ACPL-5160 and ACPL-5161 [1]

2.5-Amp Gate Drive Optocoupler with Integrated (V_{CE}) Desaturation Detection and Fault Status Feedback

Data Sheet

Description

This family of 2.5-Amp Gate Drive Optocouplers provides Integrated Desaturation (V_{CE}) Detection and Fault Status Feedback for IGBT V_{CE} fault protection in a rugged, hermetically sealed package. The devices are capable of operation and storage over the full military temperature range and can be purchased as either commercial-grade products or in fully MIL-STD compliant versions. The military standard devices are manufactured and tested on a MIL-PRF-38534 certified line to Class H specifications.

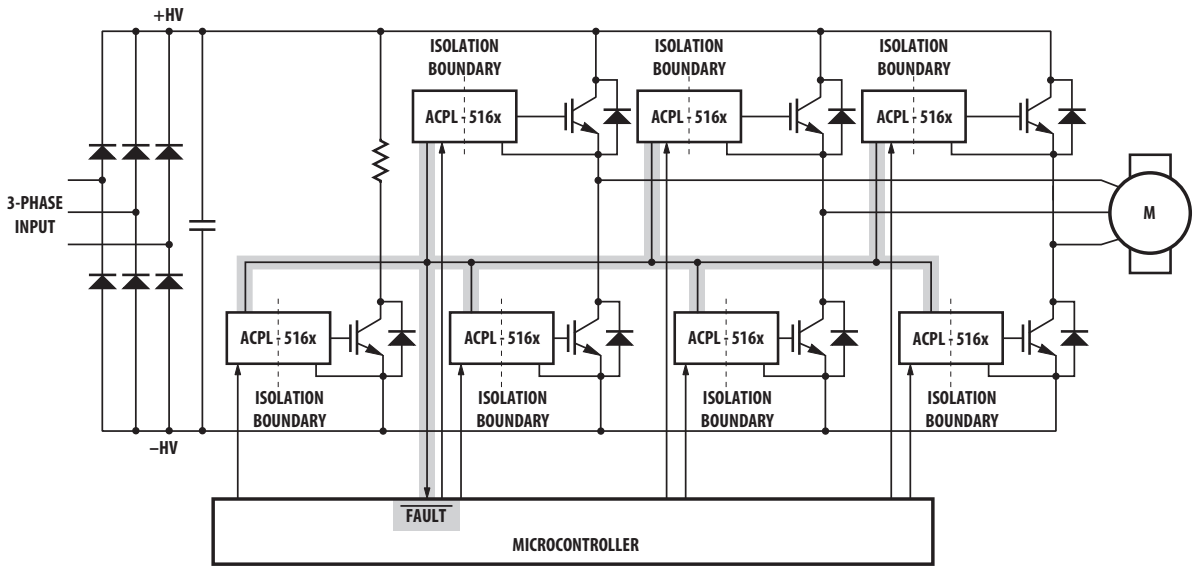
CAUTION It is advised that normal static precautions be taken in handling and assembly of this component to prevent damage and/or degradation which may be induced by ESD.

1. See Selection Guide — Lead Configuration Options for available extensions.

Features

- 2.5A maximum peak output current
- Drive IGBTs up to I_C = 150A, V_{CE} = 1200V
- Optically isolated, FAULT status feedback
- Hermetically sealed ceramic package
- CMOS/TTL compatible
- 500-ns max. switching speeds
- *Soft* IGBT turn-off
- Integrated fail-safe IGBT protection
  - DESAT (V_{CE}) detection
  - Undervoltage Lock-Out protection (UVLO) with hysteresis
- User configurable: inverting, noninverting, auto-reset, auto-shutdown
- Wide operating V_{CC} range: 15V to 30V
- –55°C to +125°C operating temperature range
- 15-kV/µs Typical Common Mode Rejection (CMR) at V_{CM} = 1000V

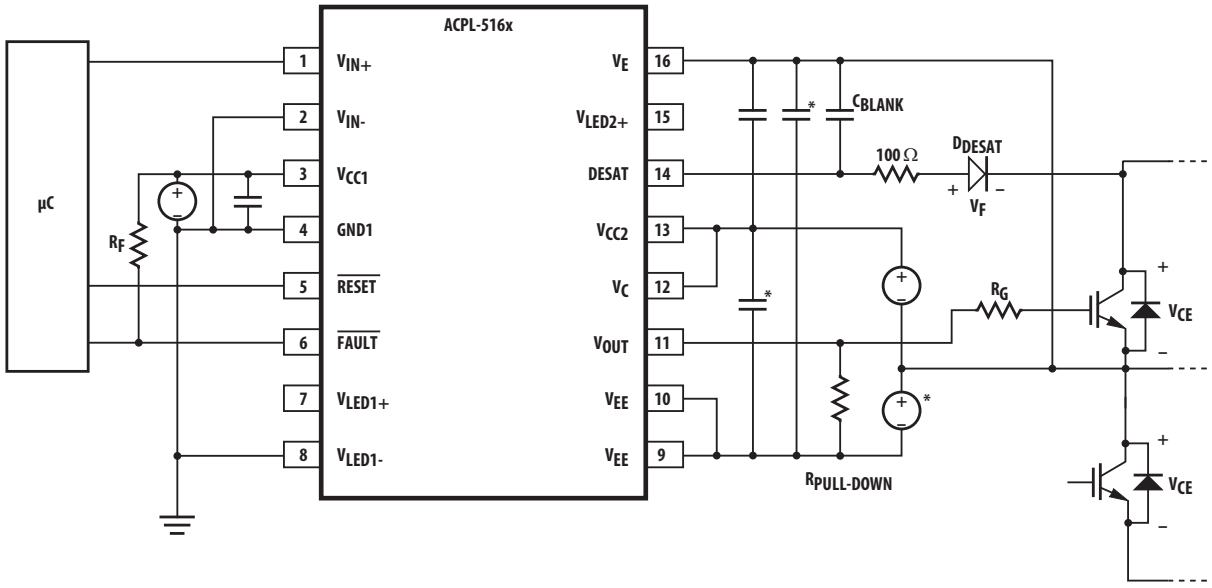
Fault Protected IGBT Gate Drive



Typical Fault Protected IGBT Gate Drive Circuit

The ACPL-516x is an easy-to-use, intelligent gate driver which makes IGBT V_{CE} fault protection compact, affordable, and easy-to-implement. Features such as user configurable inputs, integrated V_{CE} detection, undervoltage lockout (UVLO), *soft* IGBT turn-off and isolated fault feed-back provide maximum design flexibility and circuit protection.

Figure 1 Typical Desaturation Protected Gate Drive Circuit, Noninverting



Description of Operation during Fault Condition

1. DESAT terminal monitors the IGBT V_{CE} voltage through D_{DESAT}.
2. When the voltage on the DESAT terminal exceeds 7V, the IGBT gate voltage (V_{OUT}) is slowly lowered.
3. FAULT output goes low, notifying the microcontroller of the fault condition.
4. Microcontroller takes appropriate action.

Output Control

The outputs (V_{OUT} and $\overline{FAULT}$) of the ACPL-516x are controlled by the combination of V_{IN}, UVLO and a detected IGBT DESAT condition. As indicated in the following table, the ACPL-516x can be configured as inverting or noninverting using the V_{IN+} or V_{IN-} inputs respectively. When an inverting configuration is desired, V_{IN+} must be held high and V_{IN-} toggled. When a noninverting configuration is desired, V_{IN-} must be held low and V_{IN+} toggled. Once UVLO is not active ($V_{CC2} - V_E > V_{UVLO}$), V_{OUT} is allowed to go high, and the DESAT (pin 14) detection feature of the ACPL-516x will be the primary source of IGBT protection. UVLO is needed to ensure DESAT is functional. Once $V_{UVLO+} > 11.6V$, DESAT remains functional until $V_{UVLO-} < 12.4V$. Thus, the DESAT detection and UVLO features of the ACPL-516x work in conjunction to ensure constant IGBT protection.

V_{IN+}	V_{IN-}	UVLO ($V_{CC2} - V_E$)	DESAT Condition Detected on Pin 14	Pin 6 ($\overline{FAULT}$) Output	V_{OUT}
X	X	Active	X	X	Low
X	X	X	Yes	Low	Low
Low	X	X	X	X	Low
X	High	X	X	X	Low
High	Low	Not Active	No	High	High

Product Overview Description

The ACPL-516x is a highly integrated power control device that incorporates all the necessary components for a complete, isolated IGBT gate drive circuit with fault protection and feedback into one rugged, hermetically sealed package. TTL input logic levels allow direct interface with a microcontroller, and an optically isolated power output stage drives IGBTs with power ratings of up to 150A and 1200V. A high-speed internal optical link minimizes the propagation delays between the microcontroller and the IGBT while allowing the two systems to operate at very large common mode voltage differences that are common in industrial motor drives and other power switching applications. An output IC provides local protection for the IGBT to prevent damage during overcurrents, and a second optical link provides a fully isolated fault status feedback signal for the microcontroller. A built-in *watchdog* circuit monitors the power stage supply voltage to prevent IGBT damage caused by insufficient gate drive voltages. This integrated IGBT gate driver is designed to increase the performance and reliability of a motor drive without the cost, size, and complexity of a discrete design.

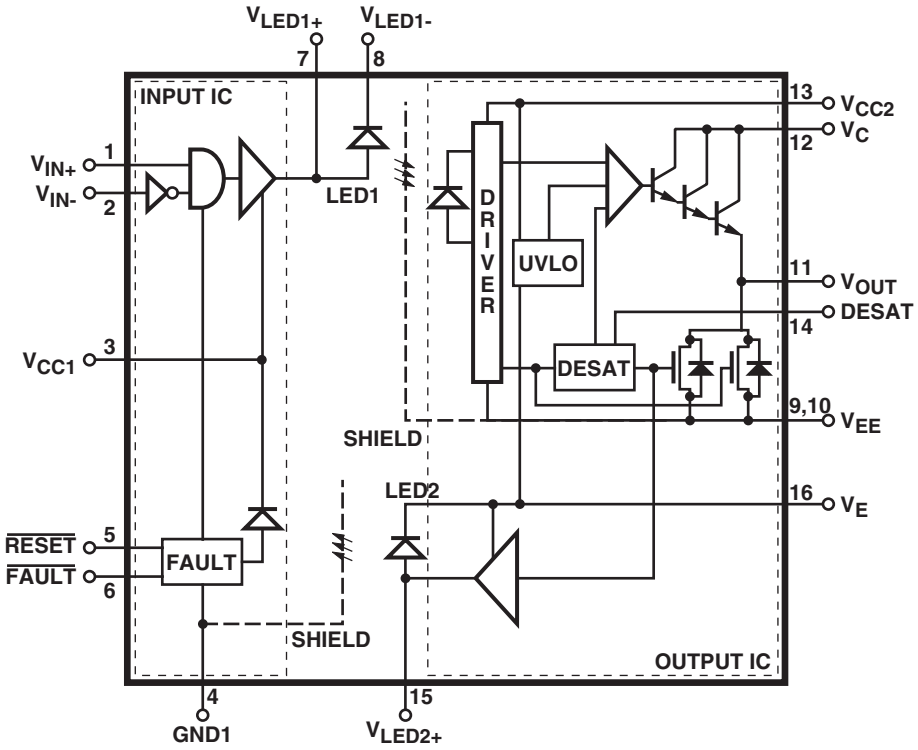
Two light emitting diodes and two integrated circuits housed in the same 16-pin ceramic package provide the input control circuitry, the output power stage, and two optical channels.

The input buffer IC is designed on a bipolar process, while the output detector IC is manufactured on a high voltage BiCMOS/Power DMOS process. The forward optical signal path, as indicated by LED1, transmits the gate control signal. The return optical signal path, as indicated by LED2, transmits the fault status feedback signal. Both optical channels are completely controlled by the input and output ICs, respectively, making the internal isolation boundary transparent to the microcontroller.

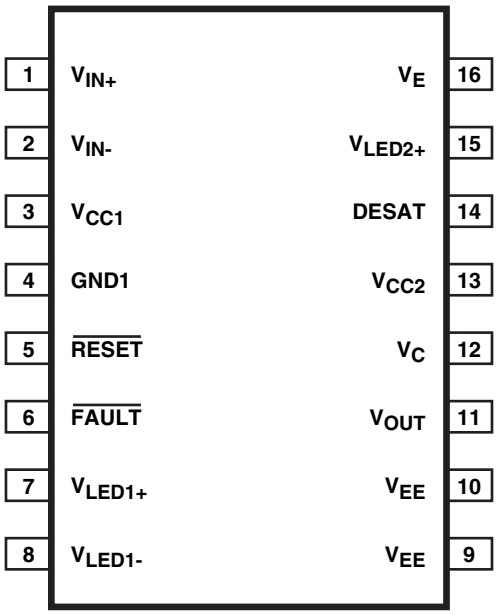
Under normal operation, the input gate control signal directly controls the IGBT gate through the isolated output detector IC. LED2 remains off and a fault latch in the input buffer IC is disabled. When an IGBT fault is detected, the output detector IC immediately begins a *soft* shutdown sequence, reducing the IGBT current to zero in a controlled manner to avoid potential IGBT damage from inductive overvoltages. Simultaneously, this fault status is transmitted back to the input buffer IC via LED2, where the fault latch disables the gate control input and the active low fault output alerts the microcontroller.

During powerup, the Undervoltage Lockout (UVLO) feature prevents the application of insufficient gate voltage to the IGBT, by forcing the ACPL-516x's output low. Once the output is in the high state, the DESAT (V_{CE}) detection feature of the ACPL-516x provides IGBT protection. Thus, UVLO and DESAT work in conjunction to provide constant IGBT protection.

Figure 2 Functional Diagram



Package Pinout



Symbol	Description	Symbol	Description
V_{IN+}	Noninverting gate drive voltage output (V_{OUT}) control input.	VE	Common (IGBT emitter) output supply voltage.
$V_{IN–}$	Inverting gate drive voltage output (V_{OUT}) control input.	V_{LED2+}	LED 2 anode. This pin must be left unconnectedfor guaranteed data sheet performance. (For optical coupling testing only.)
V_{CC1}	Positive input supply voltage (4.5V to 5.5V).	DESAT	Desaturation voltage input. When the voltage on DESAT exceeds an internal reference voltage of 7V while the IGBT is on, FAULT output is changed from a high impedance state to a logic low state within 5 µs. [a]
GND1	Input Ground.	V_{CC2}	Positive output supply voltage.
$\overline{RESET}$	$\overline{FAULT}$ reset input. A logic low input for at least 0.1 µs, asynchronously resets $\overline{FAULT}$ output high and enables V_{IN}. Synchronous control of $\overline{RESET}$ relative to V_{IN} is required. $\overline{RESET}$ is not affected by UVLO. Asserting $\overline{RESET}$ while V_{OUT} is high does not affect V_{OUT}.	V_C	Collector of output pull-up triple-darlington transistor. It is connected to V_{CC2} directly or through a resistor to limit output turn-on current.
$\overline{FAULT}$	Fault output. $\overline{FAULT}$ changes from a high impedance state to a logic low output within 5 µs of the voltage on the DESAT pin exceeding an internal reference voltage of 7V. $\overline{FAULT}$ output remains low until $\overline{RESET}$ is brought low. $\overline{FAULT}$ output is an open collector that allows the $\overline{FAULT}$ outputs from all ACPL-516x in a circuit to be connected together in a wired-OR forming a single fault bus for interfacing directly to the microcontroller.	V_{OUT}	Gate drive voltage output.
V_{LED1+}	LED 1 anode. This pin must be left unconnected for guaranteed data sheet performance. (For optical coupling testing only.)	V_{EE}	Output supply voltage.
$V_{LED1–}$	LED 1 cathode. This pin must be connected to ground.		

a. In most applications, V_{CC1} is powered up first (before V_{CC2}) and powered down last (after V_{CC2}). This is desirable for maintaining control of the IGBT gate. In applications where V_{CC2} is powered up first, it is important to ensure that V_{IN+} remains low until V_{CC1} reaches the proper operating voltage (minimum 4.5V) to avoid any momentary instability at the output during V_{CC1} ramp-up or ramp-down.

Selection Guide — Lead Configuration Options

Part Number and Options

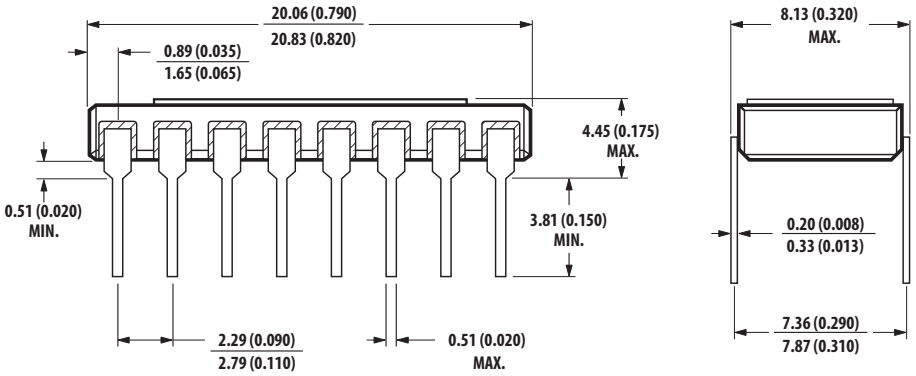
Commercial Grade	ACPL-5160
MIL-PRF-38534, Class H	ACPL-5161
Standard Lead Finish[a]	Gold Plate
Solder Dipped[b]	Option -200
Gull Wing/Soldered[b]	Option -300

a. Gold Plate lead finish: Maximum gold thickness of leads is <100 micro inches. Typical is 60 to 90 micro inches.

b. Solder lead finish: Sn63/Pb37.

