Collector of the switching device through a diode, and the R_{DSAT} resistor (see the figure below), thus the voltage at the DSAT pin is approximately the same as the voltage at the collector of the switching device with respect to VMID.

If the voltage on the DSAT pin exceeds 7 V during on-time, the Si828x shuts down the output (turn off VH PMOS after t_{FLT} delay; see the figure below) and initiates soft shutdown (after t_{DSAT} to SS delay) to protect the power switch.
The internal soft shutdown NMOS is connected to the same VH pin (see the figure below). It has higher impedance ($R_{ss} = 60 \, \Omega$) compared to the NMOS at the VL pin ($1 \, \Omega$). Due to the presence of the parasitic inductance in the loop, turning the power switch off too quickly under a high fault current condition can generate a high voltage spike at the power switch collector. To avoid the high turn-off voltage spike, only the soft shutdown NMOS is turned on under DSAT shutdown (while the VL NMOS is kept in the off state). The gate capacitance of the power switch is discharged through the RH and the internal $R_{ss}$ at a much lower rate to allow the power switch to dissipate residual energy from the circuit’s parasitic inductance under high fault current condition.

The internal $R_{ss}$ (60 $\Omega$) and a $R_{H} = 20 \, \Omega$ provide a typical soft shutdown duration of 2 $\mu$s (for 250 nC of total gate charge, $Q_g$.) If a longer soft shutdown period is required, steering diode (ES1B or similar) can be added to the VH pin to allow installation of a larger external soft shutdown resistor (see figure below).

$$t_{ssd} = 5 \times R \times C$$

where $R = R_{ss} + R_{H}$ (no steering diode) or $R = R_{ss} + R_{EX,SS}$ (with steering diode) and $C = Q_g/V_{DDB}$

**Equation 11. Soft Shutdown Timing**

![Figure 5.3. Driver Outputs](image1)

![Figure 5.4. External Soft Shutdown Resistor](image2)
Since the turn-on gate current in the above figure flows through both \( R_{EX-SS} \) and \( RH \), the value of the \( RH_{\text{adjusted}} \) can be derived from the parallel equation for \( RH \) and \( R_{EX-SS} \):

\[
RH_{\text{adjusted}} = \frac{RH \times R_{EX-SS}}{R_{EX-SS} - RH}
\]

Equation 11. Adjusted RH for Driver with SS Steering Diode

5.1 Miller Clamp

Figure 5.3 Driver Outputs on page 14 includes the internal block diagram for the CLMP pin. The internal CLMP NMOS is activated during the off period and when the voltage at the CLMP goes below 2 V. The internal CLMP comparator has hysteresis to ensure stable turn on under noisy conditions. To minimize the Miller voltage spike, it is recommended to connect the CLMP pin directly to the gate of the power switching device so that the low impedance internal NMOS CLMP driver can secure the gate directly to VSSB.
6. DSAT Fault Indication and Fault Reset

Right after the DSAT condition is detected (VDSAT > 7V), the fault indication is propagated toward the FLTb pin and arrives after the TDSAT delay (see Figure 5.2 DESAT and Soft Shutdown on page 13). The DESAT fault condition can be cleared by bringing the RSTb pin low for a minimum duration of 50 ns. Notice that the RSTb input is level-sensitive reset logic and keeping RSTb low, continues to reset the driver output. The recommended fault condition clearing sequence is to bring the input signals at the IN pins to low logic level and then pulse the RSTb pin low for a minimum duration of 50 ns. The driver is now ready to drive the power switch with input signals from the IN pins.
7. DSAT Blanking Timer Control

Due to delay when it is turned on, the VDS voltage of the power switch can stay at the high voltage VDD level for a while (see Figure 3.1 Turn-On Characteristics on page 7, t0–t2) and only transition low after the IDS current reached its maximum level. This means that the VDS voltage can be as high as a few hundreds of volts (VDD) for some time after the driver output has gone high. This high voltage at the DSAT pin can trigger the DSAT detection. To alleviate this false premature DSAT detection event, an external capacitor $C_{BL}$ can be used to set the blanking period according to Equation 12.

$$C_{BL} = t_{Blanking} \times \frac{I_{chg}}{V_{DSAT}}$$

Where:

- $V_{DSAT} = 7\, \text{V (Si828x internal comparator)}$
- $I_{chg} = 250\, \mu\text{A (Si8286)}$
- $I_{chg} = 1\, \text{mA (other Si828x)}$

Equation 12. Blanking Cap CBL

In most applications, the switching transition period is typically around a couple hundred nanoseconds. Therefore, recommended blanking period is 3 µs to provide ample delay for the VDS to reach to its low on-time voltage level. The Si828x recommended blanking capacitor values are calculated as shown in the equation below.

