Overview

This document provides design information for Smart Gate Driver Coupler operation.

This document is for reference only and should not be used as the basis for final device design.
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1. Introduction

The TLP5214A/TLP5214 product is a general-purpose gate driver coupler with additional functionality as shown in Table 1-1, including $V_{CE\text{(sat)}}$ detection, active miller clamp and fault output. It also provides protection for IGBT and MOSFET from overcurrent (typically from the inverter circuit).

<table>
<thead>
<tr>
<th></th>
<th>General-purpose gate driver coupler</th>
<th>Smart gate driver coupler</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Product name</strong></td>
<td>TLP352, TLP5701, etc.</td>
<td>TLP5214A/ TLP5214</td>
</tr>
<tr>
<td><strong>Package/internal circuit diagram</strong></td>
<td>DIP8, SO6L, etc.</td>
<td>SO16L</td>
</tr>
<tr>
<td><strong>Pins</strong></td>
<td>8pin, 6pin</td>
<td>16pin</td>
</tr>
<tr>
<td><strong>IGBT gate direct drive</strong></td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td><strong>UVLO function</strong></td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td><strong>$V_{CE\text{(sat)}}$ detection</strong></td>
<td>-</td>
<td>✓</td>
</tr>
<tr>
<td><strong>Active miller clamp</strong></td>
<td>-</td>
<td>✓</td>
</tr>
<tr>
<td><strong>FAULT output</strong></td>
<td>-</td>
<td>✓</td>
</tr>
</tbody>
</table>

TLP5214A/TLP5214 is designed for a wide range of applications, from inverter circuits used in industrial equipment (such as general-purpose inverters and power conditioners in solar power systems) to UPS and residential equipment such as home battery storage systems.
2. TLP5214A vs. TLP5214

Table 2-1 illustrates some of the key differences between the TLP5214 and TLP5214A smart gate driver coupler products from Toshiba.

<table>
<thead>
<tr>
<th>Item</th>
<th>Symbol</th>
<th>TLP5214A</th>
<th>TLP5214</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak output current (max)</td>
<td>$I_{OPH}$ / $I_{OPL}$</td>
<td>±4.0</td>
<td>±4.0</td>
<td>A</td>
</tr>
<tr>
<td>Operating temperature range</td>
<td>$T_{op}$</td>
<td>−40 to 110</td>
<td>−40 to 110</td>
<td>°C</td>
</tr>
<tr>
<td>Supply current (max)</td>
<td>$I_{CC2}$</td>
<td>3.8</td>
<td>3.5</td>
<td>mA</td>
</tr>
<tr>
<td>Supply voltage</td>
<td>$V_{CC2} - V_{EE}$</td>
<td>15 to 30</td>
<td>15 to 30</td>
<td>V</td>
</tr>
<tr>
<td>Threshold input current (max)</td>
<td>$I_{FLH}$</td>
<td>6</td>
<td>6</td>
<td>mA</td>
</tr>
<tr>
<td>Propagation delay (max)</td>
<td>$t_{PLH} / t_{PWL}$</td>
<td>150</td>
<td>150</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT threshold (typ.)</td>
<td>$V_{DESAT}$</td>
<td>6.5</td>
<td>6.5</td>
<td>V</td>
</tr>
<tr>
<td>Blanking capacitor charging current (max)</td>
<td>$I_{CHG}$</td>
<td>-0.24</td>
<td>-0.24</td>
<td>mA</td>
</tr>
<tr>
<td>Clamp pin threshold voltage (typ.)</td>
<td>$V_{tclmp}$</td>
<td>2.5</td>
<td>3.0</td>
<td>V</td>
</tr>
<tr>
<td>Propagation delay skew</td>
<td>$t_{psk}$</td>
<td>−80 to 80</td>
<td>−80 to 80</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT sense to 90% delay (max)</td>
<td>$t_{DESAT(90%)}$</td>
<td>500</td>
<td>500</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT sense to 10% delay (max)</td>
<td>$t_{DESAT(10%)}$</td>
<td>8.5</td>
<td>5</td>
<td>μs</td>
</tr>
<tr>
<td>DESAT leading edge blanking time (typ.)</td>
<td>$t_{DESAT(LEB)}$</td>
<td>1.1</td>
<td>-</td>
<td>μs</td>
</tr>
<tr>
<td>DESAT filter time (typ.)</td>
<td>$t_{DESAT(FILTER)}$</td>
<td>90</td>
<td>-</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT sense to low level FAULT signal delay (max)</td>
<td>$t_{DESAT(FAULT)}$</td>
<td>550</td>
<td>500</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT sense to low level FAULT signal delay (typ.)</td>
<td>$t_{DESAT(LOW)}$</td>
<td>200</td>
<td>200</td>
<td>ns</td>
</tr>
<tr>
<td>DESAT input mute (min)</td>
<td>$t_{DESAT(MUTE)}$</td>
<td>7</td>
<td>7</td>
<td>μs</td>
</tr>
<tr>
<td>Reset to high level FAULT signal delay</td>
<td>$t_{RESET(FAULT)}$</td>
<td>0.2 to 2</td>
<td>0.2 to 2</td>
<td>μs</td>
</tr>
<tr>
<td>Protection features</td>
<td>−</td>
<td>UVLO $V_{CE(sat)}$ detection</td>
<td>Active miller clamp Fault output</td>
<td>−</td>
</tr>
</tbody>
</table>