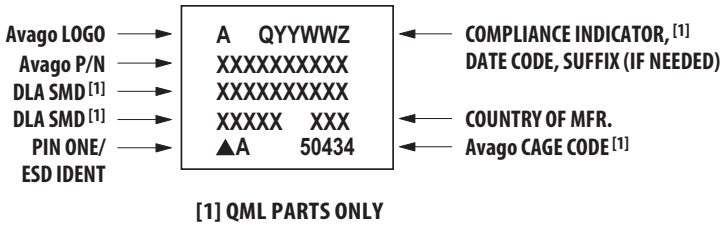
Outline Drawings

16-Pin DIP Through Hole, 2 Channels



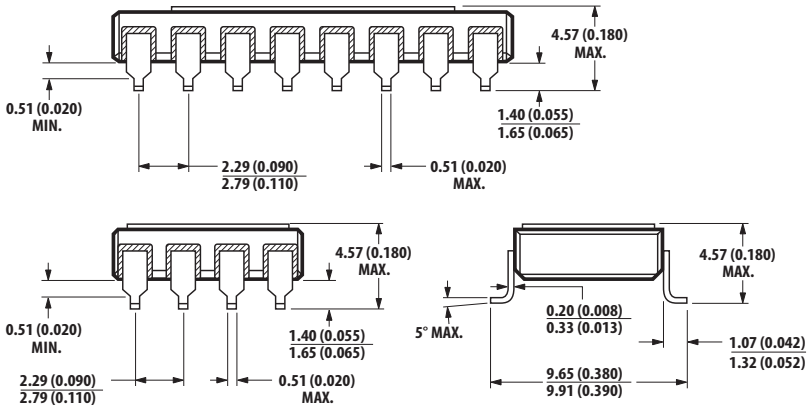
Note: Dimensions in millimeters (inches).

Device Marking



[1] QML PARTS ONLY

Hermetic Optocoupler Options

Option	Description
200	Lead finish is solder dipped rather than gold plated. This option is available on standard Commercial and Class H product.
300	Surface mountable hermetic optocoupler with leads cut and bent for gull wing assembly. This option is available on standard Commercial and Class H product. This option has solder-dipped leads. 

Absolute Maximum Ratings

Parameter	Symbol	Min	Max	Unit	Note
Storage Temperature	T_S	–65	150	°C	
Operating Temperature	T_A	–55	125	°C	
Output IC Junction Temperature	T_J	—	150	°C	a
Peak Output Current	$\lvert I_{o(peak)} \rvert$	—	2.5	A	b
Fault Output Current	I_{FAULT}	—	8.0	mA	
Positive Input Supply Voltage	V_{CC1}	–0.5	5.5	V	
Input Pin Voltages	V_{IN+}, V_{IN-}, and $V_{\overline{RESET}}$	–0.5	V_{CC1}	V	
Total Output Supply Voltage	$(V_{CC2} - V_{EE})$	–0.5	35	V	
Negative Output Supply Voltage	$(V_E - V_{EE})$	–0.5	15	V	c
Positive Output Supply Voltage	$(V_{CC2} - V_E)$	–0.5	$35 - (V_E - V_{EE})$	V	
Gate Drive Output Voltage	$V_{o(peak)}$	–0.5	V_{CC2}	V	
Collector Voltage	V_C	V_{EE} + 5 V	V_{CC2}	V	
DESAT Voltage	V_{DESAT}	V_E	V_E + 10	V	
Output IC Power Dissipation	P_O	—	600	mW	a
Input IC Power Dissipation	P_I	—	150	mW	

a. To achieve the absolute maximum power dissipation specified, pins 4, 9, and 10 require ground plane connections and may require airflow. For details on how to estimate junction temperature and power dissipation, see the Thermal Model section in the application notes at the end of this data sheet. The actual power dissipation achievable depends on the application environment (PCB layout, air flow, part placement, and so on). No power derating is required when operating below 125 °C using a high conductivity board. If a low conductivity board is used, then output IC power dissipation is derated linearly at 20 mW/°C above 120 °C. Input IC power dissipation is derated linearly at 5 mW/°C above 120°C.

b. Maximum pulse width = 10 µs, maximum duty cycle = 0.2%. This value is intended to allow for component tolerances for designs with I_O peak minimum = 2.0A. For additional details on I_{OH} peak, see the applications section. Derate linearly from 3.0A at +25°C to 2.5A at +125°C. This compensates for increased I_{OPEAK} due to changes in V_{OL} over temperature.

c. This supply is optional. Required only when negative gate drive is implemented.

Recommended Operating Conditions

Parameter	Symbol	Min	Max	Unit	Note
Operating Temperature	T_A	–55	125	°C	
Input Supply Voltage	V_{CC1}	4.5	5.5	V	a
Total Output Supply Voltage	$(V_{CC2} – V_{EE})$	15	30	V	b
Negative Output Supply Voltage	$(V_E – V_{EE})$	0	15	V	c
Positive Output Supply Voltage	$(V_{CC2} – V_E)$	15	$30 – (V_E – V_{EE})$	V	
Collector Voltage	V_C	$V_{EE} + 6$	V_{CC2}	V	

a. In most applications, V_{CC1} is powered up first (before V_{CC2}) and powered down last (after V_{CC2}). This is desirable for maintaining control of the IGBT gate. In applications where V_{CC2} is powered up first, it is important to ensure that V_{IN+} remains low until V_{CC1} reaches the proper operating voltage (minimum 4.5V) to avoid any momentary instability at the output during V_{CC1} ramp-up or ramp-down.

b. 15V is the recommended minimum operating positive supply voltage ($V_{CC2} – V_E$) to ensure adequate margin in excess of the maximum V_{UVLO+} threshold of 13.5V. For High Level Output Voltage testing, V_{OH} is measured with a DC load current. When driving capacitive loads, V_{OH} approaches V_{CC} as I_{OH} approaches zero units.

c. This supply is optional. Required only when negative gate drive is implemented.

Electrical Specifications (DC)

Unless otherwise noted, all typical values at $T_A = 25°C$, $V_{CC1} = 5V$, and $V_{CC2} – V_{EE} = 30V$, $V_E – V_{EE} = 0V$; all Minimum/Maximum specifications are at Recommended Operating Conditions.

Parameter	Symbol	Min	Typ	Max	Unit	Test Conditions	Fig	Note
Logic Low Input Voltages	VI_{N+L}, $V_{IN–L}$, $V_{\overline{RESETL}}$	—	—	0.8	V	—		
Logic High Input Voltages	V_{IN+H}, $V_{IN–H}$, $V_{\overline{RESETH}}$	2.0	—	—	V	—		
Logic Low Input Currents	I_{IN+L}, $I_{IN–L}$, $I_{\overline{RESETL}}$	–0.5	–0.36	—	mA	$V_{IN} = 0.4V$		
$\overline{FAULT}$ Logic Low Output Current	$I_{\overline{FAULTL}}$	5.0	12	—	mA	$V_{FAULT} = 0.4V$	29	
$\overline{FAULT}$ Logic High Output Current	$I_{\overline{FAULTH}}$	–40	—	—	µA	$V_{FAULT} = V_{CC1}$	30	
High Level Output Current	I_{OH}	–0.5	–1.5	—	A	$V_{OUT} = V_{CC2} – 4V$	3, 8, 31	a
		–2.0	—	—		$V_{OUT} = V_{CC2} – 15V$		b
Low Level Output Current	I_{OL}	0.5	2.0	—	A	$V_{OUT} = V_{EE} + 2.5V$	4, 9, 32	a
		2.0	—	—		$V_{OUT} = V_{EE} + 15V$		b
Low Level Output Current During Fault Condition	I_{OLF}	90	150	230	mA	$V_{OUT} – V_{EE} = 14V$	5, 33	c
High Level Output Voltage	V_{OH}	$V_C – 3.5$	$V_C – 2.5$	$V_C – 1.5$	V	$I_{OUT} = –100$ mA	6, 8, 34	d, e, f
		$V_C – 2.9$	$V_C – 2.0$	$V_C – 1.2$		$I_{OUT} = –650$ µA		
		—	—	V_C		$I_{OUT} = 0$		
Low Level Output Voltage	V_{OL}	—	0.12	0.5	V	$I_{OUT} = 100$ mA	7, 9, 35	g
High Level Input Supply Current	I_{CC1H}	—	18	22	mA	$V_{IN+} = V_{CC1} = 5.5V$, $V_{IN–} = 0V$	10, 37, 36	
Low Level Input Supply Current	I_{CCIL}	—	6.5	11	mA	$V_{IN+} = V_{IN–} = 0V$, $V_{CC1} = 5.5V$		

Parameter	Symbol	Min	Typ	Max	Unit	Test Conditions	Fig	Note
Output Supply Current	I_{CC2}	—	2.8	5	mA	V_{OUT} open	11, 12, 38, 39	f
Low Level Collector Current	I_{CL}	—	0.3	1.0	mA	$I_{OUT} = 0$	15, 58	h
High Level Collector Current	I_{CH}	—	0.3	1.3	mA	$I_{OUT} = 0$	15,56	h
		—	1.2	3.0	mA	$I_{OUT} = -650$ µA	15,57	
V_E Low Level Supply Current	I_{EL}	–0.7	–0.43	0	mA	—	14, 60	
V_E High Level Supply Current	I_{EH}	–0.5	–0.16	0	mA	—	14, 39	i
Blanking Capacitor Charging Current	I_{CHG}	–0.13	–0.26	–0.33	mA	$V_{DESAT} = 0 – 6V$	13, 40	f, j
		–0.18	–0.26	–0.33	mA	$V_{DESAT} = 0 – 6V$, $T_A = 25°C – 125°C$		
Blanking Capacitor Discharge Current	I_{DSCHG}	10	37	—	mA	$V_{DESAT} = 7V$	41	
UVLO Threshold	V_{UVLO+}	11.6	12.4	13.5	V	$V_{OUT} > 5V$	42	e, f, k
	$V_{UVLO–}$	—	11.2	12.4	V	$V_{OUT} < 5V$		e, f, l
UVLO Hysteresis	$(V_{UVLO+} – V_{UVLO–})$	0.4	1.2	—	V	—		
DESAT Threshold	V_{DESAT}	6.5	7	7.5	V	$V_{CC2} – V_E > V_{UVLO–}$	16, 43	f

a. Maximum pulse width = 50 µs, maximum duty cycle = 0.5%.

b. Maximum pulse width = 10 µs, maximum duty cycle = 0.2%. This value is intended to allow for component tolerances for designs with IO peak minimum = 2.0A. For additional details on I_{OH} peak, see the applications section. Derate linearly from 3.0A at +25°C to 2.5A at +125°C. This compensates for increased I_{OPEAK} due to changes in V_{OL} over temperature.

c. For further details, see the Slow IGBT Gate Discharge during Fault Condition section in the application notes at the end of this data sheet.

d. 15V is the recommended minimum operating positive supply voltage ($V_{CC2} – V_E$) to ensure adequate margin in excess of the maximum V_{UVLO+} threshold of 13.5V. For High Level Output Voltage testing, V_{OH} is measured with a DC load current. When driving capacitive loads, V_{OH} approaches V_{CC} as I_{OH} approaches zero units.

e. Maximum pulse width = 1.0 ms, maximum duty cycle = 20%.

f. Once V_{OUT} of the ACPL-516x is allowed to go high ($V_{CC2} – V_E > V_{UVLO}$), the DESAT detection feature of the ACPL-516x is the primary source of IGBT protection. UVLO is needed to ensure DESAT is functional. Once $V_{UVLO+} > 11.6V$, DESAT remains functional until $V_{UVLO-} < 12.4V$. Therefore, the DESAT detection and UVLO features of the ACPL-516x work in conjunction to ensure constant IGBT protection.

g. To clamp the output voltage at $V_{CC} – 3V_{BE}$, a pull-down resistor between the output and V_{EE} is recommended to sink a static current of 650A while the output is high. See the Output Pull-Down Resistor section in the application notes at the end of this data sheet if an output pull-down resistor is not used.

h. The recommended output pull-down resistor between V_{OUT} and V_{EE} does not contribute any output current when $V_{OUT} = V_{EE}$.

i. Does not include LED2 current during fault or blanking capacitor discharge current.

j. For further details, see the Blanking Time Control section in the application notes at the end of this data sheet.

k. This is the *increasing* (that is, turn-on or *positive going* direction) of $V_{CC2} – V_E$.

l. This is the *decreasing* (that is, turn-off or *negative going* direction) of $V_{CC2} – V_E$.

Switching Specifications (AC)

Unless otherwise noted, all typical values at T_A = 25°C, V_{CC1} = 5V, and $V_{CC2} - V_{EE}$ = 30V, $V_E - V_{EE}$ = 0V; all Minimum/Maximum specifications are at Recommended Operating Conditions.

Parameter	Symbol	Min	Typ	Max	Unit	Test Conditions	Fig.	Note
V_{IN} to High Level Output Propagation Delay Time	t_{PLH}	0.1	0.28	0.5	µs	Rg = 10Ω, Cg = 10 nF, f = 10 kHz, Duty Cycle = 50%	17, 18, 19, 20, 21, 22, 44, 53,54	a
V_{IN} to Low Level Output Propagation Delay Time	t_{PHL}	0.1	0.29	0.5	µs			
Pulse Width Distortion	PWD	–0.3	0.01	0.30	µs			b, c
Propagation Delay Difference Between Any Two Parts	$(t_{PHL} - t_{PLH})$ PDD	–0.35	—	0.35	µs			c, d
10% to 90% Rise Time	t_r	—	0.1	—	µs		44	
90% to 10% Fall Time	t_f	—	0.1	—	µs			
DESAT Sense to 90% V_{OUT} Delay	$t_{DESAT(90\%)}$	—	0.18	0.5	µs	Rg = 10Ω, Cg = 10 nF	23, 55	e
DESAT Sense to 10% V_{OUT} Delay	$t_{DESAT(10\%)}$	—	1.9	3.0	µs	$V_{CC2} - V_{EE}$ = 30V	24, 27, 45,55	
DESAT Sense to Low Level $\overline{FAULT}$ Signal Delay	$t_{DESAT(\overline{FAULT})}$	—	1.5	5	µs	—	25, 46, 55	f
DESAT Sense to DESAT Low Propagation Delay	$t_{DESAT(LOW)}$	—	0.25	—	µs	—	55	g
$\overline{RESET}$ to High Level $\overline{FAULT}$ Signal Delay	$t_{RESET(\overline{FAULT})}$	3	6.5	20	µs	—	26, 27, 55	h
$\overline{RESET}$ Signal Pulse Width	$PW_{\overline{RESET}}$	0.1	—	—	µs	—		
UVLO to V_{OUT} High Delay	$t_{UVLO\ ON}$	—	4.0	—	µs	V_{CC2} = 1.0 ms ramp	48	i
UVLO to V_{OUT} Low Delay	$t_{UVLO\ OFF}$	—	6.0	—	µs			j
Output High Level Common Mode Transient Immunity	$\|CM_H\|$	9	15	—	kV/µs	T_A = 25°C, V_{CM} = 1000V, V_{CC2} = 30V	49, 50, 51, 52	k
Output Low Level Common Mode Transient Immunity	$\|CM_L\|$	9	15	—	kV/µs	T_A = 25°C, V_{CM} = 1000V, V_{CC2} = 30V		l

a. This load condition approximates the gate load of a 1200V/75A IGBT.

b. Pulse Width Distortion (PWD) is defined as $|t_{PHL} - t_{PLH}|$ for any given unit.

c. As measured from V_{IN+}, V_{IN-} to V_{OUT}.

d. The difference between t_{PHL} and t_{PLH} between any two ACPL-516x parts under the same test conditions

e. Supply voltage dependent.

f. This is the amount of time from when the DESAT threshold is exceeded, until the $\overline{FAULT}$ output goes low.

g. This is the amount of time the DESAT threshold must be exceeded before V_{OUT} begins to go low, and the $\overline{FAULT}$ output to go low.

h. This is the amount of time from when $\overline{RESET}$ is asserted low, until $\overline{FAULT}$ output goes high. The minimum specification of 3 µs is the guaranteed minimum $\overline{FAULT}$ signal pulse width when the ACPL-516x is configured for Auto-Reset. For further details, see the Auto-Reset section in the application notes at the end of this data sheet.

i. This is the *increasing* (that is, turn-on or *positive going* direction) of $V_{CC2} - V_E$.

j. This is the *decreasing* (that is, turn-off or *negative going* direction) of $V_{CC2} - V_E$.

k. Common mode transient immunity in the high state is the maximum tolerable dV_{CM}/dt of the common mode pulse, V_{CM}, to assure that the output remains in the high state (that is, V_O > 15V or $\overline{FAULT}$ > 2V). A 100 pF and a 3-kΩ pull-up resistor is needed in fault detection mode.

l. Common mode transient immunity in the low state is the maximum tolerable dV_{CM}/dt of the common mode pulse, V_{CM}, to assure that the output remains in a low state (that is, V_O < 1.0V or $\overline{FAULT}$ < 0.8V).

