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Thyristor
Device Data

1991

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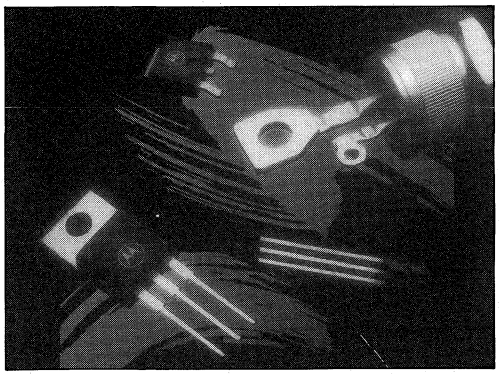

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THYRISTOR DATA

This edition of the Thyristor Data Manual has been revised extensively to reflect our current product portfolio and to incorporate new products and corrections to existing data sheets. An expanded index is intended to help the reader find information about a variety of subject material in the sections on Theory and Applications.

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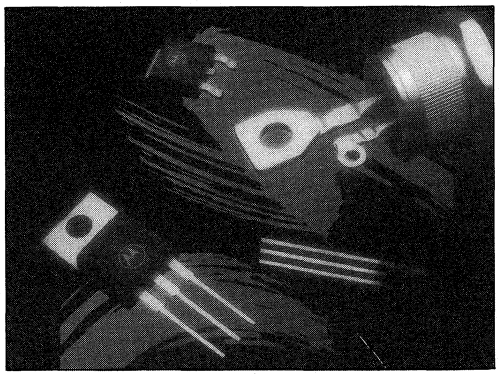

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INTRODUCTION

Thyristors can take many forms, but they have certain things in common. All of them are solid state switches which act as open circuits capable of withstanding the rated voltage until triggered. When they are triggered, thyristors become low-impedance current paths and remain in that condition until the current either stops or drops below a minimum value called the holding level. Once a thyristor has been triggered, the trigger current can be removed without turning off the device.

Silicon controlled rectifiers (SCRs) and triacs are both members of the thyristor family. SCRs are unidirectional devices where triacs are bidirectional. An SCR is designed to switch load current in one direction, while a triac is designed to conduct load current in either direction.

Structurally, all thyristors consist of several alternating layers of opposite P and N silicon, with the exact structure varying with the particular kind of device. The load is applied across the multiple junctions and the trigger current is injected at one of them. The trigger current allows the load current to flow through the device, setting up a regenerative action which keeps the current flowing even after the trigger is removed.

These characteristics make thyristors extremely useful in control applications. Compared to a mechanical switch, a thyristor has a very long service life and very fast turn on and turn off times. Because of their fast reaction times, regenerative action and low resistance once triggered, thyristors can be used as ac power controllers as well as simply turning devices on and off. Thyristors are used to control motors, incandescent lights and many other kinds of equipment.

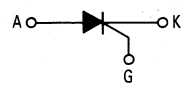
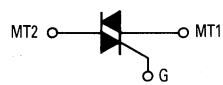
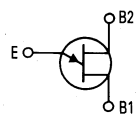
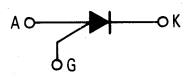
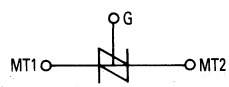
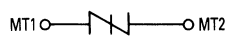
Although thyristors of all sorts are generally fairly rugged, there are several points to keep in mind when designing circuits using them. One of the most important is to respect the devices' rated limits on rate of change of voltage and current (dv/dt and di/dt). If these are exceeded, the thyristor may be damaged or destroyed. On the other hand, it is important to provide a trigger pulse large enough and fast enough to turn the gate on quickly and completely. Usually the gate trigger current should be at least three times the minimum turn-on current with a pulse rise time of less than 1 microsecond and a pulse width greater than 10 microseconds. Thyristors may be driven in many different ways, including directly from transistors or logic families, by optoisolated triac drivers, unijunction transistors (UJTs), programmable unijunction transistors (PUTs) and silicon bilateral switches (SBSs). These and other design considerations are covered in this manual.

CHAPTER 1
SYMBOLS AND TERMINOLOGY

1

SYMBOLS

The following are the most commonly used schematic symbols for thyristors:

name of device	symbol
Silicon controlled rectifier (SCR)	
Triac	
Unijunction transistor (UJT)	
Programmable unijunction transistor (PUT)	
Silicon bilateral switch (SBS)	
Sidac	

THYRISTOR TERMINOLOGY

The following terms are used in SCR and TRIAC specifications:

RATINGS

These ratings are defined as maximum values. Exceeding these values can result in permanent damage or device failure.

Terminology	Symbol	Definition
	di/dt	The maximum rate of change of current the device will withstand.
FORWARD CURRENT RMS	$I_{T(RMS)}$	The maximum value of on-state rms current the device may conduct.
FORWARD PEAK GATE CURRENT	I_{GM}, I_{GFM}	The maximum gate current which may be applied to the device to cause conduction.
PEAK FORWARD SURGE CURRENT	I_{TSM}	The maximum allowable non-repetitive surge current the device will withstand at a specified pulse width.
AVERAGE ON-STATE CURRENT	$I_{T(AV)}$	The maximum average on-state current the device may conduct under stated conditions.
PEAK GATE POWER	P_{GM}	The maximum instantaneous value of gate power dissipation between gate and cathode terminal.
FORWARD AVERAGE GATE POWER	$P_{G(AV)}$	The maximum allowable value of gate power, averaged over a full cycle, that may be dissipated between the gate and cathode terminal.
CIRCUIT FUSING CONSIDERATIONS	I^2t	The maximum forward non-repetitive overcurrent capability. Usually specified for one-half cycle of 60 Hz operation.
PEAK GATE VOLTAGE	V_{GM}	The maximum peak value of voltage allowed between the gate and cathode terminal for any bias condition.
PEAK GATE VOLTAGE FORWARD	V_{FGM}, V_{GFM}	The maximum peak value of voltage allowed between the gate and cathode terminals with these terminals forward biased.
PEAK GATE VOLTAGE REVERSE	V_{RGM}, V_{GRM}	The maximum peak value of voltage allowed between the gate and cathode with these terminals reverse biased.
PEAK REPETITIVE FORWARD BLOCKING VOLTAGE (SCR)	V_{DRM}	The maximum allowed value of repetitive forward voltage which may be applied and not switch the SCR on.
PEAK REPETITIVE REVERSE BLOCKING VOLTAGE (SCR)	V_{RRM}	The maximum allowed value of repetitive reverse voltage which may be applied to the anode terminal.
PEAK REPETITIVE OFF-STATE VOLTAGE (TRIAC)	V_{DRM}	The maximum allowed value of repetitive off-state voltage which may be applied and not switch on the triac.

CHARACTERISTICS

Terminology	Symbol	Definition
PEAK FORWARD BLOCKING CURRENT (SCR)	I_{DRM}	The maximum value of current which will flow at V_{DRM} and specified temperature.
PEAK REVERSE BLOCKING CURRENT (SCR)	I_{RRM}	The maximum value of current which will flow at V_{RRM} and specified temperature.
PEAK BLOCKING CURRENT (TRIAC)	I_{DRM}	The maximum value of current which will flow for either polarity of V_{DRM} and at specified temperature.
PEAK ON-STATE VOLTAGE	V_{TM}	The maximum voltage drop across the terminals at stated conditions.
GATE TRIGGER CURRENT	I_{GT}	The maximum value of gate current required to switch the device from the off state to the on state under specified conditions.
GATE TRIGGER VOLTAGE	V_{GT}	The gate dc voltage required to produce the gate trigger current.
HOLDING CURRENT	I_H	The value of forward anode current which allows the device to remain in conduction. Below this value the device will return to a forward blocking state at prescribed gate conditions.
CRITICAL RISE OF OFF-STATE VOLTAGE	dv/dt	The minimum value of the rate of rise of forward voltage which will cause switching from the off state to the on state.
TURN-ON TIME (SCR)	t_{gt}	The time interval between a specified point at the beginning of the gate pulse and the instant when the device voltage (current) has dropped to a specified low value during the switching of an SCR from the off state to the on state by a gate pulse.
TURN-OFF TIME (SCR)	t_q	The time interval between the instant when the SCR current has decreased to zero after external switching of the SCR voltage circuit and the instant when the thyristor is capable of supporting a specified wave form without turning on.
OPERATING JUNCTION TEMPERATURE	T_J	The junction temperature of the device as a result of ambient and load conditions.
STORAGE TEMPERATURE	T_{stg}	The temperature at which the device may be stored without harm.
CASE TEMPERATURE	T_C	The temperature of the device case under specified conditions.
AMBIENT TEMPERATURE	T_A	The air temperature measured below a device in an environment of substantially uniform temperature, cooled only by natural air currents and not materially affected by radiant and reflective surfaces.

CHARACTERISTICS

Terminology	Symbol	Definition
THERMAL RESISTANCE, CASE-TO-AMBIENT	$R_{\theta CA}$	The thermal resistance (steady-state) from the device case to the ambient.
THERMAL RESISTANCE, JUNCTION-TO-AMBIENT	$R_{\theta JA}$	The thermal resistance (steady-state) from the semiconductor junction(s) to the ambient.
THERMAL RESISTANCE, JUNCTION-TO-CASE	$R_{\theta JC}$	The thermal resistance (steady-state) from the semiconductor junction(s) to a stated location on the case.
THERMAL RESISTANCE, JUNCTION-TO-MOUNTING SURFACE	$R_{\theta JM}$	The thermal resistance (steady-state) from the semiconductor junction(s) to a stated location on the mounting surface.
TRANSIENT THERMAL IMPEDANCE, JUNCTION-TO-AMBIENT	$Z_{\theta JA(t)}$	The transient thermal impedance from the semiconductor junction(s) to the ambient.
TRANSIENT THERMAL IMPEDANCE, JUNCTION-TO-CASE	$Z_{\theta JC(t)}$	The transient thermal impedance from the semiconductor junction(s) to a stated location on the case.

CHAPTER 2
THEORY OF THYRISTOR OPERATION

To successfully apply thyristors, an understanding of their characteristics, ratings, and limitations is imperative. In this chapter, significant thyristor characteristics, the basis of their ratings, and their relationship to circuit design are discussed.

Several different kinds of thyristors are shown in Table 2.1. Silicon Controlled Rectifiers (SCRs) are the most widely used as power control elements; triacs are quite popular in lower current (under 40 A) ac power applications. Diacs, SUSs and SBSs are most commonly used as gate trigger devices for the power control elements.

Table 2.I. Thyristor Types

*JEDEC Titles	Popular Names, Types
Reverse Blocking Diode Thyristor	†Four Layer Diode, Silicon Unilateral Switch (SUS)
Reverse Blocking Triode Thyristor	Silicon Controlled Rectifier (SCR)
Reverse Conducting Diode Thyristor	†Reverse Conducting Four Layer Diode
Reverse Conducting Triode Thyristor	Reverse Conducting SCR
Bidirectional Triode Thyristor	Triac

*JEDEC is an acronym for the Joint Electron Device Engineering Councils, an industry standardization activity co-sponsored by the Electronic Industries Association (EIA) and the National Electrical Manufacturers Association (NEMA).

†Not generally available.

Before considering thyristor characteristics in detail, a brief review of their operation based upon the common two-transistor analogy of an SCR is in order.

BASIC BEHAVIOR

The bistable action of thyristors is readily explained by analysis of the structure of an SCR. This analysis is essentially the same for any operating quadrant of triac because a triac may be considered as two parallel SCRs oriented in opposite directions. Figure 2.1(a) shows the schematic symbol for an SCR, and Figure 2.1(b) shows the P-N-P-N structure the symbol represents. In the two-transistor model for the SCR shown in Figure 2.1(c), the interconnections of the two transistors are such that regenerative action occurs. Observe that if current is injected into any leg of the model, the gain of the transistors (if sufficiently high) causes this current to be amplified in another leg. In order for regeneration to occur, it is necessary for the sum of the common base current gains (α) of the two transistors to exceed unity. Therefore, because the junction leakage currents are relatively small and current gain is designed to be low at the leakage current level, the PNPN device remains off unless external current is applied. When sufficient trigger current is applied (to the gate, for example, in the case of an SCR) to raise the loop gain to unity, regeneration occurs and the on-state principal current is limited primarily by external circuit impedance. If the initiating trigger current is removed, the thyristor remains in the on state, providing the current level is high enough to meet the unity gain criteria. This critical current is called latching current.

In order to turn off a thyristor, some change in current must occur to reduce the loop gain below unity. From the model, it appears that shorting the gate to cathode would accomplish this. However in an actual SCR structure, the gate area is only a fraction of the cathode area and very little current is diverted by the short. In practice, the principal current must be reduced below a certain level, called holding current, before gain falls below unity and turn-off may commence.

In fabricating practical SCRs and Triacs, a "shorted emitter" design is generally used in which, schematically, a resistor is added from gate to cathode or gate to MT1. Because current is diverted from the N-base through the resistor, the gate trigger current, latching current and holding current all increase. One of the principal reasons for the shunt resistance is to improve dynamic performance at high temperatures. Without the shunt, leakage current on most high current thyristors could initiate turn-on at high temperatures.

Sensitive gate thyristors employ a high resistance shunt or none at all; consequently, their characteristics can be altered dramatically by use of an external resistance. An external resistance has a minor effect on most shorted emitter designs.

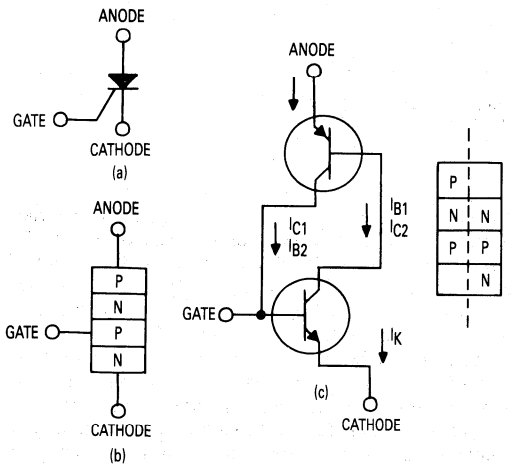


Figure 2.1. Two-transistor analogy of an SCR: (a) schematic symbol of SCR; (b) P-N-P-N structure represented by schematic symbol; (c) two-transistor model of SCR.

Junction temperature is the primary variable affecting thyristor characteristics. Increased temperatures make the thyristor easier to turn on and keep on. Consequently, circuit conditions which determine turn-on must be designed to operate at the lowest anticipated junction temperatures, while circuit conditions which are to turn off the thyristor or prevent false triggering must be designed to operate at the maximum junction temperature.

Thyristor specifications are usually written with case temperatures specified and with electrical conditions such that the power dissipation is low enough that the junction temperature essentially equals the case temperature. It is incumbent upon the user to properly account for changes in characteristics caused by the circuit operating conditions different from the test conditions.

TRIGGERING CHARACTERISTICS

Turn-on of a thyristor requires injection of current to raise the loop gain to unity. The current can take the form of current applied to the gate, an anode current resulting from leakage, or avalanche breakdown of a blocking junction. As a result, the breakover voltage of a thyristor can be varied or controlled by injection of a current at the gate terminal. Figure 2.2 shows the interaction of gate current and voltage for an SCR.

When the gate current I_g is zero, the applied voltage must reach the breakover voltage of the SCR before switching occurs. As the value of gate current is increased, however, the ability of a thyristor to support applied voltage is reduced and there is a certain value of gate current at which the behavior of the thyristor closely resembles that of a rectifier. Because thyristor turn-on, as a result of exceeding the breakover voltage, can produce high instantaneous power dissipation nonuniformly distributed over the die area during the switching transition, extreme temperatures resulting in die failure may occur unless the magnitude and rate of rise of principal current (di/dt) is restricted to tolerable levels. For normal operation, therefore, SCRs and triacs are operated at applied voltages lower than the breakover voltage, and are made to switch to the on state by gate signals high enough to assure complete turn-on

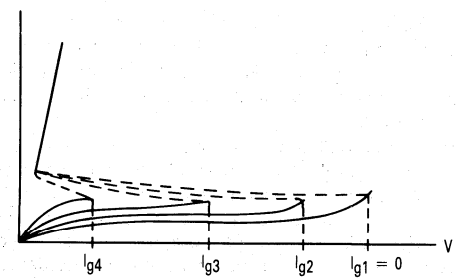


Figure 2.2. Thyristor Characteristics Illustrating Breakover as a Function of Gate Current

independent of the applied voltage. On the other hand, diacs and other thyristor trigger devices are designed to be triggered by anode breakover. Nevertheless they also have di/dt and peak current limits which must be adhered to.

A triac works the same general way for both positive and negative voltage. However since a triac can be switched on by either polarity of the gate signal regardless of the voltage polarity across the main terminals, the situation is somewhat more complex than for an SCR.

The various combinations of gate and main terminal polarities are shown in Figure 2.3. The relative sensitivity depends on the physical structure of a particular triac,

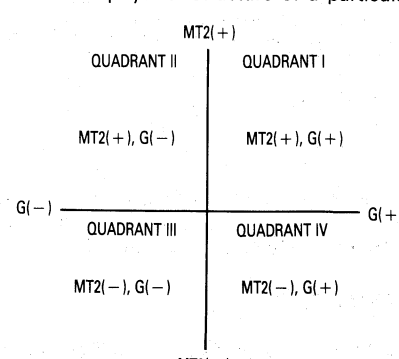


Figure 2.3. Quadrant Definitions for a Triac

but as a rule, sensitivity is highest in quadrant I and quadrant IV is generally considerably less sensitive than the others.

Gate sensitivity of a triac as a function of temperature is shown in Figure 2.4.

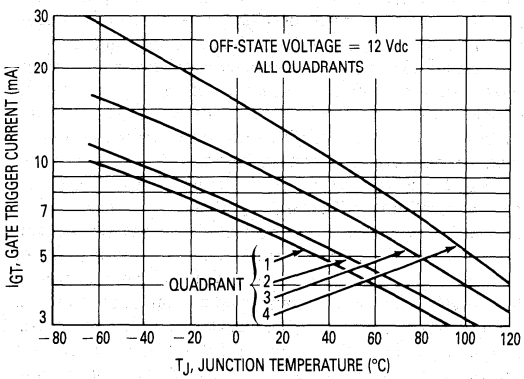


Figure 2.4. Typical Triac Triggering Sensitivity in the Four Trigger Quadrants

Since both the junction leakage currents and the current gain of the "transistor" elements increase with temperature, the magnitude of the required gate trigger current decreases as temperature increases. The gate — which can be regarded as a diode — exhibits a decreasing voltage drop as temperature increases. Thus it is important that the gate trigger circuit be designed to deliver sufficient current to the gate at the lowest anticipated temperature.

It is also advisable to observe the maximum gate current, as well as peak and average power dissipation ratings. Also in the negative direction, the maximum gate ratings should be observed. Both positive and negative gate limits are often given on the data sheets and they may indicate that protective devices such as voltage clamps and current limiters may be required in some applications. It is generally inadvisable to dissipate power in the reverse direction.

Although the criteria for turn-on have been described in terms of current, it is more basic to consider the thyristor as being charge controlled. Accordingly, as the duration of the trigger pulse is reduced, its amplitude must be correspondingly increased. Figure 2.5 shows typical behavior at various pulse widths and temperatures.

The gate pulse width required to trigger a thyristor also depends upon the time required for the anode current to reach the latching value. It may be necessary to maintain a gate signal throughout the conduction period in applications where the load is highly inductive or where the anode current may swing below the holding value within the conduction period.

When triggering an SCR with a dc current, excess leakage in the reverse direction normally occurs if the trigger signal is maintained during the reverse blocking phase of the anode voltage. This happens because the SCR operates like a remote base transistor having a gain which is generally about 0.5. When high gate drive currents are used, substantial dissipation could occur in the SCR or a significant current could flow in the load; therefore, some means usually must be provided to remove the gate signal during the reverse blocking phase.

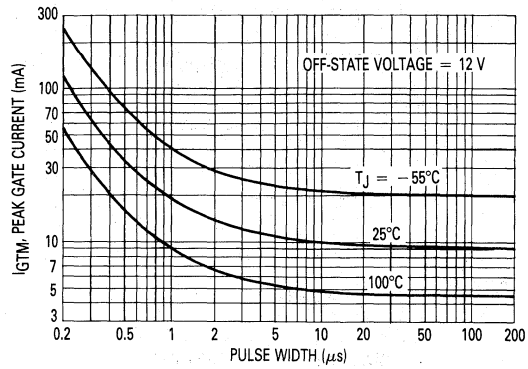


Figure 2.5. Typical Behavior of Gate Trigger Current as Pulse Width and Temperature Are Varied

LATCH AND HOLD CHARACTERISTICS

In order for the thyristor to remain in the on state when the trigger signal is removed, it is necessary to have sufficient principal current flowing to raise the loop gain to unity. The principal current level required is the latching current, I_L. Although triacs show some dependency on the gate current in quadrant II, the latching current is primarily affected by the temperature on shorted emitter structures.

In order to allow turn off, the principal current must be reduced below the level of the latching current. The current level where turn off occurs is called the holding current, I_H. Like the latching current, the holding current is affected by temperature and also depends on the gate impedance.

Reverse voltage on the gate of an SCR markedly increases the latch and hold levels. Forward bias on thyristor gates may significantly lower from the values shown in the data sheets since those values are normally given with the gate open. Failure to take this into account can cause latch or hold problems when thyristors are being driven from transistors whose saturation voltages are a few tenths of a volt.

Thyristors made with shorted emitter gates are obviously not as sensitive to the gate circuit conditions as devices which have no built-in shunt.

SWITCHING CHARACTERISTICS

When triacs or SCRs are triggered by a gate signal, the turn-on time consists of two stages: a delay time, t_d, and a rise time, t_r, as shown in Figure 2.6. The total gate controlled turn-on time, t_{gt}, is usually defined as the time interval between the 50 percent point of the leading edge

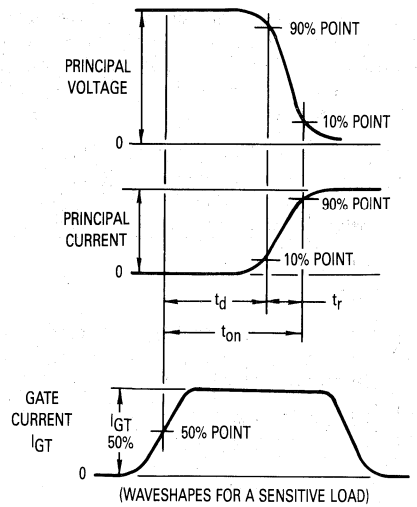


Figure 2.6. Waveshapes Illustrating Thyristor Turn-On Time For A Resistive Load

of the gate trigger voltage and 90 percent point of the principal current. The rise time t_r is the time interval required for the principal current to rise from 10 to 90 percent of its maximum value. A resistive load is usually specified.

Delay time decreases slightly as the peak off-state voltage increases. It is primarily related to the magnitude of the gate-trigger current and shows a relationship which is roughly inversely proportional.

The rise time is influenced primarily by the off-state voltage, as high voltage causes an increase in regenerative gain. Of major importance in the rise time interval is the relationship between principal voltage and current flow through the thyristor di/dt. During this time the dynamic voltage drop is high and the current density due to the possible rapid rate of change can produce localized hot spots in the die. This may permanently degrade the blocking characteristics. Therefore, it is important that power dissipation during turn-on be restricted to safe levels.

Turn-off time is a property associated only with SCRs and other unidirectional devices. (In triacs or bidirectional devices a reverse voltage cannot be used to provide circuit-commutated turn-off voltage because a reverse voltage applied to one half of the structure would be a forward-bias voltage to the other half.) For turn-off times in SCRs, the recovery period consists of two stages, a reverse recovery time and a gate or forward blocking recovery time, as shown in Figure 2.7.

When the forward current of an SCR is reduced to zero at the end of a conduction period, application of reverse voltage between the anode and cathode terminals causes reverse current flow in the SCR. The current persists until the time that the reverse current decreases to the leakage level. Reverse recovery time (t_{rr}) is usually measured from the point where the principal current changes polarity to a specified point on the reverse current waveform as indicated in Figure 2.7. During this period the anode and cathode junctions are being swept free of charge so that they may support reverse voltage. A second recovery period, called the gate recovery time, t_{gr}, must elapse for the charge stored in the forward-blocking junction to recombine so that forward-blocking voltage can be reapplied and successfully blocked by the SCR. The gate recovery time of an SCR is usually much longer than the reverse recovery time. The total time from the instant reverse recovery current begins to flow to the start of the forward-blocking voltage is referred to as circuit-commutated turn-off time t_q.

Turn-off time depends upon a number of circuit conditions including on-state current prior to turn-off, rate of change of current during the forward-to-reverse transition, reverse-blocking voltage, rate of change of reapplied forward voltage, the gate bias, and junction temperature. Increasing junction temperature and on-state current both increase turn-off time and have a more significant effect than any of the other factors. Negative gate bias will decrease the turn-off time.

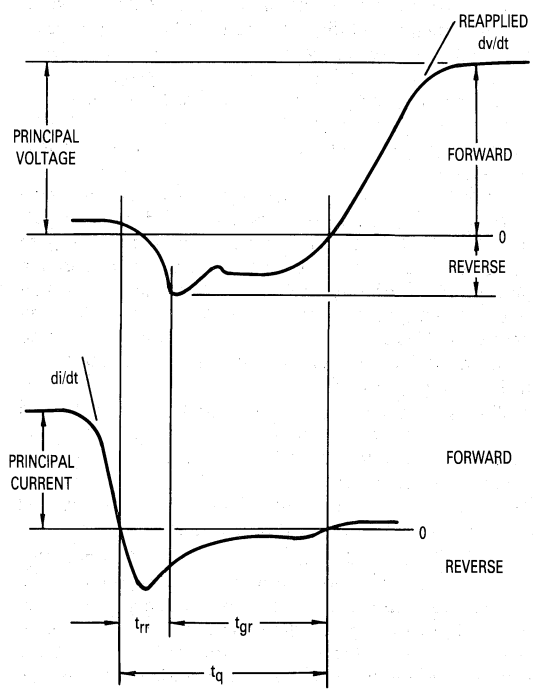


Figure 2.7. Waveshapes Illustrating Thyristor Turn-Off Time

For applications in which an SCR is used to control ac power, during the entire negative half of the sine wave a reverse voltage is applied. Turn off is easily accomplished for most devices at frequencies up to a few kilohertz. For applications in which the SCR is used to control the output of a full-wave rectifier bridge, however, there is no reverse voltage available for turn-off, and complete turn-off can be accomplished only if the bridge output is reduced close to zero such that the principal current is reduced to a value lower than the device holding current for a sufficiently long time. Turn-off problems may occur even at a frequency of 60 Hz particularly if an inductive load is being controlled.

In triacs, rapid application of a reverse polarity voltage does not cause turn-off because the main blocking junctions are common to both halves of the device. When the first triac structure (SCR-1) is in the conducting state, a quantity of charge accumulates in the N-type region as a result of the principal current flow. As the principal current crosses the zero reference point, a reverse current is established as a result of the charge remaining in the N-type region, which is common to both halves of the device. Consequently, the reverse recovery current becomes a forward current to the second half of the triac. The current resulting from stored charge causes the second half of the triac to go into the conducting state in the absence of a gate signal. Once current conduction has been established by application of a gate signal, therefore, complete loss in power control can occur as a result of interaction within the N-type base region of the triac unless sufficient time elapses or the rate of application of the reverse polarity voltage is slow enough to allow nearly all the charge to recombine in the common N-type region. Therefore, triacs are generally limited to low-frequency −60 Hz applications. Turn-off or commutation of triacs is more severe with inductive loads than with resistive loads because of the phase lag between voltage and current associated with inductive loads. Figure 2.8 shows the waveforms for an inductive load with lagging current power factor. At the time the current reaches zero crossover (Point A), the half of the triac in conduction begins to commutate when the principal current falls below the holding current. At the instant the conducting half of the triac turns off, an applied voltage opposite the current polarity is applied across the triac terminals (Point B). Because this voltage is a forward bias to the second half of the triac, the suddenly reapplied voltage in conjunction with the remaining stored charge in the high-voltage junction reduces the over-all device capability to support voltage. The result is a loss of power control to the load, and the device remains in the conducting state in absence of a gate signal. The measure of triac turn-off ability is the rate of rise of the opposite polarity voltage it can handle without remaining on. It is called commutating dv/dt (dv/dt[c]). Circuit conditions and temperature affect dv/dt(c) in a manner similar to the way t_q is affected in an SCR.

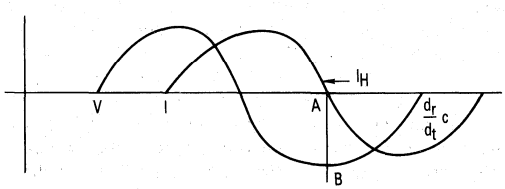


Figure 2.8. Inductive Load Waveforms

It is imperative that some means be provided to restrict the rate of rise of reapplied voltage to a value which will permit triac turn-off under the conditions of inductive load. A commonly accepted method for keeping the commutating dv/dt within tolerable levels is to use an RC snubber network in parallel with the main terminals of the triac. Because the rate of rise of applied voltage at the triac terminal is a function of the load impedance and the RC snubber network, the circuit can be evaluated under worst-case conditions of operating case temperature and maximum principal current. The values of resistance and capacitance in the snubber are then adjusted so that the rate of rise of commutating dv/dt stress is within the specified minimum limit under any of the conditions mentioned above. The value of snubber resistance should be high enough to limit the snubber capacitance discharge currents during turn-on and dampen the LC oscillation during commutation. The combination of snubber values having highest resistance and lowest capacitance that provides satisfactory operation is generally preferred.

