POWEREX

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Darlington Transistor Modules Application Information

2.0 Power Module Mounting

When mounting power transistor modules to a heatsink, care should be taken to avoid applying uneven torque to the baseplate due to one-sided tightening. It is [redacted]

Figure 2.1 Mounting Screw Fastening Pattern

2.1 Voltage Ratings

Powerex transistor modules are available in a number of different voltage ratings to accommodate the various worldwide AC supply voltages. The proper voltage rating [redacted]

The peak voltages across the transistor must be minimized. Line filters and/or voltage snubbers/clamps must be used on the DC rail to eliminate or, at least, minimize all transients. The maximum transient voltage across a transistor is usually the turn-off voltage spike shown in Figure 2.2.

This spike of voltage during turn-off can be reduced, but it cannot be totally eliminated. The spike voltage is reduced by minimizing the parasitic circuit and wiring inductance, and by using turn-off voltage snubbers. Adding voltage snubbers/clamps directly across the DC rail near the transistors will reduce the spike.

Voltage safety margin is defined as the difference between the rated transistor voltage and the peak value of the rectified AC, (or rail), voltage plus the maximum spike voltage. Voltage safety margin is dependent on the circuit and especially the snubbing and voltage clamping. The following voltage safety margins are suggested:

when:
$V_{CEO(sus)}/V_{CEV}$ =
600 volts or less, use a 50 volt safety margin.

or when:
$V_{CEO(sus)}/V_{CEV}$ =
700 to 1200 volts, use a 100 volt safety margin.

The load line during turn-on must always be within the FBSOA curve and during turn-off within the RBSOA curve. The circuit load line also determines when to use the $V_{CEO(sus)}$ rating or the V_{CEV} rating.

For example, when using a current and voltage snubber as shown in Figure 2.3, the transistor safety margin would be calculated from the maximum V_{CEV} rating. If a transistor which is rated at V_{CEV} = 1000 volts is used, then 900 volts would be the maximum peak rail voltage, including turn-off spike voltage, resulting in a 100 volt safety margin. The collector current should be less than the rated leakage current, usually less than 10 mA at the $V_{CEO(sus)}$ rating, with the voltage increasing to V_{CEV} minus 100 volts while maintaining this low leakage current level.

Although the RBSOA curve indicates that turn-off can be accomplished with higher collector currents between the $V_{CEO(sus)}$ and V_{CEV} voltages, keeping the collector current below the 10 mA level during this period will result in better reliability because the transistors are not heavily stressed in this turn-off mode. At turn-on, the inductor in the turn-on snubber of Figure 2.3 supports the rail voltage keeping the transistor within its FBSOA curve.

Figure 2.3 Snubbed Load Line, Turn-on and Turn-off to 900 Volt using a 1000 Volt Transistor

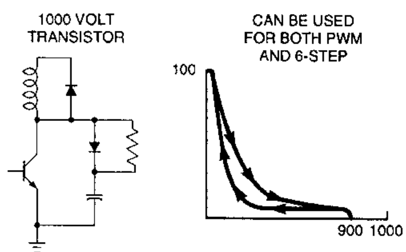
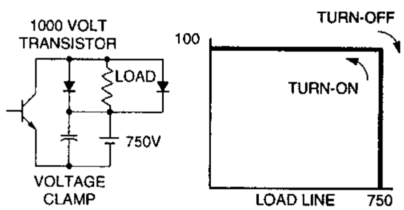


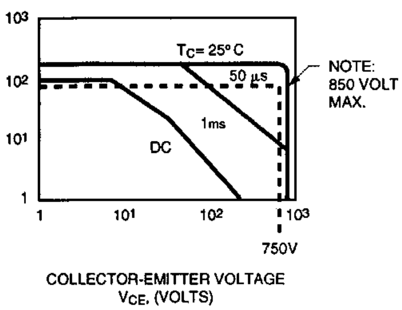
Figure 2.4 Turn-on Load Line, PWM Operation Using a 1000 Volt Transistor



As a second example, the load line for a PWM switching circuit is shown in Figure 2.4. As the transistor is pulsing on and off at a high duty cycle, there is a current flowing in a free-wheeling diode during the off period. When the transistor is on, the sudden transfer of current from the free-wheeling diode to the transistor gives the transistor a turn on load line that is very similar to a capacitive load, i.e., the current rises to its maximum before the voltage falls. Since all circuits are slightly inductive during turn-off, the turn-off load line follows the normal inductive load line, that is, the voltage rises to its maximum before the current begins to fall to zero.