Package Characteristics

Over recommended operating conditions (T_A = –55°C to +125°C) unless otherwise specified.

Parameter	Symbol	Test Conditions	Group A Subgroups	Limits			Unit	Fig	Note
				Min	Typ[a]	Max			
Input-Output Leakage Current	I_{I-O}	V_{I-O} = 1500 V_{DC}, RH ≤ 65%, t = 5 sec., T_A = 25°C	1	—	—	1.0	µA		b, c
Resistance (Input-Output)	R_{I-O}	V_{I-O} = 500 V_{DC}		—	10^{12}	—	Ω		c
Capacitance (Input-Output)	C_{I-O}	f = 1 MHz		—	2.8	—	pF		c

a. All typicals at T_A = 25°C.

b. This is a momentary withstand test, not an operating condition.

c. Device considered a two-terminal device: pins 1 to 8 shorted together and pins 9 to 16 shorted together.

Performance Plots

Figure 3 I_{OH} vs. Temperature

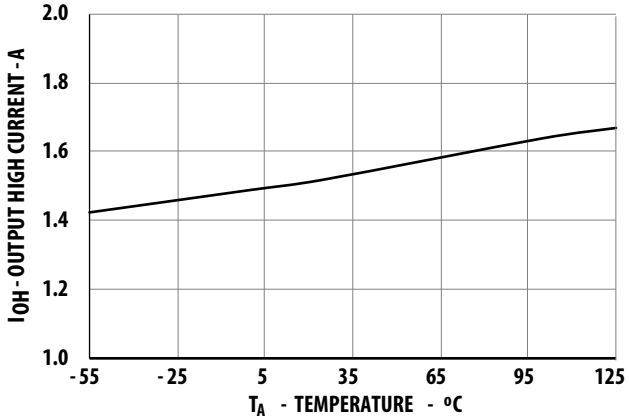


Figure 4 I_{OL} vs. Temperature

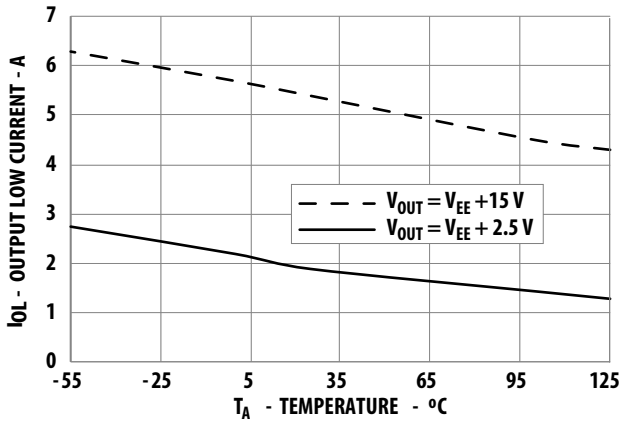


Figure 5 I_{OLF} vs. Temperature

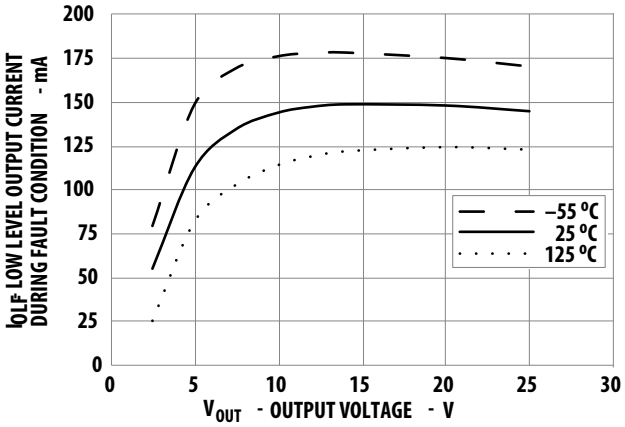


Figure 6 V_{OH} vs. Temperature

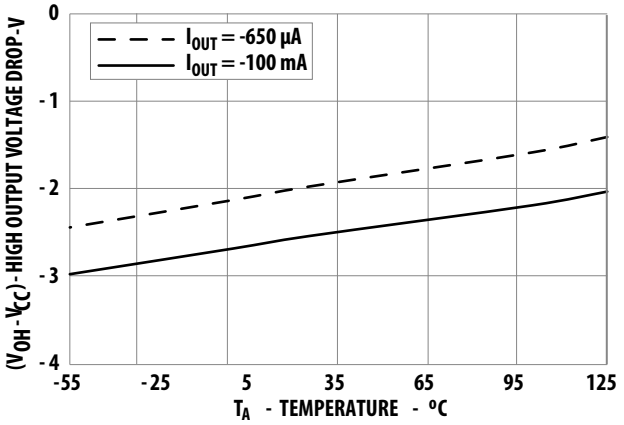


Figure 7 V_{OL} vs. Temperature

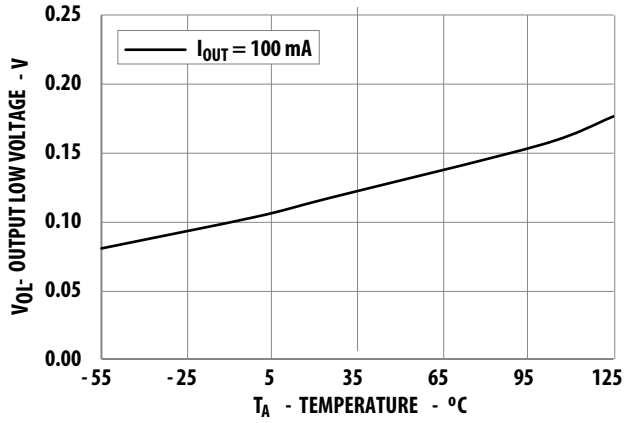


Figure 8 V_{OH} vs. I_{OH}

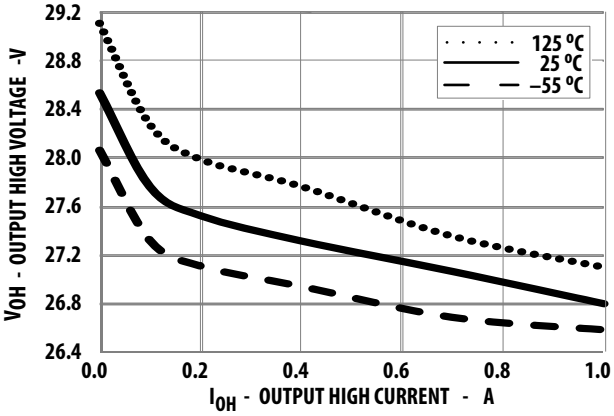


Figure 9 V_{OL} vs. I_{OL}

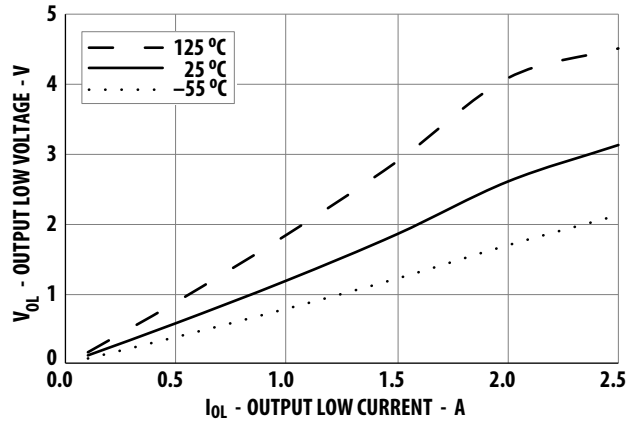


Figure 10 I_{CC1} vs. Temperature

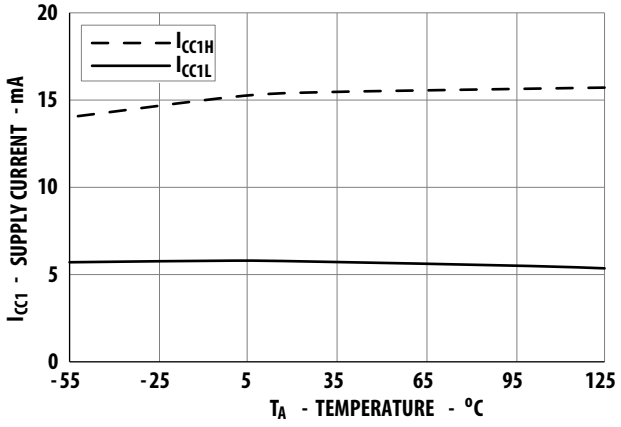


Figure 11 I_{CC2} vs. Temperature

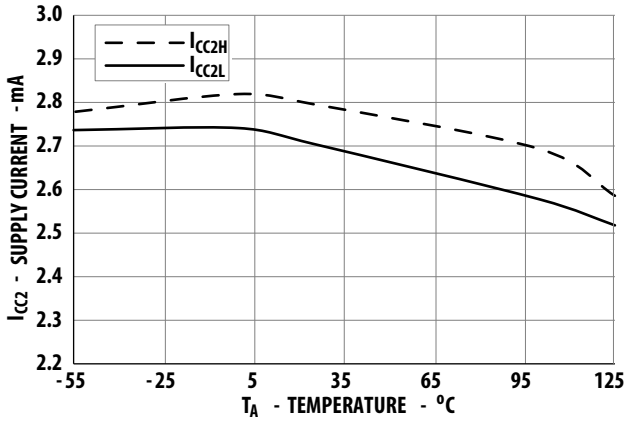


Figure 12 I_{CC2} vs. V_{CC2}

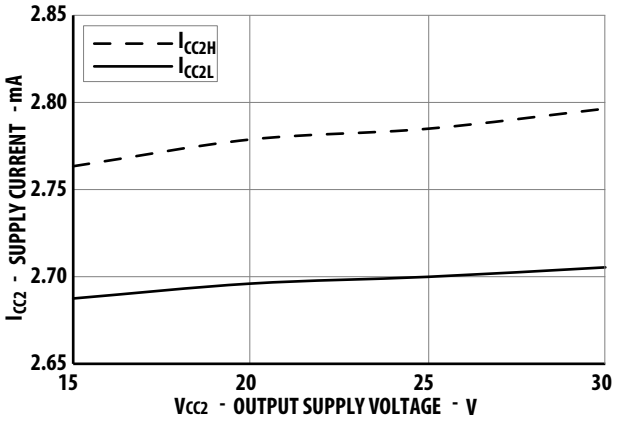


Figure 13 I_{CHG} vs. Temperature

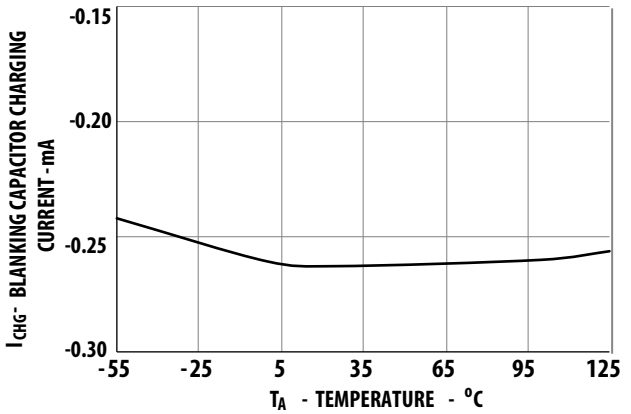


Figure 14 I_E vs. Temperature

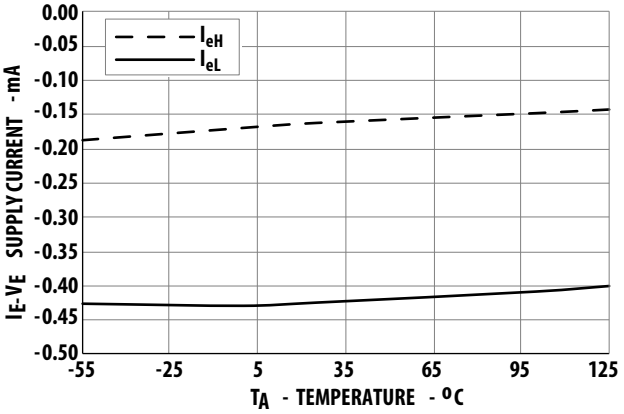


Figure 15 I_C vs. I_{OUT}

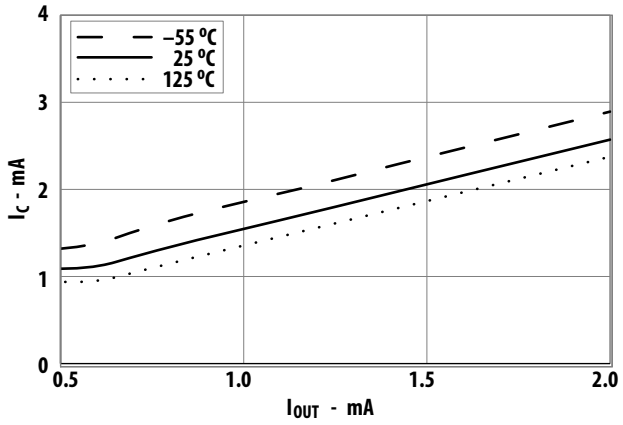


Figure 16 DESAT Threshold vs. Temperature

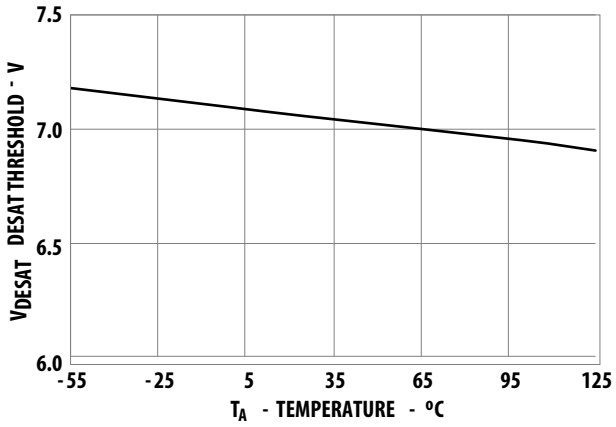


Figure 17 Propagation Delay vs. Temperature

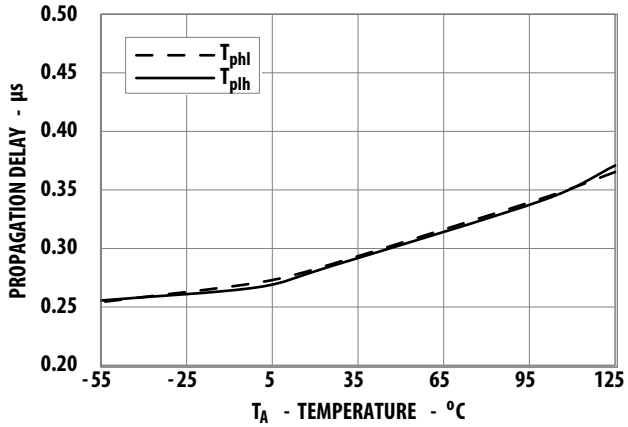


Figure 18 Propagation Delay vs. Supply Voltage

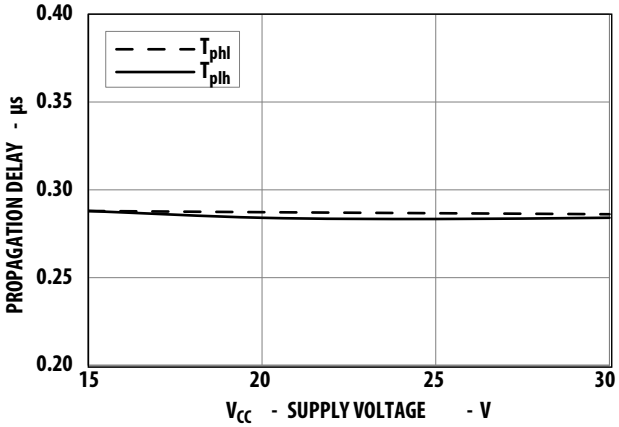


Figure 19 V_{IN} to High Propagation Delay vs. Temperature (T_{PLH})

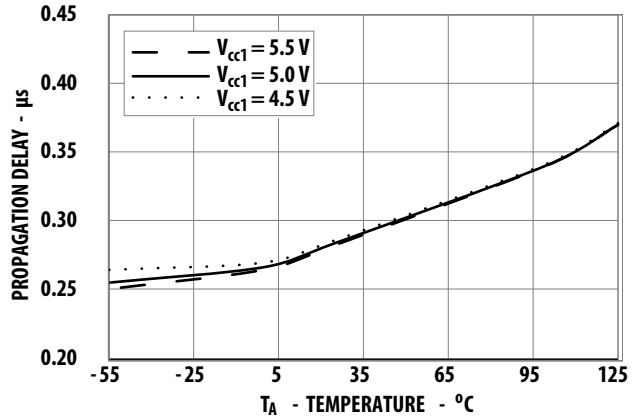


Figure 20 V_{IN} to Low Propagation Delay vs. Temperature (T_{PHL})