FALSE TRIGGERING

Circuit conditions can cause thyristors to turn on in the absence of the trigger signal. False triggering may result from:

1) A high rate of rise of anode voltage, (the dv/dt effect).
2) Transient voltages causing anode breakover.
3) Spurious gate signals.

Static dv/dt effect: When a source voltage is suddenly applied to a thyristor which is in the off state, it may switch from the off state to the conducting state. If the thyristor is controlling alternating voltage, false turn-on resulting from a transient imposed voltage is limited to no more than one-half cycle of the applied voltage because turn-off occurs during the zero current crossing. However, if the principal voltage is dc voltage, the transient may cause switching to the on state and turn-off could then be achieved only by a circuit interruption.

The switching from the off state caused by a rapid rate of rise of anode voltage is the result of the internal capacitance of the thyristor. A voltage wavefront impressed across the terminals of a thyristor causes a capacitance-charging current to flow through the device which is a function of the rate of rise of applied off-state voltage ($i = C\,dv/dt$). If the rate of rise of voltage exceeds a critical value, the capacitance charging current exceeds the gate triggering current and causes device turn-on. Operation

at elevated junction temperatures reduces the thyristor ability to support a steep rising voltage dv/dt because of increased sensitivity.

dv/dt ability can be improved quite markedly in sensitive gate devices and to some extent in shorted emitter designs by a resistance from gate to cathode (or MT1) however reverse bias voltage is even more effective in an SCR. More commonly, a snubber network is used to keep the dv/dt within the limits of the thyristor when the gate is open.

TRANSIENT VOLTAGES: — Voltage transients which occur in electrical systems as a result of disturbance on the ac line caused by various sources such as energizing transformers, load switching, solenoid closure, contractors and the like may generate voltages which are above the ratings of thyristors. Thyristors, in general, switch from the off state to the on state whenever the breakover voltage of the device is exceeded, and energy is then transferred to the load. However, unless a thyristor is specified for use in a breakover mode, care should be exercised to ensure that breakover does not occur, as some devices may incur surface damage with a resultant degradation of blocking characteristics. It is good practice when thyristors are exposed to a heavy transient environment to provide some form of transient suppression.

For applications in which low-energy, long-duration transients may be encountered, it is advisable to use thyristors that have voltage ratings greater than the highest voltage transient expected in the system. The use of voltage clipping cells (MOV or Zener) is also an effective method to hold transient below thyristor ratings. The use of an RC "snubber" circuit is effective in reducing the effects of the high-energy short-duration transients more frequently encountered. The snubber is commonly required to prevent the static dv/dt limits from being exceeded, and often may be satisfactory in limiting the amplitude of the voltage transients as well.

For all applications, the dv/dt limits may not be exceeded. This is the minimum value of the rate of rise off-state voltage applied immediately to the MT1-MT2 terminals after the principal current of the opposing polarity has decreased to zero.

SPURIOUS GATE SIGNALS: In noisy electrical environments, it is possible for enough energy to cause gate triggering to be coupled into the gate wiring by stray capacitance or electromagnetic induction. It is therefore advisable to keep the gate lead short and have the common return directly to the cathode or MT1. In extreme cases, shielded wire may be required. Another aid commonly used is to connect a capacitance on the order of 0.01 to 0.1μF across the gate and cathode terminals. This has the added advantage of increasing the thyristor dv/dt capability, since it forms a capacitive divider with the anode to gate capacitance. The gate capacitor also reduces the rate of application of gate trigger current which may cause di/dt failures if a high inrush load is present.

THYRISTOR RATINGS

To insure long life and proper operation, it is important that operating conditions be restrained from exceeding thyristor ratings. The most important and fundamental ratings are temperature and voltage which are interrelated to some extent. The voltage ratings are applicable only up to the maximum temperature ratings of a particular part number. The temperature rating may be chosen by the manufacturer to insure satisfactory voltage ratings, switching speeds, or dv/dt ability.

OPERATING CURRENT RATINGS

Current ratings are not independently established as a rule. The values are chosen such that at a practical case temperature the power dissipation will not cause the junction temperature rating to be exceeded.

Various manufacturers may choose different criteria to establish ratings. At Motorola, use is made of the thermal response of the semiconductor and worst case values of on-state voltage and thermal resistance, to guarantee the junction temperature is at or below its rated value. Values shown on data sheets consequently differ somewhat from those computed from the standard formula:

$$T_C\ (max) = T\ (rated) - R_{\theta JC} \times P_{D(AV)}$$

where

T_C (max) = Maximum allowable case temperature

T (rated) = Rated junction temperature or maximum rated case temperature with zero principal current and rated ac blocking voltage applied.

$R_{\theta JC}$ = Junction to case thermal resistance

$P_{D(AV)}$ = Average power dissipation

The above formula is generally suitable for estimating case temperature in situations not covered by data sheet information. Worst case values should be used for thermal resistance and power dissipation.

OVERLOAD CURRENT RATINGS

Overload current ratings may be divided into two types: non-repetitive and repetitive.

Non-repetitive overloads are those which are not a part of the normal application of the device. Examples of such overloads are faults in the equipment in which the devices are used and accidental shorting of the load. Non-repetitive overload ratings permit the device to exceed its maximum operating junction temperature for short periods of time because this overload rating applies following any rated load condition. In the case of a reverse blocking thyristor or SCR, the device must block rated voltage in the reverse direction during the current overload. However, no type of thyristor is required to block off-stage voltage at any time during or immediately following the overload. Thus, in the case of a triac, the device need not block in either direction during or immediately following the overload. Usually only approximately one hundred such current overloads are permitted over the life of the device. These non-repetitive overload ratings just described may be divided into two types: multicycle (which include single cycle) and subcycle. For an SCR, the multicycle overload current rating, or surge current rating as it is commonly called, is generally presented as a curve giving the maximum peak

values of half sine wave on-state current as a function of overload duration measured in number of cycles for a 60 Hz frequency.

For a triac, the current waveform used in the rating is a full sine wave. Multicycle surge curves are used to select proper circuit breakers and series line impedances to prevent damage to the thyristor in the event of an equipment fault.

The subcycle overload or subcycle surge rating curve is so called because the time duration of the rating is usually from about one to eight milliseconds which is less than the time of one cycle of a 60 Hz power source. Overload peak current is often given in curve form as a function of overload duration. This rating also applies following any rated load condition and neither off-state nor reverse blocking capability is required on the part of the thyristor immediately following the overload current. The subcycle surge current rating may be used to select the proper current-limiting fuse for protection of the thyristor in the event of an equipment fault. Since this use of the rating is so common, manufacturers simply publish the i^2t rating in place of the subcycle current overload curve because fuses are commonly rated in terms of i^2t. The i^2t rating can be approximated from the single cycle surge rating (I_{TSM}) by using:

$$i^2t = I^2_{TSM} \times t/2$$

where the time t is the time base of the overload, i.e., 8.33 ms for a 60 Hz frequency.

Repetitive overloads are those which are an intended part of the application such as a motor drive application. Since this type of overload may occur a large number of times during the life of the thyristor, its rated maximum operating junction temperature must not be exceeded during the overload if long thyristor life is required. Since this type of overload may have a complex current waveform and duty-cycle, a current rating analysis involving the use of the transient thermal impedance characteristics is often the only practical approach. In this type of analysis, the thyristor junction-to-case transient thermal impedance characteristic is added to the user's heat dissipator transient thermal impedance characteristic. Then by the superposition of power waveforms in conjunction with the composite thermal impedance curve, the overload current rating can be obtained. The exact calculation procedure is found in the power semiconductor literature.

THEORY OF SCR POWER CONTROL

The most common form of SCR power control is phase control. In this mode of operation, the SCR is held in an off condition for a portion of the positive half cycle and then is triggered into an on condition at a time in the half cycle determined by the control circuitry (in which the circuit current is limited only by the load — the entire line voltage except for a nominal one volt drop across the SCR is applied to the load).

One SCR alone can control only one half cycle of the waveform. For full wave ac control, two SCRs are connected in inverse parallel (the anode of each connected to the cathode of the other, see Figure 2.9a). For full wave dc control, two methods are possible. Two SCRs may be used in a bridge rectifier (see Figure 2.9b) or one SCR may be placed in series with a diode bridge (see Figure 2.9c).

Figure 2.10 shows the voltage waveform along with some common terms used in describing SCR operation. Delay angle is the time, measured in electrical degrees, during which the SCR is blocking the line voltage. The period during which the SCR is on is called the conduction angle.

It is important to note that the SCR is a voltage controlling device. The load and power source determine the circuit current.

Now we arrive at a problem. Different loads respond to different characteristics of the ac waveform. Some loads are sensitive to peak voltage, some to average voltage and some to rms voltage. Figure 2.11 shows the various characteristic voltages plotted against the conduction angle for half wave and full wave circuits. These voltages have been normalized to the rms of the applied voltage. To determine the actual peak, average or rms voltage for any conduction angle, we simply multiply the normalized voltage by the rms value of the applied line voltage. (These normalized curves also apply to current in a resistive circuit.) Since the greatest majority of circuits are either 115 or 230 volt power, the curves have been redrawn for these voltages in Figure 2.12.

A relative power curve has been added to Figure 2.12 for constant impedance loads such as heaters. (Incandescent lamps and motors do not follow this curve precisely since their relative impedance changes with applied voltage.) To use the curves, we find the full wave rated power of the load, then multiply by the fraction associated with the phase angle in question. For example, a 180° conduction angle in a half wave circuit provides 0.5 x full wave full-conduction power.

An interesting point is illustrated by the power curves. A conduction angle of 30° provides only three per cent of full power in a full wave circuit, and a conduction angle of 150° provides 97 per cent of full power. Thus, the control circuit can provide 94 per cent of full power control with a pulse phase variation of only 120°. Thus, it becomes pointless in many cases to try to obtain conduction angles less than 30° or greater than 150°.

CONTROL CHARACTERISTICS

The simplest and most common control circuit for phase control is a relaxation oscillator. This circuit is shown diagrammatically as it would be used with an SCR in Figure 2.13. The capacitor is charged through the resistor from a voltage or current source until the breakover voltage of the trigger device is reached. At that time, the trigger device changes to its on state, and the capacitor is discharged through the gate of the SCR. Turn-on of the

SCR is thus accomplished with a short, high current pulse. Commonly used trigger devices are unijunction and programmable unijunction transistors, silicon bilateral switches, sidacs and optically coupled thyristors. Phase control can be obtained by varying the RC time constant of a charging circuit so that trigger device turn-on occurs at varying phase angles within the controlled half cycle.

If the relaxation oscillator is to be operated from a pure dc source, the capacitor voltage-time characteristic is shown in Figure 2.14. This shows the capacitor voltage as it rises all the way to the supply voltage through several time constants. Figure 2.14(b) shows the charge characteristic in the first time constant greatly expanded. It is this portion of the capacitor charge characteristic which is most often used in SCR and Triac control circuits.

Generally, a design starting point is selection of a capacitance value which will reliably trigger the thyristor when the capacitor is discharged. Gate characteristics and ratings, trigger device properties and the load impedance play a part in the selection. Since not all of the important parameters for this selection are completely specified, experimental determination is often the best method.

Low-current loads and strongly inductive circuits sometimes cause triggering difficulty because the gate current pulse goes away before the principal thyristor current achieves the latching value. A series gate resistor can be used to introduce a RC discharge time constant in the gate circuit and lengthen trigger pulse duration allowing more time for the main terminal current to rise to the latching value. Small thyristors will require a series gate resistance to avoid exceeding the gate ratings. The discharge time constant of a snubber, if used, can also aid latching. The duration of these capacitor discharge duration currents can be estimated by

$t_{w10} = 2.3\ RC$ where t_{w10} = time for current to decay to 10% of the peak.

For example, when an 8 volt SBS is used to discharge a 0.5 μF capacitor through a 15 ohm resistor into the gate of an SCR

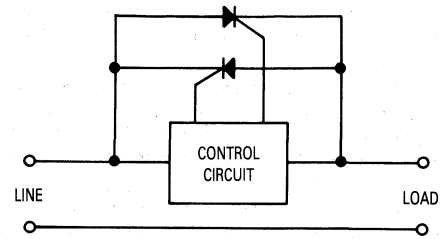
$t_{w10} = (2.3)\ (15)\ (0.5) = 17.3\ \mu s.$

Because of internal voltage drops in the SBS and SCR gates, the peak current will be somewhat less than

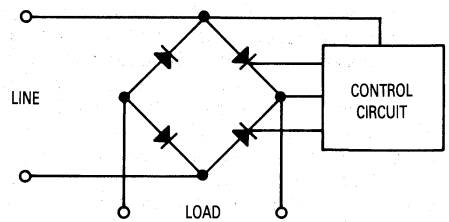
$I_{pk} = 8/15 = 0.53$ amp.

All trigger devices require some drive current to fire. Highly sensitive devices appear to be voltage operated when the current required to fire them is insignificant. The MBS4991 SBS requires that the switching current be taken into consideration. For a given RC time constant, larger capacitors allow the use of lower value timing resistors and less sensitive trigger components.

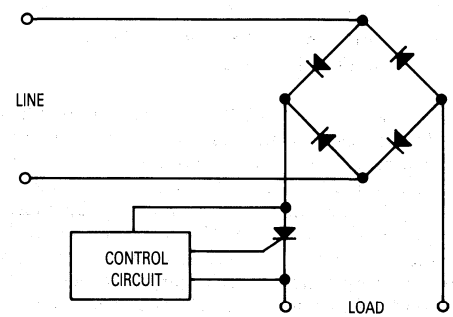
An example will demonstrate the procedure. Assume that we wish to trigger a 2N4170 SCR with an 8 volt MBS4991. We have determined that a 1 μF capacitor will supply the necessary SCR gate current magnitude and duration while not exceeding the gate ratings. Assume a 16 volt 60 Hz dc gate power supply, 30° minimum con-



(a)
ac Control



(b)
Two SCR dc Control



(c)
One SCR dc Control

Figure 2.9. SCR Connections For Various Methods Of Phase Control

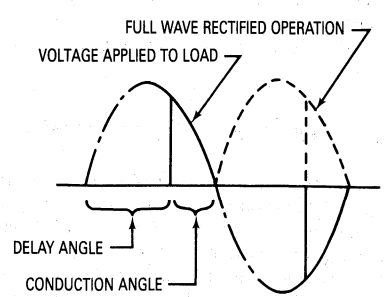


Figure 2.10. Sine Wave Showing Principles Of Phase Control

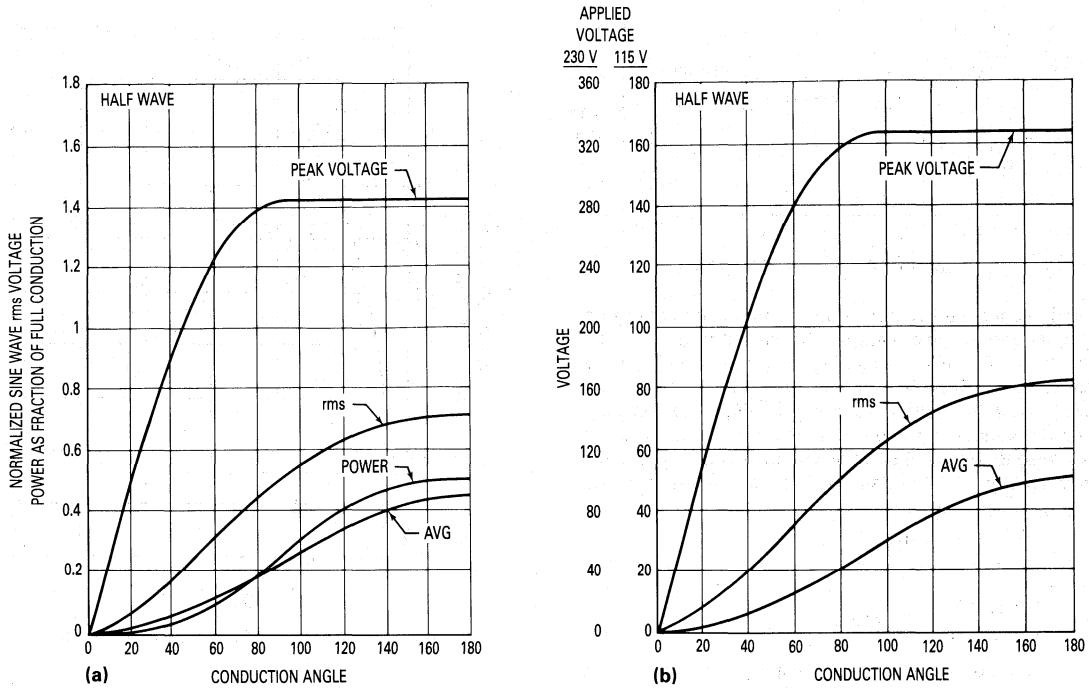


Figure 2.11. Half-Wave Characteristics Of Thyristor Power Control

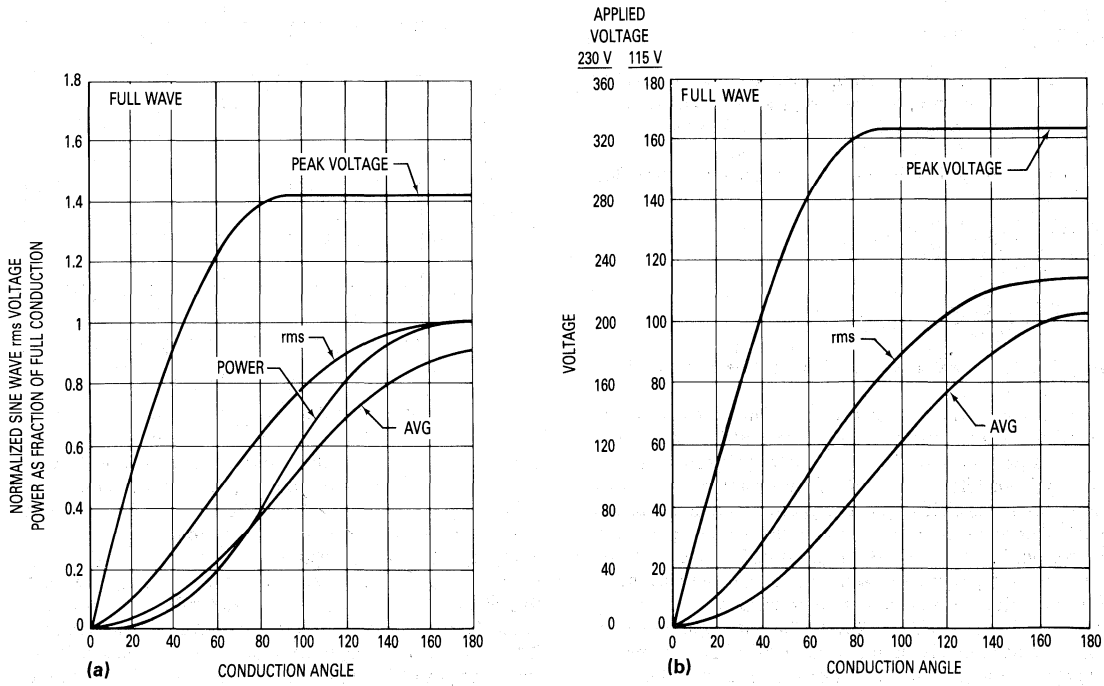


Figure 2.12. Full-Wave Characteristics Of Thyristor Power Control

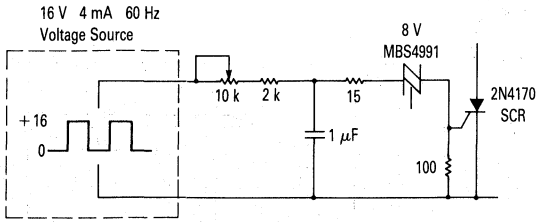


Figure 2.13. SCR Trigger Circuit

duction angle and 150° maximum conduction angle with a 60 Hz anode power source. The capacitor must charge to 8/16 or 0.5 of the available charging voltage in the desired time. Referring to Figure 2.14(b), we see that 0.5 of the charging voltage represents 0.693 time constant. The 30° conduction angle requires that the firing pulse be delayed 150° or 6.94 milliseconds (8.33 milliseconds is the period of 1/2 cycle at 60 Hz). To obtain this delay,

$$6.94 \text{ ms} = 0.693 \text{ RC}$$
$$\text{RC} = 10.01 \text{ ms.}$$

If C = 1 μF,

$$R = \frac{10 \times 10^{-3}}{1 \times 10^{-6}} = 10 \text{ k ohms.}$$

The timing resistor must be capable of supplying at least the worst case maximum SBS switching current at the peak point voltage. The available current is

$$\frac{16 \text{ V} - 8 \text{ V}}{10^4 \text{ ohms}} = 800\ \mu\text{A.}$$

This is more than the 500 μA needed by the MBS4991 at 25°C. If it were not, the design procedure would need to be repeated using larger C and smaller R. Alternatively, a more sensitive MBS4992 could be used.

To obtain minimum R, 150° conduction angle, the delay is 30° or

$$30/180 \times 8.33 = 1.39 \text{ ms}$$
$$1.39 \text{ ms} = 0.693 \text{ RC}$$
$$\text{RC} = 2.01 \text{ ms}$$

$$R = \frac{2.01 \times 10^{-3}}{1 \times 10^{-6}} = 2000 \text{ ohms.}$$

A 10 k potentiometer with a 2 k series resistor will serve this purpose.

In this application, the trigger circuit is reset by line crossing each half cycle. Consequently, SBS latching after firing is permissible. If the device were used as a free running oscillator, it would be necessary for the peak point current to be less than the minimum holding current specification of the SBS at maximum operating temperature. Timing accuracy requires the 16 V source to be capable of supplying the worst case required current. In the example, the initial instantaneous capacitor charging current will be 16 V/2 k = 8 mA. The gate load line must also enclose the peak point voltage. The SBS clamps the capacitor voltage when it breaks over causing little or no further change in the voltage across the capacitor. Consequently, all of the available current at that time (16 V–8 V)/2 k = 4 mA) diverts through the SBS causing it to fire.

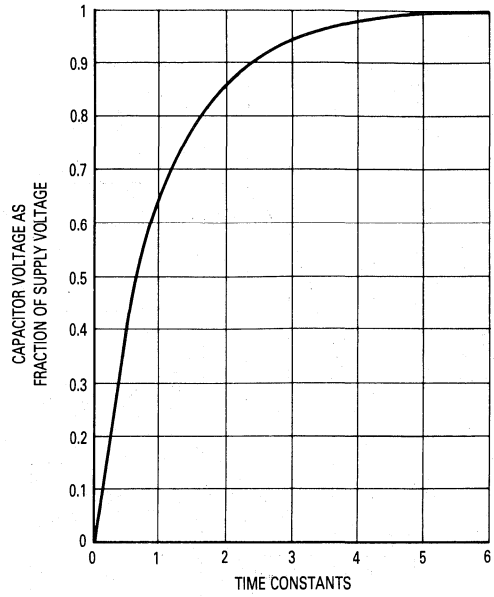


Figure 2.14(a). Capacitor Charging From dc Source

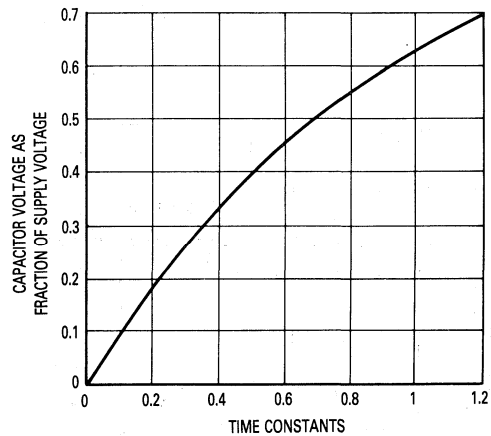


Figure 2.14(b). Expanded Scale

In many of the recently proposed circuits for low cost operation, the timing capacitor of the relaxation oscillator is charged through a rectifier and resistor using the ac power line as a source. Calculations of charging time with this circuit become exceedingly difficult, although they are still necessary for circuit design. The curves of Figure 2.15 simplify the design immensely. These curves show the voltage-time characteristic of the capacitor charged from one half cycle of a sine wave. Voltage is normalized

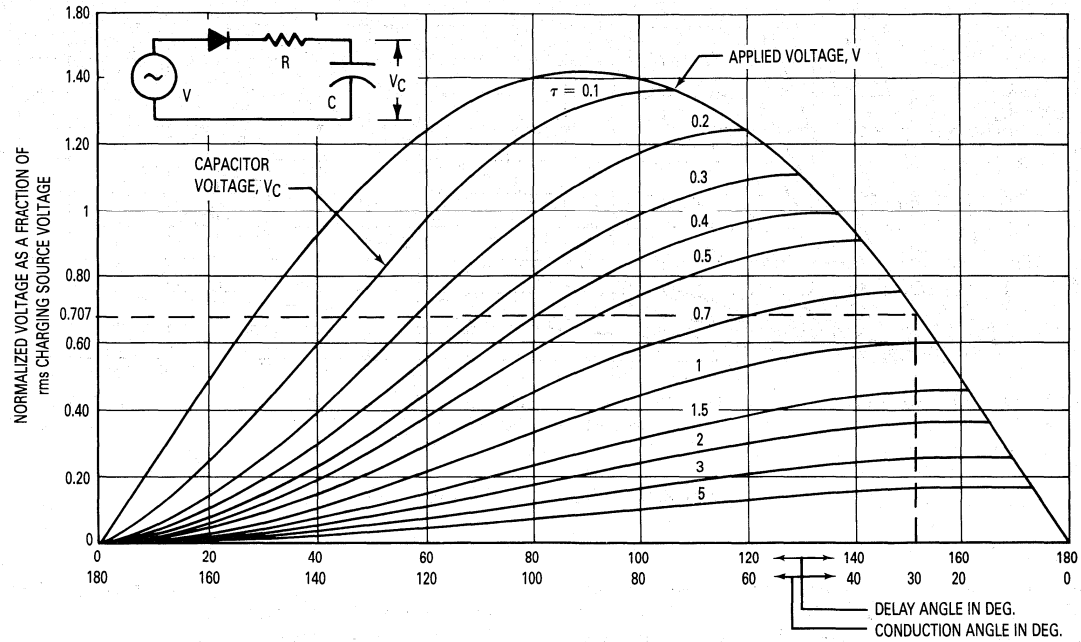


Figure 2.15(a). Capacitor Voltage When Charged

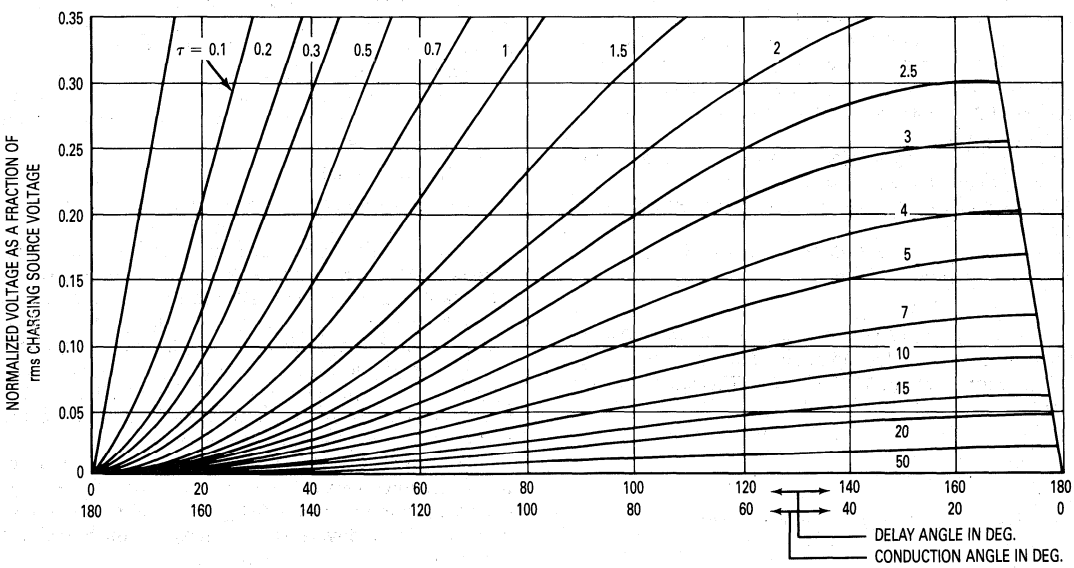


Figure 2.15(b). Expansion of Figure 2.15(a)

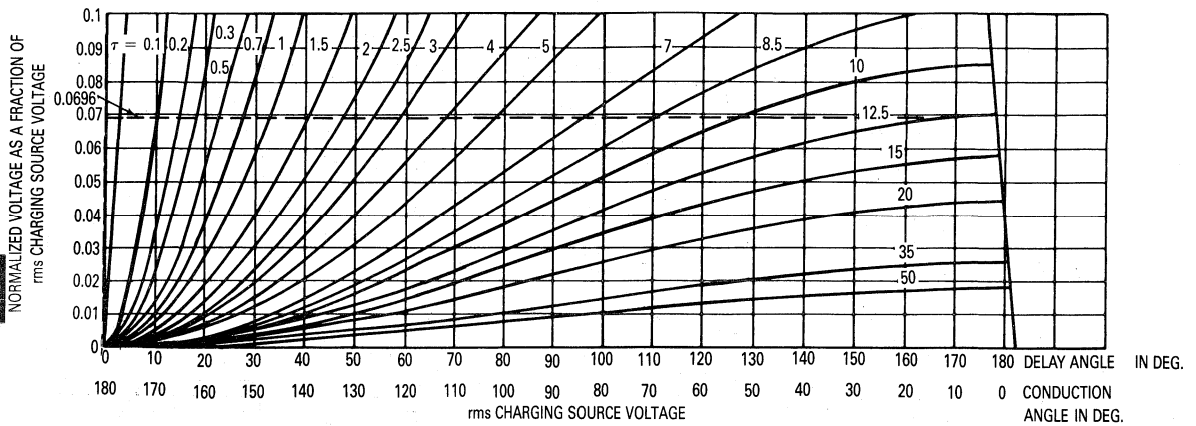


Figure 2.15(c). Expansion of Figure 2.15(b)

to the rms value of the sine wave for convenience of use. The parameter of the curves is a new term, the ratio of the RC time constant to the period of one half cycle, and is denoted by the Greek letter τ. It may most easily be calculated from the equation

$\tau = 2RCf$. Where: R = resistance in Ohms
C = capacitance in Farads
f = frequency in Hertz.