The shutdown time for the TLP5214A is slightly longer than the rise blanking time, to prevent false detection during power device startup.

The TLP5214 has no rise blanking time. It is designed for rapid switching MOSFET and other devices with instantaneous protection functionality.
3. Protection features

- **UVLO function**
  UVLO (under voltage lock-out) prevents accidental $V_{OUT}$ from the secondary internal circuit during the period while the output power voltage $V_{CC2}$ from the smart gate driver (typically in response to input power) has not yet attained the UVLO threshold voltage. Similarly, the secondary side shuts down operation to prevent accidental output in the event that the supply voltage drops below the UVLO detection voltage. UVLO resets when the supply voltage rises back up above the UVLO threshold voltage.

- **$V_{CE(sat)}$ detection**
  The $V_{CE(sat)}$ detection feature monitors the saturation level of the collector – emitter voltage $V_{CE}$ of the driver element (such as IGBT) from the DESAT terminal, and shuts down operation when overcurrent is detected. Normally, $V_{CE}$ is below the saturation voltage $V_{CE(sat)}$ (2 V approx.) when the IGBT is on. In the event of overcurrent causing non-saturation, if $V_{CE(sat)}$ increases beyond the set threshold a fault is declared and $V_{OUT}$ gradually shuts down.

- **Active miller clamp**
  The active miller clamp feature acts to minimize the increase in electrical potential associated with miller capacitance between the gate and drain of the IGBT or other component by bypassing gate resistance (or equivalent) and allowing direct connection to $V_{EE}$.

- **Fault output**
  The fault output feature outputs a fault signal to notify the primary side (the controller side) of a fault detected by the $V_{CE(sat)}$ detection feature. In the event that the $V_{CE(sat)}$ of IGBT (or other element) exceeds the standard threshold value of 6.5 V, the TLP5214A/TLP5214 initiates a two-stage overcurrent protection sequence as depicted in Figure 3.1.

  1. Soft shutdown of $V_{OUT}$ (gradual transition to OFF status) to prevent IGBT failure due to overcurrent;
  2. Fault signal sent to controller.

While most conventional devices take several microseconds to generate the fault signal to the controller and shut down the LED signal/coupler output, the TLP5214A is able to initiate the $V_{OUT}$ shutdown in less than 500 ns, and is therefore suitable for rapid-acting safety circuits.

![Figure 3-1 Overcurrent protection sequence](image-url)
**Protection circuit and reset procedure**

Once the protection circuit is triggered, the LED signal is not received for a preset period. This period is denoted $t_{DESAT(MUTE)}$. Figure 3-2 shows a connection diagram for an inverter application together with a timing chart for the protection circuit operation and reset procedure. Reset is triggered by the LED signal that follows after $t_{DESAT(MUTE)}$.

The protection circuit and reset sequence is as follows.

1. Overcurrent causes $V_{CE}$ of IGBT to exceed standard 6.5 V threshold; protection circuit initiated
2. Soft shutdown of coupler output to prevent secondary failure of IGBT due to wiring inductance
3. Signal sent to controller to reduce FAULT terminal to L level
4. New LED signal (following $t_{DESAT(MUTE)}$ protection operation) initiates reset procedure

![Figure 3-2 Protection and reset procedure—connection diagram and timing chart](image-url)
4. Application design

4.1. Parameters

Parameters for smart gate driver coupler applications are given below.

1. Gate resistance
   Gate resistance should be no greater than the maximum rated $I_{OP}$ value for the gate driver product, and should also be below the product’s maximum rated driver side current and power rating values. The power device (IGBT or MOSFET) side must also be taken into consideration since this has a bearing on turn-on and turn-off times.

2. Blanking time
   The product switches on in the presence of power and an input signal, and gate drive current is output from $V_{OUT}$. The DESAT function also operates at this time. For power devices with a longer turn-off time, it monitors the collector-emitter voltage level $V_{CE}$ and switches to shutdown mode at the point where it is about to fall.
   The timing of the voltage detection sequence can be adjusted using a blanking condenser or peripheral circuit.