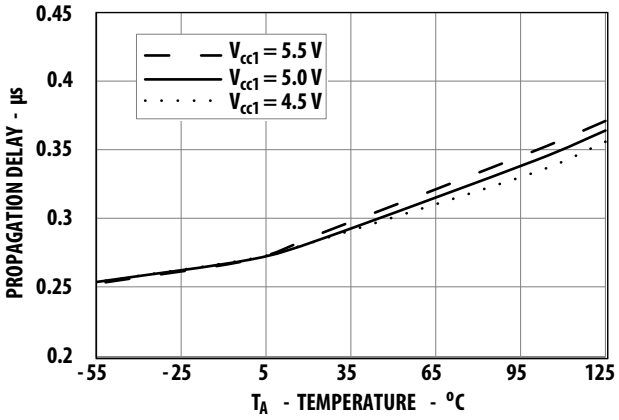


Figure 21 Propagation Delay vs. Load Capacitance

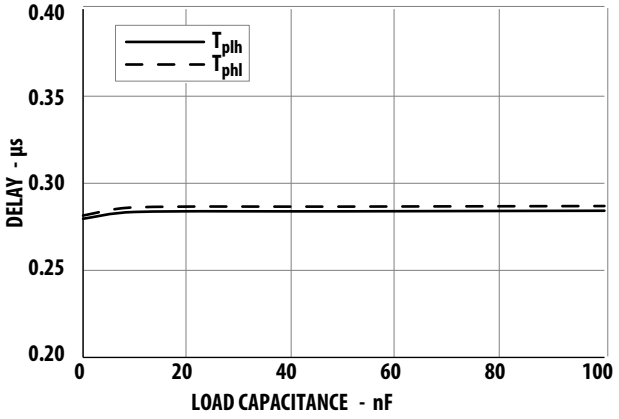


Figure 22 Propagation Delay vs. Load Resistance

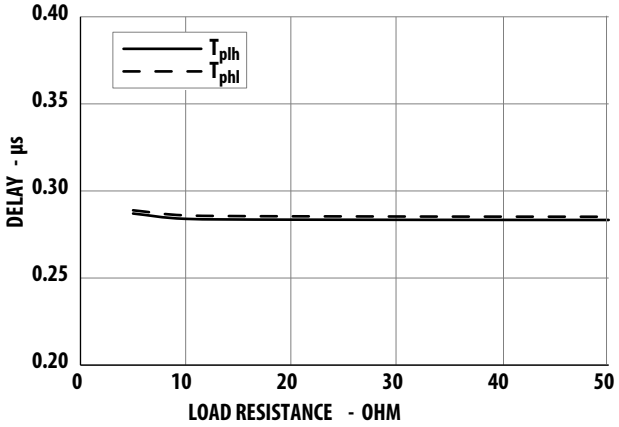


Figure 23 DESAT Sense to 90% V_{OUT} Delay vs. Temperature

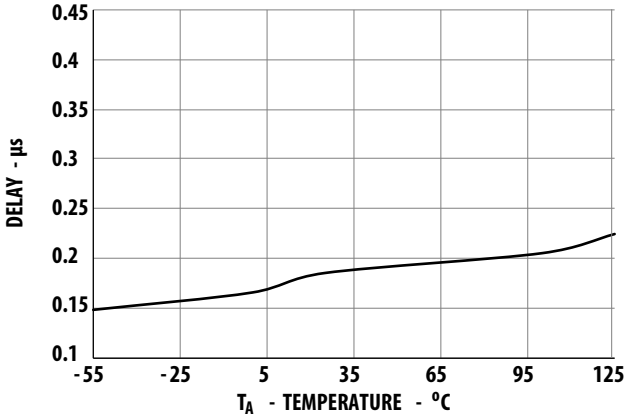


Figure 24 DESAT Sense to 10% V_{OUT} Delay vs. Temperature

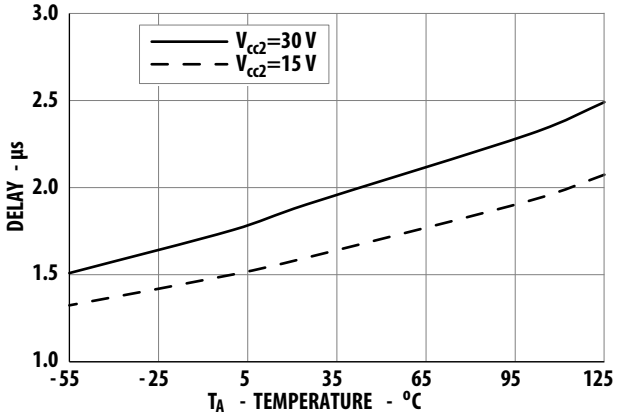


Figure 25 DESAT Sense to Low Level Fault Signal Delay vs. Temperature

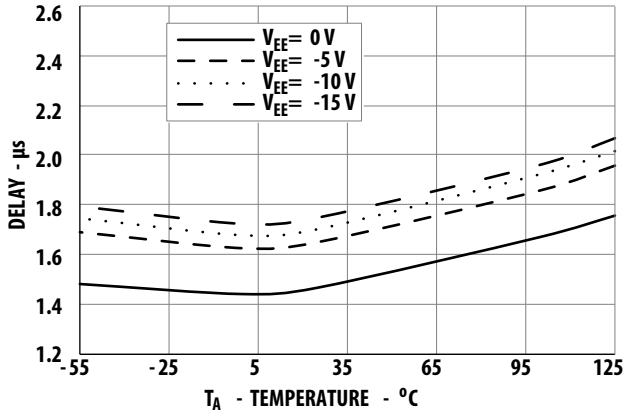


Figure 26 DESAT Sense to 10% V_{OUT} Delay vs. Load Capacitance

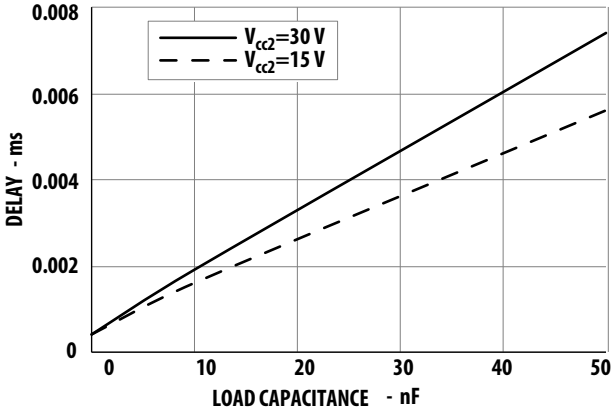


Figure 27 DESAT Sense to 10% V_{OUT} Delay vs. Load Resistance

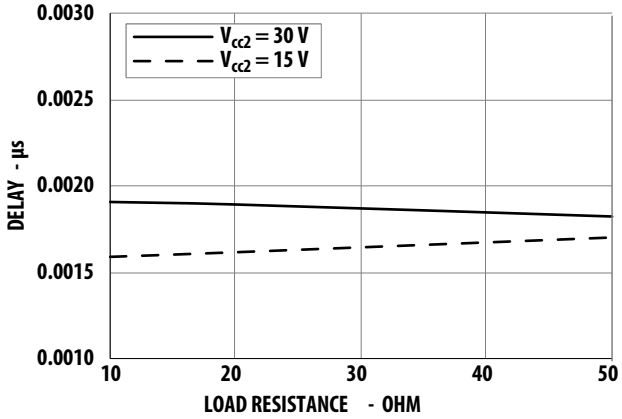
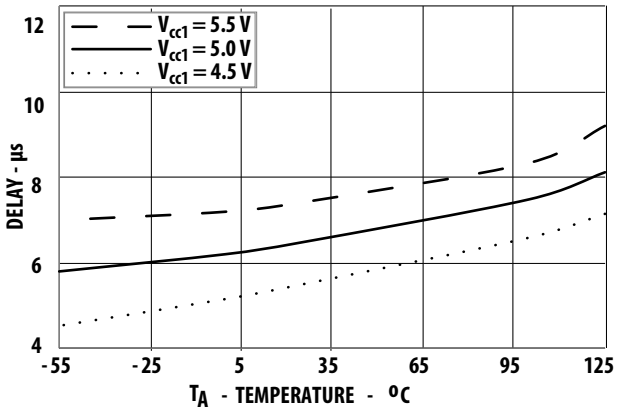


Figure 28 $\overline{\text{RESET}}$ to High Level Fault Signal Delay vs. Temperature