To use the curves when starting the capacitor charge from zero each half cycle, a line is drawn horizontally across the curves at the relative voltage level of the trigger breakdown compared to the rms sine wave voltage. The τ is determined for maximum and minimum conduction angles and the limits of R may be found from the equation for τ.

An example will again clarify the picture. Consider the same problem as the previous example, except that the capacitor charging source is the 115 Vac, 60 Hz power line.

The ratio of the trigger diode breakover voltage to the RMS charging voltage is then

$8/115 = 69.6 \times 10^{-3}$.

A line drawn at 0.0696 on the ordinate of Figure 2.15(c) shows that for a conduction angle of 30°, $\tau = 12$, and for a conduction angle of 150°, $\tau = 0.8$. Therefore, since $R = \tau/(2CF)$

$$R_{max} = \frac{12}{2(1.0\times10^{-6})60} \text{ 100 k ohms,}$$

$$R_{min} = \frac{0.8}{2(1\times10^{-6})60} = \text{6667 ohms.}$$

These values would require a potentiometer of 100 k in series with a 6.2 k minimum fixed resistance.

The timing resistor must be capable of supplying the highest switching current allowed by the SBS specification at the switching voltage.

When the conduction angle is less than 90°, triggering takes place along the back of the power line sine wave and maximum firing current thru the SBS is at the start of SBS breakover. If this current does not equal or exceed "Is" the SBS will fail to trigger and phase control will be lost. This can be prevented by selecting a lower value resistor and larger capacitor. The available current can be determined from Figure 2.15(a). The vertical line drawn from the conduction angle of 30° intersects the applied voltage curve at 0.707. The instantaneous current at breakover is then

$I = (0.707\times115^{-8})/100\text{ k} = 733\ \mu A$.

When the conduction angle is greater than 90°, triggering takes place before the peak of the sine wave. If the current thru the SBS does not exceed the switching current at the moment of breakover, triggering may still take place but not at the predicted time because of the additional delay for the rising line voltage to drive the SBS current up to the switching level. Usually long conduction angles are associated with low value timing resistors making this problem less likely. The SBS current at the moment of breakover can be determined by the same method described for the trailing edge.

It is advisable to use a shunt gate-cathode resistor across sensitive gate SCR's to provide a path for leakage currents and to insure that firing of the SCR causes turn-on of the trigger device and discharge of the gate circuit capacitor.

Figure 2.16(a) shows a simple dc full wave control circuit. R_{GK} is optional on non-sensitive gate parts. Figure 2.16(b) shows an ac control derived from that of Figure 2.16(a). Figure 2.16(c) is a double time constant circuit featuring low hysteresis.

TRIAC THEORY

The triac is a three-terminal ac semiconductor switch which is triggered into conduction when a low-energy signal is applied to its gate. Unlike the silicon controlled rectifier or SCR, the triac will conduct current in either direction when turned on. The triac also differs from the SCR in that either a positive or negative gate signal will trigger the triac into conduction. The triac may be thought of as two complementary SCRs in parallel.

The triac offers the circuit designer an economical and versatile means of accurately controlling ac power. It has several advantages over conventional mechanical switches. Since the triac has a positive "on" and a zero-current "off" characteristic, it does not suffer from the contact bounce or arcing inherent in mechanical switches. The switching action of the triac is very fast compared to conventional relays, giving more accurate control. A triac can be trigged by dc, ac, rectified ac or pulses. Because of the low energy required for triggering a triac, the control circuit can use any of a number of low-cost solid-state devices presently on the market such as transistors, unijunction transistors, bilateral switches, four-layer diodes and sensitive-gate SCRs and optically coupled drivers.

CHARACTERISTICS OF THE TRIAC

Figure 2.17(a) shows the triac symbol and its relationship to a typical package. Since the triac is a bilateral device, the terms "anode" and "cathode" used for unilateral devices have no meaning. Therefore, the terminals are simply designated by MT1, MT2, and G, where MT1 and MT2 are the current-carrying terminals, and G is the gate terminal used for triggering the triac. To avoid confusion, it has become standard practice to specify all currents and voltages using MT1 as the reference point.

The basic structure of a triac is shown in Figure 2.17(b). This drawing shows why the symbol adopted for the triac consists of two complementary SCRs with a common gate. The triac is a five-layer device with the region between MT1 and MT2 being a P-N-P-N switch (SCR) in parallel with a N-P-N-P switch (complementary SCR). Also, the structure gives some insight into the triac's ability to be triggered with either a positive or negative gate signal. The region between MT1 and G consists of two complementary diodes. A positive or negative gate signal will forward-bias one of these diodes causing the same transistor action found in the SCR. This action breaks down the blocking junction regardless of the polarity of

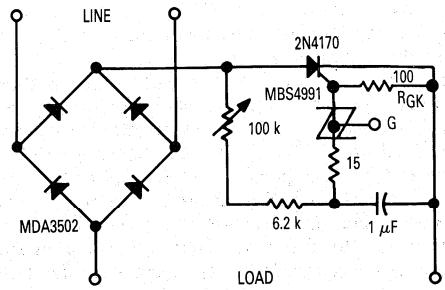


Figure 2.16(a). Simple dc Power Control Circuit

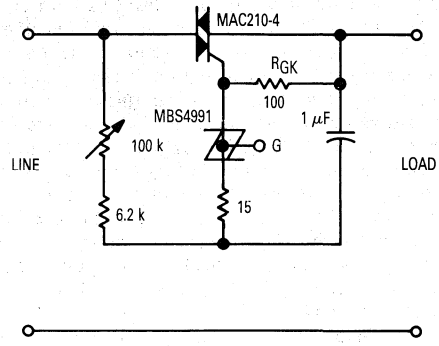


Figure 2.16(b). Simple Full-Wave Power Control

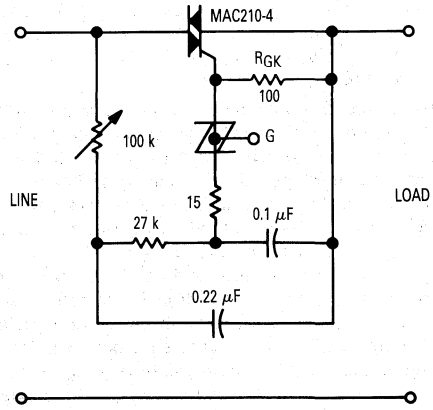


Figure 2.16(c). Full Range ac Power Control

MT1. Current flow between MT2 and MT1 then causes the device to provide gate current internally. It will remain on until this current flow is interrupted.

The voltage-current characteristic of the triac is shown in Figure 2.18 where, as previously stated, MT1 is used as the reference point. The first quadrant, Q-I, is the region where MT2 is positive with respect to MT1 and quadrant III is the opposite case. Several of the terms used in characterizing the triac are shown on the figure. V_{DRM} is the breakover voltage of the device and is the highest voltage the triac may be allowed to block in either direction. If this voltage is exceeded, even transiently, the triac may go into conduction without a gate signal. Although the triac is not damaged by this action if the current is limited, this situation should be avoided because control of the triac is lost. A triac for a particular application should have V_{DRM} at least as high as the peak of the ac waveform to be applied so reliable control can be maintained. The holding current (I_H) is the minimum value of current necessary to maintain conduction. When the current goes below I_H, the triac ceases to conduct and reverts to the blocking state. I_{DRM} is the leakage current of the triac with V_{DRM} applied from MT2 to MT1 and is several orders of magnitude smaller than the current rating of the device. The figure shows the characteristic of the triac without a gate signal applied but it should be noted that the triac can be triggered into the on state at any value of voltage up to V_{DRM} by the application of a gate signal. This important characteristic makes the triac very useful.

Since the triac will conduct in either direction and can be triggered with either a positive or negative gate signal there are four possible triggering modes (Figure 2.3):

Quadrant I; MT2(+), G(+), positive voltage and positive gate current. Quadrant II; MT2(+), G(−), positive voltage and negative gate current. Quadrant III; MT2(−), G(−), negative voltage and negative gate current. Quadrant IV; MT2(−), G(+), negative voltage and positive gate current.

Present triacs are most sensitive in quadrants I and III, slightly less so in quadrant II, and much less sensitive in quadrant IV. Therefore it is not recommended to use quadrant IV unless special circumstances dictate it.

An important fact to remember is that since a triac can conduct current in both directions, it has only a brief interval during which the sine wave current is passing through zero to recover and revert to its blocking state. For this reason, reliable operation of present triacs is limited to 60 Hz line frequency and lower frequencies.

For inductive loads, the phase-shift between the current and voltage means that at the time the current falls below I_H and the triac ceases to conduct, there exists a certain voltage which must appear across the triac. If this voltage appears too rapidly, the triac will resume conduction and control is lost. In order to achieve control with certain inductive loads, the rate of rise in voltage (dv/dt) must be limited by a series RC network across the triac. The capacitor will then limit the dv/dt across the triac. The resistor is necessary to limit the surge of current from the capacitor when the triac fires, and to damp the ringing of the capacitance with the load inductance.

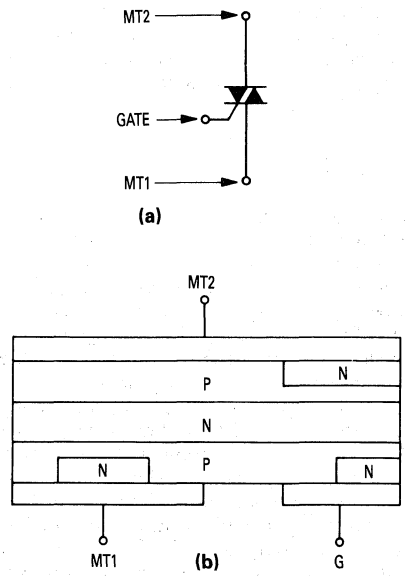


Figure 2.17. Triac Structure and Symbol

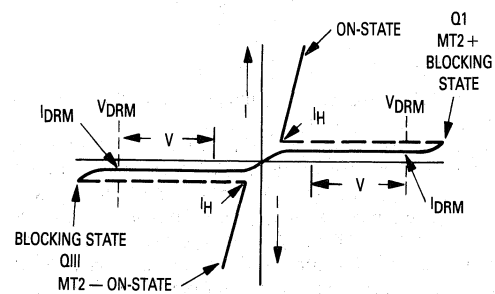


Figure 2.18. Triac Voltage-Current Characteristic

METHODS OF CONTROL

AC SWITCH

A useful application of the triac is as a direct replacement for an ac mechanical relay. In this application, the triac furnishes on-off control and the power-regulating ability of the triac is not utilized. The control circuitry for this application is usually very simple, consisting of a source for the gate signal and some type of small current switch, either mechanical or electrical. The gate signal can be obtained from a separate source or directly from the line voltage at terminal MT2 of the triac.

PHASE CONTROL

An effective and widely-used method of controlling the average power to a load through the triac is by phase control. Phase control is a method of utilizing the triac to apply the ac supply to the load for a controlled fraction of each cycle. In this mode of operation, the triac is held in an off or open condition for a portion of each positive and negative cycle, and then is triggered into an on condition at a time in the half cycle determined by the control circuitry. In the on condition, the circuit current is limited only by the load — i.e., the entire line voltage (less the forward drop of the triac) is applied to the load.

Figure 2.19 shows the voltage waveform along with some common terms used in describing triac operation. Delay angle is the angle, measured in electrical degrees, during which the triac is blocking the line voltage. The period during which the triac is on is called the conduction angle.

It is important to note that the triac is either off (blocking voltage) or fully on (conducting). When it is in the on condition, the circuit current is determined only by the load and the power source.

As one might expect, in spite of its usefulness, phase control is not without disadvantages. The main disadvantage of using phase control in triac applications is the generation of electro-magnetic interference (EMI). Each time the triac is fired the load current rises from zero to the load-limited current value in a very short time. The resulting di/dt generates a wide spectrum of noise which may interfere with the operation of nearby electronic equipment unless proper filtering is used.

ZERO POINT SWITCHING

In addition to filtering, EMI can be minimized by zero-point switching, which is often preferable. Zero-point switching is a technique whereby the control element (in this case the triac) is gated on at the instant the sine wave voltage goes through zero. This reduces, or eliminates, turn-on transients and the EMI. Power to the load is controlled by providing bursts of complete sine waves to the load as shown in Figure 2.20. Modulation can be on a random basis with an on-off control, or a proportioning basis with the proper type of proportional control.

In order for zero-point switching to be effective, it must indeed be *zero point* switching. If a triac is turned on with as little as 10 volts across it into a load of a few-hundred watts, sufficient EMI will result to nullify the advantages of adopting zero-point switching in the first place.

BASIC TRIAC AC SWITCHES

Figure 2.21 shows methods of using the triac as an on-off switch. These circuits are useful in applications where simplicity and reliability are important. As previously stated, there is no arcing with the triac, which can be very important in some applications. The circuits are for resistive loads as shown and require the addition of a dv/dt network across the triac for inductive loads. Figure 2.21(a) shows low-voltage control of the triac. When switch S1 is closed, gate current is supplied to the triac from the 10 volt battery. In order to reduce surge current failures during turn on (t_{on}), this current should be 5 to 10 times the maximum gate current (I_{GT}) required to trigger the triac.

The triac turns on and remains on until S1 is opened. This circuit switches at zero current except for initial turn on. S1 can be a very-low-current switch because it carries only the triac gate current.

Figure 2.21(b) shows a triac switch with the same characteristics as the circuit in Figure 2.21(a) except the need for a battery has been eliminated. The gate signal is obtained from the voltage at MT2 of the triac prior to turn on.

The circuit shown in Figure 2.21(c) is a modification of Figure 2.21(b). When switch S1 is in position one, the triac receives no gate current and is non-conducting. With S1 in position two, circuit operation is the same as that for Figure 2.21(b). In position three, the triac receives gate current only on positive half cycles. Therefore, the triac conducts only on positive half cycles and the power to the load is half wave.

Figure 2.21(d) shows ac control of the triac. The pulse can be transformer coupled to isolate power and control circuits. Peak current should be 10 times $I_{GT(max)}$ and the RC time constant should be 5 times $t_{on(max)}$. A high frequency pulse (1 to 5 kHz) is often used to obtain zero point switching.

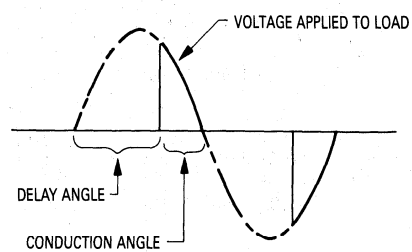


Figure 2.19. Sine Wave Showing Principles of Phase Control

ZERO POINT SWITCHING TECHNIQUES

Zero-point switches are highly desirable in many applications because they do not generate electro-magnetic interference (EMI). A zero-point switch controls sine-wave power in such a way that either complete cycles or half cycles of the power supply voltage are applied to the load as shown in Figure 2.22. This type of switching is primarily used to control power to resistive loads such as heaters. It can also be used for controlling the speed of motors if the duty cycle is modulated by having short bursts of power applied to the load and the load characteristic is primarily inertial rather than frictional. Modulation can be on a random basis with an on-off control, or on a proportioning basis with the proper type of proportioning control.

In order for zero-point switching to be effective, it must be true zero-point switching. If an SCR is turned on with an anode voltage as low as 10 volts and a load of just a few hundred watts, sufficient EMI will result to nullify the advantages of going to zero-point switching in the first place. The thyristor to be turned on must receive gate drive exactly at the zero crossing of the applied voltage.

The most successful method of zero-point thyristor control is therefore, to have the gate signal applied before the zero crossing. As soon as the zero crossing occurs, anode voltage will be supplied and the thyristor will come on. This is effectively accomplished by using a capacitor to derive a 90° leading gate signal from the power line source. However, only one thyristor can be controlled from this phase-shifted signal, and a slaving circuit is necessary to control the other SCR to get full-wave power control. These basic ideas are illustrated in Figure 2.23. The slaving circuit fires only on the half cycle after the firing of the master SCR. This guarantees that only complete cycles of power will be applied to the load. The gate signal to the master SCR receives all the control; a convenient control method is to replace the switch with a

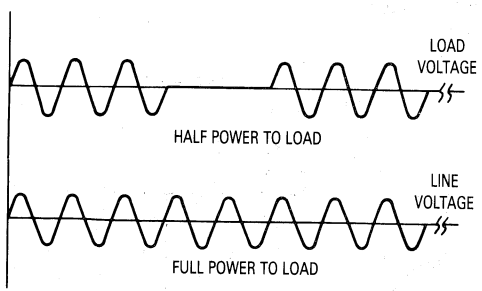
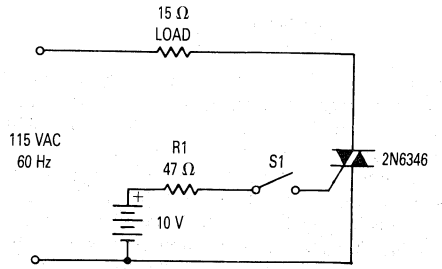
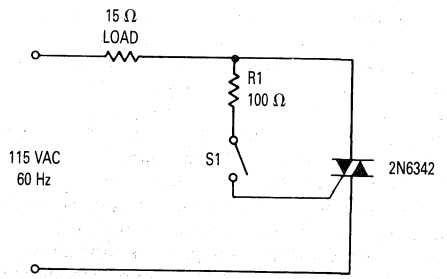


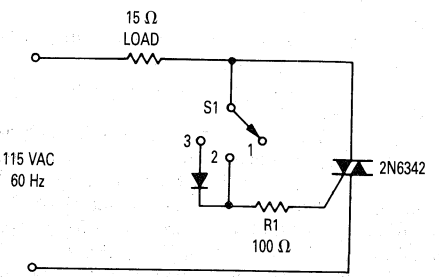
Figure 2.20. Sine Wave Showing Principles of Zero-Point Switching



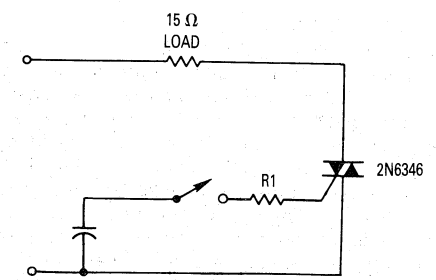
(a): Low Voltage Controlled Triac Switch



(b): Triac ac Static Contractor



(c): 3 Position Static Switch



(d): AC Controlled Triac Switch

Figure 2.21. Triac Switches

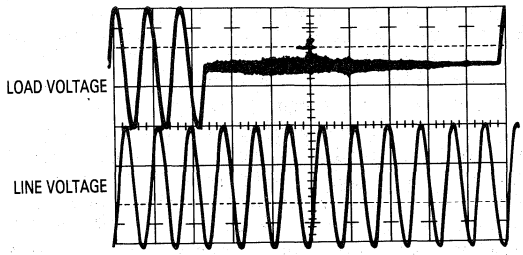


Figure 2.22. Load Voltage and Line Voltage for 25% Duty Cycle

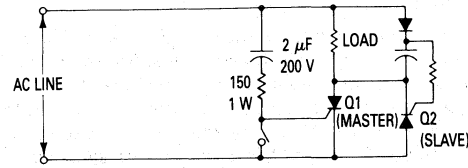


Figure 2.23. Slave and Master SCRs for Zero-Point Switching

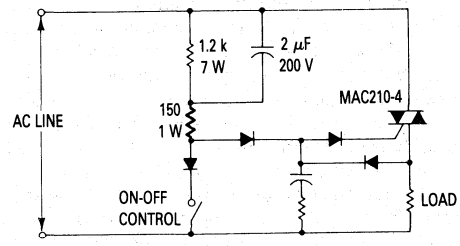


Figure 2.24. Triac Zero-Point Switch

low-power transistor, which can be controlled by bridge-sensing circuits, manually controlled potentiometers, or various other techniques.

A basic SCR is very effective and trouble free. However, it can dissipate considerable power. This must be taken into account in designing the circuit and its packaging.

In the case of triacs, a slaving circuit is also usually required to furnish the gate signal for the negative half cycle. However, triacs can use slave circuits requiring less power than do SCRs as shown in Figure 2.23. Other considerations being equal, the easier slaving will sometimes make the triac circuit more desirable than the SCR circuit.

Besides slaving circuit power dissipation, there is another consideration which should be carefully checked when using high-power zero-point switching. Since this is on-off switching, it abruptly applies the full load to the power line every time the circuit turns on. This may cause a temporary drop in voltage which can lead to erratic operation of other electrical equipment on the line (light dimming, TV picture shrinkage, etc.). For this reason, loads with high cycling rates should not be powered from the same supply lines as lights and other voltage-sensitive devices. On the other hand, if the load cycling rate is slow, say once per half minute, the loading flicker may not be objectionable on lighting circuits.

A note of caution is in order here. The full-wave zero-point switching control illustrated in Figure 2.23 should not be used as a half-wave control by removing the slave SCR. When the slave SCR in Figure 2.23 is removed, the master SCR has positive gate current flowing over approximately 1/4 of a cycle while the SCR itself is in the reverse-blocking state. This occurs during the negative half cycle of the line voltage. When this condition exists, Q1 will have a high leakage current with full voltage applied and will therefore be dissipating high power. This will cause excessive heating of the SCR and may lead to its failure. If it is desirable to use such a circuit as a half-wave control, then some means of clamping the gate signal during the negative half cycle must be devised to inhibit gate current while the SCR is reverse blocking. The circuits shown in Figures 2.25 and 2.26 do not have this disadvantage and may be used as half-wave controls.

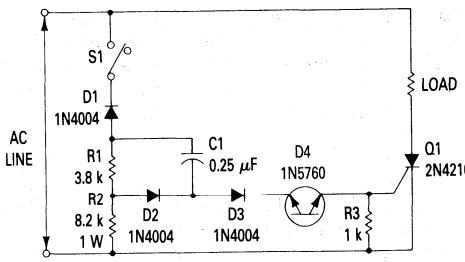


Figure 2.25. Sensitive-Gate Switch

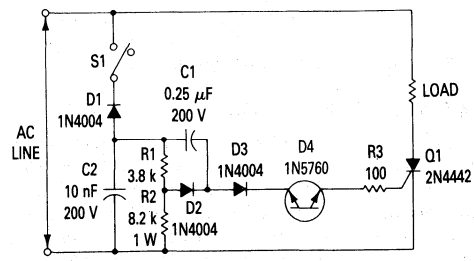


Figure 2.26. Zero-Point Switch

OPERATION

The zero-point switches shown in Figures 2.25 and 2.26 are used to insure that the control SCR turns on at the start of each positive alternation. In Figure 2.25 a pulse is generated before the zero crossing and provides a small amount of gate current when line voltage starts to go positive. This circuit is primarily for sensitive-gate SCRs. Less-sensitive SCRs, with their higher gate currents, normally require smaller values for R1 and R2 and the result can be high power dissipation in these resistors. The circuit of Figure 2.26 uses a capacitor, C2, to provide a low-impedance path around resistors R1 and R2 and can be used with less-sensitive, higher-current SCRs without increasing the dissipation. This circuit actually oscillates near the zero crossing point and provides a series of pulses to assure zero-point switching.

The basic circuit is that shown in Figure 2.25. Operation begins when switch S1 is closed. If the positive alternation is present, nothing will happen since diode D1 is reverse biased. When the negative alternation begins, capacitor C1 will charge through resistor R2 toward the limit of voltage set by the voltage divider consisting of resistors R1 and R2. As the negative alternation reaches its peak, C1 will have charged to about 40 volts. Line voltage will decrease but C1 cannot discharge because diode D2 will be reverse biased. It can be seen that C1 and three-layer diode D4 are effectively in series with the line. When the line drops to 10 volts, C1 will still be 40 volts positive with respect to the gate of Q1. At this time D4 will see about 30 volts and will trigger. This allows C1 to discharge through D3, D4, the gate of Q1, R2, and R1. This discharge current will continue to flow as the line voltage crosses zero and will insure that Q1 turns on at the start of the positive alternation. Diode D3 prevents reverse gate-current flow and resistor R3 prevents false triggering.

The circuit in Figure 2.26 operates in a similar manner up to the point where C1 starts to discharge into the gate. The discharge path will now be from C1 through D3, D4, R3, the gate of Q1, and capacitor C2. C2 will quickly charge from this high pulse of current. This reduces the voltage across D4 causing it to turn off and again revert to its blocking state. Now C2 will discharge through R1 and R2 until the voltage on D4 again becomes sufficient to cause it to break back. This repetitive exchange of charge from C1 to C2 causes a series of gate-current pulses to flow as the line voltage crosses zero. This means that Q1 will again be turned on at the start of each positive alternation as desired. Resistor R3 has been added to limit the peak gate current.

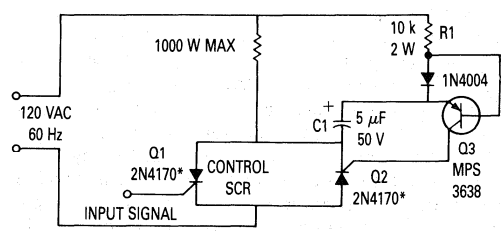
AN SCR SLAVING CIRCUIT

An SCR slaving circuit will provide full-wave control of an ac load when the control signal is available to only one of a pair of SCRs. An SCR slaving circuit is commonly used where the master SCR is controlled by zero-point switching. Zero-point switching causes the load to receive a full cycle of line voltage whenever the control signal is applied. The duty cycle of the control signal therefore determines the average amount of power supplied to the load. Zero-point switching is necessary for large loads such as electric heaters because conventional phase-shift techniques would generate an excessive amount of electro-magnetic interference (EMI).

This particular slaving circuit has two important advantages over standard RC discharge slaving circuits. It derives these advantages with practically no increase in price by using a low-cost transistor in place of the current-limiting resistor normally used for slaving. The first advantage is that a large pulse of gate current is available at the zero-crossing point. This means that it is not necessary to select sensitive-gate SCRs for controlling power. The second advantage is that this current pulse is reduced to zero within one alternation. This has a couple of good effects on the operation of the slaving SCR. It prevents gate drive from appearing while the SCR is reverse-biased, which would produce high power dissipation within the device. It also prevents the slaved SCR from being turned on for additional half cycles after the drive is removed from the control SCR.

OPERATION

The SCR slaving circuit shown in Figure 2.27 provides a single power pulse to the gate of SCR Q2 each time SCR Q1 turns on, thus turning Q2 on for the half cycle following the one during which Q1 was on. Q2 is therefore turned on only when Q1 is turned on, and the load can be controlled by a signal connected to the gate of Q1 as shown in the schematic. The control signal can be either dc or a power pulse. If the control signal is synchronized with the power line, this circuit will make an excellent zero-point switch. During the time that Q1 is on, capacitor C1 is charged through R1, D1 and Q1. While C1 is being charged, D1 reverse-biases the base-emitter junction of Q3, thereby holding it off. The charging time constant, R1, C1, is set long enough that C1 charges for practically the entire half cycle. The charging rate of C1 follows an "S" shaped curve, charging slowly at first, then faster as the supply voltage peaks, and finally slowly again as the supply voltage decreases. When the supply voltage falls below the voltage across C1, diode D1 becomes reverse biased and the base-emitter of Q3 becomes forward biased. For the values shown, this occurs approximately 6° before the end of the half cycle con-



*1000 WATT LOAD. SEE TEXT.