Figure 2.5 continues this example with the load line drawn on the FBSOA curve. This FBSOA curve is for the KS221K10 Darlington transistor which is rated for 1000 volts $V_{CEV(sus)}$ and I_C =100 amps. Since there is no turn-on snubber in this example, to keep the load line within the FBSOA curve, voltage safety margin must be calculated based upon the $V_{CEO(sus)}$ rating and not the V_{CEV} rating. Note that although this transistor is rated for V_{CEV} = 1000 volts, the FBSOA curve is terminated at 850 volts.

Figure 2.5 Turn-on Load Line Superimposed on Forward Bias Safe Operating Area Curve



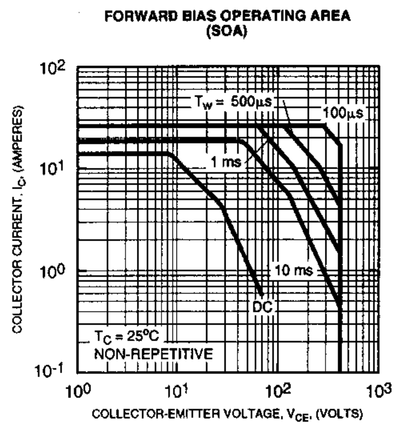
This voltage is the $V_{CEO(sus)}$ rating. The load line drawn here is 750 volts and with the addition of the 100 volt safety margin, the result is equal to the 850 volts $V_{CEO(sus)}$ rating of this transistor.

2.2 Current Ratings

Although the maximum voltage ratings should never be exceeded, exceeding the maximum collector current can be done safely for certain conditions. Figure 2.6 is the FBSOA curve for the KS621K30. For the DC curve, the maximum current is limited to 300 amps and therefore, the RMS current limit is 300 amperes. The internal transistor wires are the limiting factor.

The 600 ampere maximum current limit is set on the one millisecond curve and according to JEDEC guidelines, the transistor must be capable of operating with one millisecond pulses at 50% duty cycle. It is possible to operate this transistor with a smaller pulse, less than 50% duty cycle and not exceed the RMS bonding wire limit and also not exceed the maximum

Figure 2.6 Forward Bias Safe Operating Area Curve Provides Device Current Ratings



junction temperature. Therefore, it is safe under certain conditions to exceed the maximum collector current ratings.

2.3 Base Drive

The factors which determine the choice of forward base current, I_{B1}, are the gain, saturation voltage, switching speed and short circuit capability.The lower limit of I_{B1} is determined by the gain of the the device. To minimize conduction losses, it is desirable to increase I_{B1} to lower $V_{CE(sat)}$. The practical upper limitation on I_{B1} occurs when further increases of I_{B1} do not significantly reduce $V_{CE(sat)}$ at the maximum operating collector current, i.e. hard saturation. Increasing I_{B1} has the disadvantages that storage time is lengthened and short circuit capability is compromised.

The choice between operating in hard saturation or quasi-saturation depends upon operating frequency. At low frequencies, conduction losses dominate favoring operation in hard saturation. At higher frequencies, quasi-saturation operation provides improved switching performance. The baker clamp circuit and proportional transformer drive are two popular circuits for quasi-saturation operation.

When using the transistor in a switching mode the forward base current is generally selected by the following equation.

$$I_{B1} = (1.5 \text{ to } 2.0) \bullet (I_C/h_{FE})$$

In this equation, h_{FE} is the minimum gain specification at the desired operating $V_{CE(sat)}$.

The open circuit voltage of the forward base drive circuit must be sufficient to overcome the $V_{BE(sat)}$ which can be in the 3 or 4 volt range for triple Darlingtons.