Test Circuit Diagrams

Figure 29 $I_{\overline{FAULTL}}$ Test Circuit

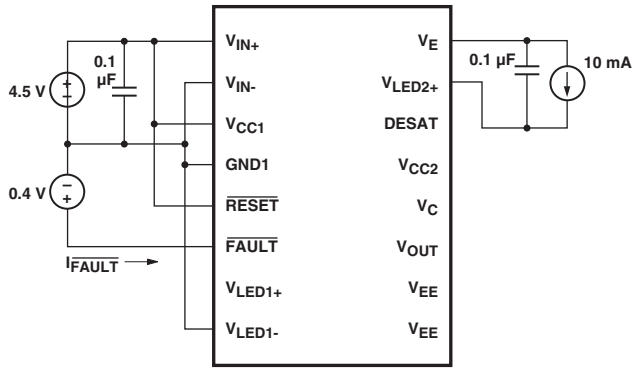


Figure 30 $I_{\overline{FAULTH}}$ Test Circuit

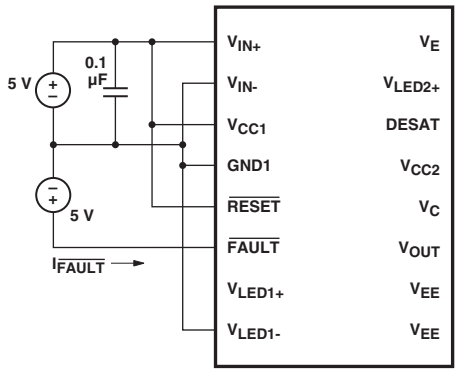


Figure 31 I_{OH} Pulsed Test Circuit

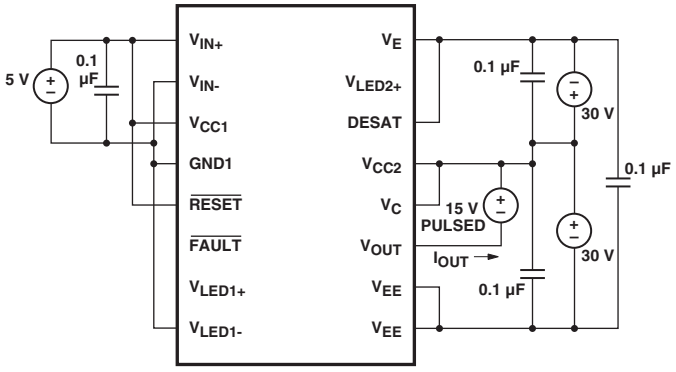


Figure 32 I_{OL} Pulsed Test Circuit

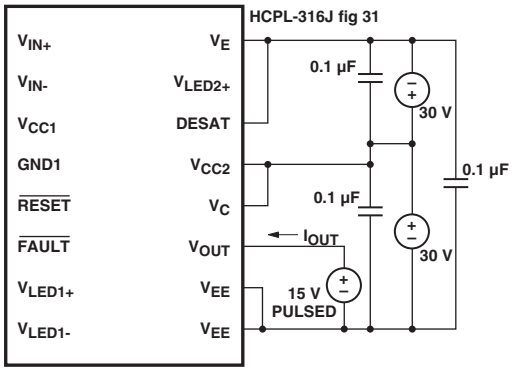


Figure 33 I_{OLF} Test Circuit

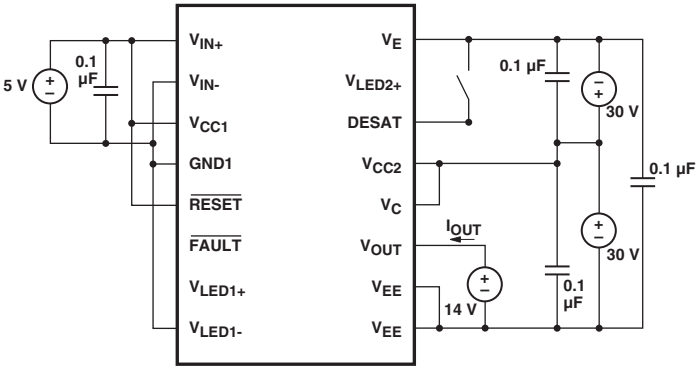


Figure 34 V_{OH} Pulsed Test Circuit

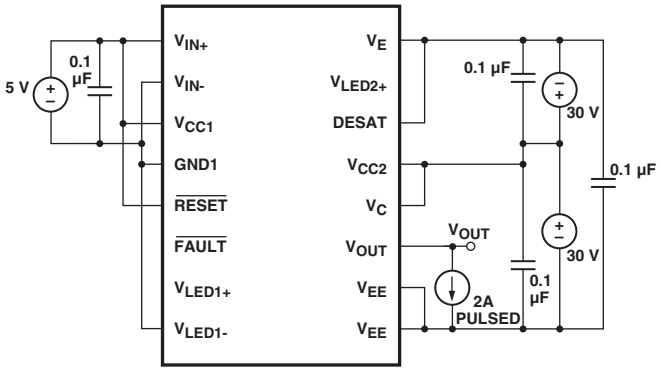


Figure 35 V_{OL} Test Circuit

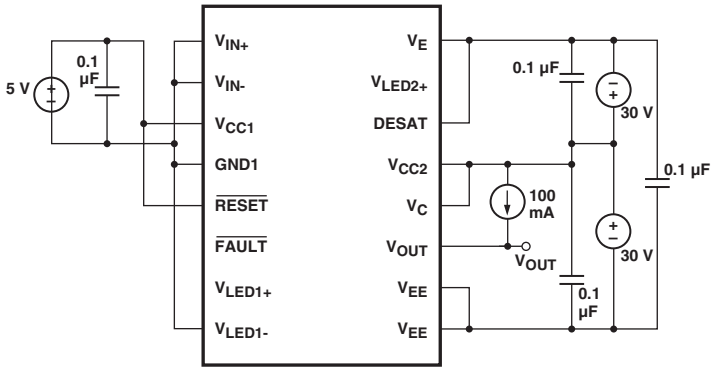


Figure 36 I_{CC1H} Test Circuit

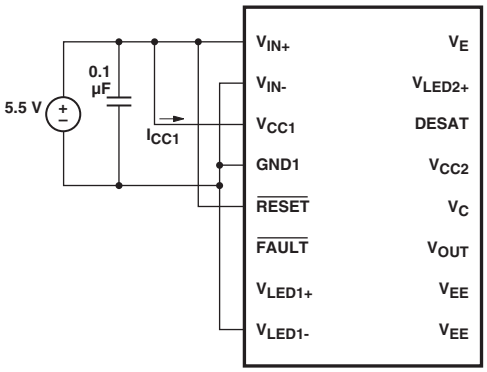


Figure 37 I_{CC1L} Test Circuit

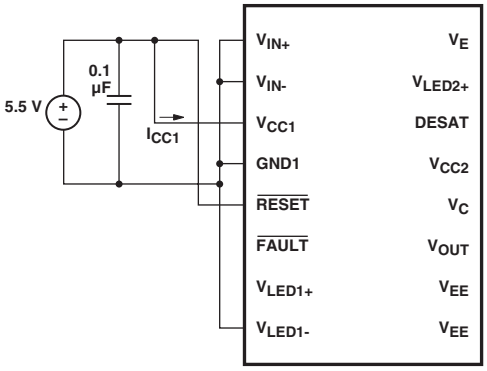


Figure 38 I_{CC2H} Test Circuit

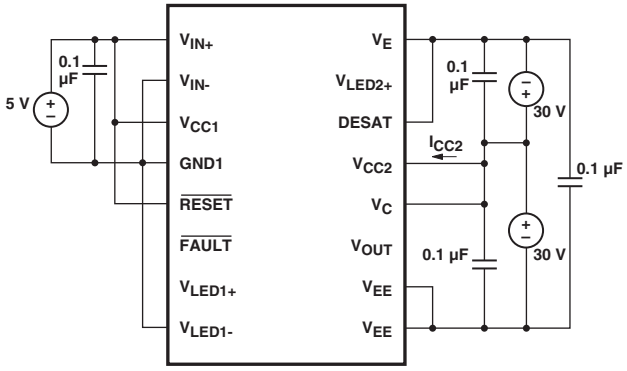


Figure 39 I_{CC2L} Test Circuit

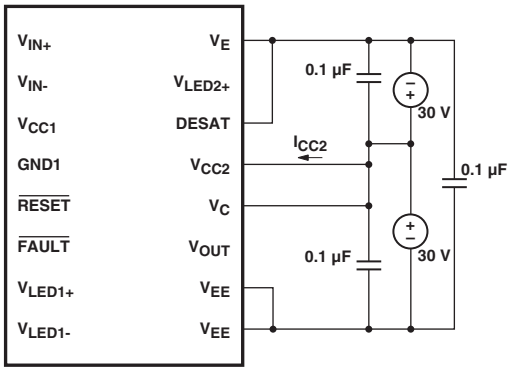


Figure 40 I_{CHG} Pulsed Test Circuit

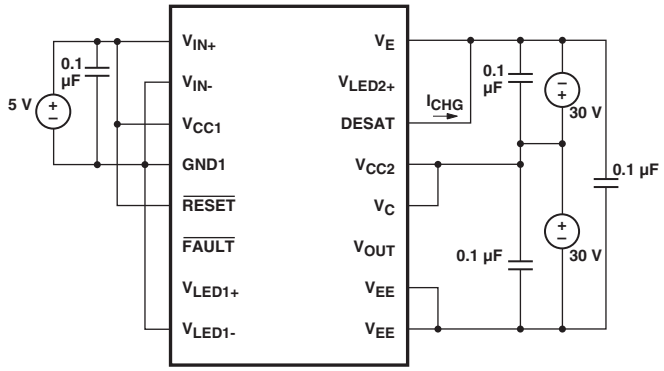


Figure 41 I_{DSCHG} Test Circuit

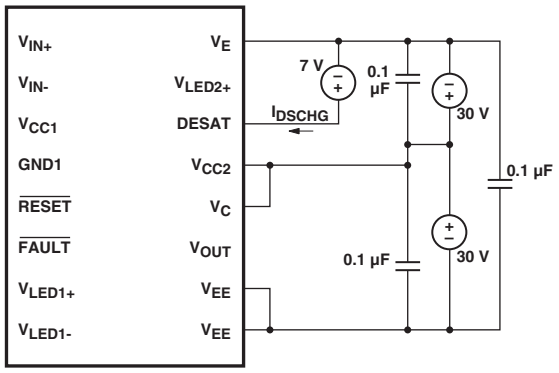


Figure 42 UVLO Threshold Test Circuit

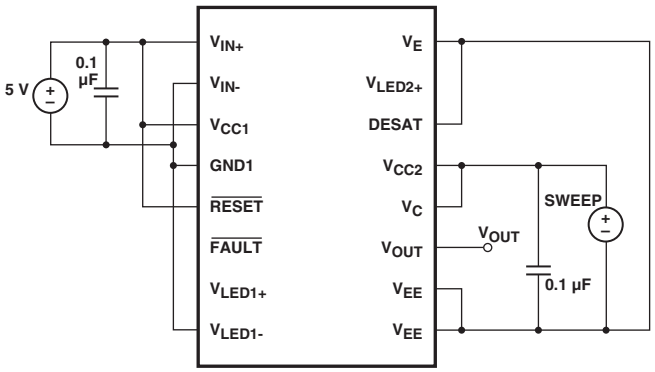


Figure 43 DESAT Threshold Test Circuit

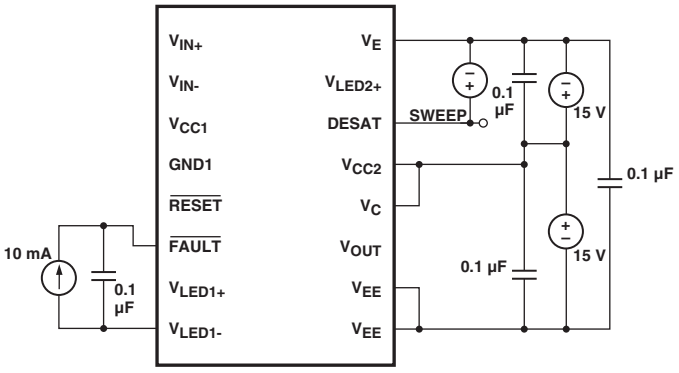


Figure 44 t_{PLH}, t_{PHL}, t_r, t_f Test Circuit

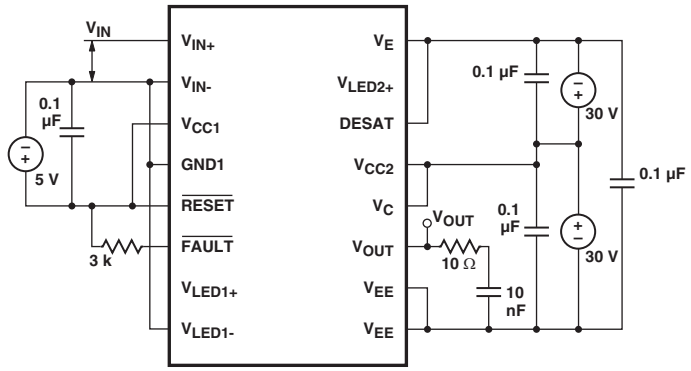


Figure 45 $t_{DESAT(10\%)}$ Test Circuit

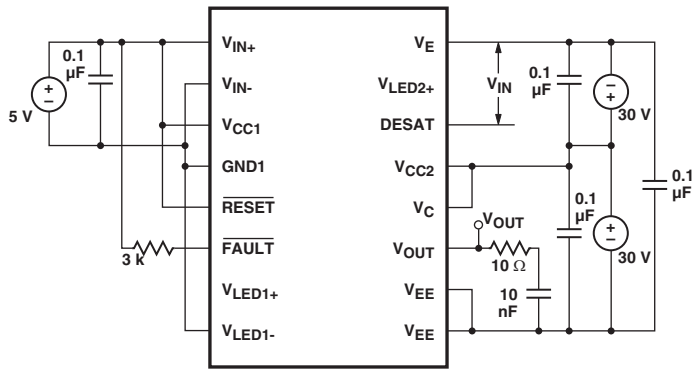


Figure 46 $t_{DESAT(\overline{FAULT})}$ Test Circuit

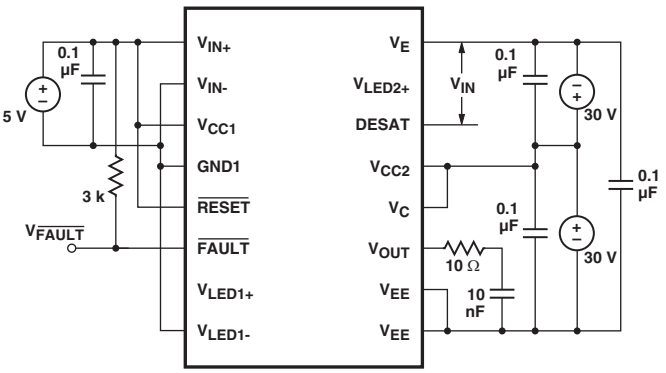


Figure 47 $t_{\overline{RESET(FAULT)}}$ Test Circuit

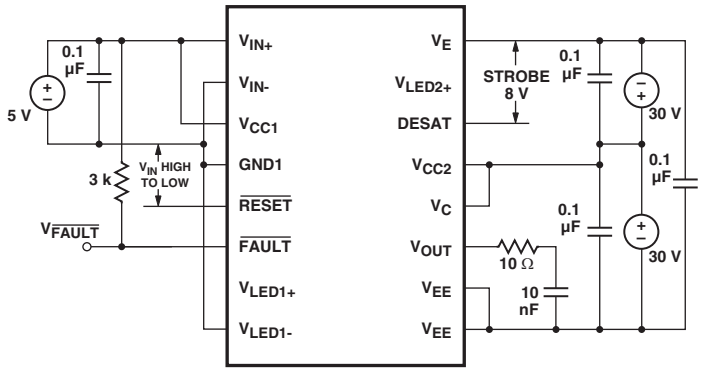


Figure 48 UVLO Delay Test Circuit

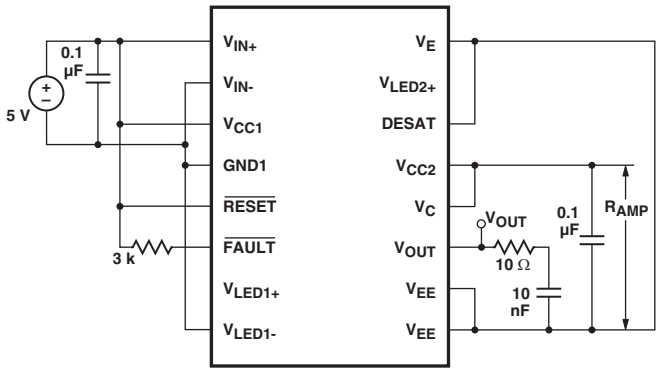


Figure 49 CMR Test Circuit, LED2 Off

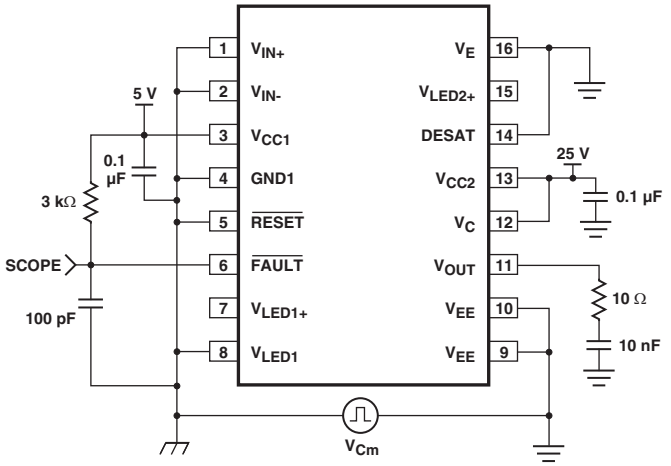


Figure 50 CMR Test Circuit, LED2 On

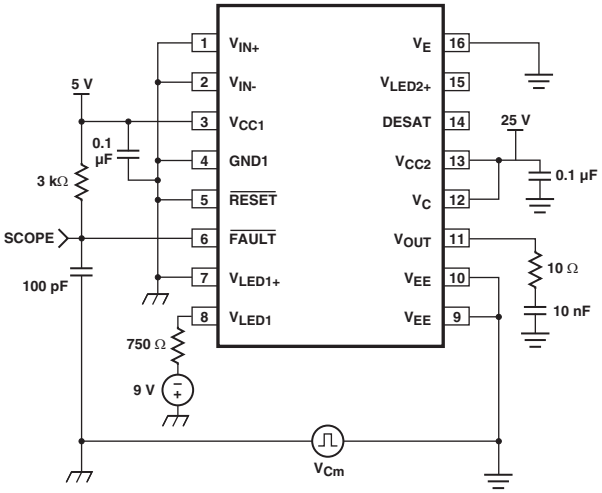


Figure 51 CMR Test Circuit, LED1 Off

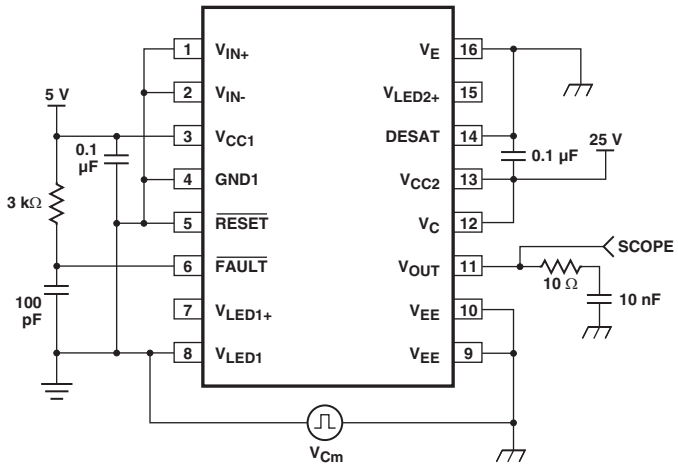


Figure 52 CMR Test Circuit, LED1 On

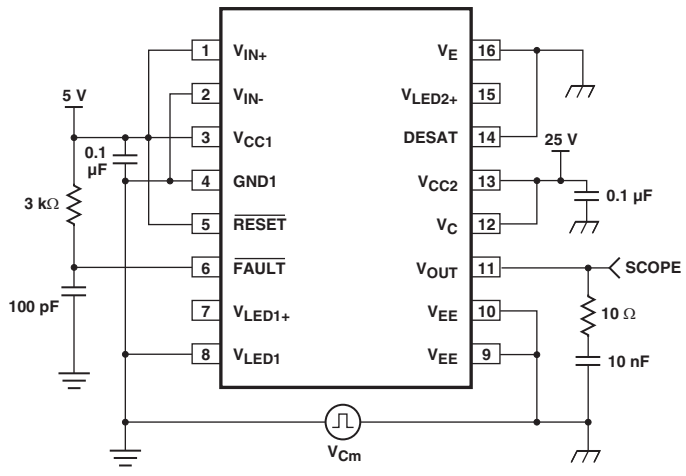


Figure 53 V_{OUT} Propagation Delay Waveforms, Noninverting Configuration

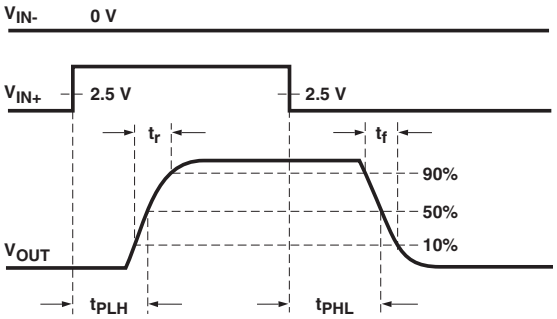


Figure 54 V_{OUT} Propagation Delay Waveforms, Inverting Configuration

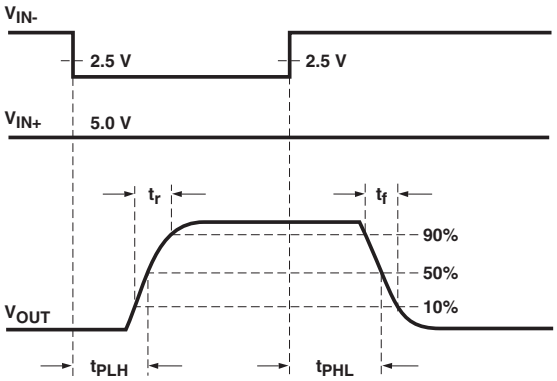


Figure 55 DESAT, $V_{OUT,}$ Fault, Reset Delay Waveforms

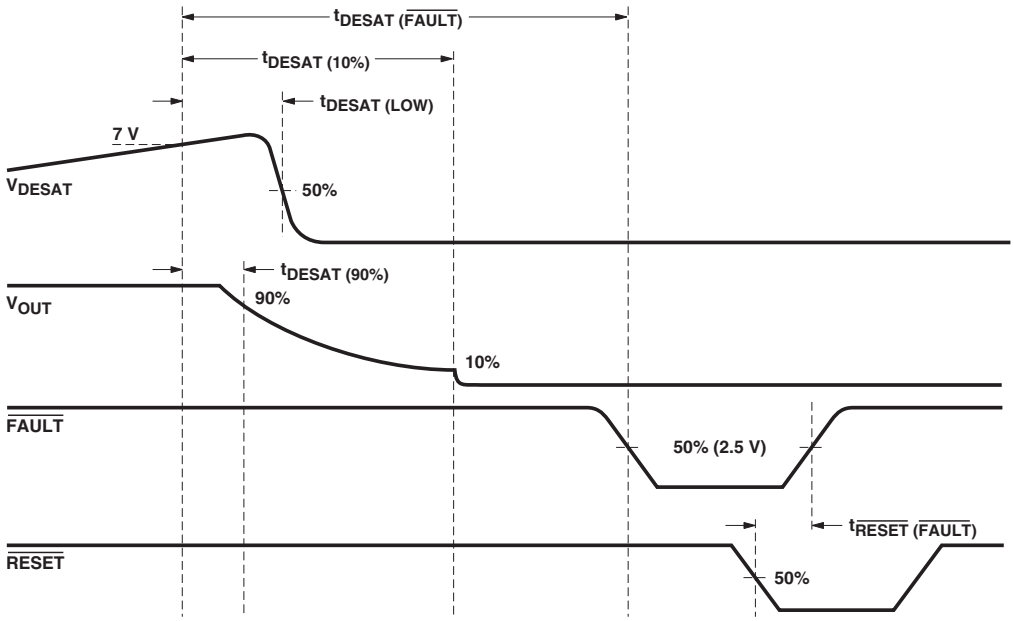


Figure 56 I_{CH} Test Circuit

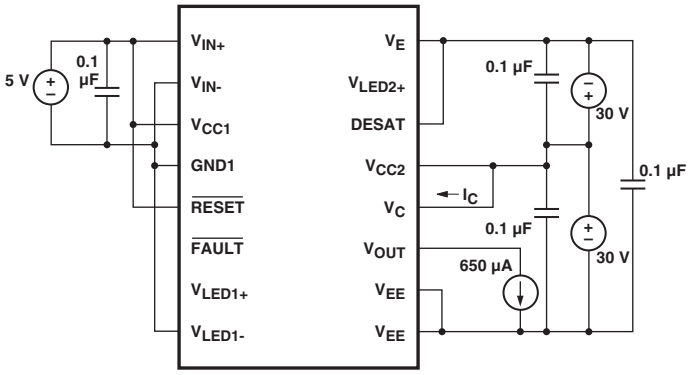


Figure 57 I_{CH} Test Circuit

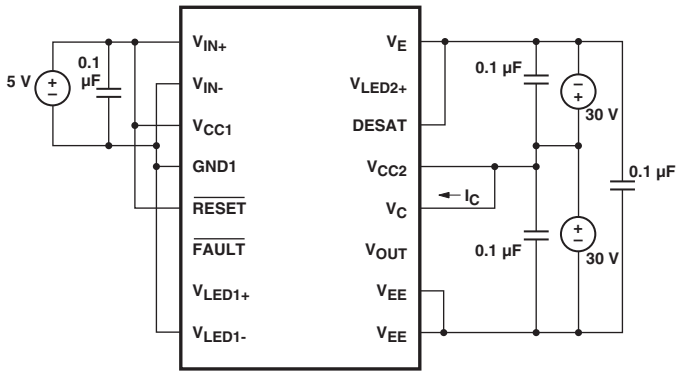


Figure 58 I_{CL} Test Circuit

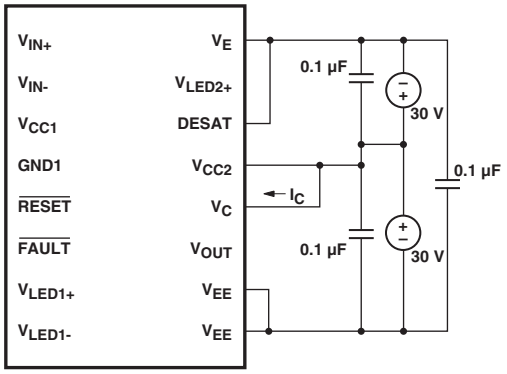


Figure 59 I_{EH} Test Circuit

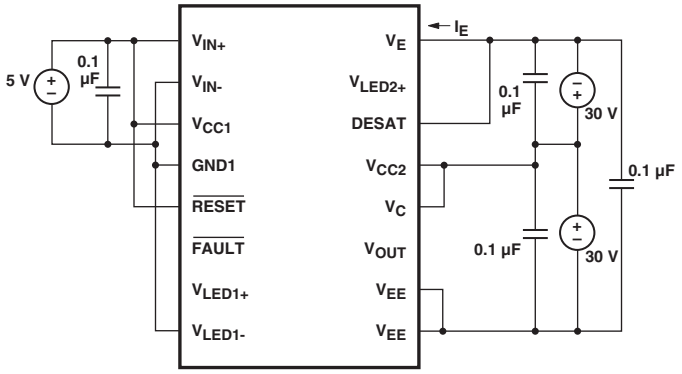
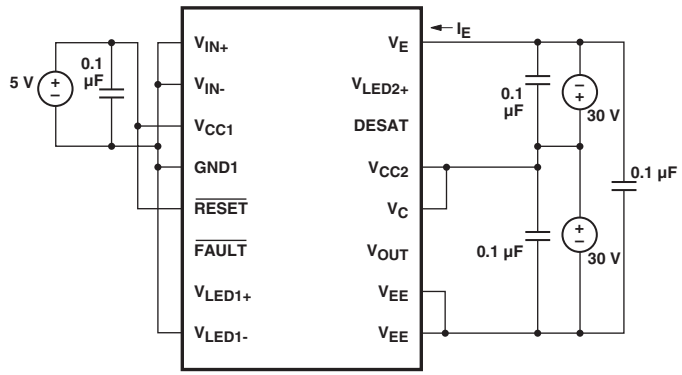