Figure 2.27. SCR Slave Circuit

duction of Q1. The base current is derived from the energy stored in C1. This turns on Q3, discharging C1 through Q3 and into the gate of Q2. As the voltage across C1 decreases, the base drive of Q3 decreases and somewhat limits the collector current. The current pulse must last until the line voltage reaches a magnitude such that latching current will exist in Q2. The values shown will deliver a current pulse which peaks at 100 mA and has a magnitude greater than 50 mA when the anode-cathode voltage of Q2 reaches plus 10 volts. This circuit completely discharges C1 during the half cycle that Q2 is on. This eliminates the possibility of Q2 being slaved for additional half cycles after the drive is removed from Q1. The peak current and the current duration are controlled by the values of R1 and C1. The values chosen provide sufficient drive for "shorted emitter" SCRs which typically require 10 to 20 mA to fire. The particular SCR used must be capable of handling the maximum current requirements of the load to be driven; the 8 ampere, 200 V SCRs shown will handle a 1000 watt load.

CHAPTER 3
THYRISTOR DRIVERS AND TRIGGERING

1

Triggering a thyristor requires meeting its gate energy specifications and there are many ways of doing this. In general, the gate should be driven hard and fast to ensure complete gate turn on and thus minimize di/dt effects. Usually this means a gate current of at least three times the gate turn on current with a pulse rise of less than one microsecond and a pulse width greater than 10 microseconds. The gate can also be driven by a dc source as long as the average gate power limits are met.

Some of the methods of driving the gate include:

1) Direct drive from logic families of transistors
2) Opto triac drivers
3) Unijunction transistors (UJTs) and programmable unijunction transistors (PUTs)
4) Silicon bilateral switches (SBSs)
5) SIDACs

In this chapter we will discuss all of these, as well as some of the important design and application considerations in triggering thyristors in general. In the chapter on applications, we will also discuss some additional considerations relating to drivers and triggers in specific applications.

PULSE TRIGGERING OF SCRs

GATE TURN-ON MECHANISM

The turn-on of PNPN devices has been discussed in many papers where it has been shown that the condition of switching is given by $\frac{dv}{di} = 0$ (i.e., $\alpha_1 + \alpha_2 = 1$, where α_1 and α_2 are the current amplification factors of the two "transistors." However, in the case of an SCR connected to a reverse gate bias, the device can have $\alpha_1 + \alpha_2 = 1$ and still stay in the blocking state. The condition of turn-on is actually $\alpha_1 + \alpha_2 > 1$.

The current amplification factor, α, increases with emitter current; some typical curves are shown in Figure 3.1. The monotonical increase of α with I_E of the device in the blocking state makes the regeneration of current (i.e., turn-on) possible.

Using the two transistor analysis, the anode current, I_A, can be expressed as a function of gate current, I_G, as:

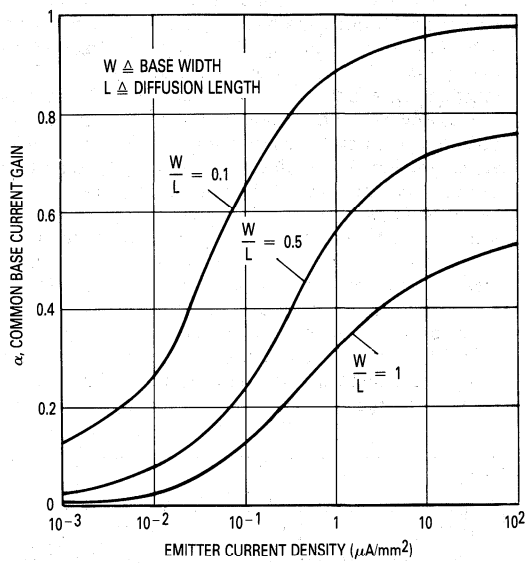


Figure 3.1. Typical Variation of Transistor α with Emitter Current Density

$$I_A = \frac{\alpha_2 I_G + I_{CS1} + I_{CS2}}{1 - \alpha_1 - \alpha_2} \quad (1)$$

Definitions and derivations are given in Appendix I. Note that the anode current, I_A, will increase to infinity as $\alpha_1 + \alpha_2 = 1$. This analysis is based upon the assumption that no majority carrier current flows out of the gate circuit. When no such assumption is made, the condition for turn-on is given by:

$$\frac{I_K}{I_A} = \frac{1 - \alpha_1}{\alpha_2} \quad (2)$$

which corresponds to $\alpha_1 + \alpha_2 > 1$ (see Appendix I).

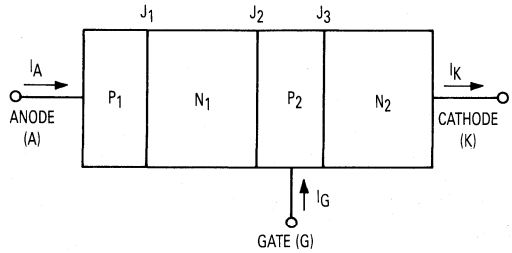


Figure 3.2. Schematic Structure of an SCR, Positive Currents Are Defined as Shown by the Arrows

Current regeneration starts when charge or current is introduced through the gate (Figure 3.2). Electrons are injected from the cathode across J_3; they travel across the P_2 "base" region to be swept out by the collector junction, J_2, and thrown into the N_1 base. The increase of majority carrier electrons in region N_1 decreases the potential in region N_1, so that holes from P_1 are injected across the junction J_1, into the N_1 "base" region to be swept across J_2, and thrown into the P_2 "base" region. The increase in the potential of region P_2 causes more electrons to be injected into P_2, thereby repeating the cycle. Since α increases with the emitter current, an increase of regeneration takes place until $\alpha_1 + \alpha_2 > 1$. Meanwhile, more carriers are collected than emitted from either of the emitters. The continuity of charge flow is violated and there is an electron build-up on the N_1 side of J_2, and a hole build-up on the P_2 side. When the inert impurity charges are compensated for by injected majority carriers, the junction J_2 becomes forward biased. The collector emits holes back to J_1 and electrons to J_3 until a steady state continuity of charge is established.

During the regeneration process, the time it takes for a minority carrier to travel across a base region is the transit time, t, which is given approximately as:

$$t_1 = \frac{W^2_i}{2D_i} \quad \text{where } W_i = \text{base width}, \; D_i = \text{diffusion length} \tag{3}$$

(The subscript "i" can be either 1 or 2 to indicate the appropriate base.) The time taken from the start of the gate trigger to the turn-on of the device will be equal to some multiple of the transit time.

CURRENT PULSE TRIGGERING

Current pulse triggering is defined as supplying current through the gate to compensate for the carriers lost by recombination in order to provide enough current to sustain increasing regeneration. If the gate is triggered with a current pulse, shorter pulse widths require higher currents as shown by Figure 3.3(a). Figure 3.3(a) seems to indicate there is a constant amount of charge required to trigger on the device when I_G is above a threshold level. When the charge required for turn-on is plotted versus pulse current or pulse width, there is an optimum range of current levels or pulse widths for which the charge is minimum, as shown in region A of Figure 3.3(b) and (c). Region C shows that for lower current levels (i.e., longer minimum pulse widths) more charge is required to trigger on the device. Region B shows increasing charge required as the current gets higher and the pulse width smaller.

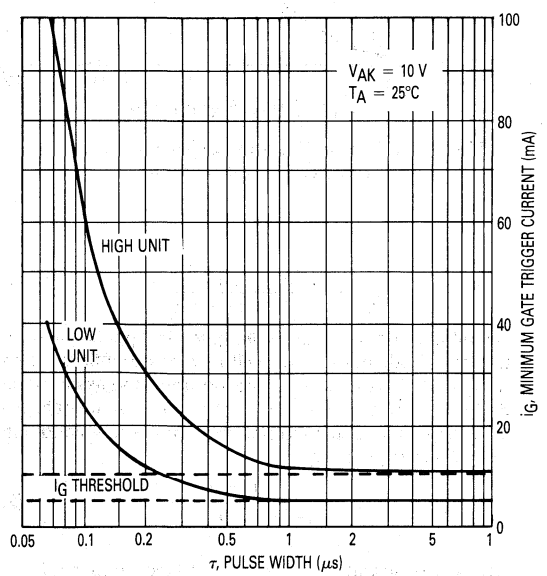


Figure 3.3(a). Typical Variation of Minimum Gate Current Required to Trigger

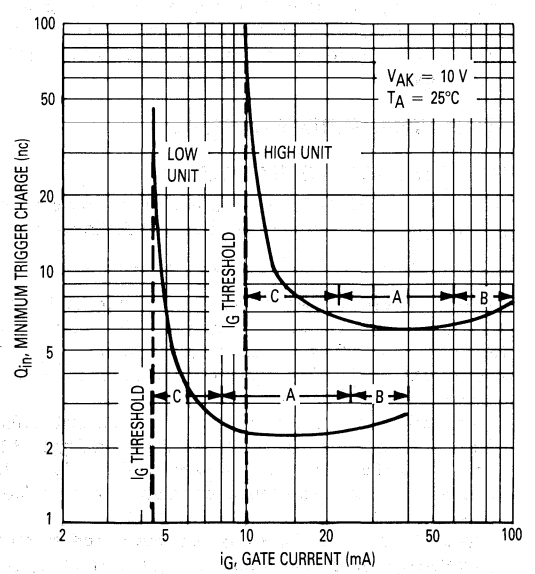


Figure 3.3(b). Variation of Charge versus Gate Current

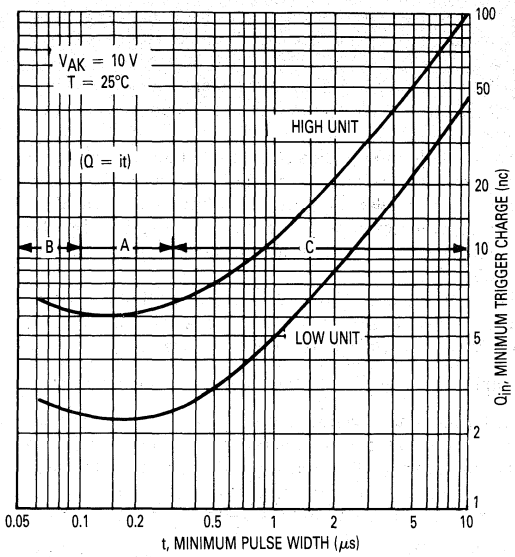


Figure 3.3(c). Variation of Charge versus Minimum Pulse Width

The charge characteristic curves can be explained qualitatively by the variation of current amplification (α_T) with respect to emitter current. A typical variation of α_1 and α_2 for a thyristor is shown in Figure 3.4(a). From Figure 3.4(a), it can be deduced that the total current amplification factor, $\alpha_T = \alpha_1 + \alpha_2$, has a characteristic curve as shown in Figure 3.4(b). (The data does not correspond to the data of Figure 3.3 — they are taken for different types of devices.)

The gate current levels in region A of Figure 3.3 correspond to the emitter (or anode) currents for which the slope of the α_T curve is steepest (Figure 3.4(b)). In region A the rate that α_T builds up with respect to changes of I_E (or I_A) is high, little charge is lost by recombination, and therefore, a minimum charge is required for turn-on.

In region C of Figure 3.3, lower gate current corresponds to small I_E (or I_A) for which the slope of α_T, as well as α_T itself, is small. It takes a large change in I_E (or I_A) in order to build up α_T. In this region, a lot of the charge supplied through the gate is lost by recombination. The charge required for turn-on increases markedly as the gate current is decreased to the threshold level. Below this threshold, the device will not turn on regardless of how long the pulse width becomes. At this point, the slope of α_T is equal to zero; all of the charge supplied is lost completely in recombination or drained out through gate-cathode shunt resistance. A qualitative analysis of variation of charge with pulse width at region A and C is discussed in Appendix II.

In region B, as the gate current level gets higher and the pulse width smaller, there are two effects that contribute to an increasing charge requirement to trigger-on the device: (1) the decreasing slope of α_T and, (2) the transit time effect. As mentioned previously, it takes some multiple of the transit time for turn-on. As the gate pulse width decreases to N ($t_{N1} + t_{P2}$) or less, (where N is a positive real number, t_{N1} = transit time of base N_1, and

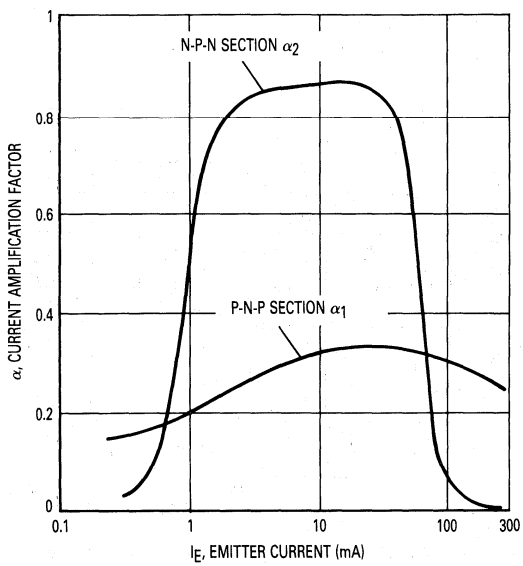


Figure 3.4(a). The Variation of α_1 and α_2 with Emitter Current for the Two Sections of Two Typical Silicon Controlled Rectifiers

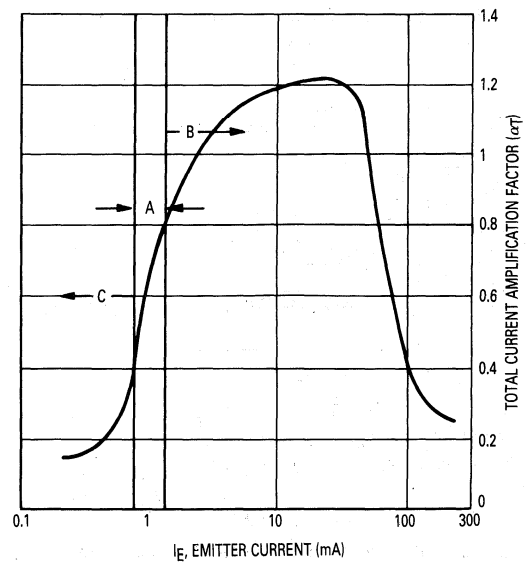


Figure 3.4(b). Typical Variation of α_T versus Emitter Current

t_{P2} = transit time of base P2) the amount of current required to turn-on the device should be large enough to flood the gate to cathode junction nearly instantaneously with a charge which corresponds to I_E (or I_A) high enough to give $\alpha_T > 1$.

CAPACITANCE CHARGE TRIGGERING

Using a gate trigger circuit as shown in Figure 3.5, the charge required for turn-on increases with the value of capacitance used as shown in Figure 3.7. Two reasons may account for the increasing charge characteristics:

1) An effect due to threshold current.
2) An effect due to variation of gate spreading resistance.

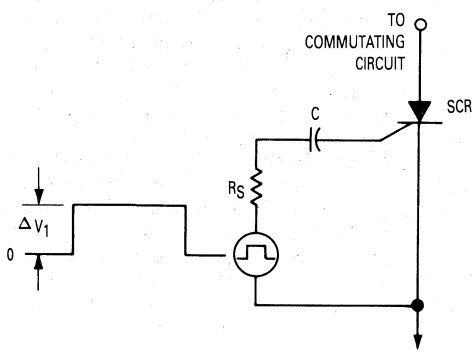


Figure 3.5. Gate Circuit of Capacitance Charge Triggering

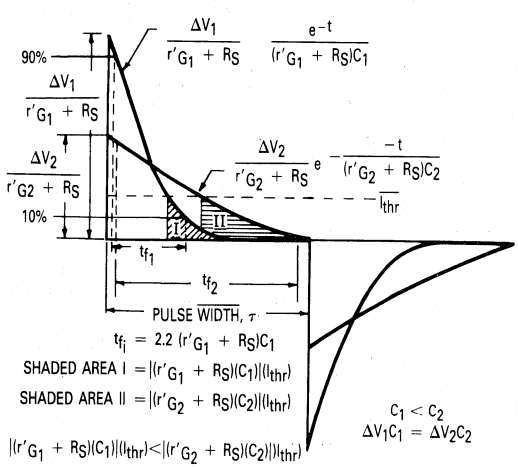


Figure 3.6. Gate Current Waveform in Capacitance Charge Triggering

Consider the gate current waveform in Figure 3.6; the triggering pulse width is made large enough such that $\tau >> t_{fi}$; the threshold trigger current is shown as I_{thr}. All of the charge supplied at a transient current level less than I_{thr} is lost by recombination, as shown in the shaded regions.

The gate spreading resistance (r'_G) of the gate junction varies inversely with peak current; the higher the peak current, the smaller the gate spreading resistance. Variation of gate spreading resistance measured by the method of Time Domain Reflectometry is plotted in Figure 3.8.

From the data of Figure 3.7, it is clear that for larger values of capacitance a lower voltage level is required for turn-on. The peak current of the spike in Figure 3.6 is given by $I_{pk} = \frac{\Delta V}{R_s + r'_G}$; the smaller ΔV, the smaller I_{pk}. Smaller I_{pk} in turn yields large r'_G, so that r'_G is dependent on the value of capacitance used in capacitance charge triggering. This reasoning is confirmed by measuring the fall time of the gate trigger voltage and calculating the transient gate spreading resistance, r'_G, from: $R_s + r'_G = \frac{t_f}{2.2\,C}$. Results are plotted in Figure 3.9. As expected, r'_G increases with increasing values of capacitance used. Referring back to Figure 3.6, for the same amount of charge ($C\,\Delta V$), the larger the $(R_s + r'_G)C$ time constant of the current spike, the more charge under the threshold level is lost in recombination. Increasing the value of C will increase the time constant more rapidly than if r'_G were invariant. Therefore, increasing the value of C should increase the charge lost as shown in Figure 3.7. Note that a two order of magnitude increase in capacitance increased the charge by less than 3:1.

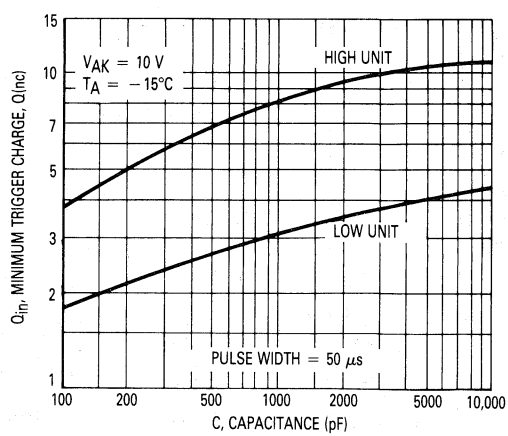


Figure 3.7. Variation of Trigger Charge versus Capacitance Used

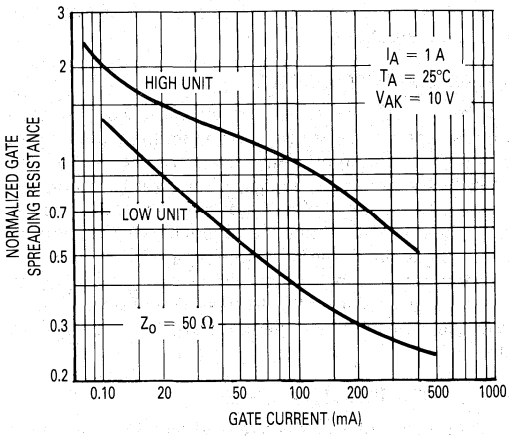


Figure 3.8. Variation of Gate Spreading Resistance versus Gate Peak Current

EFFECT OF TEMPERATURE

The higher the temperature, the less charge required to turn on the device, as shown in Figure 3.10. At the range of temperatures where the SCR is operated the life time of minority carriers increases with temperature; therefore less charge into the gate is lost in recombination.

As analyzed in Appendix II, there are three components of charge involved in gate triggering: (1) Q_r, charge lost in recombination, (2) Q_{dr}, charge drained out through the built-in gate-cathode shunt resistance, (3) Q_{tr}, net charge for triggering. All of them are temperature dependent. Since the temperature coefficient of voltage across a p-n junction is small, Q_{dr} may be considered invariant of temperature. At the temperature range of operation, the temperature is too low to give rise to significant impurity gettering, lifetime increases with temperature causing Q_r to decrease with increasing temperature. Also, Q_{tr} decreases with increasing temperature because at a constant current the α_T of the device in the blocking state increases with temperature;[7] in other words, to attain $\alpha_T = 1$ at an elevated temperature, less anode current, hence gate current [see equation (3) of Appendix I], is needed; therefore, Q_{tr} decreases. The input charge, being equal to the sum of Q_{tr}, Q_r, and Q_{dr}, decreases with increasing temperature.

The minimum current trigger charge decreases roughly exponentially with temperature. Actual data taken on an MCR729 deviate somewhat from exponential trend (Figure 3.10). At higher temperatures, the rate of decrease is less; also for different pulse widths the rates of decrease of Q_{in} are different; for large pulse widths the recombination charge becomes more significant than that of small pulse widths. As the result, it is expected and Figure 3.10 shows that Q_{in} decreases more rapidly with temperature at high pulse widths. These effects are analyzed in Appendix II [equation (10)]. The theory and experiment agree reasonably well.

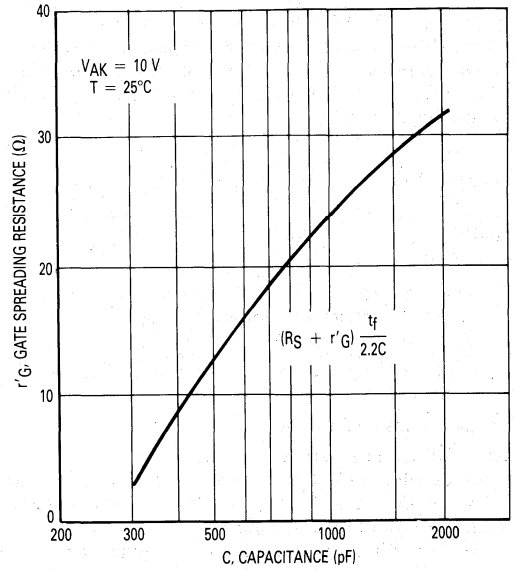


Figure 3.9. Variation of Transient Base Spreading Resistance versus Capacitance

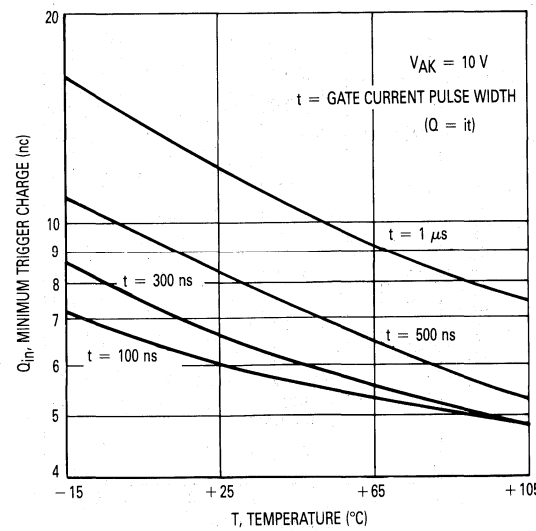


Figure 3.10. Variation of Q versus Temperature

EFFECT OF BLOCKING VOLTAGE

An SCR is an avalanche mode device; the turn-on of the device is due to multiplication of carriers in the middle collector junction. The multiplication factor is given by the empirical equation

$$M = \frac{1}{1 - (\frac{V}{V_B})^n} \qquad (6)$$

where

M ≡ Multiplication factor

V ≡ Voltage across the middle "collector" junction (voltage at which the device is blocking prior to turn-on)

V_B ≡ Breakdown voltage of the middle "collector" junction

n ≡ Some positive number

Note as V is increased, M also increases and in turn α increases (the current amplification factor $\alpha = \gamma\delta\beta M$ where γ ≡ Emitter efficiency, β ≡ Base transport factor, and δ ≡ Factor of recombination).

The larger the V, the larger is α_T. It would be expected for the minimum gate trigger charge to decrease with increasing V. Experimental results show this effect (see Figure 3.11). For the MCR729, the gate trigger charge is only slightly affected by the voltage at which the device is blocking prior to turn-on; this reflects that the exponent, n, in equation (6) is small.

EFFECT OF GATE CIRCUIT

As mentioned earlier, to turn on the device, the total amplification factor must be greater than unity. This means that if some current is being drained out of the gate which bleeds the regeneration current, turn-on will be affected. The higher the gate impedance, the less the gate trigger charge. Since the regenerative current prior to turn-on is small, the gate impedance only slightly affects the required minimum trigger charge; but in the case of over-driving the gate to achieve fast switching time, the gate circuit impedance will have noticeable effect.

EFFECT OF INDUCTIVE LOAD

The presence of an inductive load tends to slow down the change of anode current with time, thereby causing the required charge for triggering to increase with the value of inductance. For dc or long pulse width current triggering, the inductive load has little effect, but its effect increases markedly at short pulse widths, as shown in Figure 3.12. The increase in charge occurs because at short pulse widths, the trigger signal has decreased to a negligible value before the anode current has reached a level sufficient to sustain turn-on.

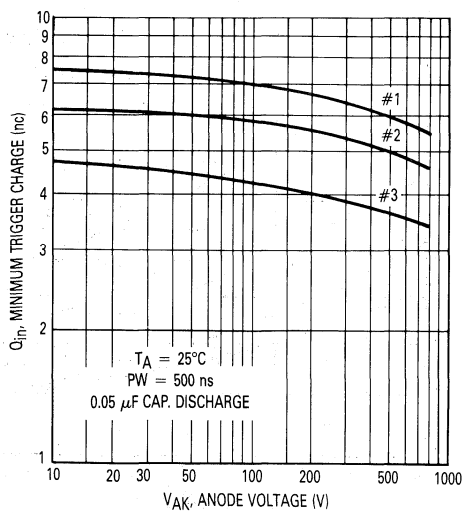


Figure 3.11. Variation of Current Trigger Charge versus Blocking Voltage Prior to Turn-On

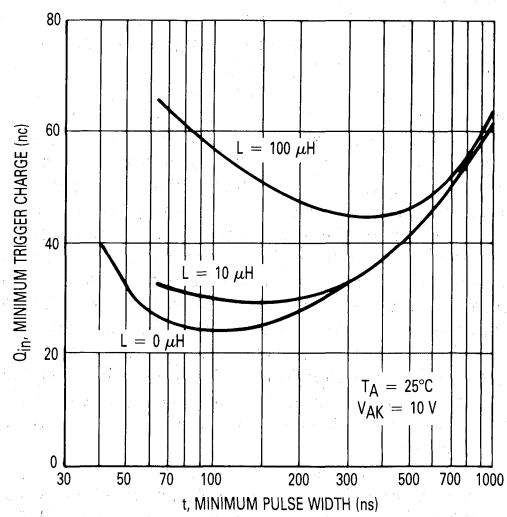


Figure 3.12. Effect of Inductance Load on Triggering Charge

USING NEGATIVE BIAS AND SHUNTING

Almost all SCR's exhibit some degree of turn-off gain. At normal values of anode current, negative gate current will not have sufficient effect upon the internal feedback loop of the device to cause any significant change in anode current. However, it does have a marked effect at low anode current levels; it can be put to advantage by using it to modify certain device parameters. Specifically, turn-off time may be reduced and hold current may be increased. Reduction of turn-off time and increase of hold current are useful in such circuits as inverters or in full-wave phase control circuits in which inductance is present.

Negative gate current may, of course, be produced by use of an external bias supply. It may also be produced by taking advantage of the fact that during conduction the gate is positive with respect to the cathode and providing an external conduction path such as a gate-to-cathode resistor. All Motorola SCR's, with the exception of sensitive gate devices, are constructed with a built in gate-to-cathode shunt, which produces the same effect as negative gate current. Further change in characteristics can be produced by use of an external shunt. Shunting does not produce as much of a change in characteristics as does negative bias, since the negative gate current, even with an external short circuit, is limited by the lateral resistance of the base layer. When using external negative bias the current must be limited, and care must be taken to avoid driving the gate into the avalanche region.

The effects of negative gate current are not shown on the device specification sheets. The curves in Figure 3.13 represent measurements made on a number of SCRs, and should therefore not be considered as spec limits. They do, however, show definite trends. For example, all of the SCRs showed an improvement in turn-off time of about one-third by using negative bias up to the point where no further significant improvement was obtained.

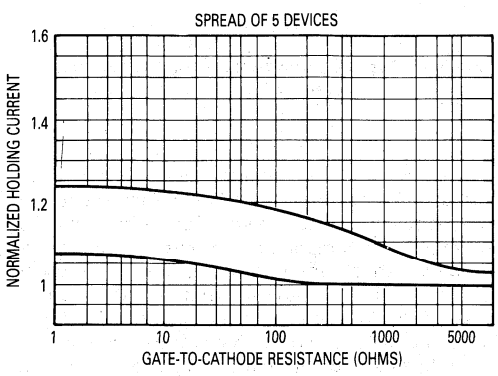


Figure 3.13(a). Normalized Holding Current versus Gate-to-Cathode Resistance

The increase in hold current by use of an external shunt resistor ranged typically between 5 and 75 percent, whereas with negative bias, the range of improvement ran typically between 2-1/2 and 7 times the open gate value. Note that the holding current curves are normalized and are referred to the open gate value.