3. Short-circuit monitoring
   The DESAT terminal is monitored constantly to detect power device faults such as short circuits. If the voltage exceeds the standard threshold value of 6.5 V, the product is switched to shutdown mode. The DESAT diode can be augmented with a Zener diode or SBD to further reduce the short-circuit threshold voltage for the power device.

4. Primary fault signal pull-up resistance
   The open collector feedback circuit output is connected to a pull-up resistor. Recovery time after a fault is governed by the resistance value of the pull-up resistor, and should be tailored to the system requirements and input power.

5. Preventing malfunction
   After all settings have been entered, it may be necessary to insert additional components to prevent malfunction.
4.2. Blanking time settings and adjustment method

4.2.1. Blanking time

Figure 4-1 shows a typical applied circuit with IGBT drive, while Figure 4-2 shows the timing chart for the switching sequence. When the LED input current $I_F$ switches from off to on, the increase in voltage at the output terminal $V_{OUT}$ causes the external IGBT to turn on. At the same time, the blanking capacitance charging current $I_{CHG}$ is output from the DESAT terminal for the purpose of monitoring the collector-emitter voltage $V_{CE}$ for the external IGBT, and the voltage at the DESAT terminal begins rising. TLP5214A has a $t_{DESAT(LEB)}$ setting designed to prevent malfunction associated with DESAT terminal rise.

When the IGBT switches on normally, the DESAT protection circuit needs to be disabled until $V_{CE}$ reaches $V_{th(IGBT)}$, the short-circuit threshold voltage for the IGBT; otherwise a malfunction will occur. Figure 4-2 illustrates the time from the $I_F$ rise until the DESAT voltage reaches the standard threshold of 6.5 V, known as the blanking time ($t_{BLANK}$). The blanking time depends on the value of the condenser ($C_{BLANK}$) between the DESAT and VE terminals.

Normally, $t_{BLANK}$ should be longer than the time $t_{th}$ required for $V_{CE}$ to reach $V_{th(IGBT)}$ but shorter than the IGBT short-circuit tolerance interval $t_{SC}$.

Figure 4-1 TLP5214A circuit diagram during $C_{BLANK}$ charging

Figure 4-2 TLP5214A timing chart for LED off → on sequence
4.2.2. Calculating the blanking time

t\(_{\text{BLANK}}\) is expressed in terms of \(C_{\text{BLANK}}, V_{\text{DESAT}}, I_{\text{CHG}}\) and \(t_{\text{DESAT}(\text{LEB})}\) as follows:

\[
t_{\text{BLANK}} = C_{\text{BLANK}} \times \frac{V_{\text{DESAT}}}{I_{\text{CHG}}} + t_{\text{DESAT}(\text{LEB})}
\]

where

\[
V_{\text{DESAT}} = 6.5 \text{ V (standard value)} \\
I_{\text{CHG}} = 240 \, \mu\text{A (standard value)} \\
t_{\text{DESAT}(\text{LEB})} = 1.1 \, \mu\text{s (standard value for TLP5214A)}
\]

Since \(C_{\text{BLANK}} = 200 \, \text{pF}\), \(t_{\text{BLANK}}\) is expressed as follows:

\[
t_{\text{BLANK}} = 200 \times 10^{-12} \text{ F} \times 6.5 \text{ V} / (240 \times 10^{-6}) \text{ A} + 1.1 \, \mu\text{s} = 6.5 \, \mu\text{s}
\]

Figure 4-3 shows the relationship between \(C_{\text{BLANK}}\) and \(t_{\text{BLANK}}\). As Figure 4-4 shows, a higher \(C_{\text{BLANK}}\) value lengthens the delay time until overcurrent protection is enabled by altering the gradient of the voltage rise time between the DESAT terminals. Note that in real-world applications, \(t_{\text{BLANK}}\) is also influenced by other factors such as the parasitic capacitance of connected diode(s).

The \(C_{\text{BLANK}}\) value can be used to adjust the \(t_{\text{BLANK}}\) period to prevent malfunction associated with overcurrent. Note that this period is shorter than the power device short-circuit tolerance interval.
Reference: Impact of $C_{BLANK}$ on representative waveform

Figure 4-5 shows actual waveform observations at DESAT and $V_{OUT}$ terminals when the TLP5214A LED is on.