Figure 60 I_{EL} Test Circuit



Typical Application/Operation

Introduction to Fault Detection and Protection

The power stage of a typical three phase inverter is susceptible to several types of failures, most of which are potentially destructive to the power IGBTs. These failure modes can be grouped into four basic categories: phase and/or rail supply short circuits due to user misconnect or bad wiring, control signal failures due to noise or computational errors, overload conditions induced by the load, and component failures in the gate drive circuitry. Under any of these fault conditions, the current through the IGBTs can increase rapidly, causing excessive power dissipation and heating. The IGBTs become damaged when the current load approaches the saturation current of the device, and the collector to emitter voltage rises above the saturation voltage level. The drastically increased power dissipation very quickly overheats the power device and destroys it. To prevent damage to the drive, fault protection must be implemented to reduce or turn off the overcurrents during a fault condition.

A circuit providing fast local fault detection and shutdown is an ideal solution, but the number of required components, board space consumed, cost, and complexity have until now limited its use to high performance drives. The features which this circuit must have are high speed, low cost, low resolution, low power dissipation, and small size.

Applications Information

The ACPL-516x satisfies these criteria by combining a high speed, high output current driver, high voltage optical isolation between the input and output, local IGBT desaturation detection and shut down, and an optically isolated fault status feedback signal into a single 16-pin DIP package.

The fault detection method, which is adopted in the ACPL-516x, is to monitor the saturation (collector) voltage of the IGBT and to trigger a local fault shutdown sequence if the collector voltage exceeds a predetermined threshold. A small gate discharge device slowly reduces the high short circuit IGBT current to prevent damaging voltage spikes. Before the dissipated energy can reach destructive levels, the IGBT is shut off. During the off state of the IGBT, the fault detect circuitry is simply disabled to prevent false fault signals.

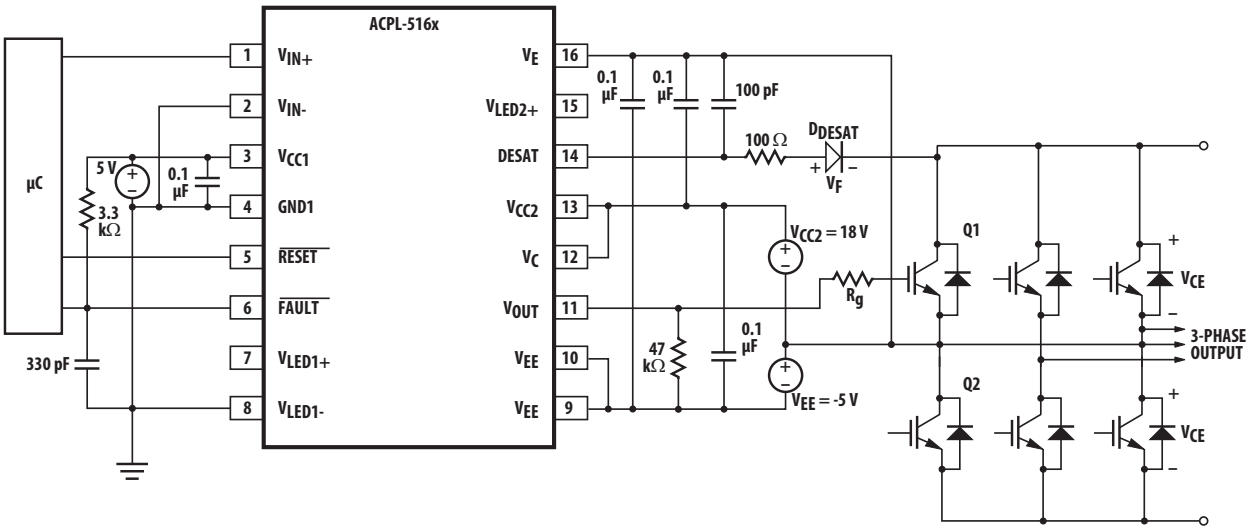
The alternative protection scheme of measuring IGBT current to prevent desaturation is effective if the short circuit capability of the power device is known, but this method will fail if the gate drive voltage decreases enough to only partially turn on the IGBT. By directly measuring the collector voltage, the ACPL-516x limits the power dissipation in the IGBT even with insufficient gate drive voltage. Another more subtle advantage of the desaturation detection method is that power dissipation in the IGBT is monitored, while the current sense method relies on a preset current threshold to predict the safe limit of operation. Therefore, an overly conservative overcurrent threshold is not needed to protect the IGBT.

Recommended Application Circuit

The ACPL-516x has both inverting and noninverting gate control inputs, an active low reset input, and an open collector fault output suitable for wired OR applications. The recommended application circuit shown in Figure 61 illustrates a typical gate drive implementation using the ACPL-516x.

The four supply bypass capacitors (0.1 µF) provide the large transient currents necessary during a switching transition. Because of the transient nature of the charging currents, a low current (5 mA) power supply suffices. The DESAT diode and 100-pF capacitor are the necessary external components for the fault detection circuitry. The gate resistor (10Ω) serves to limit gate charge current and indirectly control the IGBT collector voltage rise and fall times. The open collector fault output has a passive 3.3-kΩ pull-up resistor and a 330-pF filtering capacitor. A 47-kΩ pull-down resistor on V_{OUT} provides a more predictable high level output voltage (V_{OH}). In this application, the IGBT gate driver will shut down when a fault is detected and will not resume switching until the microcontroller applies a reset signal.

Figure 61 Recommended Application Circuit



Description of Operation/Timing

Figure 62 illustrates input and output waveforms under the conditions of normal operation, a DESAT fault condition, and normal reset behavior.

Normal Operation

During normal operation, V_{OUT} of the ACPL-516x is controlled by either V_{IN+} or V_{IN-}, with the IGBT collector-to-emitter voltage being monitored through D_{DESAT}. The $\overline{FAULT}$ output is high and the $\overline{RESET}$ input should be held high. See Figure 62.

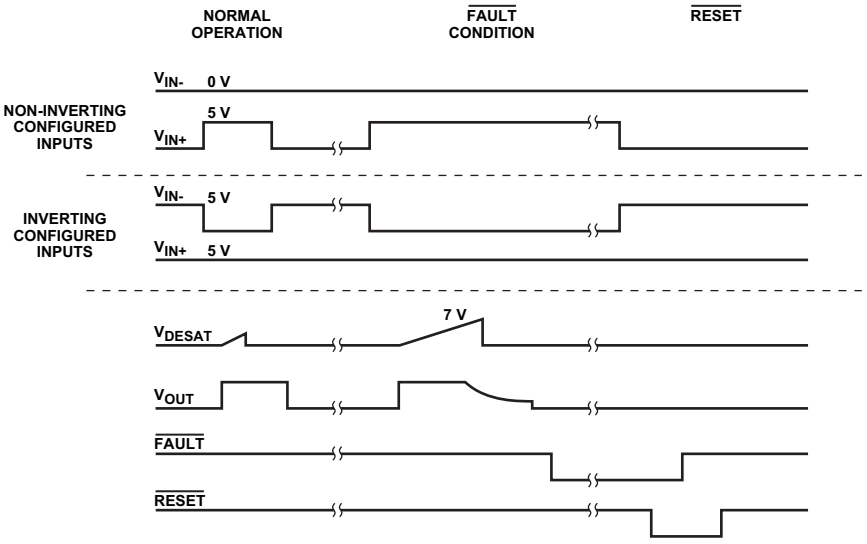
Fault Condition

When the voltage on the DESAT pin exceeds 7V while the IGBT is on, V_{OUT} is slowly brought low in order to *softly* turn off the IGBT and prevent large di/dt induced voltages. Also activated is an internal feedback channel which brings the $\overline{FAULT}$ output low for the purpose of notifying the microcontroller of the fault condition. See Figure 62.

Reset

The $\overline{FAULT}$ output remains low until $\overline{RESET}$ is brought low. See Figure 62. While asserting the $\overline{RESET}$ pin (LOW), the input pins must be asserted for an output low state (V_{IN+} is LOW or V_{IN-} is HIGH). This can be accomplished either by software control (i.e., of the microcontroller) or hardware control (see Figure 72 through Figure 75).

Figure 62 Timing Diagram



Slow IGBT Gate Discharge during Fault Condition

When a desaturation fault is detected, a weak pull-down device in the ACPL-516x output drive stage will turn on to *softly* turn off the IGBT. This device slowly discharges the IGBT gate to prevent fast changes in drain current that could cause damaging voltage spikes due to lead and wire inductance. During the slow turn off, the large output pull-down device remains off until the output voltage falls below V_{EE} + 2V, at which time the large pull down device clamps the IGBT gate to V_{EE}.

DESAT Fault Detection Blanking Time

The DESAT fault detection circuitry must remain disabled for a short time period following the turn-on of the IGBT to allow the collector voltage to fall below the DESAT threshold. This time period, called the DESAT blanking time, is controlled by the internal DESAT charge current, the DESAT voltage threshold, and the external DESAT capacitor. The nominal blanking time is calculated in terms of external capacitance (C_{BLANK}), $\overline{FAULT}$ threshold voltage (V_{DESAT}), and DESAT charge current (I_{CHG}) as $t_{BLANK} = C_{BLANK} \times V_{DESAT} / I_{CHG}$. The nominal blanking time with the recommended 100-pF capacitor is 100 pF × 7V/250 µA = 2.8 µs. The capacitance value can be scaled slightly to adjust the blanking time, though a value smaller than 100 pF is not recommended. This nominal blanking time also represents the longest time it will take for the ACPL-516x to respond to a DESAT fault condition. If the IGBT is turned on while the collector and emitter are shorted to the supply rails (switching into a short), the soft shutdown sequence begins after approximately 3 µs. If the IGBT collector and emitter are shorted to the supply rails *after the IGBT is already on*, the response time is much quicker due to the parasitic parallel capacitance of the DESAT diode. The recommended 100-pF capacitor should provide adequate blanking as well as fault response times for most applications.

Undervoltage Lockout

The ACPL-516x Undervoltage Lockout (UVLO) feature is designed to prevent the application of insufficient gate voltage to the IGBT by forcing the ACPL-516x output low during power-up. IGBTs typically require gate voltages of 15V to achieve their rated $V_{CE(ON)}$ voltage. At gate voltages below 13V typically, their on-voltage increases dramatically, especially at higher currents. At very low gate voltages (below 10V), the IGBT might operate in the linear region and quickly overheat. The UVLO function causes the output to be clamped whenever insufficient operating supply (V_{CC2}) is applied. Once V_{CC2} exceeds V_{UVLO+} (the positive-going UVLO threshold), the UVLO clamp is released to allow the device output to turn on in response to input signals. As V_{CC2} is increased from 0V (at some level below V_{UVLO+}), first the DESAT protection circuitry becomes active. As V_{CC2} is further increased (above V_{UVLO+}), the UVLO clamp is released. Before the time the UVLO clamp is released, the DESAT protection is already active. Therefore, the UVLO and DESAT FAULT DETECTION features work together to provide seamless protection regardless of supply voltage (V_{CC2}).

Behavioral Circuit Schematic

The functional behavior of the ACPL-516x is represented by the logic diagram in Figure 63, which fully describes the interaction and sequence of internal and external signals in the ACPL-516x.

Input IC

In the normal switching mode, no output fault has been detected, and the low state of the fault latch allows the input signals to control the signal LED. The fault output is in the open-collector state, and the state of the Reset pin does not affect the control of the IGBT gate. When a fault is detected, the $\overline{\text{FAULT}}$ output and signal input are both latched. The fault output changes to an active low state, and the signal LED is forced off (output LOW). The latched condition persists until the Reset pin is pulled low.

Output IC

Three internal signals control the state of the driver output: the state of the signal LED, as well as the UVLO and FAULT signals. If no fault on the IGBT collector is detected, and the supply voltage is above the UVLO threshold, the LED signal controls the driver output state. The driver stage logic includes an interlock to ensure that the pull-up and pull-down devices in the output stage are never on at the same time. If an undervoltage condition is detected, the output is actively pulled low by the 50x DMOS device, regardless of the LED state. If an IGBT desaturation fault is detected while the signal LED is on, the Fault signal will latch in the high state. The triple darlington AND the 50x DMOS device are disabled, and a smaller 1x DMOS pull-down device is activated to slowly discharge the IGBT gate. When the output drops below 2V, the 50x DMOS device again turns on, clamping the IGBT gate firmly to V_{EE}. The FAULT signal remains latched in the high state until the signal LED turns off.

Figure 63 Behavioral Circuit Schematic

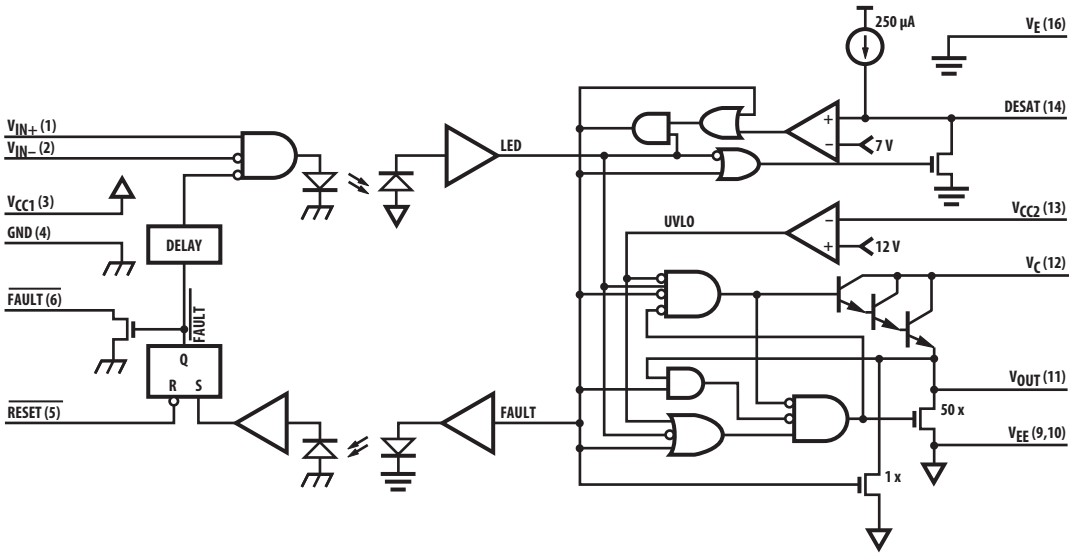


Figure 64 Output Pull-Down Resistor

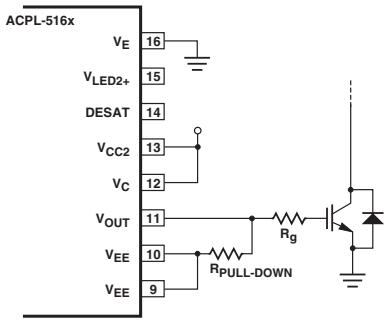


Figure 65 DESAT Pin Protection

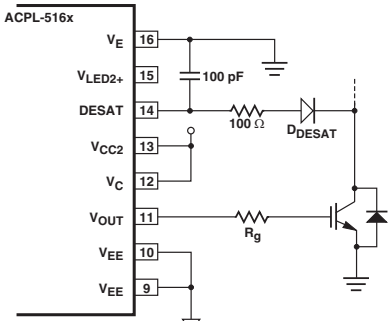
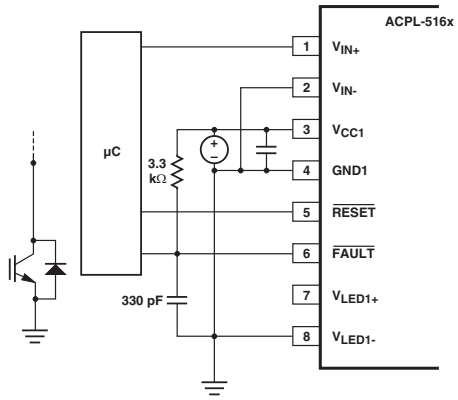


Figure 66 $\overline{\text{FAULT}}$ Pin CMR Protection



Other Recommended Components

The application circuit in Figure 61 includes an output pull-down resistor, a DESAT pin protection resistor, a $\overline{\text{FAULT}}$ pin capacitor (330 pF), and a $\overline{\text{FAULT}}$ pin pull-up resistor.