- Si8286: $C_{BL} = 3 \times 10^{-6} \times \frac{250 \times 10^{-6}}{7} = 100\, \text{pF}$
- Other Si828x: $C_{BL} = 3 \times 10^{-6} \times \frac{1 \times 10^{-2}}{7} = 390\, \text{pF}$

Note: It is possible to change the $C_{BL}$ value to accommodate a specific blanking delay.
8. Design Example

Design a half-bridge power circuit using N-Channel MOSFETs with the following operating conditions:

1. $I_{\text{max}} = 20 \text{ A}$, $HVDC = 400 \text{ V}$
2. $V_{\text{DS-on}} < 1 \text{ V}$ at $20 \text{ A}$ to minimize loss
3. Rise time $= 400 \text{ ns}$, Fall time $= 200 \text{ ns}$
4. $F_{\text{sw}} = 200 \text{ kHz}$

Solution:

1. Select a MOSFET with $V_{\text{DSS}} = 650\text{V}$ for an approximate $40\%$ voltage margin while operating at $HVDC = 400\text{V}$. The MOSFET data sheet provides the $V_{\text{GS}}$ vs total gate charge as shown in Figure 3.2 $V_{\text{GS}}$ vs. Gate Charge on page 8 and a similar figure as shown below provides the MOSFET output characteristic $V_{\text{DS}}$ vs. $I_{\text{DS}}$.

2. Select the Si8285 driver with separated $V_H$ and $V_L$ pin to handle independent $400 \text{ ns}$ rise and $200 \text{ ns}$ fall times.

![MOSFET Output Characteristics @ Tj = 125 °C](image)

3. Applying the given information from items 1 and 2 to the figure above, we see that the circuit would need $V_{\text{DDB}} = V_{\text{GS}} = 15 \text{ V}$ to achieve $20 \text{ A}$ with voltage drop $V_{\text{DS}} = 1 \text{ V}$.

4. Figure 3.1 Turn-On Characteristics on page 7 indicates that it requires a total switching charge of $Q_g = 250 \text{ nC}$ to drive the $V_{\text{GS}}$ from $0 \text{ V}$ to $15 \text{ V}$.

5. Calculations.

From Equation 2:

$$I_{\text{g\_on}} = \frac{Q_g}{t_{\text{rise}}}; I_{\text{g\_off}} = \frac{Q_g}{t_{\text{fall}}}$$

$$I_{\text{g\_on}} = \frac{250 \text{ nC}}{400 \text{ ns}} = 0.625 \text{ A}; I_{\text{g\_off}} = \frac{250 \text{ nC}}{200 \text{ ns}} = 1.25 \text{ A}$$

From Equation 3, calculate $R_H$:

$$R_H = \frac{V_{\text{DDB}}}{I_{\text{g\_on}}}$$

$$R_H = \frac{15 \text{ V}}{0.625 \text{ A}} = 24 \Omega$$
From Equation 6, calculate RH power dissipation:

\[ P_{RH} = \frac{1}{2} f Q_g V_{DDDB} \left( \frac{R_{OH} + RH}{2.48 \Omega + 24 \Omega} \right) \]

\[ P_{RH} = \frac{1}{2} \times 200 \text{ kHz} \times 250 \text{ nC} \times 15 \text{ V} \times \left\{ \frac{24}{2.48 \Omega + 24 \Omega} \right\} = 340 \text{ mW} \]

Select resistor package SMD 1210 for RH resistor to handle up to 0.5 W of power.

From Equation 4, calculate RL:

\[ R_L = \frac{V_{DDDB}}{I_{g_{off}}} \]

\[ R_L = \frac{15 \text{ V}}{1.25 \text{ A}} = 12 \Omega \]

From Equation 7, calculate RL power dissipation:

\[ P_{RH} = \frac{1}{2} 200 \text{ kHz} \times 250 \text{ nC} \times 15 \text{ V} \times \left\{ \frac{12 \Omega}{0.84 \Omega + 12 \Omega} \right\} = 350 \text{ mW} \]

Select resistor package SMD 1210 for RL resistor to handle up to 0.5 W of power

From Equation 9, calculate Si8285 power dissipation:

\[ PD = (V_{DDA})(IDDA) + (V_{DDDB})(IDDB) + f \times Q_{int} \times V_{DDDB} + \frac{1}{2} f \left( Q_{IGBT} \right) V_{DDDB} \left[ \frac{R_{OH}}{R_{OH} + RH} + \frac{R_{OL}}{R_{OL} + RL} \right] \]

Note the following information from the Si8285 data sheet:
- PD is the total Si828x device power dissipation (W).
- IDDA is the input-side maximum bias current (= 6.5 mA).
- IDDB is the driver die maximum bias current (= 4.5 mA).
- Q_{int} is the internal parasitic charge (= 3 nC).
- VDDA is the input-side VDD supply voltage (= 3.3 V).
- VDDDB is the total driver-side supply voltage (= VGS = 15 V).
- f is the IGBT switching frequency (Hz) (= 200 kHz).
- RH is the VH external gate resistor (= 24 Ω and 12 Ω).
- R_{OH} is the RDS(ON) of the driver pull-up switch (= 2.48 Ω).
- R_{OL} is the RDS(ON) of the driver pull-down switch (= 0.86 Ω).