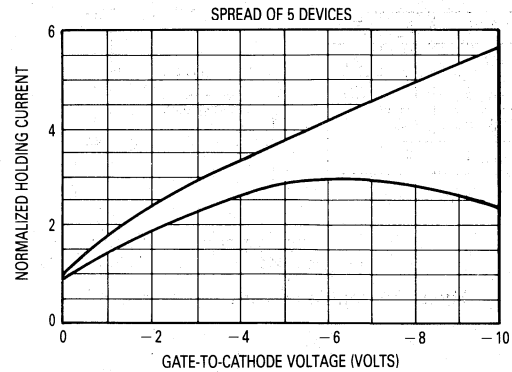


Figure 3.13(b). Normalized Holding Current versus Gate-to-Cathode Voltage

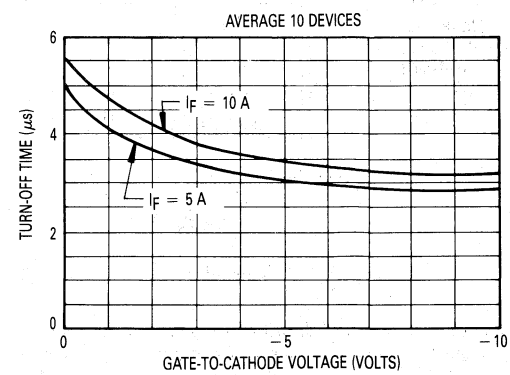


Figure 3.13(c). Turn-Off Time versus Bias

REDUCING di/dt — EFFECT FAILURES

Figure 3.14 shows a typical SCR structural cross section (not to scale). Note that the collector of transistor 1 and the base of transistor 2 are one and the same layer. This is also true for the collector of transistor 2 and the base of transistor 1. Although for optimum performance as an SCR the base thicknesses are great compared to a normal transistor, nevertheless, base thickness is still small compared to the lateral dimensions. When applying positive bias to the gate, the transverse base resistance, spreading resistance or r_b' will cause a lateral voltage drop which

will tend to forward bias those parts of the transistor 1 emitter-junction closest to the base contact (gate) more heavily, or sooner than the portions more remote from the contact area. Regenerative action, consequently will start in an area near the gate contact, and the SCR will turn on first in this area. Once on, conduction will propagate across the entire junction.

LAYER	T_2	T_1
NO. 4		(E)
NO. 3	(C)	(B)
NO. 2	(B)	(C)
NO. 1	(E)	

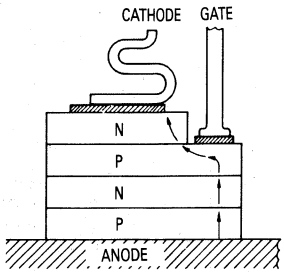


Figure 3.14(a). Construction of Typical SCR

TYPICAL SCR CONSTRUCTION SHOWING THE DIE IN PROPER SCALE . . .

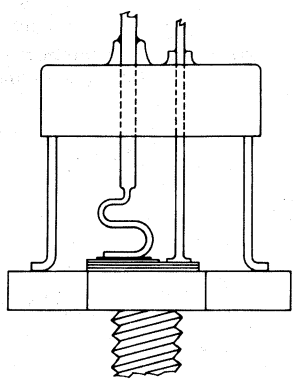


Figure 3.14 (b)

The phenomenon of di/dt failure is related to the turn-on mechanism. Let us look at some of the external factors involved and see how they contribute. Curve 3.15(a) shows the fall of anode-to-cathode voltage with time. This fall follows a delay time after the application of the gate bias. The delay time and fall time together are called turn-on time, and, depending upon the device, will take anywhere from tens of nanoseconds up to a few microseconds. The propagation of conduction across the entire junction requires a considerably longer time. The time required for propagation or equalization of conduction is represented approximately by the time required for the anode-to-cathode voltage to fall from the 10 percent point to its steady state value for the particular value of anode current under consideration (neglecting the change due to temperature effects). It is during the interval of time between the start of the fall of anode-to-cathode voltage and the final equalization of conduction that the SCR is most susceptible to damage from excessive current.

Let us superimpose a current curve (b) on the anode-to-cathode voltage versus time curve to better understand this. If we allow the current to rise rapidly to a high value we find by multiplying current and voltage that the instantaneous dissipation curve (c) reaches a peak which may be hundreds of times the steady state dissipation level for the same value of current.

At the same time it is important to remember that the dissipation does not take place in the entire junction, but is confined at this time to a small volume. Since temperature is related to energy per unit volume, and since the energy put into the device at high current levels may be very large while the volume in which it is concentrated is very small, very high spot temperatures may be achieved. Under such conditions, it is not difficult to attain temperatures which are sufficient to cause localized melting of the device.

Even if the peak energy levels are not high enough to be destructive on a single-shot basis, it must be realized that since the power dissipation is confined to a small area, the power handling capabilities of the device are lessened. For pulse service where a significant percentage of the power per pulse is dissipated during the fall-time interval, it is not acceptable to extrapolate the steady state power dissipation capability on a duty cycle basis to obtain the allowable peak pulse power.

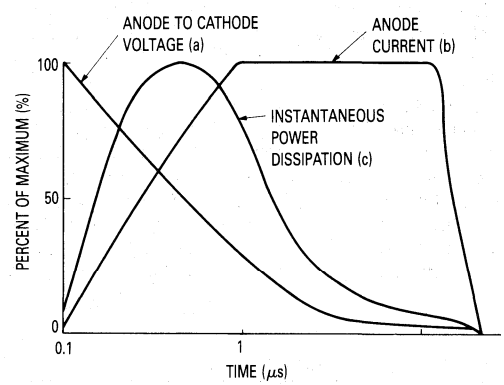


Figure 3.15. Typical Conditions — Fast-Rise, High Current Pulse

The final criterion for the limit of operation is junction temperature. For reliable operation the instantaneous junction temperature must always be kept below the maximum junction temperature as stated on the manufacturer's data sheet. Some SCR data sheets at present include information on how to determine the thermal response of the junction to current pulses. This information is not useful, however, for determining the limitations of the device before the entire junction is in conduction, because they are based on measurements made with the entire junction in conduction.

At present, there is no known technique for making a reasonably accurate measurement of junction temperature in the time domain of interest. Even if one were to devise a method for switching a sufficiently large current in a short enough time, one would still be faced with the problem of charge storage effects in the device under test masking the thermal effects. Because of these and other problems, it becomes necessary to determine the device limitations during the turn-on interval by destructive testing. The resultant information may be published in a form such as a maximum allowable current versus time, or simply as a maximum allowable rate of rise of anode current (di/dt).

Understanding the di/dt failure mechanism is part of the problem. To the user, however, a possible cure is infinitely more important. There are three approaches that should be considered.

Because of the lateral base resistance the portion of the gate closest to the gate contact is the first to be turned on because it is the first to be forward biased. If the minimum gate bias to cause turn-on of the device is used, the spot in which conduction is initiated will be smallest in size. By increasing the magnitude of the gate trigger pulse to several times the minimum required, and applying it with a very fast rise time, one may considerably increase the size of the spot in which conduction starts. Figure 3.16(a) illustrates the effect of gate drive on voltage fall time and Figure 3.16(b) shows the improvement in instantaneous dissipation. We may conclude from this that overdriving the gate will improve the di/dt capabilities of the device, and we may reduce the stress on the device by doing so.

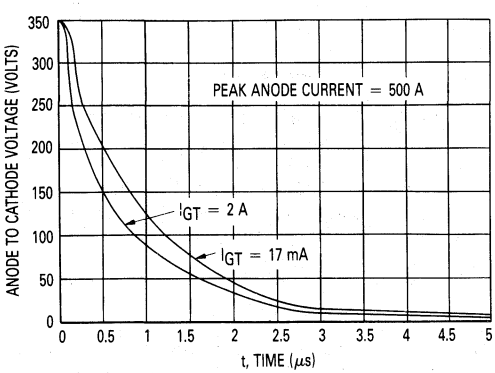


Figure 3.16(a). Effect of Gate Drive on Fall Time

A very straightforward approach is to simply slow down the rate of rise of anode current to insure that it stays within the device ratings. This may be done simply by adding some series inductance to the circuit.

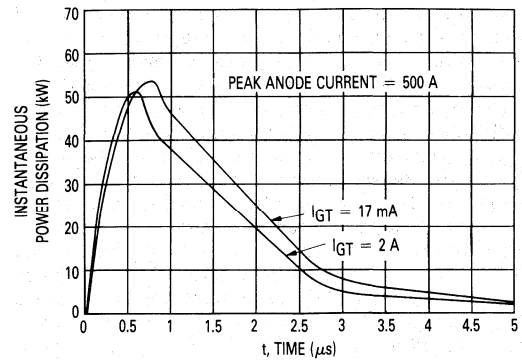


Figure 3.16(b). Effect of Gate Drive On Turn-On Dissipation

If the application should require a rate of current rise beyond the rated di/dt limit of the device, then another approach may be taken. The device may be turned on to a relatively low current level for a sufficient time for a large part of the junction to go into conduction; then the current level may be allowed to rise much more rapidly to very high levels. This might be accomplished by using a delay reactor as shown in Figure 3.17. Such a reactor would be wound on a square loop core so that it would have sharp saturation characteristic and allow a rapid current rise. It is also possible to make use of a separate saturation winding. Under these conditions, if the delay is long enough for the entire junction to go into conduction, the power handling capabilities of the device may be extrapolated on a duty cycle basis.

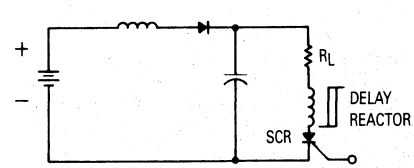


Figure 3.17. Typical Circuit Use of a Delay Reactor

WHY AND HOW TO SNUB THYRISTORS

Inductive loads (motors, solenoids, etc.) present a problem for the power triac because the current is not in phase with the voltage. An important fact to remember is that since a triac can conduct current in both directions, it has only a brief interval during which the sine wave current is passing through zero to recover and revert to its blocking state. For inductive loads, the phase shift between voltage and current means that at the time the current of the power handling triac falls below the holding current and the triac ceases to conduct, there exists a certain voltage which must appear across the triac. If this voltage appears too rapidly, the triac will resume con-

duction and control is lost. In order to achieve control with certain inductive loads, the rate of rise in voltage (dv/dt) must be limited by a series RC network placed in parallel with the power triac as shown in Figure 3.18. The capacitor C_S will limit the dv/dt across the triac.

The resistor R_S is necessary to limit the surge current from C_S when the triac conducts and to damp the ringing of the capacitance with the load inductance L_L. Such an RC network is commonly referred to as a "snubber."

Figure 3.19 shows current and voltage waveforms for the power triac. Commutating dv/dt for a resistive load is typically only 0.13 V/μs for a 240 V, 50 Hz line source and 0.063 V/μs for a 120 V, 60 Hz line source. For inductive loads the "turn-off" time and commutating dv/dt stress are more difficult to define and are affected by a number of variables such as back EMF of motors and the ratio of inductance to resistance (power factor). Although it may appear from the inductive load that the rate or rise is extremely fast, closer circuit evaluation reveals that the commutating dv/dt generated is restricted to some finite value which is a function of the load reactance L_L and the device capacitance C but still may exceed the triac's critical commutating dv/dt rating which is about 50 V/μs. It is generally good practice to use an RC snubber network across the triac to limit the rate of rise (dv/dt) to a value below the maximum allowable rating. This snubber network not only limits the voltage rise during commutation but also suppresses transient voltages that may occur as a result of ac line disturbances.

There are no easy methods for selecting the values for R_S and C_S of a snubber network. The circuit of Figure 3.18 is a damped, tuned circuit comprised of R_S, C_S, R_L and L_L, and to a minor extent the junction capacitance of the triac. When the triac ceases to conduct (this occurs every half cycle of the line voltage when the current falls below the holding current), the triac receives a step impulse of line voltage which depends on the power factor of the load. A given load fixes R_L and L_L; however, the circuit designer can vary R_S and C_S. Commutating dV/dt can be lowered by increasing C_S while R_S can be increased to decrease resonant "over ringing" of the tuned circuit.

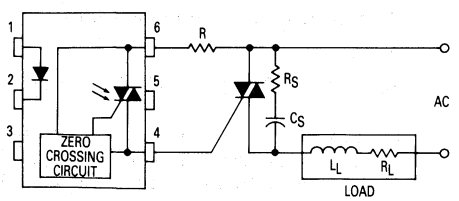


Figure 3.18. Triac Driving Circuit — with Snubber

BASIC CIRCUIT ANALYSIS

Figure 3.20 shows an equivalent circuit used for analysis, in which the triac has been replaced by an ideal switch. When the triac is in the blocking or non-conducting state, represented by the open switch, the circuit is a standard RLC series network driven by an ac voltage source. The following differential equation can be obtained by summing the voltage drops around the circuit:

$$(R_L + R_S)\, i(t) + L\frac{di(t)}{dt} + \frac{q_c(t)}{C_S} = V_M \sin(\omega t + \phi) \qquad (2)$$

in which i(t) is the instantaneous current after the switch opens, $q_c(t)$ is the instantaneous charge on the capacitor, V_M is the peak line voltage, and ϕ is the phase angle by which the voltage leads the current prior to opening of the switch. After differentiation and rearrangement, the equation becomes a standard second-order differential equation with constant coefficients.

With the imposition of the boundary conditions that $i(o) = 0$ and $q_c(o) = 0$ and with selected values for R_L, L, R_S and C_S, the equation can be solved, generally by the use of a computer. Having determined the magnitude

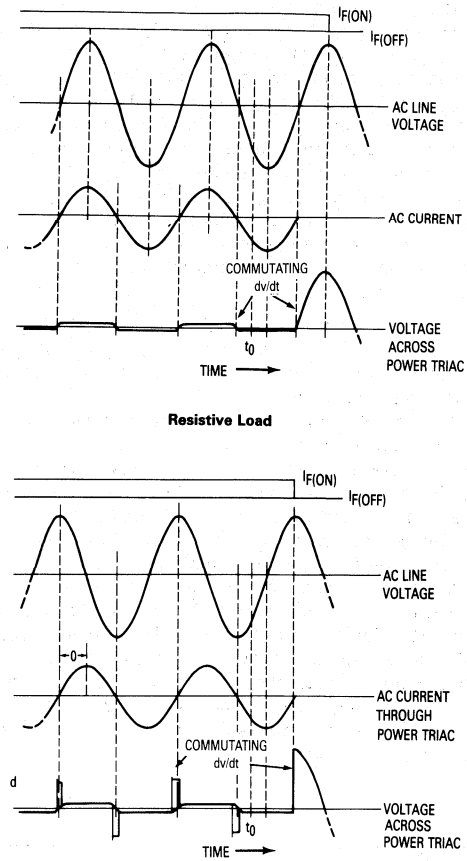


Figure 3.19. Current and Voltage Waveforms During Commutation

and time of occurrence of the peak voltage across the thyristor, it is then possible to calculate the values and times of the voltages at 10% and 63% of the peak value. This is necessary in order to compute the dv/dt stress as defined by the following equation:

$$\frac{dv}{dt} = \frac{V_2 - V_1}{t_2 - t_1}$$

where V_1 and t_1 are the voltage and time at the 10% point and V_2 and t_2 are the voltage and time at the 63% point.

Solution of the differential equation for assumed load conditions will give the circuit designer a starting point for selecting R_S and C_S.

Because the design of a snubber is contingent on the load, it is almost impossible to simulate and test every possible combination under actual operating conditions. It is advisable to measure the peak amplitude and rate of rise of voltage across the triac by use of an oscilloscope, then make the final selection of R_S and C_S experimentally. Additional comments about circuit values for SCRs and Triacs are made in Chapter 6.

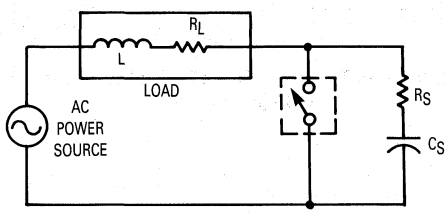


Figure 3.20. Equivalent Circuit used for Analysis

USING SENSITIVE GATE SCRs

In applications of sensitive gate SCRs such as the Motorola 2N6237, the gate-cathode resistor, R_{GK} (Figure 3.21) is an important factor. Its value affects, in varying degrees, such parameters as I_{GT}, V_{DRM}, dv/dt, I_H, leakage current, and noise immunity.

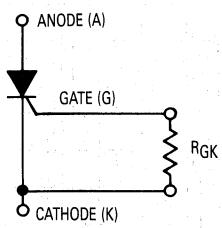


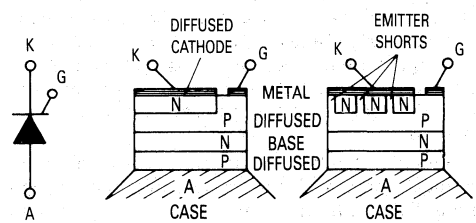
Figure 3.21. Gate-Cathode Resistor, R_{GK}

SCR CONSTRUCTION

The initial step in making an SCR is the creation, by diffusion, of P-type layers in N-type silicon base material. Prior to the advent of the all-diffused SCR, the next step was to form the gate-cathode P-N junction by alloying in a gold-antimony foil. This produced a silicon P-N junction of the regrown type over most of the junction area. However, a resistive rather than semiconductor junction would form where the molten alloy terminated at the surface. This formed an internal R_{GK}, looking in at the gate-cathode terminals, that reduced the "sensitivity" of the SCR.

Modern practice is to produce the gate-cathode junction by masking and diffusing, a much more controllable process. It produces a very clean junction over the entire junction area with no unwanted resistive paths. Good dv/dt performance by larger SCRs, however, requires resistive paths distributed over the junction area. These are diffused in as emitter shorts and naturally desensitize the device. Smaller SCRs may rely on an external R_{GK} for adequate dv/dt performance; an all-diffused SCR without emitter shorts will have sensitive gate characteristics. Figure 3.22(a) shows a cross-section of the simple construction found in small sensitive gate SCRs. Figure 3.22(b) depicts shorted emitter construction.

The sensitive gate SCR, therefore, is an all-diffused design with no emitter shorts. It has a very high impedance path in parallel with the gate-cathode P-N diode; the better the process is the higher this impedance, until a very good device cannot block voltage in the forward direction without an external R_{GK}. This is so simple because thermally generated leakage currents flowing from the anode into the gate junction are sufficient to turn on the SCR. The value for R_{GK} is usually one kilohm and its presence and value affects many other parameters.



(a). Simple Construction **(b). Shorted Emitter Construction**

Figure 3.22. Sensitive Gate SCR Construction

FORWARD BLOCKING VOLTAGE AND CURRENT, V_{DRM} AND I_{DRM}

The 2N6237 family is specified to have an I_{DRM}, or anode-to-cathode leakage current, of less than 200 μA at maximum operating junction temperature and rated V_{DRM}. This leakage current increases if R_{GK} is omitted and, in fact, the device may well be able to regenerate and turn on. Tests were run on several 2N6239 devices to establish the dependency of the leakage current on R_{GK} and to determine its relationship with junction temperature, T_J, and forward voltage V_{AK} (Figure 3.23a).

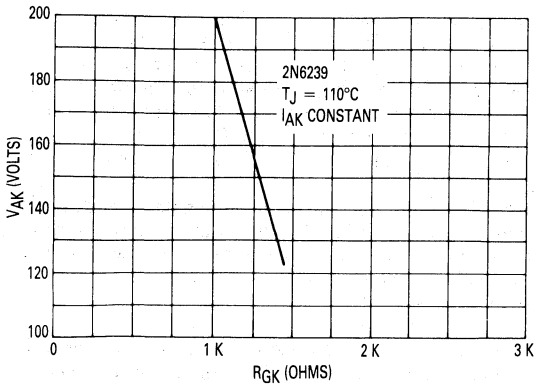


Figure 3.23(a). V_{AK} versus R_{GK} (Typical) for Constant Leakage Current

Figure 3.23(d) is a plot of V_{AK}, forward voltage, versus R_{GK} taken at the maximum rated operating junction temperature of 110°C. With each device the leakage current, I_{AK}, is set for a V_{AK} of 200 V, then V_{AK} reduced and R_{GK} varied to re-establish the same leakage current. The plot shows that the leakage current is *not* strongly voltage dependent or, conversely, R_{GK} may *not* be increased for derate.

While the leakage current is not voltage dependent, it is very temperature dependent. The plot is Figure 3.23(b) of T_J, junction temperature, versus R_{GK} taken at V_{DRM}, the maximum forward blocking voltage shows this dependence. For each device (2N6239 again) the leakage current, I_{AK}, was measured at the maximum operating junction temperature of 110°C, then the junction temperature was reduced and R_{GK} varied to re-establish that same leakage current. The plot shows that the leakage current is strongly dependent on junction temperature. Conversely R_{GK} may be increased for derated temperature.

To summarize, the leakage current in a sensitive gate SCR is much more temperature sensitive than voltage sensitive. Operation at lower junction temperatures allows an increase in the gate-cathode resistor which makes the SCR-resistor combination more "sensitive."

RATE-OF-RISE OF ANODE VOLTAGE, dv/dt

An SCR's junctions exhibit capacitance due to the separation of charge when the device is in a blocking state. If an SCR is subjected to forward dv/dt, this capacitance can couple sufficient current into the SCR's gate to turn it on, as shown in Figure 3.23(c). R_{GK} acts as a diversionary path for the dv/dt current. (In larger SCRs, where the lateral gate resistance of the device limits the influence of R_{GK}, this path is provided by the resistive emitter shorts mentioned previously.) The gate-cathode resistor, then, might be expected to have some effect on the dv/dt performance of the SCR. Using the 2N6241 as an example, a plot of dv/dt versus R_{GK}, in Figure 3.23(d), shows a very strong relationship indeed. If R_{GK} is increased, dv/dt performance suffers and if R_{GK} is reduced, the dv/dt performance improves. Reverse biasing the gate also improves dv/dt performance as might be expected and, on the 2N6241 the improvement is by a factor of about 50 when the gate is reverse biased by one volt.

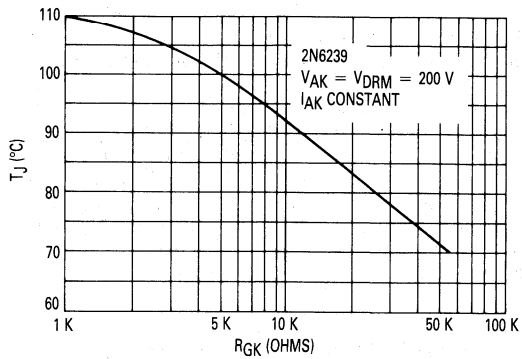


Figure 3.23(b). T_J versus R_{GK} (Typical) for Constant Leakage Current

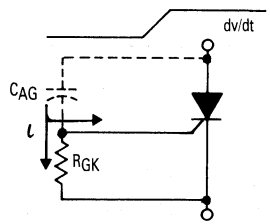


Figure 3.23(c). dv/dt Firing of an SCR

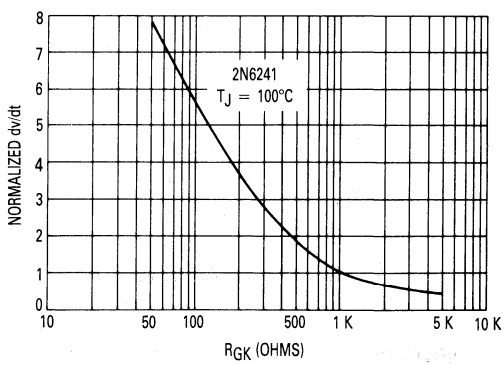


Figure 3.23(d). dv/dt versus R_{GK} (Typical)

GATE CURRENT, I_{GT}

The total gate current that a gating circuit must supply is the sum of the current that the device itself requires to fire and the current flowing to circuit ground through R_{GK}, as shown in Figure 3.24. I_{GT}, the current required by the device so that it may fire, is usually specified by the device manufacturer as a maximum at some temperature (for the 2N6237 series it is 500 μA maximum at −40°C). The current flowing through R_{GK} is defined by the resistor value and by the *gate-to-cathode voltage* that the SCR needs to fire. This is 1 V maximum at −40°C for the 2N6236 series, for example.

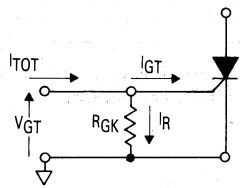


Figure 3.24. SCR and R_{GK} "Gate" Currents

The gate-cathode junction is a P-N silicon junction and thus has roughly the same temperature coefficient as a silicon diode, −2 mV/°C. Figure 3.25 is a typical plot of V_{GT} versus temperature for sensitive gate SCRs.

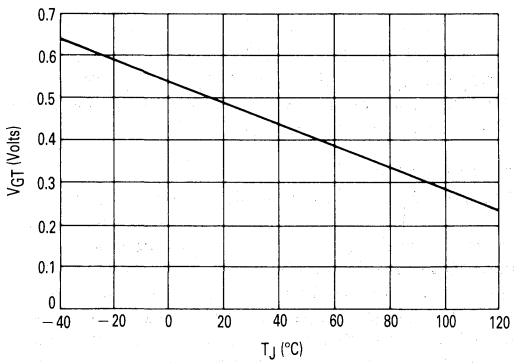


Figure 3.25. V_{GT} versus T_J (Typical)

GATE CURRENT, $I_{GT(min)}$

SCR manufacturers sometimes get requests for a sensitive-gate SCR specified with an $I_{GT(min)}$, that is, the maximum gate current that will *not* fire the device. This requirement conflicts with the basic function of a sensitive gate SCR, which is to fire at zero or very low gate current, $I_{GT(max)}$. Production of devices with a measurable $I_{GT(min)}$ is at best difficult and deliveries can be sporadic!

One reason for an $I_{GT(min)}$ requirement might be some measurable off-state gating circuit leakage current, perhaps the collector leakage of a driving transistor. Such current can readily be bypassed by a suitably chosen R_{GK}. The V_{GT} of the SCR at the temperature in question can be estimated from Figure 3.25, an Ohm's Law calculation made, and the resistor installed to define this "won't fire" current. This is a repeatable design well in the control of the equipment designer.

HOLDING CURRENT, I_H

The holding current of an SCR is the minimum anode current required to maintain the device in the on state. It is usually specified as a maximum for a series of devices (for instance, 5 mA maximum at 25°C for the 2N6237 series). A particular device will turn off somewhere between this maximum and zero anode current and there is perhaps a 20-to-1 spread in each lot of devices.

Figure 3.26 shows to holding current increasing with decreasing R_{GK} as the resistor siphones off more and more of the regeneratively produced gate current when the device is in the latched condition.

NOISE IMMUNITY

Changes in electromagnetic and electrostatic fields coupled into wires or printed circuit lines can trigger these sensitive devices, as can logic circuit glitches. The result is more serious than with a transistor since an SCR will latch on. Careful wire harness design (twisted pairs and adequate separation from high-power wiring) and printed circuit layout (gate and return runs adjacent to one another) can minimize potential problems. Another help is the gate-cathode resistor since, with a one-kilohm resistor, 100 μA to 1 mA of noise current is required to generate sufficient voltage to fire the device. To serve this purpose R_{GK} must be mounted right at the gate-cathode terminals of the SCR.

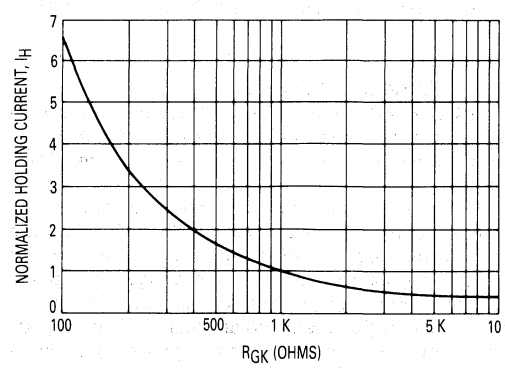
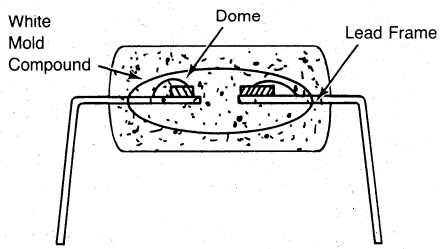


Figure 3.26. I_H versus R_{GK} (Typical)

DRIVERS: THE MOC3011 NON-ZERO CROSSING TRIAC DRIVER

The MOC3011 non-zero crossing triac driver consists of a gallium arsenide infrared LED optically exciting a silicon detector chip, which is especially designed to drive triacs controlling loads on a 115 Vac power line. The detector chip is a complex device which functions in much the same manner as a small triac, generating the signals necessary to drive the gate of a larger triac. The MOC3011 shows a low power exciting signal to drive a high power load with a very small number of components, and at the same time provides practically complete isolation of the driving circuitry from the power line.

The construction of the MOC3011 follows the same highly successful coupler technology used in Motorola's broad line of plastic couplers (Figure 3.27). The planar lead frame with a plastic "dome" undermold provides a stable dielectric capable of sustaining 7.5 kV between the input and output sides of the device. The detector chip is passivated with silicon nitride and uses Motorola's annular ring to maintain stable breakdown parameters.



Co-Planar Isolator Package

Figure 3.27. Motorola "Dome" Coupler Package

BASIC ELECTRICAL DESCRIPTION

The GaAs LED has nominal 1.3 V forward drop at 10 mA and a reverse breakdown voltage greater than 3 V. The maximum current to be passed through the LED is 50 mA.