It can be seen that the $C_{BLANK}$ value has a direct impact on $t_{BLANK}$. In real-world situations, we also have to consider the capacitance of connected diodes on the DESAT line as well as stray capacitance in the circuit, so these are incorporated into the design estimates. The example above assumes substrate capacitance of approximately 340 pF. (The capacitance of the probe used for waveform observation is also included.)
4.3. Blanking time vs. switching time

The switching time is the period from when the smart gate driver coupler LED comes on to when the IGBT is turned on (see Figure 4-6). It should be no greater than $t_{\text{BLANK}}$, thus:

$$t_{\text{pLH}} \text{ for TLP5214A + IGBT } t_{\text{ON}}^* = \text{switching time} < t_{\text{BLANK}}$$

* assuming $t_{\text{th}} \approx t_{\text{ON}}$

where

$t_{\text{pLH}} = 150 \text{ ns (max)}$ (from TLP5214A data sheet)

$t_{\text{ON}}$ is calculated as follows:

$$t_{\text{ON}} = \frac{Q_g}{I_O} \text{ (for IGBT)} / I_O \text{ (TLP5214A output current)}$$

For the purpose of this example we assume $V_{GE} = 15 \text{ V}$ and $I_O = 1.5 \text{ A}$ for IGBT GT30J341 switching.

Based on the $V_{CE}$ and $V_{GE} - Q_s$ characteristics on the data sheet we have $Q_g = 130 \text{ nC}$ (see Figure 4-7). Thus:

$$t_{\text{ON}} = \frac{130 \text{ nC}}{1.5 \text{ A}} \approx 87 \text{ ns}$$

So the switching time is given by:

$$t_{\text{ON}} = 150 \text{ ns} + 87 \text{ ns}$$

$$= 237 \text{ ns} < 6.5 \mu\text{s}$$

which is less than $t_{\text{BLANK}}$ as required.

![Figure 4-7 Switching time](image1)

![Figure 4-6 VCE - Q_s curve for power device](image2)
Reference: Timing sheet for fault event ON

Figure 4-8 shows waveforms at each terminal (I_F, V_O, I_O, V_CE and DESAT) during normal operation and during a fault event.

After input I_F to the smart gate driver coupler, V_O output drives the IGBT gate. If the IGBT switches normally, the V_CE voltage drops to the IGBT saturation voltage. The DESAT terminal voltage monitoring the V_CE terminal likewise drops down to the sum total of the saturation voltage and the DESAT diode V_F.

When an alarm short (or equivalent fault) ① occurs, depending on the nature of the fault, the current I_C (denoted by blue line) increases, leading to overcurrent ②. The increase in IGBT current forces up the IGBT V_CE(sat) ③ as shown in Figure 4-10. The DESAT terminal voltage also rises simultaneously ④. When the DESAT terminal voltage exceeds the threshold value (6.5 V standard), the coupler is deemed to have shorted and protection mode engages ⑤. The coupler shuts down V_O and I_O ⑥ and also initiates a gradual shutdown to prevent any noise associated with the abrupt change to off status ⑦. The coupler responsible for detecting the fault notifies the input side by switching on an internal LED and forwarding the fault status.

![Figure 4-8 Timing chart for normal operation and fault event (reference)](image)

When an alarm short (or equivalent fault) ① occurs, depending on the nature of the fault, the current I_C (denoted by blue line) increases, leading to overcurrent ②. The increase in IGBT current forces up the IGBT V_CE(sat) ③ as shown in Figure 4-10. The DESAT terminal voltage also rises simultaneously ④. When the DESAT terminal voltage exceeds the threshold value (6.5 V standard), the coupler is deemed to have shorted and protection mode engages ⑤. The coupler shuts down V_O and I_O ⑥ and also initiates a gradual shutdown to prevent any noise associated with the abrupt change to off status ⑦. The coupler responsible for detecting the fault notifies the input side by switching on an internal LED and forwarding the fault status.

![Figure 4-9 IGBT turn-off waveform](image)

![Figure 4-10 IGBT saturation voltage vs. collector current](image)
4.4. Setting the time using an external blanking circuit (R_B)

If we increase the C_BLANK value to boost noise tolerance during switching, this lengthens the charging time, raising the possibility that the protection feature may not engage during the t_SC period. Instead, we can use an external resistor as shown in Figure 4-11 to boost the C_BLANK charging current and ensure that protection remains enabled. The external resistor R_B between the DESAT terminals draws external current I_B from the output V_OUT which supplements I_CHG.

R_B allows greater control over the C_BLANK charging current and therefore greater design flexibility in regards to the blanking time.

![Figure 4-11 Suggested external blanking](image)

The voltage applied to the blanking condenser is expressed as follows:

\[
V_I = V_{OUT} - V_E \\
= R_B \times i(t) + 1 / C_{BLANK} / (I_{CHG} + i(t)dt) \\
i(t) = (V_I/R_B + I_{CHG}) \exp(-t/(C_{BLANK} \times R_B)) - I_{CHG} \\
V_{DESAT}(t) = V_I - R_B \times i(t) \\
= V_I - (V_I + R_B \times I_{CHG}) \exp(-t / (C_{BLANK} \times R_B)) + R_B \times I_{CHG}
\]

Thus blanking time is given by:

\[
t_{BLANK} = -C_{BLANK} \times R_B \times \log(1 - V_{DESAT} / (V_I + R_B \times I_{CHG}))
\]