Output Pull-Down Resistor

During the output high transition, the output voltage rapidly rises to within 3 diode drops of V_{CC2}. If the output current then drops to zero due to a capacitive load, the output voltage slowly rises from roughly $V_{CC2} - 3(V_{BE})$ to V_{CC2} within a period of several microseconds. To limit the output voltage to $V_{CC2} - 3(V_{BE})$, a pull-down resistor between the output and V_{EE} is recommended to sink a static current of several 650 µA while the output is high. Pull-down resistor values are dependent on the amount of positive supply and can be adjusted according to the formula, $R_{pull\text{-}down} = [V_{CC2} - 3 \times (V_{BE})]/650\ \mu A$.

DESAT Pin Protection

The freewheeling of flyback diodes connected across the IGBTs can have large instantaneous forward voltage transients that greatly exceed the nominal forward voltage of the diode. This can result in a large negative voltage spike on the DESAT pin, which draws substantial current out of the IC if protection is not used. To limit this current to levels that will not damage the IC, a 100Ω resistor should be inserted in series with the DESAT diode. The added resistance will not alter the DESAT threshold or the DESAT blanking time.

Capacitor on FAULT Pin for High CMR

Rapid common mode transients can affect the fault pin voltage while the fault output is in the high state. A 330-pF capacitor (Figure 65) should be connected between the fault pin and ground to achieve adequate CMOS noise margins at the specified CMR value of 15 kV/µs. The added capacitance does not increase the fault output delay when a desaturation condition is detected.

Pull-up Resistor on $\overline{\text{FAULT}}$ Pin

The $\overline{\text{FAULT}}$ pin is an open-collector output and therefore requires a pull-up resistor to provide a high-level signal.

Driving with Standard CMOS/TTL for High CMR

Capacitive coupling from the isolated high voltage circuitry to the input referred circuitry is the primary CMR limitation. This coupling must be accounted for to achieve high CMR performance. The input pins V_{IN+} and V_{IN-} must have active drive signals to prevent unwanted switching of the output under extreme common mode transient conditions. Input drive circuits that use pull-up or pull-down resistors, such as open collector configurations, should be avoided. Standard CMOS or TTL drive circuits are recommended.

User-Configuration of the ACPL-516x Input Side

The V_{IN+}, V_{IN-}, $\overline{FAULT}$, and $\overline{RESET}$ input pins make a wide variety of gate control and fault configurations possible, depending on the motor drive requirements. The ACPL-516x has both inverting and non-inverting gate control inputs, an open collector fault output suitable for wired-OR applications, and an active low reset input.

Driving Input pf ACPL-516x in Noninverting/Inverting Mode

The Gate Drive Voltage Output of the ACPL-516x can be configured as inverting or noninverting using the V_{IN-} and V_{IN+} inputs. As shown in Figure 67, when a noninverting configuration is desired, V_{IN-} is held low by connecting it to GND1 and V_{IN+} is toggled. As shown in Figure 68, when an inverting configuration is desired, V_{IN+} is held high by connecting it to V_{CC1} and V_{IN-} is toggled.

Local Shutdown, Local Reset

As shown in Figure 69, the fault output of each ACPL-516x gate driver is polled separately, and the individual reset lines are asserted low independently to reset the motor controller after a fault condition.

Global-Shutdown, Global Reset

As shown in Figure 70, when configured for inverting operation, the ACPL-516x can be configured to shutdown automatically in the event of a fault condition by tying the $\overline{FAULT}$ output to V_{IN+}. For high reliability drives, the open collector $\overline{FAULT}$ outputs of each ACPL-516x can be wire-ORed together on a common fault bus, forming a single fault bus for interfacing directly to the microcontroller. When any of the six gate drivers detects a fault, the fault output signal disables all six ACPL-516x gate drivers simultaneously and thereby provides protection against further catastrophic failures.

Figure 67 Typical Input Configuration, Noninverting

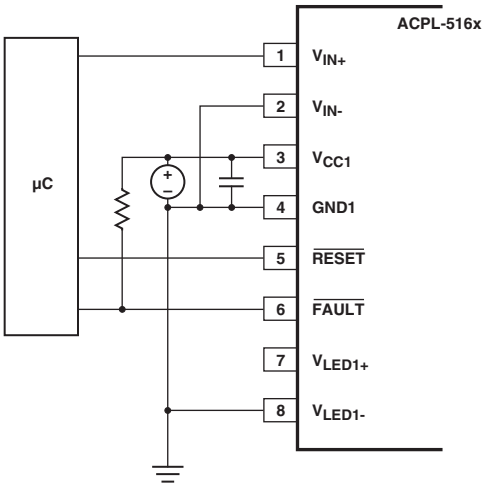


Figure 68 Typical Input Configuration, Inverting

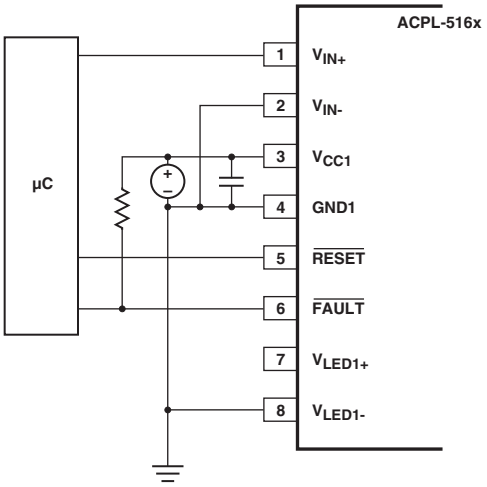


Figure 69 Local Shutdown, Local Reset Configuration

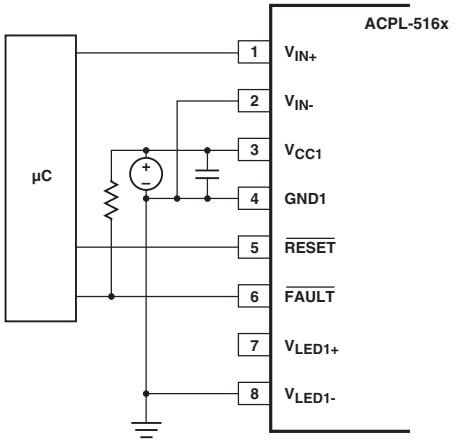
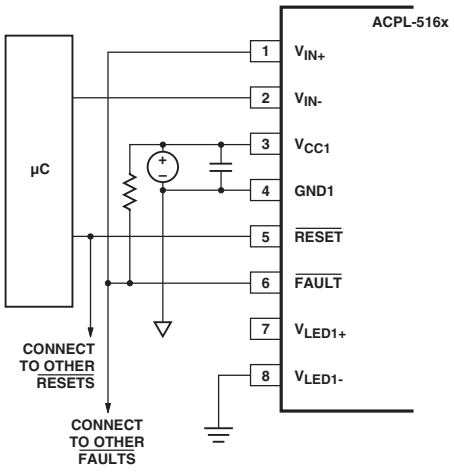


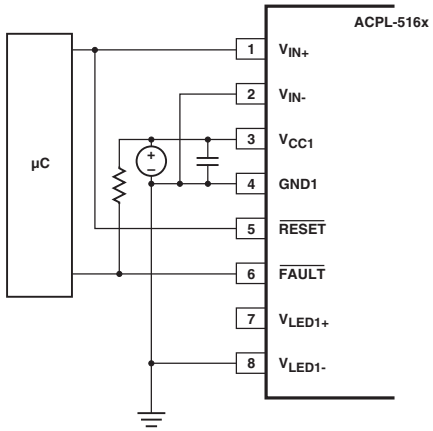
Figure 70 Global-Shutdown, Global Reset Configuration



Auto-Reset

As shown in Figure 71, when the inverting V_{IN-}. input is connected to ground (noninverting configuration), the ACPL-516x can be configured to reset automatically by connecting $\overline{RESET}$ to V_{IN+}. In this case, the gate control signal is applied to the noninverting input as well as the reset input to reset the fault latch every switching cycle. During normal operation of the IGBT, asserting the reset input low has no effect. Following a fault condition, the gate driver remains in the latched fault state until the gate control signal changes to the *gate low* state and resets the fault latch. If the gate control signal is a continuous PWM signal, the fault latch is always reset by the next time the input signal goes high. This configuration protects the IGBT on a cycle-by-cycle basis and automatically resets before the next *on* cycle. The fault outputs can be wire-ORed together to alert the microcontroller, but this signal would not be used for control purposes in this (Auto-Reset) configuration. When the ACPL-516x is configured for Auto-Reset, the guaranteed minimum $\overline{FAULT}$ signal pulse width is 3 µs.

Figure 71 Auto-Reset Configuration



Resetting Following a Fault Condition

To resume normal switching operation following a fault condition ($\overline{FAULT}$ output low), the $\overline{RESET}$ pin must first be asserted low in order to release the internal fault latch and reset the $\overline{FAULT}$ output (high). Prior to asserting the $\overline{RESET}$ pin low, the input (V_{IN}) switching signals must be configured for an output (V_{OL}) low state. This can be handled directly by the microcontroller or by hardwiring to synchronize the $\overline{RESET}$ signal with the appropriate input signal. Figure 72 shows how to connect the $\overline{RESET}$ to the V_{IN+} signal for safe automatic reset in the noninverting input configuration. Figure 73 shows how to configure the V_{IN+}/$\overline{RESET}$ signals so that a $\overline{RESET}$ signal from the microcontroller causes the input to be in the *output-off* state. Similarly, Figure 74 and Figure 75 show automatic $\overline{RESET}$ and microcontroller $\overline{RESET}$ safe configurations for the inverting input configuration.

Figure 72 Safe Hardware Reset for Noninverting Input Configuration (Automatically Resets for Every V_{IN+} Input)

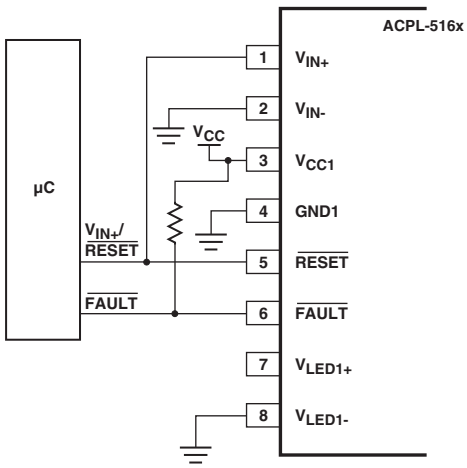


Figure 73 Safe Hardware Reset for Noninverting Input Configuration

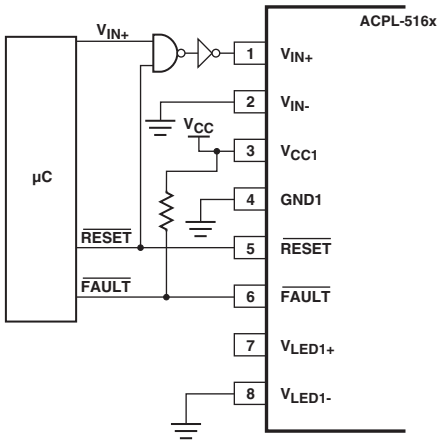


Figure 74 Safe Hardware Reset for Inverting Input Configuration

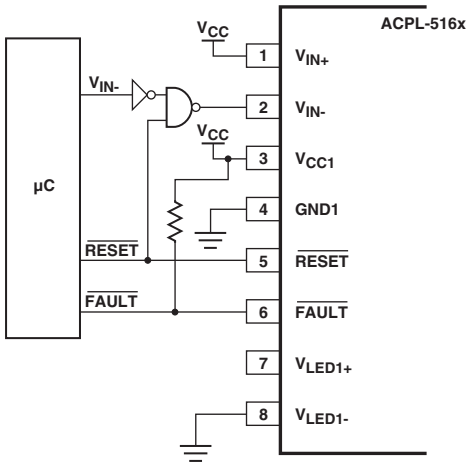
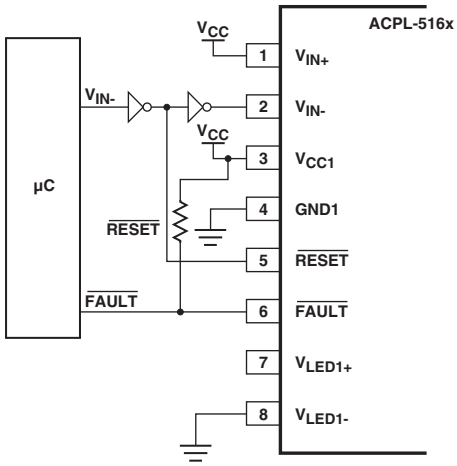


Figure 75 Safe Hardware Reset for Inverting Input Configuration (Automatically Resets for Every V_{IN-} Input)



User-Configuration of the ACPL-516x Output Side RG and Optional Resistor RC

The value of the gate resistor R_G (along with V_{CC2} and V_{EE}) determines the maximum amount of gate-charging/discharging current ($I_{ON,PEAK}$ and $I_{OFF,PEAK}$) and thus should be carefully chosen to match the size of the IGBT being driven. Often it is desirable to have the peak gate charge current be somewhat less than the peak discharge current ($I_{ON,PEAK} < I_{OFF,PEAK}$). For this condition, an optional resistor (R_C) can be used along with R_G to independently determine $I_{ON,PEAK}$ and $I_{OFF,PEAK}$ without using a steering diode. As an example, refer to Figure 76. Assuming that R_G is already determined and that the design $I_{OH,PEAK}$ = 0.5A, the value of R_C can be estimated in the following way:

$$R_C + R_G = \frac{[V_{CC2} - V_{OH} - (V_{EE})]}{I_{OH,PEAK}}$$

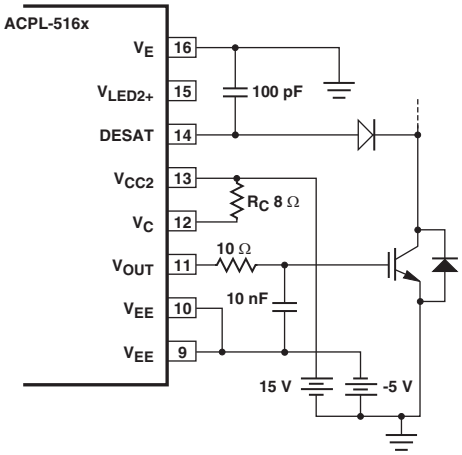
$$= \frac{[4V - (-5V)]}{0.5A}$$

$$= 18\Omega$$

$$R_C = 8\Omega$$

See Power/Layout Considerations section for more information on calculating value of RG.

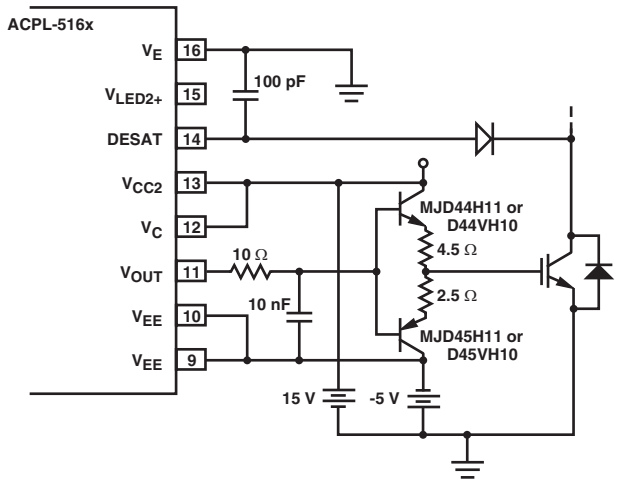
Figure 76 Use of R_C to Further Limit $I_{ON,PEAK}$



Higher Output Current Using an External Current Buffer

To increase the IGBT gate drive current, a noninverting current buffer (such as the npn/pnp buffer shown in Figure 77) can be used. Inverting types are not compatible with the desaturation fault protection circuitry and should be avoided. To preserve the slow IGBT turn-off feature during a fault condition, a 10-nF capacitor should be connected from the buffer input to V_{EE} and a 10Ω resistor inserted between the output and the common npn/pnp base. The MJD44H11/MJD45H11 pair is appropriate for currents up to 8A maximum. The D44VH10/ D45VH10 pair is appropriate for currents up to 15A maximum.