The detector has a minimum blocking voltage of 250 Vdc in either direction in the off state. In the on state, the detector will pass 100 mA in either direction with less than 3 V drop across the device. Once triggered into the on (conducting) state, the detector will remain there until the current drops below the holding current (typically 100 μA) at which time the detector reverts to the off (non-conducting) state. The detector may be triggered into the on state by exceeding the forward blocking voltage, by voltage ramps across the detector at rates exceeding the static dv/dt rating, or by photons from the LED. The LED is guaranteed by the specifications to trigger the detector into the on state when 10 mA or more is passed through the LED. A similar device, the MOC3010, has exactly the same characteristics except it requires 15 mA to trigger.

Since the MOC3011 looks essentially like a small optically triggered triac, we have chosen to represent it as shown on Figure 3.28.

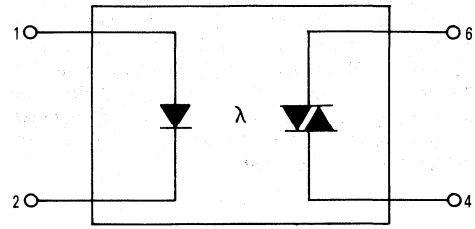


Figure 3.28. Schematic Representation of MOC3011 and MOC3010

TRIAC DRIVING REQUIREMENTS

Figure 3.29 shows a simple triac driving circuit using the MOC3011. The maximum surge current rating of the MOC3011 sets the minimum value of R1 through the equation:

$$R1(min) = V_{in(pk)}/1.2\ A$$

If we are operating on the 115 Vac nominal line voltage, $V_{in(pk)} = 180$ V, then

$$R1(min) = V_{in(pk)}/1.2\ A = 150 \text{ ohms.}$$

In practice, this would be a 150 or 180 ohm resistor. If the triac has $I_{GT} = 100$ mA and $V_{GT} = 2$ V, then the voltage V_{in} necessary to trigger the triac will be given by $V_{inT} = R1 \cdot I_{GT} + V_{GT} + V_{TM} = 20$ V.

RESISTIVE LOADS

When driving resistive loads, the circuit of Figure 3.29 may be used. Incandescent lamps and resistive heating elements are the two main classes of resistive loads for which 115 Vac is utilized. The main restriction is that the triac must be properly chosen to sustain the proper inrush loads. Incandescent lamps can sometimes draw a peak current known as "flashover" which can be extremely high, and the triac should be protected by a fuse or rated high enough to sustain this current.

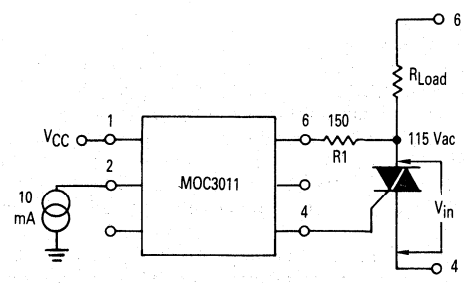


Figure 3.29. Simple Triac Gating Circuit

LINE TRANSIENTS — STATIC dv/dt

Ocassionally transient voltage disturbance on the ac line will exceed the static dv/dt rating of the MOC3011. In this case, it is possible that the MOC3011 and the associated triac will be triggered on. This is usually not a problem, except in unusually noisy environments, because the MOC3011 and its triac will commute off at the next zero crossing of the line voltage, and most loads are not noticeably affected by an occasional single half-cycle of applied power. See Figure 3.31 for typical dv/dt versus temperature curves.

INDUCTIVE LOADS — COMMUTATING dv/dt

Inductive loads (motors, solenoids, magnets, etc.) present a problem both for triacs and for the MOC3011 because the voltage and current are not in phase with each other. Since the triac turns off at zero current, it may be trying to turn off when the applied current is zero but the applied voltage is high. This appears to the traic like a sudden rise in applied voltage, which turns on the triac if the rate of rise exceeds the commutating dv/dt of the triac or the static dv/dt of the MOC3011.

SNUBBER NETWORKS

The solution to this problem is provided by the use of snubber networks to reduce the rate of voltage rise seen by the device. In some cases, this may require two snubbers — one for the triac and one for the MOC3011. The triac snubber is dependent upon the triac and load used. In many applications the snubber used for the MOC3011 will also adequately protect the triac.

In order to design a snubber properly, one should know the power factor of the reactive load, which is defined as the cosine of the phase shift caused by the load. Unfortunately, this is not always known, and this makes snubbing network design somewhat empirical. However a method of designing a snubber network may be defined, based upon a typical power factor. This can be used as a "first cut" and later modified based upon experiment.

Assume an inductive load with a power factor of PF = 0.1 is to be driven. The triac might be trying to turn off when the applied voltage is given by

$$V_{to} = V_{pk}\sin\phi \approx V_{pk} \approx 180\ V$$

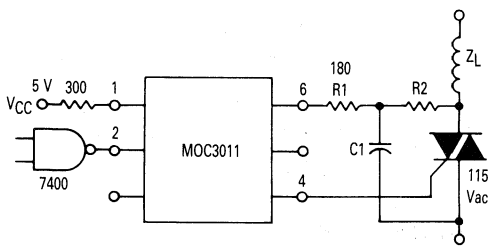
First, one must choose R1 (Figure 3.30) to limit the peak capacitor discharge current through the MOC3011. This resistor is given by

$$R1 = V_{pk}/I_{max} = 180/1.2\ A = 150\ \Omega$$

A standard value, 180 ohm resistor can be used in practice for R1.

It is necessary to set the time constant for $\tau = R_2C$. Assuming that the triac turns off very quickly, we have a peak rate of rise at the MOC3011 given by

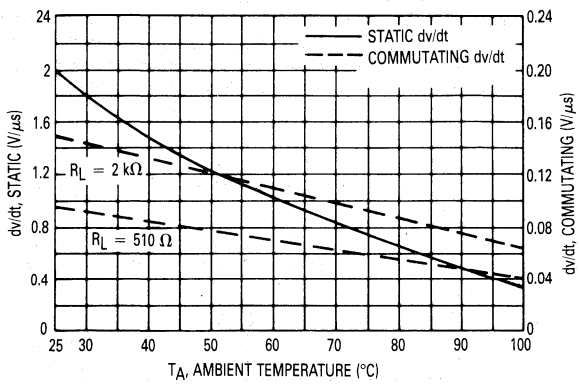
$$dv/dt = V_{to}/\tau = V_{to}/R_2C$$

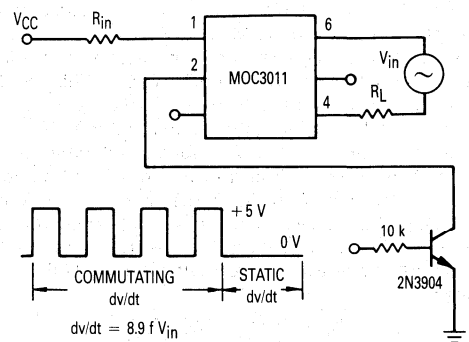


NOTE: CIRCUIT SUPPLIES 25 mA DRIVE TO GATE OF TRIAC AT V_{in} = 25 V AND $T_A \leq 70°C$.

TRIAC		
I_{GT}	R2	C
15 mA	2400	0.1
30 mA	1200	0.2
50 mA	800	0.3

Figure 3.30. Logic to Inductive Load Interface





1) For a more thorough discussion of snubbers, see page 1-3-9.

Figure 3.31. dv/dt versus Temperature

Setting this equal to the worst case dv/dt (static) for the MOC3011 which we can obtain from Figure 3.33 and solving for R_2C:

$$dv/dt\ (T_J = 70°C) = 0.8\ V/\mu s = 8 \times 10^3$$
$$R_2C = V_{to}/(dv/dt) = 180/(8 \times 10^5) \approx 230 \times 10^{-6}$$

The largest value of R2 available is found, taking into consideration of triac gate requirements. If a sensitive gate triac is used, such as 2N6071B, I_{GT} = 15 mA @ −40°C. If the triac is to be triggered when $V_{in} \leq 40$ V

$$(R1 + R2) \approx V_{in}/I_{GT} \approx 40/0.015 \approx 2.3\ k$$

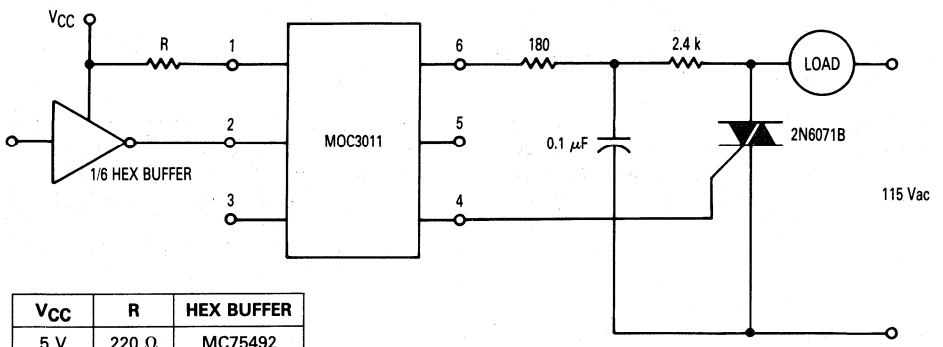
If we let R2 = 2400 ohms and C1 = 0.1 μF, the snubbing requirements are met. Triacs having less sensitive gates will require that R2 be lower and C1 be correspondingly higher as shown in Figure 3.30.

INCREASING INPUT SENSITIVITY

In some cases, the logic gate may not be able to source or sink 15 mA directly. CMOS, for example, is specified to have only 0.5 mA output, which must then be increased to drive the MOC3011. There are numerous ways to increase this current to a level compatible with the MOC3011 input requirements; an efficient way is to use athe MC14049B shown in Figure 3.32. Since there are six such buffers in a single package, the user can have a small package count when using several MOC3011's in one system.

INPUT PROTECTION CIRCUITS

In some applications, such as solid state relays, in which the input voltage varies widely, the designer may want to limit the current applied to the LED of the MOC3011. The circuit shown in Figure 3.33 allows a non-critical range of input voltages to properly drive the MOC3011 and at the same time protects the input LED from inadvertent application of reverse polarity.



V_{CC}	R	HEX BUFFER
5 V	220 Ω	MC75492
10 V	600 Ω	MC75492
15 V	910 Ω	MC14049B

Figure 3.32. MOS to ac Load Interface

INPUT CIRCUITRY

RESISTOR INPUT

When the input conditions are well controlled, as for example when driving the MOC3011 from a TTL, DTL, or HTL gate, only a single resistor is necessary to interface the gate to the input LED of the MOC3011. This resistor should be chosen to set the current into the LED to be a minimum of 10 mA but no more than 50 mA. 15 mA is a suitable value, which allows for considerable degradation of the LED over time, and assures a long operating life for the coupler. Currents higher than 15 mA do not improve performance and may hasten the aging process inherent in LED's. Assuming the forward drop to be 1.5 V at 15 mA allows a simple formula to calculate the input resistor.

$$R_i = (V_{CC} - 1.5)/0.015$$

Examples of resistive input circuits are seen in Figures 3.28 and 3.32.

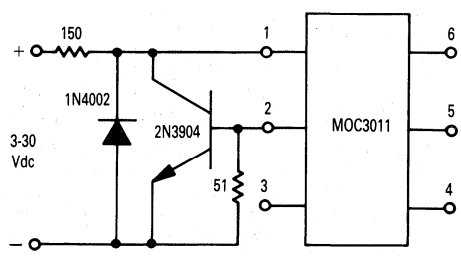


Figure 3.33. MOC3011 Input Protection Circuit

LED LIFETIME

All light emitting diodes slowly decrease in brightness during their useful life, an effect accelerated by high temperatures and high LED currents. To allow a safety margin and insure long service life, the MOC3011 is actually tested to trigger at a value lower than the specified 10 mA input threshold current. The designer can therefore design the input circuitry to supply 10 mA to the LED and still be sure of satisfactory operation over a long operating lifetime. On the other hand, care should be taken to insure that the maximum LED input current (50 mA) is not exceeded or the lifetime of the MOC3011 may be shortened.

TEST RESULTS ON THE MOC3011 AND MOC3021

Figure 3.34(a) shows the commutating dv/dt test circuit, and Figure 3.34(b) commutating dv/dt versus load resistance and ambient temperature for the non-zero crossing triac drivers.

The static dv/dt test circuit is shown in Figure 3.35(a). Static dv/dt versus ambient temperature and load resistance is graphed in Figures 3.35(a) and 3.35(b) for the MOC3011 and MOC3021, respectively. The static dv/dt is of primary importance when the triac driver is being used in conjunction with a triac, to control loads on the ac line.

Common-mode dv/dt is another significant parameter with non-zero crossing triac drivers. A common-mode spike on either the LED or detector could trigger the triac drivers. Thus, a large common-mode dv/dt rating is essential when using these devices in an electronically noisy environment. The test circuit used to measure this parameter is shown in Figure 3.36. Typical values exceed 5000 V/μs at 100°C, which is attained through the internal structure of the triac drivers.

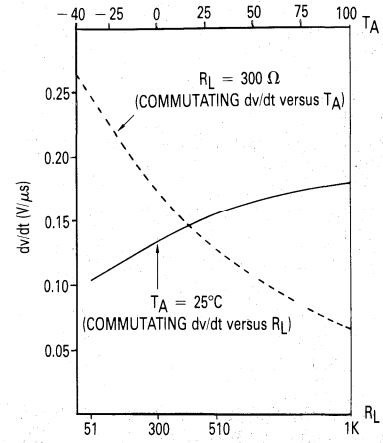
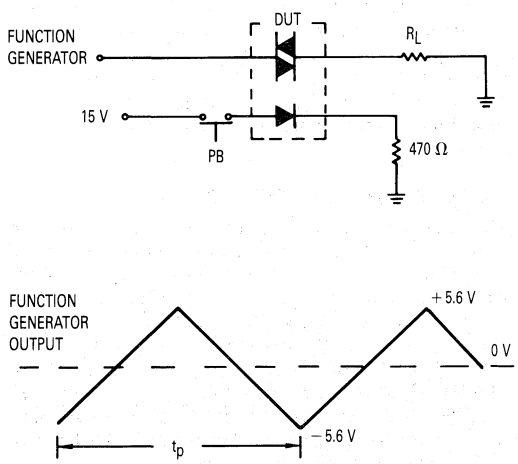


Figure 3.34(b). Commutating dv/dt versus Ambient Temperature and Load Resistance for the MOC3011 and MOC3021

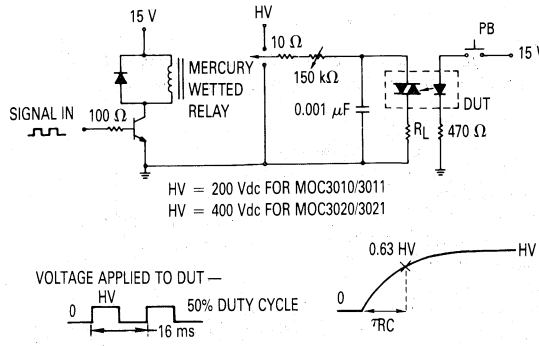


TEST PROCEDURE —

THE FREQUENCY OF THE TRAINGULAR WAVEFORM IS INCREASED UNTIL THE DUT REMAINS ON AFTER BEING TRIGGERED BY THE PUSHBUTTON. THE FREQUENCY IS THEN DECREASED UNTIL THE DUT TURNS OFF AND t_p IS MEASURED AT THIS POINT.

$$dv/dt = 22.4\ V/t_p$$

Figure 3.34(a). Commutating dv/dt Test Circuit



TEST PROCEDURE —

TURN THE DUT ONE WHILE APPLYING SUFFICIENT dv/dt TO ENSURE THAT IT REMAINS ON EVEN AFTER THE TRIGGER CURRENT IS REMOVED. THEN DECREASE dv/dt UNTIL THE DUT TURNS OFF. MEASURE τ_{RC}, THE TIME IT TAKES TO RISE TO 0.63 HV, AND DIVIDE 0.63 HV BE τ_{RC} TO GET dv/dt.

Figure 3.35(a). Static dv/dt Test Circuit

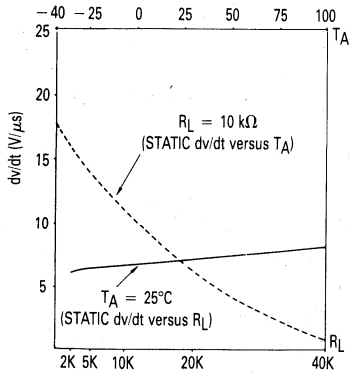


Figure 3.35(b). Static dv/dt versus Ambient Temperature and Load Resistance for the MOC3011

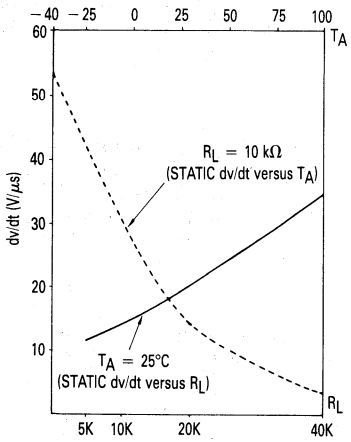
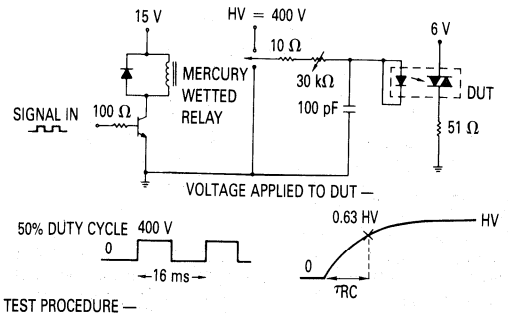


Figure 3.35(c). Static dv/dt versus Ambient Temperature and Load Resistance for the MOC3021



TEST PROCEDURE —
APPLY SUFFICIENT COMMON-MODE dv/dt TO TRIGGER THE DUT. MEASURE τ_{RC}, THE TIME IT TAKES TO RISE TO 0.63 HV AND DIVIDE 0.63 HV BY τ_{RC} TO GET dv/dt.

Figure 3.36. Common-Mode dv/dt Test Circuit

DRIVERS: THE MOC3061 AND MOC3081 ZERO-CROSSING TRIAC DRIVERS

Many new applications in the power control field were made possible by Motorola's introduction of the MOC3011 and MOC3021 optically isolated triac drivers. These six-pin, solid-state devices, with an input current of as little as 15 mA, can supply up to 100 mA drive current to switch triacs on either a 115 Vac or 230 Vac power line. Thus, high power ac loads can be controlled from low power circuitry, while an isolation voltage of 7.5 kV is maintained from input to output.

Increased flexibility in power control is possible with the MOC3061 and MOC3081 optically isolated, zero-crossing triac drivers. These new devices offer a unique zero-crossing feature, guaranteeing the triac driver will only 'switch on' between −25 V and +25 V. This insures lower generated noise and inrush currents and extends the life of incandescent lamp filaments. The circuitry necessary for zero-crossing is also responsible for improved device parameters, allowing the use of the MOC3061 and MOC3081 in a wide range of applications.

Zero-crossing triac drivers are currently being used in electrically noisy industrial environments, where control signals may travel several hundred yards before reaching the MOC3061 or MOC3081 triac driver. This is due to the improved common-mode noise immunity of these devices. Their large static dv/dt rating eliminates the need for snubber networks and guarantees that the zero-crossing triac drivers cannot be unintentionally triggered 'on,' by ac line noise. This is particularly important when the triac driver is used to control industrial equipment where inadvertent operation may cause damage or injury.

ELECTRICAL CHARACTERISTICS

The GaAs LED has a maximum forward voltage drop of 1.5 V at 30 mA and a reverse breakdown voltage of 3 V or more. The recommended LED current to trigger the detector (either the MOC3061 or MOC3081) is 15 mA. The minimum LED pulse width capable of triggering the detector is typically 10 microseconds at both 15 mA for the MOC3061 and MOC3081.

The MOC3061 detector has a minimum blocking voltage of 600 Vdc in either direction in the 'off' state whereas the minimum blocking voltage of the MOC3081 is 800 Vdc. Once triggered 'on,' either detector will pass 100 mA with less than 3 V drop across the device. Both will remain in this conducting state until the current drops below the holding current (typically 100 µA for the MOC3061 and the MOC3081) at which time the detector returns to the non-conducting state.

It may also be possible to inadvertently trigger the detector into the 'on' state by exceeding the forward blocking voltage or by voltage ramps across the detector exceeding the static dv/dt rating. These voltage ramps usually result from disturbances on the ac power line. However, since the typical static dv/dt rating of the zero-crossing triac driver is in excess of 1000 V/microsecond

at 100°C, ac line noise poses little or no problem for these devices. As a result, the need for snubber networks on the triac drivers, to reduce the speed of voltage ramps, is eliminated. Figure 3.35 shows the static dv/dt test circuit.

Once the triac controlled by a triac driver is switched to the conducting state, very little voltage appears across MT1 and MT2, the main terminals of the triac (Figure 3.37). Since there is also a small voltage drop from the gate to T2, the total voltage across the triac driver terminals T1 and T2 is very small — typically less than 2 V. Thus, the detector of the triac driver will conduct very little current and will, for practical purposes, be 'off.' This condition has very important implications when a triac driver and triac are being used to control an inductive load.

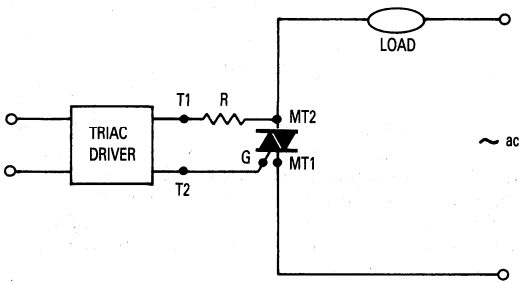


Figure 3.37. Typical ac Power Line Control Circuit

Commutating dv/dt, a phenomena associated with inductive loads, presents a problem for both triacs and zero-crossing triac drivers. In the steady state, voltage and current are 90 degrees out of phase for an inductive load. Therefore, when a triac attempts to turn 'off' at zero current, the applied voltage is high. This sudden rise in voltage will trigger the triac back 'on' if it exceeds the commutating dv/dt of the triac, since it has been on; or, the static dv/dt of the triac driver, since it has been basically off. Because the static dv/dt of the zero-crossing triac driver is so large, the commutating dv/dt of the triac will be the limiting factor. Therefore, commutating dv/dt only becomes a limiting factor when the zero-crossing triac drivers are being used in direct drive applications. Figure 3.38 shows commutating dv/dt versus ambient temperature and load resistance for the zero-crossing triac drivers.

Another important parameter associated with the zero-crossing triac drivers, is common-mode dv/dt. A large common-mode dv/dt rating is essential when using these devices in a noisy environment, due to the possibility of triggering the triac drivers 'on' with a common-mode spike on either the LED or detector. The test circuit used to measure this quantity is shown in Figure 3.36. Typical values exceed 5000 V/μs at 100°C. This value is attained through the inhibiting action of the internal circuitry of the zero-crossing triac drivers.

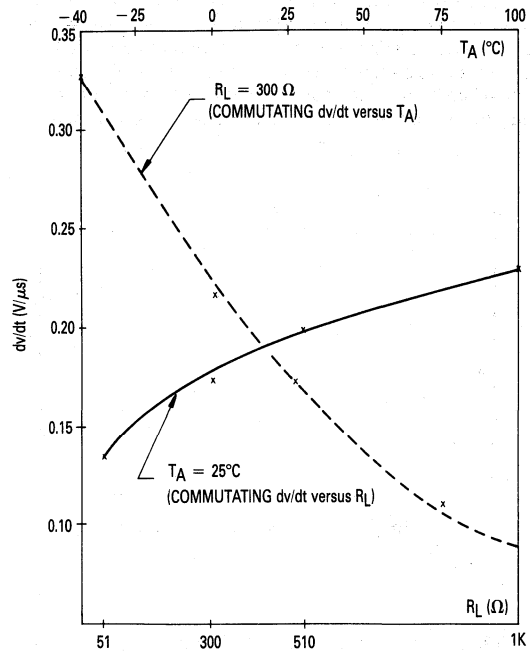


Figure 3.38. Commutating dv/dt versus Ambient Temperature and Load Resistance

ZERO CROSSING VERSUS NON-ZERO CROSSING TRIAC DRIVERS

Several of the benefits made possible by the zero-crossing circuitry have already been discussed. The large common-mode dv/dt makes these devices insensitive to common-mode EMI. In the same manner, the typical static dv/dt rating of 1000 V/microsecond at 100°C virtually eliminates the possibility of the zero-crossing triac drivers being unintentionally triggered on by ac line noise. This also eliminates the need for snubber networks on the triac drivers, along with their associated component cost and printed circuit board space.

In addition to the improvement in device parameters caused by the zero-crossing circuitry, the basic concept of switching the load 'on' only near zero voltage has many benefits.

There has been some controversy as to whether a peak-crossing triac driver might be more appropriate than a zero-crossing triac driver when controlling inductive loads. This is because the voltage and current are 90

degrees out of phase in an inductive load. At first glance it would seem logical to switch an inductive load 'on' at peak voltage to minimize the inrush current. However, voltage and current are only 90 degrees out of phase in an inductive load in the steady state. When switching an inductive load 'on,' an initial transient is also present. This transient must be added to the steady state response to fully characterize current flow in an inductive load. To illustrate this behavior, analyze the circuit of Figure 3.39, which can be used to model the response of an inductive load to the sudden application of a sinusoidal voltage.

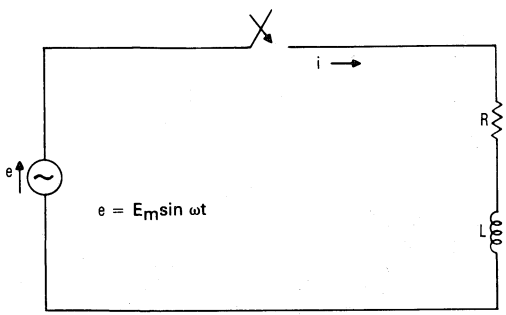


Figure 3.39. Circuit Model for Inductive Load Response

The switch is closed at t = 0. The applied voltage will be zero at this instant, analogous to the use of a zero-crossing triac driver. A small series resistance (R) has been included, since all inductors have some finite resistance.

The governing differential equation is

$$e = iR + L\frac{di}{dt} \tag{1}$$

with the initial condition $i(0-) = i(0+) = 0$, due to the inductor and no current flowing in the circuit with the switch open.

The total solution to this differential equation is the sum of a transient and steady state response

$$i_{TOT} = i_{TRAN} + i_{SS} \tag{2}$$

where

$$i_{TRAN} = A_1 e - Rt/L \tag{3}$$

and

$$i_{SS} = A_2 \sin \omega t + A_3 \cos \omega t \tag{4}$$

Substituting these solutions into the governing differential equation and satisfying the initial condition gives

$$i_{TOT} = \overbrace{\frac{E_m L\omega}{R^2+L^2\omega^2} e^{-Rt/L}}^{i_{TRAN}} + \tag{5}$$

$$\overbrace{\frac{E_m R}{R^2+L^2\omega^2} \sin \omega t - \frac{E_m L\omega}{R^2+L^2\omega^2} \cos \omega t}^{i_{SS}}$$

This solution also gives insight into another important parameter associated with triacs — the di/dt rating. If the maximum di/dt rating of a triac is exceeded, the triac's characteristics can be adversely affected or the triac permanently damaged. In the case of the zero-crossing triac driver, the di/dt applied to the triac is easily calculated from the previous solution.

$$\frac{di_{TOT}}{dt} = \frac{-R}{L}\frac{E_m L\omega}{R^2+L^2\omega^2} e^{-Rt/L} + \frac{E_m R\omega}{R^2+L^2\omega^2} \cos \omega t$$

$$+ \frac{E_m L\omega^2}{R^2+L^2\omega^2} \sin \omega t \tag{6}$$

$$\left.\frac{di_{TOT}}{dt}\right|_{t=0} = 0$$

Thus, there is little possibility of damage to the triac when using a zero-crossing triac driver. If a peak-crossing triac driver were used, the voltage source shown in Figure 3.39 would become $e = E_m \cos \omega t$ to model the peak voltage impressed across the load at t = 0. The initial condition of $i(0-) = i(0+) = 0$ remains the same and the solution becomes

$$i_{TOT} = \overbrace{\frac{-E_m R}{R^2+L^2\omega^2} e^{-Rt/L}}^{i_{TRAN}} +$$

$$\overbrace{\frac{E_m L\omega}{R^2+L^2\omega^2} \sin \omega t + \frac{E_m R}{R^2+L^2\omega^2} \cos \omega t}^{i_{SS}} \tag{7}$$

With a peak-crossing triac driver

$$\frac{di_{TOT}}{dt} = \frac{R}{L}\frac{E_m R}{R^2+L^2\omega^2} e^{-Rt/L}$$

$$+ \frac{E_m L\omega^2}{R^2+L^2\omega^2} \cos \omega t - \frac{E_m R\omega}{R^2+L^2\omega^2} \sin \omega t \tag{8}$$

$$\left.\frac{di_{TOT}}{dt}\right|_{t=0} = \frac{E_m}{R^2+L^2\omega^2}\left[\frac{R^2}{L} + L\omega^2\right] = \frac{E_m}{L} \tag{9}$$

This value can become very large depending upon the peak applied voltage and inductance of the load. If it exceeds the maximum di/dt rating of the triac, serious damage can result.