For the TLP5214A, blanking time at DESAT rise is included so we add t_{DESAT(LEB)} to the above. Given that C_{BLANK} = 300 pF, R_B = 30 kΩ, V_OUT = 17 V, V_EE = -10 V and V_E ≈ 0 V, and from the data sheet we know that V_{DESAT} = 6.5 V, I_{CHG} = 0.25 mA and t_{DESAT(LEB)} = 1.1 μs, we have:

\[
t = -300 \times 10^{-12} \times 30 \times 10^3 \times \log(1 - 6.5 / (17 + 30 \times 10^3 \times 250 \times 10^{-6})) + 1.1 \times 10^{-6} \\
= -9000 \times 10^{-9} \times \log (1 - 6.5 / (17 + 7.5)) + 1.1 \times 10^{-6} \\
= -9 \times 10^{-6} \times \log (0.7346) + 1.1 \times 10^{-6} \\
= 2.774 \times 10^{-6} + 1.1 \times 10^{-6} \quad \text{which gives us } t_{BLANK} = 3.9 \mu s.
\]
4.4.1. Impact of $R_B$ on $V_{OUT}$ waveform

Figure 4-12 shows waveform observations for the circuit shown in Figure 4-13, with and without a 30 kΩ $R_B$ resistor, where $C_{BLANK} = 440$ pF (external 100 pF + test substrate capacitance of 340 pF), $V_{CC2} = 17$ V and $V_{EE} = -10$ V.

$t_{BLANK}$ is 11 μs without $R_B$ and 4.5 μs with $R_B$. Blanking time is shorter due to current $I_B$ flowing through $R_B$.

Thus a higher $C_{BLANK}$ value will not exceed the shorting protection time.

$R_B = 30$ kΩ

**Figure 4-12** $V_{OUT}$ waveform with no $R_B$ (top) and $R_B = 30kΩ$ (bottom)

Note: Test substrate capacitance = 340 pF approx. (including SBD and Zener diode)

**Figure 4-13** Test circuit with $R_B$
4.4.2. Varying $C_{BLANK}$ while keeping $R_B$ constant

Figure 4-14 shows the impact of $C_{BLANK}$ when $R_B$ is constant, using three values for $C_{BLANK}$: 100 pF, 330 pF and 680 pF.

![Figure 4-14 Impact of CBLANK on waveform (RB constant)](image)

Figure 4-15 shows the test circuit. Waveforms were observed with $R_B = 30 \, k\Omega$ (constant), $V_{CC2} = 17$ V and $V_{EE} = -10$ V, and $C_{BLANK}$ varying in the range 100 – 3,000 pF (substrate capacitance = 340 pF).

![Figure 4-15 Test circuit with RB constant and CBLANK varying](image)

Note: Test substrate capacitance = 340 pF approx. (including SBD and Zener diode)

$t_{BLANK}$ is given by the following expression:

$$t_{BLANK} = -C_{BLANK} \times R_B \times \log(1 - V_{DESAT}/(V_I + R_B \times I_{CHG}))$$

(see page 15)

Figure 4-16 plots the calculated $t_{BLANK}$ values against the observed values.

For $C_{BLANK}$ values up to approximately 2,000 pF, the calculated values are consistent with the observations. Note that larger condensers can be limited by the IGBT short-circuit withstand time, so it is important to consider the appropriate capacity for the condenser.
4.4.3. Varying $R_B$ while keeping $C_{BLANK}$ constant

Figure 4-17 shows the impact of $R_B$ when $C_{BLANK}$ is fixed.

![Figure 4-17 Impact of $R_B$ on waveform ($C_{BLANK}$ constant)](image1)

Figure 4-18 shows the test circuit. Waveforms were observed with external $C_{BLANK} = 330$ pF (excluding substrate capacitance), $V_{CC2} = 17$ V and $V_{EE} = -10$ V, and $R_B$ varying in the range $330\,\Omega$ to $30\,k\Omega$.

![Figure 4-18 Test circuit with $C_{BLANK}$ constant and $R_B$ varying](image2)

Note: Test substrate capacitance = 340 pF approx. (including SBD and Zener diode)

Figure 4-19 plots the observed $t_{BLANK}$ values against the estimates calculated using the expression on page 15.

While a lower resistance value can be used to keep $t_{BLANK}$ short, during $V_{OUT}$ output the higher current flowing to $R_B$ boosts current consumption.