\[ PD = (3.3)(0.0065) + (15)(0.0045) + 200,000 \times 3 \times 15 + \frac{1}{2} \left( 200,000 \right) \left( \frac{2.6}{2.6 + 2.4} + \frac{0.8}{0.8 + 12} \right) \]

\[ PD = 158 \text{ mW} \]

From Equation 10, calculate the junction temperature at the maximum device’s recommended ambient temperature of 125 °C:

\[ T_{j_{max}} = PD \times \theta_{ja} + TA \]

\[ T_{j_{max}} = (0.158 \times 60) + 125 = 134.5 \text{ °C} \]

This junction temperature is below the maximum Si828x junction temperature of 150 °C, and thus it is safe to operate at this power level.

DSAT Blanking Circuit: Using the standard 3 µs DSAT blanking circuit for Si8285 driver with CBL = 390 pF, R_DSAT = 100 Ω, and ES1J for 600 V D_DSAT diode.
Figure 8.2. Half-Bridge MOSFET Gate Example Circuit
9. Conclusion

In conclusion, this application note has provided users with an overview of gate power circuit and step-by-step calculations for the values of the external passive components, RH, RL, CBL, etc. used with the Si828x ISODrivers. The gate resistor (RH, RL) values are calculated based on the total gate charge required to transition the power switch between the ON and OFF states. Since the total gate charge value in the power device data sheet is for a typical device, it does not account for the variation between the power switches as well as the specific PCB trace impedances. Nevertheless, the gate resistor calculations in this application note provides a good approximation, and some fine tuning may be necessary to achieve optimized switching performance. The various control and protection features of the Si828x family have been discussed, and a half-bridge power circuit example has been provided.
Disclaimer
Silicon Labs intends to provide customers with the latest, accurate, and in-depth documentation of all peripherals and modules available for system and software implementers using or intending to use the Silicon Labs products. Characterization data, available modules and peripherals, memory sizes and memory addresses refer to each specific device, and “Typical” parameters provided can and do vary in different applications. Application examples described herein are for illustrative purposes only. Silicon Labs reserves the right to make changes without further notice to the product information, specifications, and descriptions herein, and does not give warranties as to the accuracy or completeness of the included information. Without prior notification, Silicon Labs may update product firmware during the manufacturing process for security or reliability reasons. Such changes will not alter the specifications or the performance of the product. Silicon Labs shall have no liability for the consequences of use of the information supplied in this document. This document does not imply or expressly grant any license to design or fabricate any integrated circuits. The products are not designed or authorized to be used within any FDA Class III devices, applications for which FDA premarket approval is required or Life Support Systems without the specific written consent of Silicon Labs. A “Life Support System” is any product or system intended to support or sustain life and/or health, which, if it fails, can be reasonably expected to result in significant personal injury or death. Silicon Labs products are not designed or authorized for military applications. Silicon Labs products shall under no circumstances be used in weapons of mass destruction including (but not limited to) nuclear, biological or chemical weapons, or missiles capable of delivering such weapons. Silicon Labs disclaims all express and implied warranties and shall not be responsible or liable for any injuries or damages related to use of a Silicon Labs product in such unauthorized applications.

Note: This content may contain offensive terminology that is now obsolete. Silicon Labs is replacing these terms with inclusive language wherever possible. For more information, visit www.silabs.com/about-us/inclusive-lexicon-project

Trademark Information
Silicon Laboratories Inc.®, Silicon Laboratories®, Silicon Labs®, SiLabs® and the Silicon Labs logo®, Bluegiga®, Bluegiga Logo®, Clockbuilder®, CMEMS®, DSPLL®, EFM®, EFM32®, EFR, Ember®, Energy Micro, Energy Micro logo and combinations thereof, “the world’s most energy friendly microcontrollers”, Ember®, EZLink®, EZRadio®, EZRadioPRO®, Gecko®, Gecko OS, Gecko OS Studio, ISOModem®, Precision32®, ProSLIC®, Simplicity Studio®, SiPHY®, Telegesis, the Telegesis Logo®, USBXpress®, Zentri, the Zentri logo and Zentri DMS, Z-Wave®, and others are trademarks or registered trademarks of Silicon Labs. ARM, CORTEX, Cortex-M3 and THUMB are trademarks or registered trademarks of ARM Holdings. Keil is a registered trademark of ARM Limited. Wi-Fi is a registered trademark of the Wi-Fi Alliance. All other products or brand names mentioned herein are trademarks of their respective holders.