It is easy to conclude that, when controlling inductive loads using a triac driver and triac, a zero-crossing triac driver is a far more desirable device than a peak-crossing triac driver.

DRIVERS: PROGRAMMABLE UNIJUNCTION TRANSISTORS

The programmable unijunction transistor (PUT) is a four layer device similar to an SCR except that the anode gate rather than the cathode gate is brought out. It is normally used in conventional unijunction transistor (UJT) circuits. The characteristics of both devices are similar, but the triggering voltage of the PUT is programmable and can be set by an external resistive voltage divider network. The PUT is faster and more sensitive than the UJT. It finds limited application as a phase control element and is most often used in long duration timer circuits. In general, the PUT is more versatile and is a more economical device than the UJT and will replace it in many applications.

OPERATION OF THE PUT

The PUT has three terminals, an anode (A), gate (G), and cathode (K). The symbol and a transistor equivalent circuit are shown in Figure 3.40. As can be seen from the equivalent circuit, the device is actually an anode-gated SCR. This means that if the gate is made negative with respect to the anode, the device will switch from a blocking state to its on state.

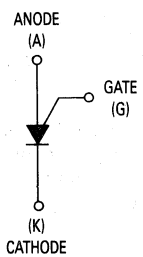


Figure 3.40(a). PUT Symbol

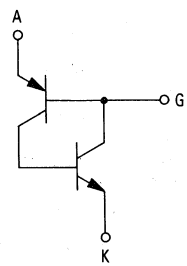


Figure 3.40(b). Transistor Equivalent

Since the PUT is normally used as a unijunction transistor (UJT), the UJT terminology is used to describe its parameters. In order to operate this device as a UJT, an external reference voltage must be maintained at the gate terminal. A typical relaxation type oscillator circuit is shown in Figure 3.41(a). The voltage divider shown is a typical way of obtaining the gate reference. In this circuit, the characteristic curve looking into the anode-cathode terminals would appear as shown in Figure 3.41(b). The peak and valley points are stable operating points at either end of a negative resistance region. The peak point voltage (V_P) is essentially the same as the external gate reference, the only difference being the gate diode drop. Since the reference is circuit and not device dependent, it may be varied, and in this way, V_P is programmable. This feature is the most significant difference between the UJT and the PUT.

In characterizing the PUT, it is convenient to speak of the Thevenin equivalent circuit for the external gate voltage (V_S) and the equivalent gate resistance (R_G). The parameters are defined in terms of the divider resistors (R1 and R2) and supply voltage as follows:

$$V_S = R1\ V1/(R1 + R2)$$
$$R_G = R1\ R2/(R1 + R2)$$

Most device parameters are sensitive to changes in V_S and R_G. For example, decreasing R_G will cause peak and valley currents to increase. This is easy to see since R_G actually shunts the device and will cause its sensitivity to decrease.

CHARACTERISTICS OF THE PUT

Table 3.I is a list of typical characteristics of Motorola's MPU131 series of programmable unijunction transistors. The test circuits and test conditions shown are essentially the same as for the data sheet characteristics. The data presented here defines the static curve shown in Figure 3.41(b) for a 10 V gate reference (V_S) with various gate resistances (R_G). It also indicates the leakage currents of these devices and describes the output pulse. Values given are for 25°C unless otherwise noted.

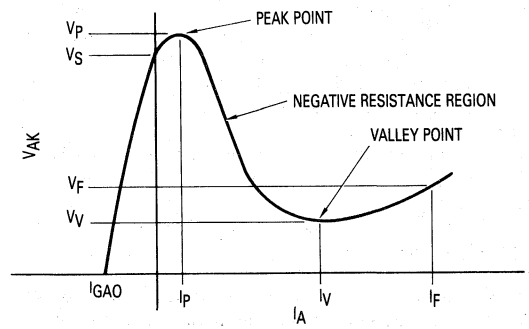


Figure 3.41(a). Static Characteristics

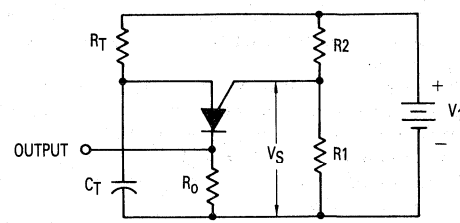


Figure 3.41(b). Typical Oscillator Circuit

Table 3.I. Typical PUT Characteristics

Symbol	Test Circuit Figure	Test Conditions	MPU131	MPU132	MPU133	Unit
I_P	3.64	R_G = 1 mΩ R_G = 10 kΩ	1.25 4	0.19 1.20	0.08 0.70	μA μA
I_V	3.64	R_G = 1 MΩ R_G = 10 kΩ	18 270	18 270	18 270	μA μA
V_{AG}		(See Figure 3.46)				
I_{GAO}		V_S = 40 V	See Figure 3.47			
I_{GKS}		V_S = 40 V	5	5	5	nA
V_F	Curve Tracer Used	I_F = 50 mA	0.8	0.8	0.8	V
V_O	3.67		16	16	16	V
t_r	3.68		40	40	40	ns

1

PEAK POINT CURRENT, (I_P)

The peak point is indicated graphically by the static curve. Reverse anode current flows with anode voltages less than the gate voltage (V_S) because of leakage from the bias network to the charging network. With currents less than I_P, the device is in a blocking state. With currents above I_P, the device goes through a negative resistance region to its on state.

The charging current, or the current through a timing resistor, must be greater than I_P at V_P to insure that a device will switch from a blocking to an on state in an oscillator circuit. For this reason, maximum values of I_P are given on the data sheet. These values are dependent on V_S temperature, and R_G. Typical curves on the data sheet indicate this dependence and must be consulted for most applications.

The test circuit in Figure 3.42 is a sawtooth oscillator which uses a 0.01 μF timing capacitor, a 20 V supply, an adjustable charging current, and equal biasing resistors (R). The two biasing resistors were chosen to give an equivalent R_G of 1 MΩ and 10 kΩ. The peak point current was measured with the device off just prior to oscillation as detected by the absence of an output voltage pulse. The 2N5270 held effect transistor circuit is used as a current source. A variable gate voltage supply was used to control this current.

VALLEY POINT CURRENT, (I_V)

The valley point is indicated graphically in Figure 3.40. With currents slightly less than I_V, the device is in an unstable negative resistance state. A voltage minimum occurs at I_V and with higher currents, the device is in a stable on state.

When the device is used as an oscillator, the charging current or the current through a timing resistor must be less than I_V at the valley point voltage (V_V). For this reason, minimum values for I_V are given on the data sheet for R_G = 10 kΩ. With R_G = 1 MΩ, a reasonable "low" is 2 μA for all devices.

When the device is used as an SCR in the latching mode, the anode current must be greater than I_V. Maximum values for I_V are given with R_G = 1 MΩ. All devices have a reasonable "high" of 400 μA I_V with R_G = 10 kΩ.

LATCHING AND HOLDING CURRENT, (I_L, I_H)

Using the test circuit in Figure 3.42, an attempt was made to differentiate between latching current (I_L), holding current (I_H), and valley current. With the device latched on, reducing the current causes a voltage minimum which is the valley point. The device does remain on at lower currents until holding current (I_H) is reached. The holding current is measured as detected by the absence of an output voltage pulse just before oscillation occurs. The latching current is generally higher than I_H and is measured by increasing the current while the device is oscillating and recording the value at which oscillations stop. With the PUT, there was no measurable difference between these parameters. This is not necessarily true with a unijunction transistor.

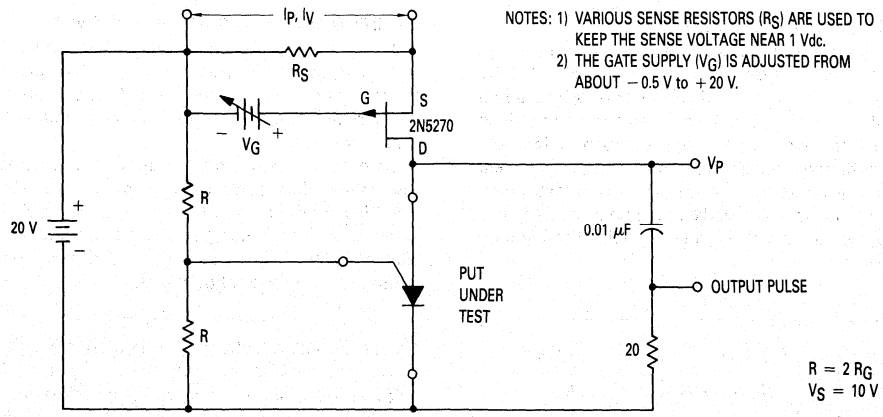
The valley current does vary with circuit parameters and temperature as was true of I_P. Typical data sheet curves identify this dependence and are frequently used to approximate actual variations of I_V.

PEAK POINT VOLTAGE, (V_P)

The unique feature of the PUT is that the peak point voltage can be determined externally. This programmable feature gives this device the ability to function in voltage controlled oscillators or similar applications. The triggering or peak point voltage is approximated by

$$V_P \approx V_T + V_S,$$

where V_S is the unloaded divider voltage and V_T is the offset voltage. The actual offset voltage will always be higher than the anode-gate voltage V_{AG}, because I_P flows out of the gate just prior to triggering. This makes $V_T = V_{AG} + I_P R_G$. A change in R_G will affect both V_{AG} and $I_P R_G$ but in opposite ways. First, as R_G increases, I_P decreases and causes V_{AG} to decrease. Second, since I_P does not decrease as fast as R_G increases, the $I_P R_G$ product will increase and the actual V_T will increase. These second order effects are difficult to predict and measure. Allowing V_T to be 0.5 V as a first order approximation gives sufficiently accurate results for most applications.



NOTES: 1) VARIOUS SENSE RESISTORS (R_S) ARE USED TO KEEP THE SENSE VOLTAGE NEAR 1 Vdc.
2) THE GATE SUPPLY (V_G) IS ADJUSTED FROM ABOUT −0.5 V to +20 V.

Figure 3.42. Test Circuit for I_P, V_P and I_V

The peak point voltage was tested using the circuit in Figure 3.42 and a scope with 10 MΩ input impedance across the PUT. A Tektronix, Type W plug-in was used to determine this parameter.

FORWARD ANODE-GATE VOLTAGE, (V_{AG})

The forward anode-to-gate voltage drop affects the peak point voltage as was previously discussed. The drop is essentially the same as a small signal silicon diode and is plotted in Figure 3.43. The voltage decreases as current decreases, and the change in voltage with temperature is greater at low currents. At 10 nA the temperature coefficient is about −2.4 V/°C and it drops to about −1.6 mV/°C at 10 mA. This information is useful in applications where it is desirable to temperature compensate the effect of this diode.

GATE-CATHODE LEAKAGE CURRENT, (I_{GKS})

The gate-to-cathode leakage current is the current that flows from the gate to the cathode with the anode shorted to the cathode. It is actually the sum of the open circuit gate-anode and gate-cathode leakage currents. The shorted leakage represents current that is shunted away from the voltage divider.

GATE-ANODE LEAKAGE CURRENT, (I_{GAO})

The gate-to-anode leakage current is the current that flows from the gate to the anode with the cathode open. It is important in long duration timers since it adds to the charging current flowing into the timing capacitor. The typical leakage currents measured at 40 V are shown in Figure 3.44. Leakage at 25°C is approximately 1 nA and the current appears to double for about every 10°C rise in temperature.

FORWARD VOLTAGE, (V_F)

The forward voltage (V_F) is the voltage drop between the anode and cathode when the device is biased on. It is the sum of an offset voltage and the drop across some internal dynamic impedance which both tend to reduce the output pulse. The typical data sheet curve shows this impedance to be less than 1 ohm for up to 2 A of forward current. This is essentially an order of magnitude better than the UJT which is closer to 20 ohms.

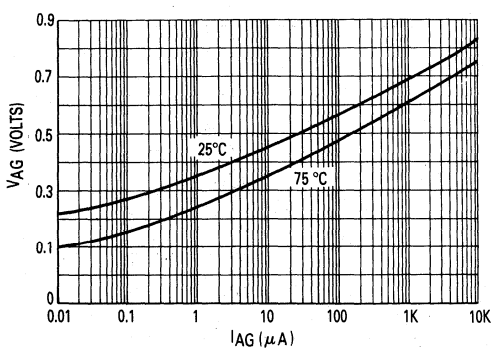


Figure 3.43. Voltage Drop of MPU131 Series

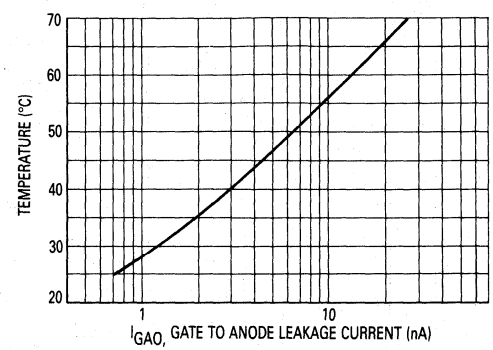


Figure 3.44. Typical Leakage Current of the MPU131, 132 and 133 Reverse Voltage Equals 40 V

PEAK OUTPUT VOLTAGE, (V_O)

The peak output voltage is not only a function of V_P, V_F and dynamic impedance, but is also affected by switching speed. This is particularly true when small capacitors (less than 0.01 μF) are used for timing since they lose part of their charge during the turn on interval. The use of a relatively large capacitor (0.2 μF) in the test circuit of Figure 3.45 tends to minimize this last effect. The output voltage is measured by placing a scope across the 20 ohm resistor which is in series with the cathode lead.

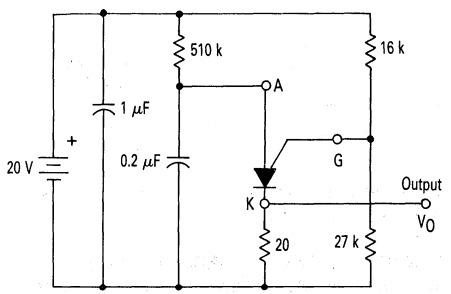


Figure 3.45. PUT Test Circuit for Peak Output Voltage (V_O)

RISE TIME, (t_r)

Rise time is a useful parameter in pulse circuits that use capacitive coupling. It can be used to predict the amount of current that will flow between these circuits. Rise time is specified using a sampling scope and measuring between 0.6 V and 6 V on the leading edge of the output pulse. Even fast scopes (100 MHz bandpass) degrade this measurement and readings must be corrected by calculations. It is preferable to use a 1000 pF capacitor and a sampling scope as shown in Figure 3.46 to read the 10% to 90% points directly. The resulting typical rise times of 40 ns are significantly better than those of the UJT which are about 100 ns.

MINIMUM AND MAXIMUM FREQUENCY

In actual tests with devices whose parameters are known, it is possible to establish minimum and maximum values of timing resistors that will guarantee oscillation. The circuit under discussion is a conventional RC relaxation type oscillator.

To obtain maximum frequency, it is desirable to use low values of capacitance (1000 pF) and to select devices and bias conditions to obtain high I_V. It is possible to use stray capacitance but the results are generally unpredictable. The minimum value of timing resistance is obtained using the following rule of thumb:

$$R_{(min)} = 2(V_1 - V_V)/I_V$$

where the valley voltage (V_V) is often negligible.

To obtain minimum frequency, it is desirable to use high values of capacitance (10 μF) and to select devices and bias conditions to obtain low I_P. It is important that the capacitor leakage be quite low. Glass and mylar dielectrics are often used for these applications. The maximum timing resistor is as follows:

$$R_{(max)} = (V_I - V_P)/2I_P$$

In a circuit with a fixed value of timing capacitance, our most sensitive PUT, the MPU133, offers the largest dynamic frequency range. Allowing for capacitance and bias changes, the approximate frequency range of a PUT is from 0.003 Hz to 2.5 kHz.

TEMPERATURE COMPENSATION

The PUT with its external bias network exhibits a relatively small frequency change with temperature. The uncompensated RC oscillator shown in Figure 3.47 was tested at various frequencies by changing the timing resistor R_T. At discrete frequencies of 100, 200, 1000 and 2000 Hz, the ambient temperature was increased from 25° to 60°C. At these low frequencies, the negative temperature coefficient of V_{AG} predominated and caused a consistent 2% increase in frequency. At 10 kHz, the frequency remained within 1% over the same temperature range. The storage time phenomenon which increases the length of the output pulse as temperature increases is responsible for this result. Since this parameter has

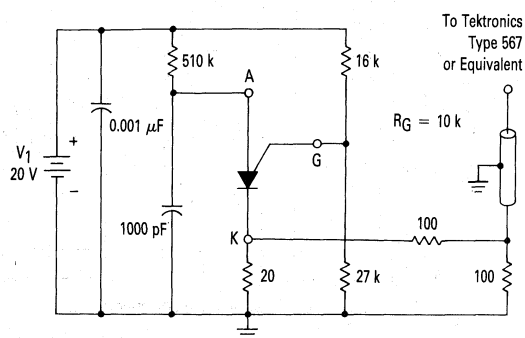


Figure 3.46. t_r Test Circuit for PUTs

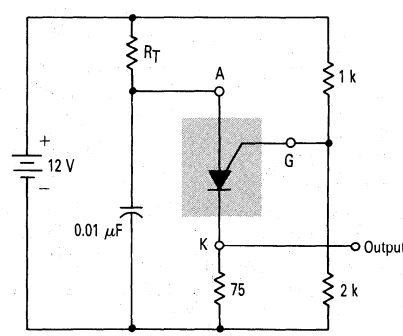


Figure 3.47. Uncompensated Oscillator

not been characterized, it is obvious that temperature compensation is more practical with relatively low frequency oscillators.

Various methods of compensation are shown in Figure 3.48. In the low cost diode-resistor combination of 3.48(a), the diode current is kept small to cause its temperature coefficient to increase. In 3.48(b), the bias current through the two diodes must be large enough so that their total coefficient compensates for V_{AG}. The transistor approach in 3.48(c) can be the most accurate since its temperature coefficient can be varied independently of bias current.

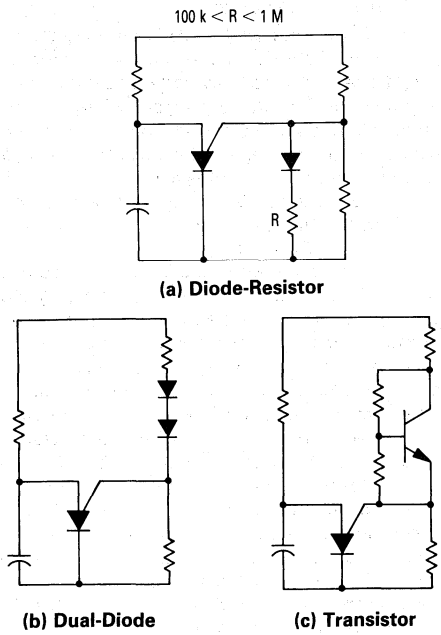


Figure 3.48. Temperature Compensation Techniques

THE SILICON BILATERAL SWITCH

The silicon bilateral switch (SBS) is an advanced semiconductor with negative resistance switching characteristics similar to the 3-layer diode, 4-layer diode and unijunction transistor (UJT). The latter devices have seen wide application, especially in triggering circuits where they can supply the fast rising, high-current gate pulse necessary for the proper operation of power thyristors. In such applications, the SBS can improve circuit performance and reduce cost at the same time.

These devices are not just an improved version of a PNPN diode. They are actually fabricated as simple integrated circuits consisting of transistors, diodes and resistors, connected as two anti-parallel, regenerative switches. Since the device is fabricated as an IC, the components are well matched resulting in an asymmetry, or difference of positive V_S and negative V_S, of less than 0.5 volts.

A third lead, designated the Gate, has been brought out for increased circuit flexibility. Since these devices are a regenerative switch, they may also be designed into many low power latching circuits.

The equivalent circuit diagram of an SBS and its symbol are shown in Figure 3.49. The device is actually a simple IC and consists of two halves of a PNP and an NPN transistor, a 6.8 volt zener diode and a 15 kΩ resistor, R_B. Unlike existing 4-layer diodes which use a stacked structure, the SBS is constructed using annular techniques. The result is a device with better stability and control of its electrical parameters.

Electrical characteristics are shown in Figure 3.50 and the parameters are defined as follows: V_S, the switching voltage, is the maximum forward voltage the device can sustain without switching to the conducting state; I_S, the

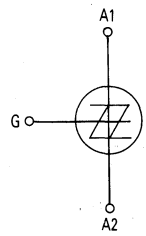


Figure 3.49(a). SBS Equivalent Circuit and Symbol

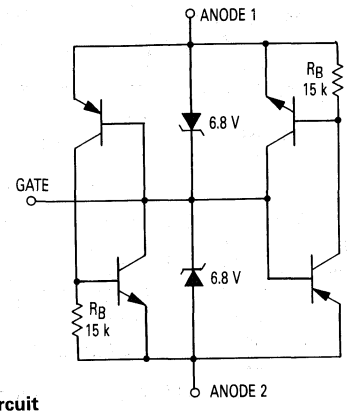


Figure 3.49(b)

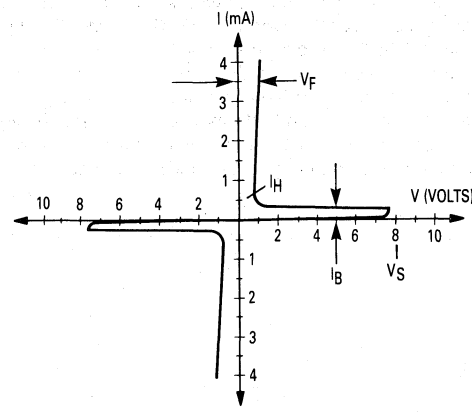


Figure 3.50. SBS Anode 1-Anode 2 V-I Characteristics

switching current, is the current through the device when V_S is applied; V_F, the forward voltage, is the voltage drop across the device when it is in the conducting state and passing a specified current; I_H, the holding current, is the current necessary to sustain conduction; I_B is the leakage current through the device with five volts bias.

Operation of the SBS can be best understood by referring to Figures 3.50 and 3.51. Consider an adjustable source of voltage with a current limiting resistor in series supplying a voltage to a device anode 1 that is five volts positive with respect to anode 2. Since this voltage is less than the sum of V_{BE} of the PNP transistor and V_Z of the 6.8 volt zener diode, only a very small leakage current will flow and the device in the off or blocking state. As the supply voltage is increased, a point will be reached (near V_S) where a small increase in voltage results in a substantial increase in current flow. The PNP transistor purposely has high current gain and most of this increased current flows out of its collector and produces a voltage drop across R_B.

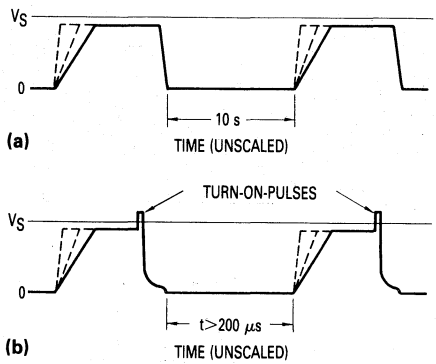


Figure 3.51. Waveforms for dv/dt Test

The two transistors are connected in a positive feedback loop similar to the equivalent circuit for an SCR where the collector current of one is the base current for the other. When the voltage across R_B is sufficient to turn the NPN transistor on and the loop gain exceeds unity, both transistors are driven into saturation, the voltage across the device abruptly drops and the current through it is limited mainly by the external circuitry. The device has now switched to the on or conducting state.

The 6.8 volt zener diode has a positive temperature voltage coefficient which is opposite to that of V_{BE} of the PNP transistor. The net result is good temperature stability of V_S, typically +0.02%/°C.

V_F, the forward voltage across the device, remains relatively low even if the current through it is greatly increased, rising approximately 3 volts/ampere. The device will remain on until the current through it is reduced below the holding current value. One method of insuring turn-off is to apply a reverse voltage less than V_R, the maximum reverse voltage. After a few microseconds (turn-off time) have elapsed, the transistors will have recovered from saturation and the device will again block a forward voltage up to V_S.

A third lead, the gate, can be used to modify the characteristics of the SBS. As an example, connecting a 9.4 volt zener diode from gate to cathode would lower V_S to approximately 4.6 volts. Connecting a 20 kΩ resistor from gate to anode and a similar resistor from gate to cathode will lower V_S to approximately 4 volts at the expense of increased current around the device prior to switching. Also, if a voltage less than V_S is applied to an SBS it can be "gated" on by drawing a small current out of the gate lead.

Like other regenerative switches, the SBS has a tendency to switch on in the presence of rapidly rising anode voltage. The dv/dt rating of the SBS is difficult to define and the method of measurement may produce erroneous results. A test ramp of voltage with adjustable dv/dt as shown in Figure 3.51(a) may be applied to the device if not repeated more frequently than every 10 seconds. The device may switch to the on state when the dv/dt is in the range of 1 to 10 volts/microsecond. The repetitive waveform of (b) may be applied much more frequently (convenient for an oscilloscope display) providing only that the time interval between turn-off and the next ramp is longer than the turn-off time of the device. The turn-on pulse in (c) is necessary to discharge internal capacitance which can accumulate a charge and give false indication of very high dv/dt capability.

Sweeping an SBS in either direction will yield similar results. However, when an SBS has been conducting in one direction and the anode voltage is rapidly reversed, the dv/dt must be limited to approximately 0.1 volt/microsecond. This is necessary because if the transistors in the conducting half of the device have not recovered from saturation, they will provide a path for a current to turn the opposite side on.

CHAPTER 4
THE SIDAC, A NEW HIGH VOLTAGE BILATERAL TRIGGER

1

The SIDAC is a high voltage bilateral trigger device that extends the trigger capabilities to significantly higher voltages and currents than have been previously obtainable, thus permitting new, cost-effective applications. Being a bilateral device, it will switch from a blocking state to a conducting state when the applied voltage of either polarity exceeds the breakover voltage. As in other trigger devices, (SBS, Four Layer Diode), the SIDAC switches through a negative resistance region to the low voltage on-state (Figure 4.1) and will remain on until the main terminal current is interrupted or drops below the holding current.

SIDAC's are available in the large MK1V series and economical, easy to insert, small MKP9V series axial lead packages. Breakdown voltages ranging from 104 to 280 V are available. The MK1V devices feature bigger chips and provide much greater surge capability along with somewhat higher RMS current ratings.

The high-voltage and current ratings of SIDACs make them ideal for high energy applications where other trigger devices are unable to function alone without the aid of additional power boosting components.

The basic SIDAC circuit and waveforms, operating off of ac are shown in Figure 4.2. Note that once the input voltage exceeds $V_{(BO)}$, the device will switch on to the forward on-voltage V_{TM} of typically 1.1 V and can conduct as much as the specified repetitive peak on-state current I_{TRM} of 20 A (10 μs pulse, 1 kHz repetition frequency).

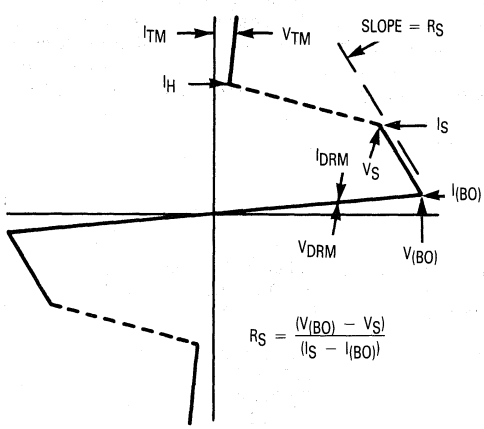


Figure 4.1(a). Idealized SIDAC V-I Characteristics

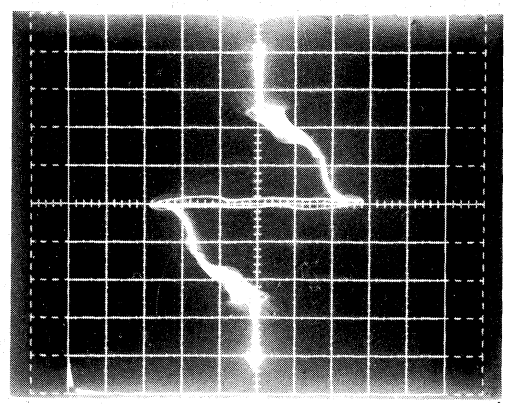


Figure 4.1(b). Actual MKP9V130 V-I Characteristic. Horizontal: 50 V/Division. Vertical: 20 mA/Division. (0,0) at Center. R_L = 14 k Ohm.