![Figure 4-19 $t_{BLANK}$ estimates vs. observed values ($C_{BLANK}$ fixed)](image3)
4.5. Modifying the IGBT short detection threshold voltage

The DESAT terminal monitors the voltage at the terminal during $I_F$ input. If the terminal voltage $V_{\text{DESAT}}$ exceeds the standard threshold of 6.5 V, the DESAT circuit engages and the product goes into protection mode. Note that the power device $V_{\text{CE}}$ value extracted from the diode or resistor may differ slightly from the observed IGBT $V_{\text{CE}}$ value. Figure 4-20 shows how to regulate variation in the short detection threshold voltage when a diode or equivalent is present.

Figure 4-21 shows how we can add multiple DESAT diodes in order to either engage protection at a lower voltage or reduce the short detection threshold voltage based on the IGBT side voltage $V_{\text{th(IGBT)}}$ in line with the safe operating range of the IGBT. By reducing the voltage through $V_F$ for multiple devices, we can bring down $V_{\text{th(IGBT)}}$ and set it as a new $V_{\text{th(IGBT)}}$ value. This is method ①. The other method ②, using multiple Zener diodes, offers a greater degree of precision.

Method ①: $\text{New } V_{\text{th(IGBT)}} = V_{\text{DESAT}} - (n \times V_F + R_{\text{DESAT}} \times I_{\text{CHG}})$ where $n$ is the number of diodes

Method ②: $\text{New } V_{\text{th(IGBT)}} = V_{\text{DESAT}} - (V_F + V_Z + R_{\text{DESAT}} \times I_{\text{CHG}})$ where $V_Z$ is the Zener voltage

For example, with Method ①, if we use three diodes at $V_F = 0.4 \text{ V}$ and $240 \mu\text{A}$ and $R_{\text{DESAT}} = 100 \Omega$, we get: $\text{New } V_{\text{th(IGBT)}} = 6.5 - (3 \times 0.4 \text{ V} + 100 \Omega \times 240 \mu\text{A}) \approx 5.3 \text{ V}$

In normal operation, forward current flowing to the DESAT diode is used to monitor the IGBT $V_{\text{CE}}$ voltage. In high-power applications, elements such as reverse recovery current may be generated during switching and these can lead to false detection of DESAT voltage. An FRD with low parasitic capacitance can be used to keep reverse recovery current to a minimum.
4.6. Gate capacitance, gate resistance and propagation delay

Figure 4-22 illustrates the relationships between propagation delay $t_{PLH}$ / $t_{PHL}$ and $C_g$ and between $t_{PLH}$ / $t_{PHL}$ and $R_g$. The measuring circuit is shown in Figure 4-23 and the waveform observation point in Figure 4-24. It can be seen that $C_g$ and $R_g$ have negligible impact on propagation delay.
4.7. Gate capacitance, output power voltage and soft turn-off time

The soft turn-off time of the protection circuit \( t_{\text{DESAT}(10\%)} \) is governed by gate capacitance \( C_g \) and output power voltage \( V_{CC2} \). Figure 4-25 illustrates how \( C_g \) and gate resistance \( R_g \) affect soft turn-off time. Unlike normal switching operation, there is a soft shutdown followed by a gradual decline in electric potential, as shown in Figure 4-26. Clearly the soft turn-off time is impacted by both the power supply voltage and the gate capacitance.

![Figure 4-25 C\(_g\) and R\(_g\) vs. soft turn-off time](image)

![Figure 4-26 Soft turn-off time (observed)](image)
4.8. Bypass condenser and spare terminals

The smart gate driver coupler is a high-performance IC coupler. This means that malfunctions may occur if power supply noise and spare terminals are not dealt with correctly. Figure 4-27 shows how to employ a bypass condenser and what to do with terminals that are not in use.

① Install 1 µF bypass condensers between $V_E$ and $V_{CC2}$, and between $V_{CC2}$ and $V_{EE}$, as close as possible to the terminals. If the circuit uses a negative supply, another bypass condenser is needed between the $V_E$ and $V_{EE}$ terminals.

② Install a 0.1 µF bypass condenser between the $V_{CC1}$ and $V_S$ terminals, as close as possible to the terminals.

③ The LED terminal (pin 15) is a test pin and should not be connected to anything.

④ If the $V_{CLAMP}$ terminal is unused (i.e. when there is no need for an active miller clamp), short it to the $V_{EE}$ terminal.

⑤ If the DESAT terminal is unused, short it to the $V_E$ terminal and isolate from protection.

![Figure 4-27 Sample configuration for bypass condensers and unused terminals](image-url)
4.9. Protecting the DESAT terminal from voltage spikes during IGBT switching

A reverse recovery spike from an external IGBT freewheeling diode can cause the DESAT terminal to fall below the electrical potential of ground, generating forward current and damaging the DESAT terminal. It is important to protect the DESAT terminal by adding a Zener diode or Schottky diode (SBD) between the DESAT and \( V_E \) terminals as shown in Figure 4-28. Ensure that the diode has the correct rated value.