Figure 77 Current Buffer for Increased Drive Current



DESAT Diode and DESAT Threshold

The DESAT diode's function is to conduct forward current, allowing sensing of the IGBT's saturated collector-to-emitter voltage, V_{CESAT}, (when the IGBT is *on*) and to block high voltages (when the IGBT is *off*). During the short period of time when the IGBT is switching, there is commonly a very high dV_{CE}/dt voltage ramp rate across the IGBT's collector-to-emitter. This results in I_{CHARGE} (= $C_{D\text{-}DESAT}$ x dV_{CE}/dt) charging current which charges the blanking capacitor, C_{BLANK}. In order to minimize this charging current and avoid false DESAT triggering, it is best to use fast response diodes. Listed in the below table are fast-recovery diodes that are suitable for use as a DESAT diode (D_{DESAT}). In the recommended application circuit shown in Figure 61, the voltage on pin 14 (DESAT) is $V_{DESAT} = V_F + V_{CE}$, (where V_F is the forward ON voltage of D_{DESAT} and V_{CE} is the IGBT collector-to-emitter voltage). The value of V_{CE}, which triggers DESAT to signal a FAULT condition, is nominally 7V – V_F. If desired, this DESAT threshold voltage can be decreased by using multiple DESAT diodes in series. If *n* is the number of DESAT diodes, then the nominal threshold value becomes $V_{CE,FAULT(TH)} = 7V - n \times V_F$. In the case of using two diodes instead of one, diodes with half of the total required maximum reverse-voltage rating can be chosen.

Part Number	Manufacturer	t_{rr} (ns)	Max. Reverse Voltage Rating, V_{RRM} (V)	Package Type
MUR1100E	Motorola	75	1000	59-04 (axial leaded)
MURS160T3	Motorola	75	600	Case 403A (surface mount)
UF4007	General Semi.	75	1000	DO-204AL (axial leaded)
BYM26E	Philips	75	1000	SOD64 (axial leaded)
BYV26E	Philips	75	1000	SOD57 (axial leaded)
BYV99	Philips	75	600	SOD87 (surface mount)

Power/Layout Considerations

Operating Within the Maximum Allowable Power Ratings (Adjusting Value of R_G)

When choosing the value of R_G, it is important to confirm that the power dissipation of the ACPL-516x is within the maximum allowable power rating.

The steps for doing this are:

1. Calculate the minimum desired R_G.
2. Calculate total power dissipation in the part referring to Figure 79. (Average switching *energy supplied to ACPL-516x per cycle vs. R_G* plot.)
3. Compare the input and output power dissipation calculated in step 2 to the maximum recommended dissipation for the ACPL-516x. (If the maximum recommended level has been exceeded, it might be necessary to raise the value of RG to lower the switching power and repeat step 2.)

As an example, the total input and output power dissipation can be calculated given the following conditions:

- $I_{ON, MAX} \sim 2.0A$
- $V_{CC2} = 18V$
- $V_{EE} = -5V$
- $f_{CARRIER} = 15$ kHz

Step 1: Calculate RG minimum from IOL peak specification:

To find the peak charging I_{OL} assume that the gate is initially charged the steady-state value of V_{EE}. Therefore apply the following relationship:

$$R_G = \frac{[V_{OH}\text{at } 650\ \mu A - (V_{OL} + V_{EE})]}{I_{OL,PEAK}}$$

$$= \frac{[V_{CC2} - 1 - (V_{OL} + V_{EE})]}{I_{OL,PEAK}}$$

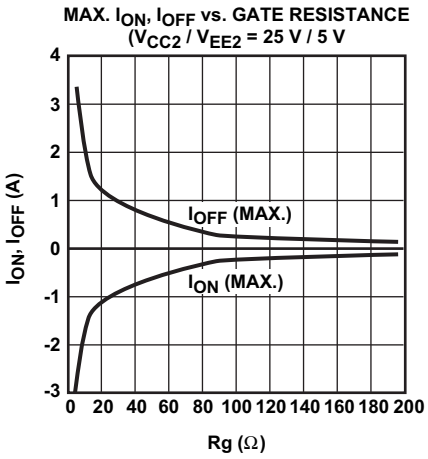
$$= \frac{18V - 1V - (1.5V + (-5V))}{2.0A}$$

$$= 10.25\Omega$$

$$\approx 10.5\Omega \text{ (for a 1\% resistor)}$$

Note from Figure 78 that the real value of I_{OL} may vary from the value calculated from the simple model shown.

Figure 78 Typical Peak I_{ON} and I_{OFF} Currents vs. R_G (for ACPL-516x Output Driving an IGBT Rated at 600V/100A)



Step 2: Calculate total power dissipation in the ACPL-516x:

The ACPL-516x total power dissipation (P_T) is equal to the sum of the input-side power (P_I) and output-side power (P_O):

$$P_T = P_I + P_O$$

$$P_I = I_{CC1} \times V_{CC1}$$

$$P_O = P_{O(BIAS)} + P_{O,SWTICH}$$

$$= I_{CC2} \times (V_{CC2} - V_{EE}) + E_{SWITCH} \times f_{SWITCH}$$

Where,

$P_{O(BIAS)}$ = steady-state power dissipation in the ACPL-516x due to biasing the device.

$P_{O(SWITCH)}$ = transient power dissipation in the ACPL-516x due to charging and discharging power device gate.

E_{SWITCH} = Average Energy dissipated in ACPL-516x due to switching of the power device over one switching cycle (µJ/cycle).

f_{SWITCH} = average carrier signal frequency.

For $R_G = 10.5$, the value read from Figure 79 is $E_{SWITCH} = 6.05$ µJ. Assume a worst-case average $I_{CC1} = 16.5$ mA (which is given by the average of I_{CC1H} and I_{CC1L}). Similarly the average $I_{CC2} = 5.5$ mA.

$$P_I = 16.5\text{ mA} \times 5.5V = 90.8\text{ mW}$$

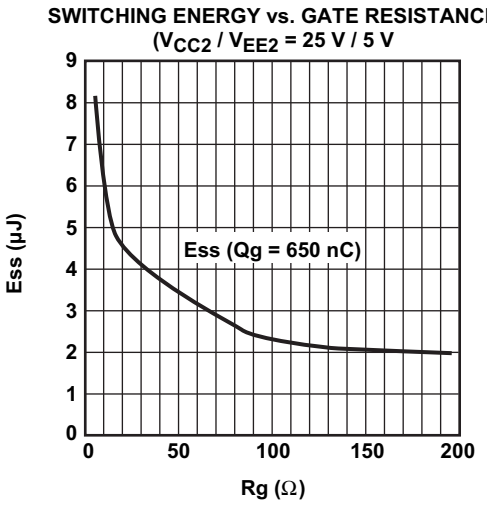
$$P_O = P_{O(BIAS)} + P_{O,SWITCH}$$

$$= 5.5\text{ mA} \times (18\text{ V} - (-5\text{ V})) + 6.051\ \mu J \times 15\text{ kHz}$$

$$= 126.5\text{ mW} + 90.8\text{ mW}$$

$$= 217.3\text{ mW}$$

Figure 79 Switching Energy Plot for Calculating Average P_{SWITCH} (for ACPL-516x Output Driving an IGBT Rated at 600V/100A)



Step 3: Compare the calculated power dissipation with the absolute maximum values for the ACPL-516x:

For the example,

P_I = 90.8 mW < 150 mW (abs. max.) **OK**

P_O= 217.3 mW < 600 mW (abs. max.) **OK**

Therefore, the power dissipation absolute maximum rating has not been exceeded for the example.

For an explanation on how to calculate the maximum junction temperature of the ACPL-516x for a given PC board layout configuration, refer to the following Thermal Model section.

Thermal Model

R_{11}, R_{12}, R_{13}, R_{14}, R_{21}, R_{22}, R_{23}, R_{24}, R_{31}, R_{32}, R_{33}, R_{34}, R_{41}, R_{42}, R_{43}, R_{44}: Thermal Resistances in °C/W.

- R_{11}: Thermal Resistance of LED1 due to heating of LED1.
- R_{12}: Thermal Resistance of LED1 due to heating of INPUT IC.
- R_{13}: Thermal Resistance of LED1 due to heating of LED2.
- R_{14}: Thermal Resistance of LED1 due to heating of OUTPUT IC.
- R_{21}: Thermal Resistance of INPUT IC due to heating of LED1.
- R_{22}: Thermal Resistance of INPUT IC due to heating of INPUT IC.
- R_{23}: Thermal Resistance of INPUT IC due to heating of LED2.
- R_{24}: Thermal Resistance of INPUT IC due to heating of OUTPUT IC.
- R_{31}: Thermal Resistance of LED2 due to heating of LED1.
- R_{32}: Thermal Resistance of LED2 due to heating of INPUT IC.
- R_{33}: Thermal Resistance of LED2 due to heating of LED2.
- R_{34}: Thermal Resistance of LED2 due to heating of OUTPUT IC.
- R_{41}: Thermal Resistance of OUTPUT IC due to heating of LED1.
- R_{42}: Thermal Resistance of OUTPUT IC due to heating of INPUT IC.
- R_{42}: Thermal Resistance of OUTPUT IC due to heating of LED2.
- R_{42}: Thermal Resistance of OUTPUT IC due to heating of OUTPUT IC.

Description

This thermal model assumes that the ACPL-516x optocoupler is mounted onto a 76.2 mm × 76.2 mm low and high conductivity printed circuit board (PCB) per JEDEC standard. The PCB boards are made of FR-4 material and the thickness of the copper traces is per JEDEC standards for low/high conductivity board.

The ACPL-516x is a hybrid device with four die: an input LED1, an input buffer IC, an output feedback LED2, and an output detector IC. The temperature at the LEDs and the ICs of the optocoupler can be calculated by using the following equations:

$\Delta T_{1A} = R_{11}P_1 + R_{12}P_2 + R_{13}P_3 + R_{14}P_4$

$\Delta T_{2A} = R_{21}P_1 + R_{22}P_2 + R_{23}P_3 + R_{24}P_4$

$\Delta T_{3A} = R_{31}P_1 + R_{32}P_2 + R_{33}P_3 + R_{34}P_4$

$\Delta T_{4A} = R_{41}P_1 + R_{42}P_2 + R_{43}P_3 + R_{44}P_4$

where:

ΔT_{1A} = Temperature difference between ambient and LED1.

ΔT_{2A} = Temperature difference between ambient and INPUT IC.

ΔT_{3A} = Temperature difference between ambient and LED2.

ΔT_{4A} = Temperature difference between ambient and OUTPUT IC.

P_1 = Power dissipation from LED1.

P_2 = Power dissipation from INPUT IC.

P_3 = Power dissipation from LED2.

P_4 = Power dissipation from OUTPUT IC.

Thermal Coefficient Data (Units in °C/W)

High Conductivity Board															
R_{11}	R_{12}	R_{13}	R_{24}	R_{21}	R_{22}	R_{23}	R_{24}	R_{31}	R_{32}	R_{33}	R_{34}	R_{41}	R_{42}	R_{43}	R_{44}
111	26	28	26	24	66	30	23	23	29	79	25	27	26	26	35

Low Conductivity Board															
R_{11}	R_{12}	R_{13}	R_{24}	R_{21}	R_{22}	R_{23}	R_{24}	R_{31}	R_{32}	R_{33}	R_{34}	R_{41}	R_{42}	R_{43}	R_{44}
125	37	41	32	41	70	47	30	36	38	93	28	41	35	40	38

Junction Temperature Calculation

Assume maximum power dissipation, Pmax(buffer) = 0.15W, Pmax(detector) = 0.6 W, P(LED) ~ 0.02W. If the ambient temperature is 125°C, the calculated junction temperature for a high conductivity board is:

$T2 = (R_{21} \times P_1 + R_{22} \times P_2 + R_{23} \times P_3 + R_{24} \times P_4) + T_a$

$= (24 \times 0.02 + 66 \times 0.15 + 30 \times 0.02 + 23 \times 0.6) + 125 \sim 150°C$

$T4 = (R_{41} \times P_1 + R_{42} \times P_2 + R_{43} \times P_3 + R_{44} \times P_4) + T_a$

$= (27 \times 0.02 + 26 \times 0.15 + 26 \times 0.02 + 35 \times 0.6) + 125 \sim 150°C$

NOTE

- The junction temperatures of the input and output IC is ~150°C when operating at 125°C.
- No power derating is required when operating below 125°C using a high conductivity board.

If low conductivity board is used, the calculated junction temperature is:

$T2 = (R_{21} \times P_1 + R_{22} \times P_2 + R_{23} \times P_3 + R_{24} \times P_4) + T_a$

$= (41 \times 0.02 + 70 \times 0.15 + 47 \times 0.02 + 30 \times 0.6) + 125 \sim 155°C$

$T4 = (R_{41} \times P_1 + R_{42} \times P2 + R_{43} \times P_3 + R_{44} \times P_4) + T_a$

$= (41 \times 0.02 + 35 \times 0.15 + 40 \times 0.02 + 38 \times 0.6) + 125 \sim 155°C$

NOTE

- The junction temperatures of the input and output IC exceeded the absolute maximum junction temperature of 150°C.
- Power derating is required so that the junction temperatures do not exceed 150°C.
- Output IC power dissipation is derated linearly at 20 mW/°C above 120°C.
- Input IC power dissipation is derated linearly at 5 mW/°C above 120°C.

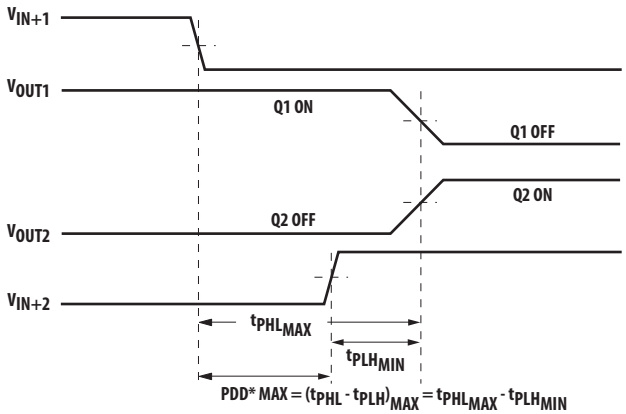
System Considerations

Propagation Delay Difference (PDD)

The ACPL-516x includes a Propagation Delay Difference (PDD) specification intended to help designers minimize *dead time* in their power inverter designs. Dead time is the time period during which both the high and low side power transistors (Q1 and Q2 in Figure 61) are off. Any overlap in Q1 and Q2 conduction results in large currents flowing through the power devices between the high and low voltage motor rails, a potentially catastrophic condition that must be prevented.

To minimize dead time in a given design, the turn-on of the ACPL-516x driving Q2 should be delayed (relative to the turn-off of the ACPL-516x driving Q1) so that under worst-case conditions, transistor Q1 has just turned off when transistor Q2 turns on, as shown in Figure 80. The amount of delay necessary to achieve this condition is equal to the maximum value of the propagation delay difference specification, PDD_{MAX}, which is specified to be 400 ns over the operating temperature range of –55°C to +125°C.

Figure 80 Minimum LED Skew for Zero Dead Time

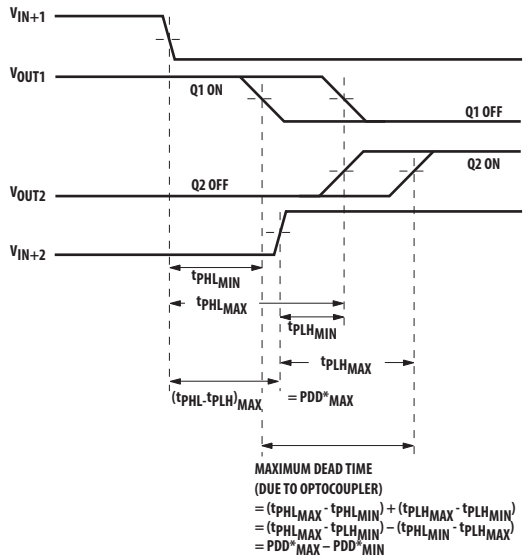


*PDD = PROPAGATION DELAY
NOTE: FOR PDD CALCULATIONS THE PROPAGATION DELAYS ARE TAKEN AT THE SAME TEMPERATURE AND TEST CONDITIONS.

Delaying the ACPL-516x turn-on signals by the maximum propagation delay difference ensures that the minimum dead time is zero, but it does not tell a designer what the maximum dead time will be. The maximum dead time is equivalent to the difference between the maximum and minimum propagation delay difference specifications as shown in Figure 81. The maximum dead time for the ACPL-516x is 800 ns (= 400 ns – (–400 ns)) over an operating temperature range of–55 °C to +125 °C.

Note that the propagation delays used to calculate PDD and dead time are taken at equal temperatures and test conditions since the optocouplers under consideration are typically mounted in close proximity to each other and are switching identical IGBTs.

Figure 81 Waveforms for Dead Time Calculation



*PDD = PROPAGATION DELAY DIFFERENCE
NOTE: FOR DEAD TIME AND PDD CALCULATIONS ALL PROPAGATION DELAYS ARE TAKEN AT THE SAME TEMPERATURE AND TEST CONDITIONS.