The switching performance of power Darlingtons depends on a correctly shaped base drive circuit. Turn-on depends on the rise time and magnitude of I_{B1}. Turn-on time and turn-on switching losses can be significantly reduced by using a high rise time, high initial magnitude base current, i.e. base peaking. Base peaking also reduces the dynamic saturation effect, (discussed later in power dissipation section). Circuits which employ turn-on snubbers that limit turn-on di/dt do not benefit from base peaking. Base peaking is also undesirable in an application that could turn on into a short circuit.

For high frequency operation, turn-off time should be as short as possible to minimize circuit dead time and reduce turn off switching losses. Reduction of storage and current fall times can be obtained by increasing the magnitude of the reverse base drive current, I_{B2}.

However, too high a value of I_{B2} can cause current fall time to increase. Thus, hard turn-off, i.e. switching the base emitter directly to a stiff reverse voltage source is not recommended. Increasing I_{B2} has the disadvantage that the device RBSOA capability is reduced.

When using the transistor in a switching mode, I_{B2} is generally made equal in magnitude to I_{B1} unless limited to a lesser value by the RBSOA curve. A reverse base to emitter voltage of 5 volts is recommended.

Figure 2.7 illustrates the typical base drive current waveform recommendation. As noted above, this waveform is not optimum for all situations.

Figure 2.8 provides an example of a basic drive circuit for a Darlington transistor. A complete base drive circuit will often include extra circuits such as desaturation monitoring, overcurrent shutdown, bias supply undervoltage, and minimum/maximum on/off time circuits.

One aspect of base drive design often overlooked by beginners is provision for fail safe operation. A well designed base drive circuit will always default to an off-bias condition in the absence of a control signal and it will not allow an on condition until bias power supplies are in a stable condition with sufficient voltage to supply adequate base current to the power transistor. Note in Figure 2.8 that a forward base current cannot be supplied until a control current is supplied to the optocoupler illustrating the principle of fail safe design.

Figure 2.7 Typical Recommended Base Drive Current Waveform

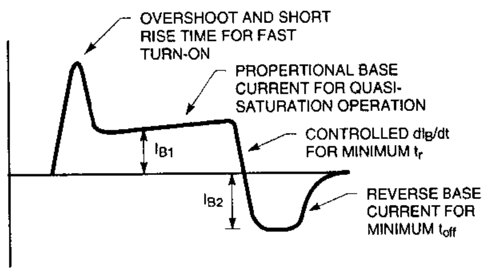
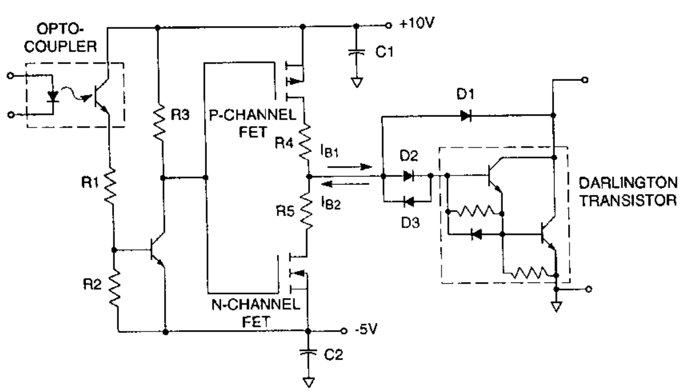
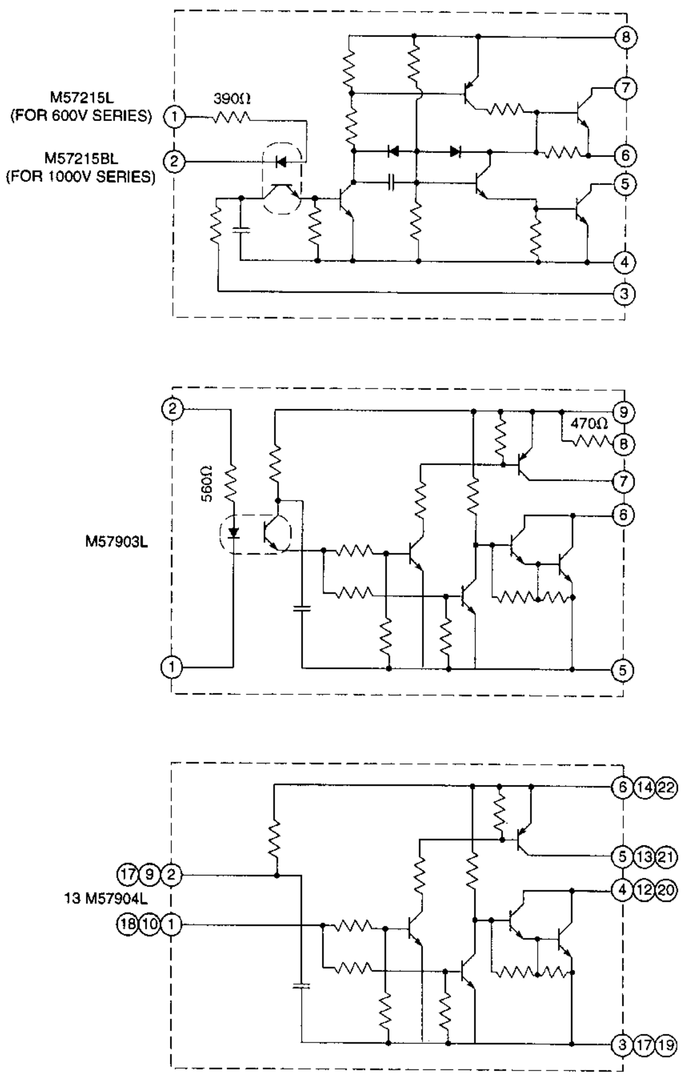