The Zener diode (\( V_Z = 7 \) to 8 V) protects the DESAT terminal from positive overvoltage while the Schottky diode prevents forward bypass by the parasitic diode at the DESAT terminal. Since adding diodes to prevent false detection will increase the capacitance between the DESAT and \( V_E \) terminals, it may be necessary to modify the \( C_{BLANK} \) setting.

![Figure 4-28 Preventing DESAT false detection](image-url)
4.10. Buffer transistor

Maximum output current from the TLP5214A/TLP5214 is 4 A. In the event of insufficient IGBT gate drive current, a buffer transistor can be added. A condenser should also be installed between the buffer input terminal and VEE to allow IGBT soft turn-off when protection engages. The capacitance will depend on the type of circuit and the soft turn-off time. For a $t_{\text{DESAT}(10\%)}$ value of 7 $\mu$s, a 25 nF condenser is recommended. A resistor $R_g$ is also required between the TLP5214A output and NPN/PNP base; the size of the resistor will depend on the maximum rated value of the product. Where the $V_{\text{CLAMP}}$ terminal is not used (typically due to negative supply), connect to the VEE terminal instead.

If the IGBT requires gate drive current greater than 4 A, consider using a transistor such as TTC3710B or TTA1452B as shown in Table 4-1. If using a DESAT diode, we recommend an FRD with voltage resistance equivalent to the IGBT.

<table>
<thead>
<tr>
<th>Product code</th>
<th>Absolute maximum rating</th>
<th>Package</th>
</tr>
</thead>
<tbody>
<tr>
<td>NPN</td>
<td>PNP</td>
<td></td>
</tr>
<tr>
<td>TTC3710B</td>
<td>TTA1452B</td>
<td>TO-220SIS</td>
</tr>
<tr>
<td>TPCP8902</td>
<td></td>
<td>PS8</td>
</tr>
</tbody>
</table>

Table 4-1 Transistor range

The IGBT gate current and transistor collector current are limited by the gate resistance, as per the expression below. Circuit design should take into consideration the maximum rating for the $V_{\text{OUT}}$ terminal and the rated values for the IGBT and the transistor.

$$\text{gate current} = \frac{(V_{\text{OH}} - V_{\text{OL}})}{(R_g + rg)}$$

where $rg$ is IGBT internal gate resistance.
4.11. LED signal waveform shaping

Where the control substrate and motor controller substrate are separate and there is considerable
distance between the TLP5214A and the CPU, inductance effects from the wiring can affect input
signal inclination.

Figure 4-30 shows how a hysteresis buffer inserted before the TLP5214A input terminal can be
used to shape the waveform of the input signal. Table 4-2 lists recommended CMOS logic buffer
products.

![Figure 4-30 Sample input signal waveform shaping configuration](image)

| Product code  | Function               | VCC(opr)   | |OH| / |IOL|  | tpd          | Package     |
|---------------|------------------------|------------|----------------------|-----------------|--------------|-------------|
| 74VHC244FT    | Octal Schmitt Bus Buffer | 1.8 to 5.5V | 16 mA                | 3.9 ns (typ.)   | TSSOP20B     |
4.12. Pull-up resistance $R_F$ for primary side fault signal

The product output is in open collector configuration, and requires pull-up resistance $R_F$ as shown in Figure 4-30 to be used as voltage signal.

- The secondary side LED for the fault signal is approximately less than 10 mA < (ILED = 8.5 mA min). In the event of a fault, the sink current at the fault signal output terminal is no less than 5 mA (reference value). Assuming $V_{CC1} = 5$ V, and allowing a 50% margin for chronological variation and temperature fluctuation, we get
  
  $$R_F = \frac{5 \text{ V}}{2.5 \text{ mA}} = 2 \text{ k}\Omega$$

  Thus, pull-up resistance should be no less than 2 kΩ.

  Around 10 kΩ is recommended in order to reduce current consumption. Note that FAULT terminal recovery times are longer for a higher $R_F$ value, so this should be taken into consideration.

- During normal operation, the open collector $T_r$ is off so the FAULT terminal has high impedance. At $V_{CC1} = 5$ V, $R_F$ should not be less than 2 kΩ, even for a long fault signal cable subject to external noise interference. Where terminals for multiple devices are connected together, leading to the possibility of simultaneous fault signals, the supply current should also be taken into consideration.

  Buffers may be used to reduce noise and/or supplement the supply current.

![Figure 4-31 Circuit diagram for pull-up resistor $R_F$ in primary side fault signal](image-url)
4.13. During a protection operation

The reset function that cancels a smart gate driver coupler protection operation is triggered by LED input.

When protection mode is engaged due to a fault, the LED has to be switched off and on again to reset the protection operation.

Figure 4-32 shows the internal block diagram for the TLP5214A, while Figure 4-33 shows the timing sequence from the moment protection is engaged until the fault output is canceled via an LED reset signal. When the TLP5214A enters protection mode, the feedback LED (i.e., the fault output LED) lights and the fault outputs to report an IGBT fault. If fault mode persists, the secondary side fault output LED lights and a current of approximately 10 mA flows between the VCC2 and VE terminals, leading to increased power loss on the secondary side. Bootstrap circuits with IC power are liable to sudden discharge from the condensers, so the possibility of voltage drop must be taken into consideration. If the protection function engages, the system should be stopped and restarted as soon as possible.

When VCC2 = 30 V and fault mode is engaged with LED current of 10 mA, IC consumption is approximately 28 V (Vdrop). The loss at IC is given by:

\[ P = V_{\text{drop}} \times I_{\text{LED}} \]
\[ = 28 \text{ V} \times 10 \text{ mA} \]
\[ = 280 \text{ mW} \]

Given the thermal resistance of the product Rth(j-a) = 70°C/W (from Table 4-3) we can calculate the temperature as follows:

\[ \Delta T_j = 70 \times 0.28 = 19.6^\circ\text{C} \]

So high temperatures should be avoided.

Table 4-3 TLP5214A thermal resistance
Test substrate: standard JEDEC

<table>
<thead>
<tr>
<th>Reference value</th>
<th>TLP5214A</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rth(j-a)</td>
<td>70°C/W</td>
</tr>
</tbody>
</table>
4.14. Miller capacitance error and response

Malfunctions associated with miller capacitance $C_{CG}$ between the IGBT collector and gate are typical of the problems that can be caused by inverter switching noise. Figure 4-34 illustrates a miller capacitance malfunction in a typical coupler configuration on the lower arm of the inverter circuit.

When the IGBT for the upper arm of the inverter circuit is on, the electrical potential at the mid-point abruptly increases, while the displacement current $I_S (= C_{CG} \times (dV_{CG}/dt))$ flows via the lower arm IGBT $C_{CG}$ in the photocoupler output direction. As it passes through the circuit gate resistor $R_g$ the voltage drops and the gate voltage rises, potentially causing a false ON at the IGBT and shorting out the upper and lower arms.

Three strategies can be employed to prevent a miller capacitance malfunction.

1. **Use negative power**
   A negative power supply puts the gate at negative potential with the IGBT is off, preventing malfunction (see Figure 4-35).

2. **Change the gate resistance**
   Lower gate resistance combined with diodes in parallel suppresses the gate’s contribution to the voltage increase (see Figure 4-36).

3. **Add a miller clamp circuit**
   The smart gate driver coupler includes an active miller clamp circuit that creates a short-circuit between the IGBT gate and emitter. When photocoupler output switches from high to low and the gate voltage falls below 3 V (approximately), the MOSFET between $V_{CLAMP}$ and $V_{EE}$ switches on and the gate is clamped on the emitter $V_{EE}$, as shown in Figure 4-37.
5. Key design considerations

The smart gate driver is a high-performance IC coupler with a number of built-in features. It should be noted that peripheral drivers can sometimes cause the smart gate driver to malfunction. The following considerations should also be taken into account at the design stage.

1. Gate resistance $R_g$
   A larger $R_g$ will help to reduce the surge voltage at switching as well as the likelihood of $dv/dt$ striking error, but the higher resistance may also exacerbate losses due to longer switching times for power devices. The gate resistance value should take into account peripheral circuits and power devices.

2. Separation between driver circuit and power device
   Excessive separation between the driver circuit (coupler) and the power device can add noise to and cause oscillation of the gate signal. The potential for a malfunction can be minimized by designing the driver circuit and power device as close together as possible; connecting them with the thickest possible wiring; and using a higher gate resistance value and/or one that is close to that of the power device.
   Since the wiring for the DESAT terminal that monitors the power device saturation voltage can affect the blanking time, it should be kept as far as possible from $V_E$ so that it does not create capacitance.

3. Bootstrap circuit diode
   Given that the GND potential of the high side IGBT/MOSFET can vary anywhere between zero and 600 V or higher, depending on the application, power for the driver coupler has to come from a floating power supply or bootstrap circuit. If the bootstrap option is used, a high-speed diode is recommended, one that is designed for the same voltages as the power device (at least 600 V).

4. Gate-emitter resistance $R_{GE}$
   The IGBT can fail if voltage is applied to the collector-emitter when gate-emitter is open. This can be prevented by either adding in up to 10 kΩ of resistance or changing the power supply input order so that the gate is served first.

5. Power device in parallel
   Power devices for high-capacitance inverters are often wired in parallel. In this case it is important that the circuit is designed to provide all devices with the same level of current. This will prevent oscillation, which can occur if current is unbalanced and becomes concentrated in a single device.
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