Figure 2.8 Typical Base Drive Circuit for Power Darlingtons



D1 & D2- BAKER CLAMP DIODES
D3- REQUIRED TO SHUNT I_{B2} AROUND D2
R4- SETS VALUE OF I_{B1}
R5- SETS VALUE OF I_{B2}
C1 & C2- LOCAL DECOUPLING OF BASE DRIVE SUPPLIES
OPTO COUPLER- ISOLATES DRIVE CIRCUIT FROM CONTROL ELECTRONICS

Figure 2.9 Schematic Representations of Powerex Hybrid Base Drive ICs



2.4 Hybrid Base Drive IC

Powerex offers a series of hybrid base drive integrated circuits (ICs) which can directly interface TTL level logic signals to the base of a power Darlington module. These hybrid ICs can speed design in time of power Darlingtons while providing a base drive solution with proven reliability in a compact, economical package with low power consumption. Figure 2.9 illustrates schematically three different types of hybrid base drive ICs available.

The M57215L is a dual power supply type with an internal optocoupler for isolation of control circuits from the base drive circuit. The M57215BL is identical to the M57214L, except it has a higher isolation voltage rating suited for application with 1000 volt rated Darlingtons. The M57903L is a single power supply type with an internal optocoupler. The M57903L does include a pull down transistor which when used in conjunction with an external capacitor can supply a transient reverse base current at turn-off. The M57904L is similar to the M57903L except it does not include an internal optocoupler. The M57904L contains three independent identical circuits and is suitable for operating transistor modules in a three phase inverter. Figure 2.10 illustrates application of the M57903L and M57904L and Figure 2.11 illustrates application of the M57215L and M57950L.

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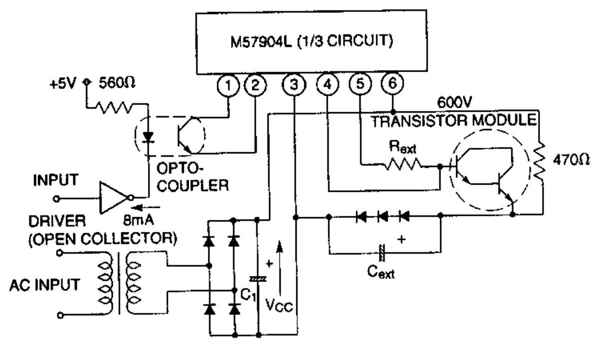
Figure 2.10 Application of the M57903L and M57904L Hybrid Base Drive ICs

	Module Rating			
	20A	30A	50A	Unit
V_{CC}	10	10	10	Volts
V_{IN}	4 ~ 5	4 ~ 5	4 ~ 5	Volts
R_{ext}	20	15	8.7	Ω
C_{ext}	22	33	47	μF
C_1	2200	3300	4700	μF
f	2	2	2	kHz



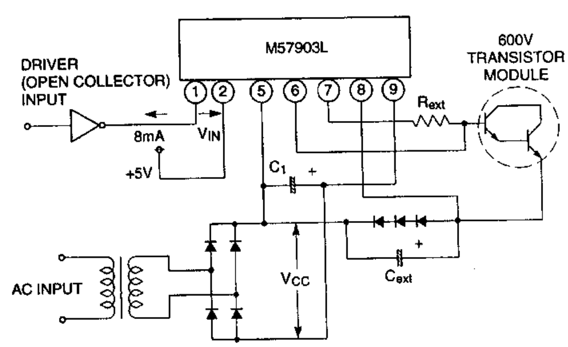
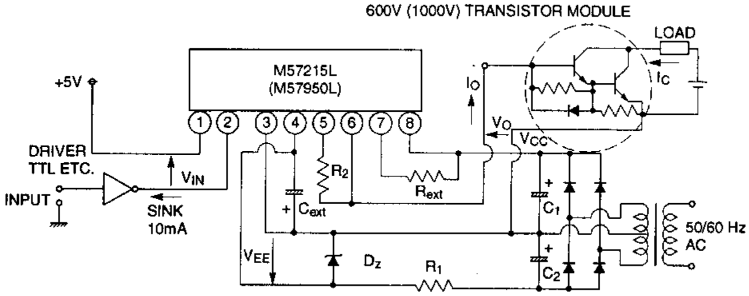


Figure 2.11 Application of the M57215L and M57215BL Hybrid Base Drive ICs

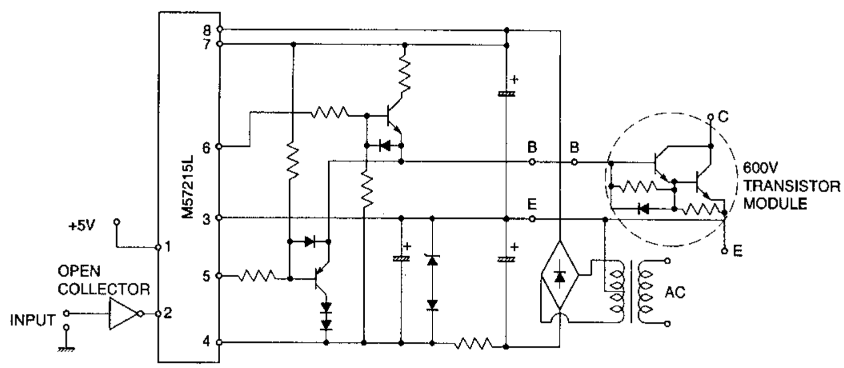
(A) Circuit configuration for 20-75 Aamp 600 and 1000 Volt Darlingtons

Note: This circuit and circuit value can be used for 1000V transistor modules with the M57215BL.



	Module Rating				
	20A	30A	50A	75A	Unit
V_{CC}	10	10	10	10	Volts
V_{EE}	-3	-3	-3	-3	Volts
V_{IN}	4~5	4~5	4~5	4~5	Volts
R_{ext}	27	20	12	9	Ω
R_2	3.3	2.2	1	1	Ω
f	2	2	2	2	kHz
R1	150	150	150	150	Ω
DZ	IN4372A	IN4372A	IN4372A	IN4372A	IN4372A
C_{ext}	10	22	22	47	μF
C_1	2200	3300	4700	4700	μF
C_2	470	470	470	470	μF

(B) Circuit configuration for 100-300 Amp 600 Volt Darlingtons



(C) Circuit configuration for 100-300 Amp 1000 Volt Darlingtons