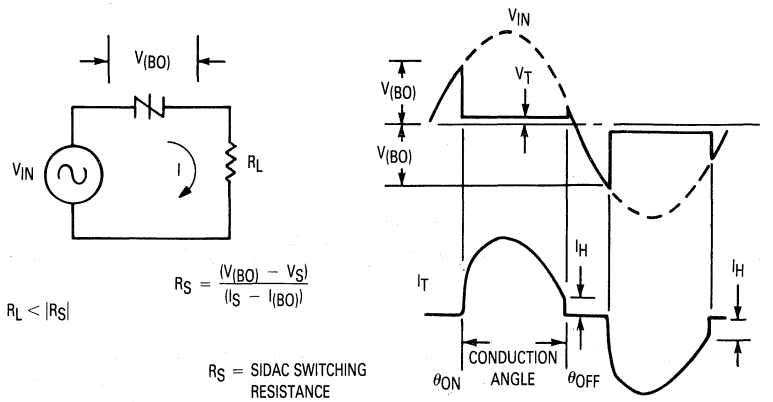


Figure 4.2. Basic SIDAC Circuit and Waveforms

Operation from an AC line with a resistive load can be analyzed by superimposing a line with slope $= -1/R_L$ on the device characteristic. When the power source is AC, the load line can be visualized as making parallel translations in step with the instantaneous line voltage and frequency. This is illustrated in Figure 4.3 where v_1 through v_5 are the instantaneous open circuit voltages of the AC generator and i_1 through i_5 are the corresponding short circuit currents that would result if the SIDAC was not in the circuit. When the SIDAC is inserted in the circuit, the current that flows is determined by the intersection of the load line with the SIDAC characteristic. Initially the SIDAC blocks, and only a small leakage current flows at times 1 through 4. The SIDAC does not turn-on until the load line supplies the breakover current ($I_{(BO)}$) at the breakover voltage ($V_{(BO)}$).

If the load resistance is less than the SIDAC switching resistance, the voltage across the device will drop quickly as shown in Figure 4.2. A stable operating point (V_T, I_T) will result if the load resistor and line voltage provide a current greater than the latching value. The SIDAC remains in an "on" condition until the generator voltage causes the current through the device to drop below the holding value (I_H). At that time, the SIDAC switches to the point (V_{off}, I_{off}) and once again only a small leakage current flows through the device.

Figure 4.4 illustrates the result of operating a SIDAC with a resistive load greater than the magnitude of its switching resistance. The behavior is similar to that described in Figures 4.2 and 4.3 except that the turn-on and turn-off of the SIDAC is neither fast nor complete. Stable operating points on the SIDAC characteristics

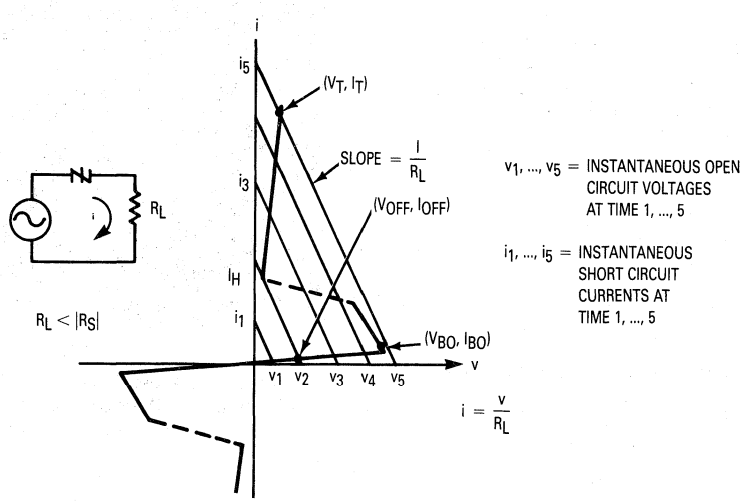


Figure 4.3. Load Line for Figure 4.2. (1/2 Cycle Shown.)

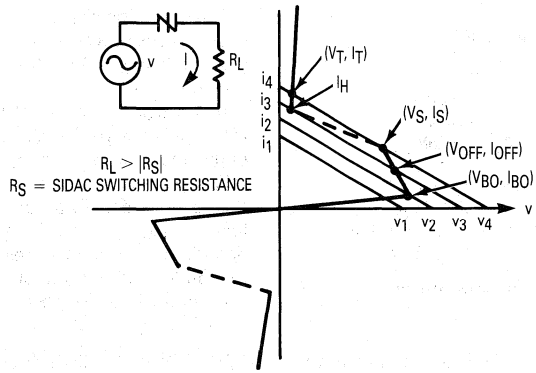


Figure 4.4. High Resistance Load Line with Incomplete Switching

between $(V_{(BO)}, I_{(BO)})$ and (V_S, I_S) result as the generator voltage increases from v_2 to v_4. The voltage across the SIDAC falls only partly as the loadline sweeps through this region. Complete turn-on of the SIDAC to (V_T, I_T) does not occur until the load line passes through the point (V_S, I_S). The load line illustrated in Figure 4.4 also results in incomplete turn-off. When the current drops below I_H, the operating point switches to (V_{off}, I_{off}) as shown on the device characteristic.

The switching current and voltage can be 2 to 3 orders of magnitude greater than the breakover current and on-state voltage. These parameters are not as tightly specified as V_{BO} and I_{BO}. Consequently operation of the SIDAC in the state between fully on and fully off is undesirable because of increased power dissipation, poor efficiency, slow switching, and tolerances in timing.

Figure 4.5 illustrates a technique which allows the use of the SIDAC with high impedance loads. A resistor can be placed around the load to supply the current required to latch the SIDAC. Highly inductive loads slow the current rise and the turn-on of the SIDAC because of their L/R time constant. The use of shunt resistor around the load will improve performance when the SIDAC is used with inductive loads such as small transformers and motors.

The SIDAC can be used in oscillator applications. If the load line intersects the device characteristic at a point where the total resistance $(R_L + R_S)$ is negative, an unstable operating condition with oscillation will result. The resistive load component determines steady-state behavior. The reactive components determine transient behavior. Figure 4.10 shows a SIDAC relaxation oscillator application. The wide span between I_{BO} and I_H makes the SIDAC easy to use. Long oscillation periods can be achieved with economical capacitor sizes because of the low device $I_{(BO)}$.

Z1 is typically a low impedance. Consequently the SIDAC's switching resistance is not important in this application. The SIDAC will switch from a blocking to full on-state in less than a fraction of a microsecond.

The timing resistor must supply sufficient current to fire the SIDAC but not enough current to hold the SIDAC in an on-state. These conditions are guaranteed when the timing resistor is selected to be between R_{max} and R_{min}.

For a given time delay, capacitor size and cost is minimized by selecting the largest allowable timing resistor. R_{max} should be determined at the lowest temperature of operation because $I_{(BO)}$ increases then. The load line corresponding to R_{max} passes through the point $(V_{(BO)}, I_{(BO)})$ allowing the timing resistor to supply the needed breakover current at the breakover voltage. The load line for a typical circuit design should enclose this point to prevent sticking in the off state.

Requirements for higher oscillation frequencies and greater stored energy in the capacitor result in lower values for the timing resistor. R_{min} should be determined at the highest operating temperature because I_H is lower then. The load line determined by R and V_{in} should pass below I_H on the device characteristic or the SIDAC will stick in the on-state after firing once. I_H is typically more than 2 orders of magnitude greater than I_{BO}. This makes the SIDAC well suited for operation over a wide temperature span.

SIDAC turn-off can be aided when the load is an underdamped oscillatory CRL circuit. In such cases, the SIDAC current is the sum of the currents from the timing resistor and the ringing decay from the load. SIDAC turn-off behavior is similar to that of a TRIAC where turn-off will not occur if the rate of current zero crossing is high. This is a result of the stored charge within the volume of the device. Consequently, a SIDAC cannot be force commuted like an SCR. The SIDAC will pass a ring wave of sufficient amplitude and frequency. Turn-off requires the device current to approach the holding current gradually. This is a complex function of junction temperature, holding current magnitude, and the current wave parameters.

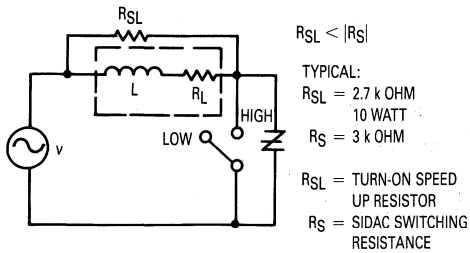


Figure 4.5. Inductive Load Phase Control

1

How can the SIDAC be used? One application is to replace the combination of a small-signal trigger and TRIAC with the SIDAC, as shown in Figure 4.6. In this example, the trigger — an SBS (Silicon Bidirectional Switch) that conducts at about 8 V — will fire the TRIAC by dumping the charge from the capacitor into the gate of the TRIAC. This circuit is amenable to phase controlling the TRIAC, if so required, as the RC time constant can be readily varied.

The simple SIDAC circuit can also supply switchable load current. However, the conduction angle is not readily controllable, being a function of the peak applied voltage and the breakover voltage of the SIDAC. As an example, for peak line voltage of about 170 V, at $V_{(BO)}$ of 115 V and a holding current of 100 mA, the conduction angle would be about 130°. With higher peak input voltages (or lower breakdown voltages) the conduction angle would correspondingly increase. For non-critical conduction angle, 1 A rms switching applications, the SIDAC is a very cost-effective device.

Figure 4.7 shows an example of a SIDAC used to phase control an incandescent lamp. This is done in order to lower the RMS voltage to the filament and prolong the life of the bulb. This is particularly useful when lamps are used in hard to reach locations such as outdoor lighting in signs where replacement costs are high. Bulb life span can be extended by 1.5 to 5 times depending on the type of lamp, the amount of power reduction to the filament, and the number of times the lamp is switched on from a cold filament condition.

The operating cost of the lamp is also reduced because of the lower power to the lamp; however, a higher wattage bulb is required for the same lumen output. The maximum possible energy reduction is 50% if the lamp wattage is not increased. The minimum conduction angle is 90° because the SIDAC must switch on before the peak of the line voltage. Line regulation and breakover voltage tolerances will require that a conduction angle longer than 90° be used, in order to prevent lamp turn-off under low line voltage conditions. Consequently, practical conduction angles will run between 110° and 130° with corresponding power reductions of 10% to 30%.

In Figure 4.2 and Figure 4.7, the SIDAC switching angles are given by:

$$\theta_{ON} = \sin^{-1}(V_{(BO)}/V_{pk})$$

where V_{pk} = Maximum Instantaneous Line Voltage

$$\theta_{OFF} = 180 - \sin^{-1}\left(\frac{(J_H \cdot R_L) + V_T}{V_{pk}}\right)$$

where θ_{ON}, θ_{OFF} = Switching Angles in degrees
V_T = 1 V = Main Terminal Voltage at $I_T = I_H$

Generally the load current is much greater than the SIDAC holding current. The conduction angle then becomes 180° minus $\theta_{(on)}$.

Rectifiers have also been used in this application to supply half wave power to the lamp. SIDAC's prevent the flicker associated with half-wave operation of the lamp. Also, full wave control prevents the introduction of a DC component into the power line and improves the color temperature of the light because the filament has less time to cool during the off time.

The fast turn-on time of the SIDAC will result in the generation of RFI which may be noticeable on AM radios operated in the vicinity of the lamp. This can be prevented by the use of an RFI filter. A possible filter design is shown in Figure 4.5. This filter causes a ring wave of current through the SIDAC at turn-on time. The filter inductor must be selected for resonance at a frequency above the upper frequency limit of human hearing and as low below the start of the AM broadcast band as possible for maximum harmonic attenuation. In addition, it is important that the filter inductor be non-saturating to prevent dI/dT damage to the SIDAC. For additional information on filter design see page 1-5-30 and Figure 5.34.

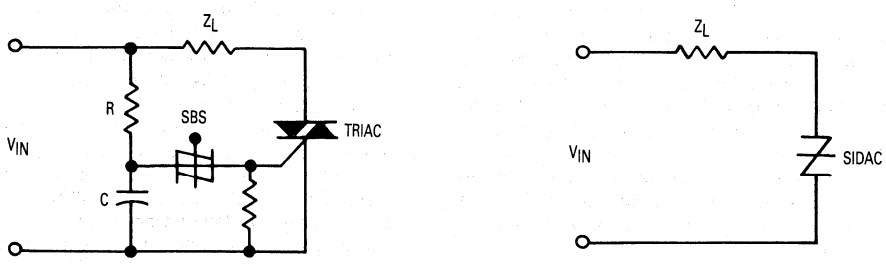


Figure 4.6. Comparison of a TRIAC and SIDAC Circuits

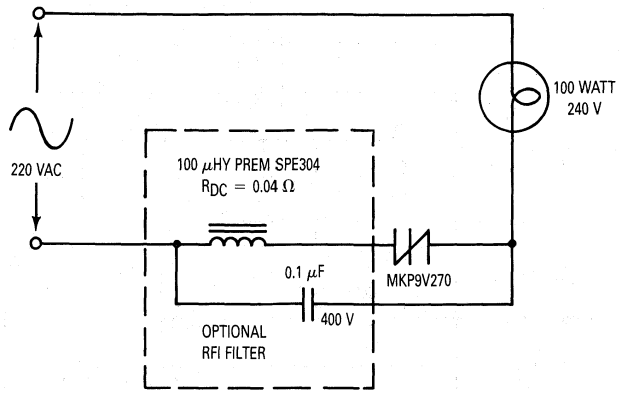


Figure 4.7. Long-Life Circuit for Incandescent Lamp

The sizing of the SIDAC must take into account the RMS current of the lamp, thermal properties of the SIDAC, and the cold start surge current of the lamp which is often 10 to 20 times the steady state load current. When lamps burn out, at the end of their operating life, very high surge currents which could damage the SIDAC are possible because of arcing within the bulb. The large MK1V device is recommended if the SIDAC is not to be replaced along with the bulb.

Since the MK1V series of SIDACs have relatively tight $V_{(BO)}$ tolerances (104 V to 115 V for the −115 device), other possible applications are over-voltage protection (OVP) and detection circuits. An example of this, as illustrated in Figure 4.8, is the SIDAC as a transient protector in the transformer-secondary of the medium voltage power supply, replacing the two more expensive back-to-back zeners or an MOV. The device can also be used across the output of the regulator (<100 V) as a simple OVP, but for this application, the regulator must have current foldback or a circuit breaker (or fuse) to minimize the dissipation of the SIDAC.

Another example of OVP is the telephony applications as illustrated in Figure 4.9. To protect the Subscriber Loop Interface Circuit (SLIC) and its associated electronics from voltage surges, two SIDACs and two rectifiers are used for secondary protection (primary protection to 1,000 V is provided by the gas discharge tube across the lines). As an example, if a high positive voltage transient appeared on the lines, rectifier D1 (with a P.I.V. of 1,000 V) would block it and SIDAC D4 would conduct the surge to ground. Conversely, rectifier D2 and SIDAC D3 would protect the SLIC for negative transients. The SIDACs will not conduct when normal signals are present.

Being a negative resistance device, the SIDAC also can be used in a simple relaxation oscillator where the frequency is determined primarily by the RC time constant (Figure 4.10). Once the capacitor voltage reaches the SIDAC breakover voltage, the device will fire, dumping the charged capacitor. By placing the load in the discharge path, power control can be obtained; a typical load could be a transformer-coupled xeon flasher, as shown in Figure 4.12.

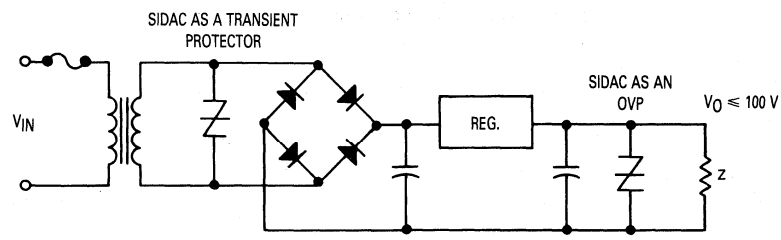


Figure 4.8. Typical Application of SIDACs as a Transient Protector and OVP in a Regulated Power Supply

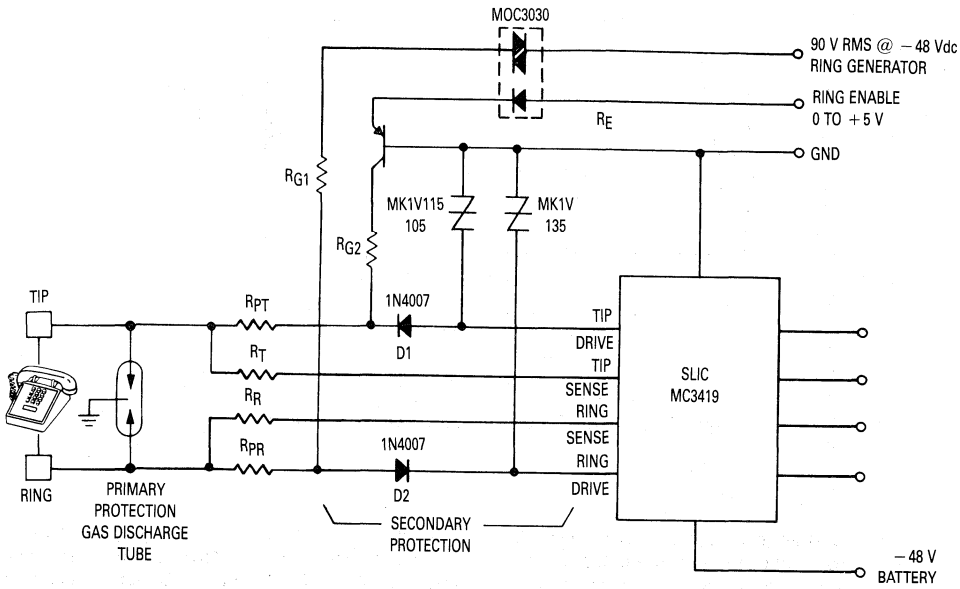


Figure 4.9. SIDACs Used for OVP in Telephony Applications

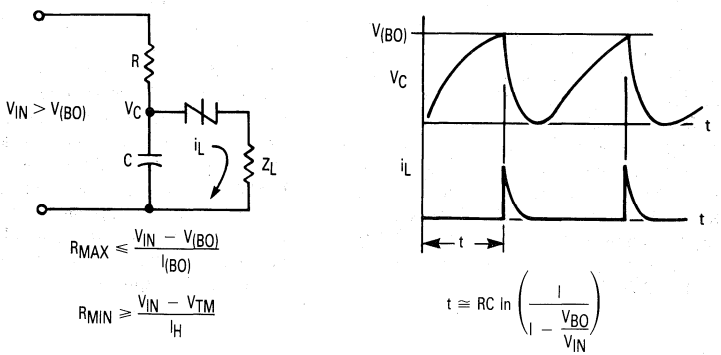


Figure 4.10. Relaxation Oscillator Using a SIDAC

SIDAC's provide an economical means for starting high intensity high pressure gas discharge lamps. These lamps are attractive because of their long operating life and high efficiency. They are widely used in outdoor lighting for these reasons.

Figure 4.13 illustrates how SIDAC's can be used in sodium vapor lamp starters. In these circuits, the SIDAC is used to generate a short duration (1 to 20 μs) high-voltage pulse of several KV or more which is timed by means of the RC network across the line to occur near the peak of the AC input line voltage. The high voltage pulse strikes the arc which lights the lamp.

In these circuits, an inductive ballast is required to provide a stable operating point for the lamp. The lamp is a negative resistance device whose impedance changes with current, temperature, and time over the first few minutes of operation. Initially, before the lamp begins to conduct, the lamp impedance is high and the full line voltage appears across it. This allows C to charge to the breakover voltage of the SIDAC, which then turns on discharging the capacitor through a step-up transformer generating the high voltage pulse. When the arc strikes, the voltage across the lamp falls reducing the available charging voltage across RC to the point where V_C no longer exceeds $V_{(BO)}$ and the SIDAC remains off. The low duty cycle lowers average junction temperature improving SIDAC reliability. Normal operation approximates non-repetitive conditions. However, if the lamp fails or is removed during replacement, operation of the SIDAC will be at the 60 Hz line frequency. The design of the circuit should take into account the resulting steady state power dissipation.

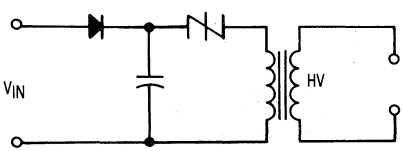


Figure 4.11. Typical Capacitor Discharge SIDAC Circuit

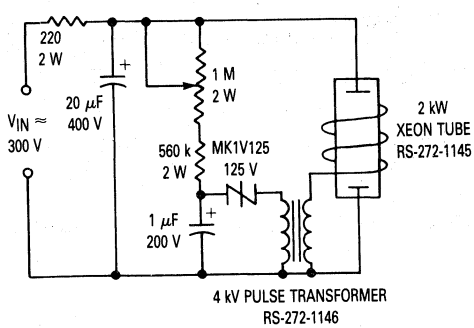
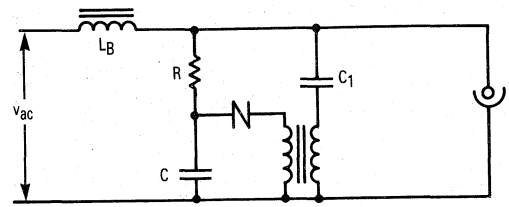
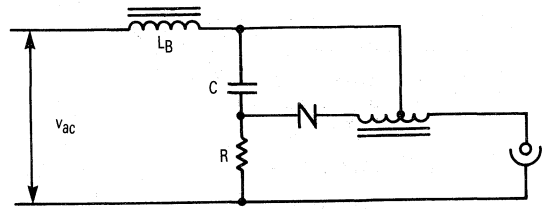


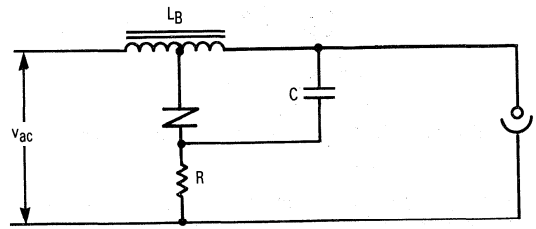
Figure 4.12. Xeon Flasher Using a SIDAC



(a). Conventional HV Transformer



(b). H.V. Auto-Transformer

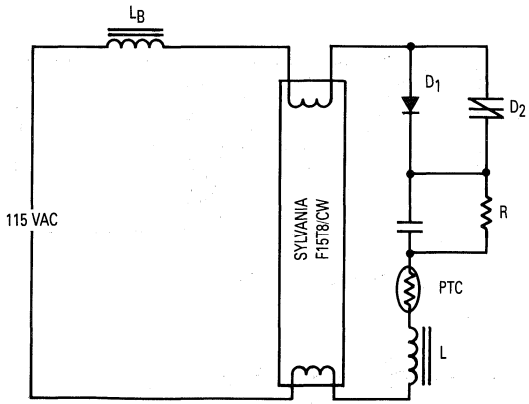


(c). Tapped Ballast Auto Transformer

Figure 4.13. Sodium Vapor Lamp Starter Circuits

Figure 4.14 illustrates a solid state fluorescent lamp starter using the SIDAC. In this circuit the ballast is identical to that used with the conventional glow-tube starter shown in Figure 4.15.

The glow tube starter consists of a bimetallic switch placed in series with the tube filaments which closes to energize the filaments and then opens to interrupt the current flowing through the ballast inductor thereby generating the high-voltage pulse necessary for starting. The mechanical glow-tube starter is the circuit component most likely to cause unreliable starting.



L_B	UNIVERSAL MFG CORP CAT200-H2 14-15-20-22 WATT BALLAST 325 mHY 28.9 Ω DCR
D1	1N4005 RECTIFIER
D2	MKP9V270 SIDAC
C	3 VFD 400 V
R	68 k OHMS 112 WATT
PTC	KEYSTONE CARBON COMPANY RL3006-50-40-25-PTO 50 OHMS/25°C
L	MICROTRAN QIL 50-F 50 mHY 11 OHMS

Figure 4.14. Fluorescent Starter Using SIDAC

The heating of the filaments causes thermonic emission of electrons from them. These electrons are accelerated along the length of the tube causing ionization of the argon gas within the tube. The heat generated by the starting current flow through the tube vaporizes the mercury droplets within the tube which then become ionized themselves causing the resistance and voltage across the tube to drop significantly. The drop in voltage across the tube is used to turn off the starting circuit and prevent filament current after the lamp is lit.

The SIDAC can be used to construct a reliable starter circuit providing fast, positive lamp ignition. The starter shown in Figure 4.14 generates high voltage by means of a series CRL charging circuit. The circuit is roughly analogous to a TRIAC snubber used with an inductive load, except for a lower damping factor and higher Q. The size of C determines the amount of filament heating current by setting the impedance in the filament circuit before ionization of the tube.

The evolution of this circuit can be understood by first considering an impractical circuit (Figure 4.16).

If L_B and C are adjusted for resonance near 60 Hz, the application of the AC line voltage will result in a charging current that heats the filaments and a voltage across the capacitor and tube that grows with each half-cycle of the AC line until the tube ionizes. Unfortunately, C is a large capacitor which can suddenly discharge through the tube causing high current pulses capable of destroying the tube filament. Also C provides a permanent path for filament current after starting. These factors cause short tube operating life and poor efficiency because of filament power losses. The impractical circuit must be modified to:

(1) Switch off the filament current after starting.
(2) Limit capacitor discharge current spikes.

In Figure 4.14 a parallel connected rectifier and SIDAC have been added in series with the capacitor C. The breakover voltage of the SIDAC is higher than the peak of the line voltage. Diode D1 is therefore necessary to provide a current path for charging C.

On the first half-cycle, C resonant charges through diode D1 to a peak voltage of about 210 V, and remains at that value because of the blocking action of the rectifier and SIDAC. During this time, the bleeder resistor R has negligible effect on the voltage across C because the RC time constant is long in comparison to the line period. When the line reverses, the capacitor voltage boosts the voltage across the SIDAC until breakover results. This results in a sudden step of voltage across the inductor L, causing resonant charging of the capacitor to a higher voltage on the 2nd half-cycle.

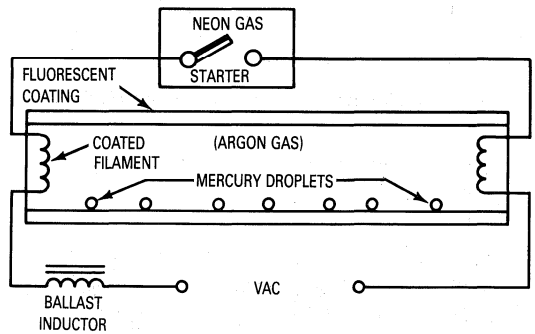


Figure 4.15. Fluorescent Lamp with Glow Tube Starter

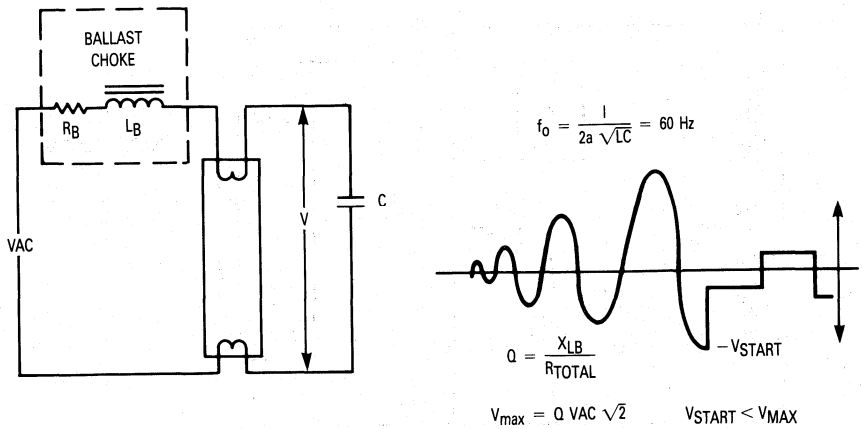
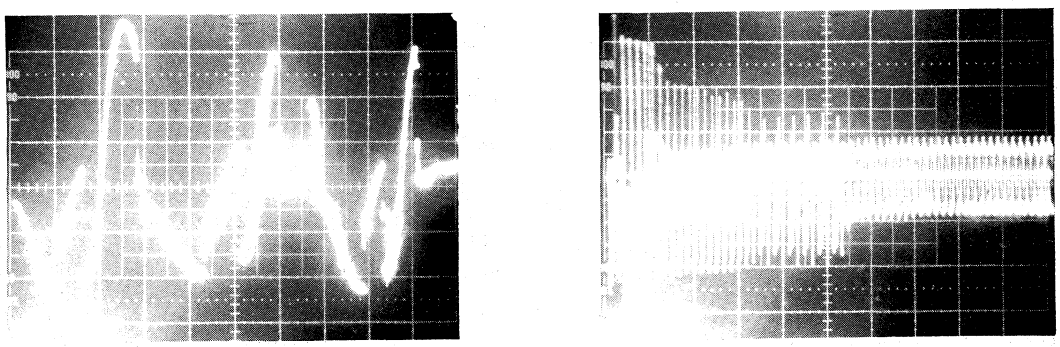


Figure 4.16. Impractical Starter Circuit



(a). 5 ms/DIVISION **(b). 100 ms/DIVISION**

Figure 4.17. Starting Voltage Across Fluorescent Tube 100 V/DIV
0 V AT CENTER
V_{Line} = 110 V

Several cycles of operation are necessary to approach steady state operating conditions. Figure 4.17 shows the starting voltage waveform across the tube.

The components R, PTC, and L serve the dual role of guarantying SIDAC turn-off and preventing capacitor discharge currents through the tube.

SIDAC's can also be used with auto-transformer ballasts. The high voltage necessary for starting is generated by the leakage autotransformer. The SIDAC is used to turn-on the filament transformer initially and turn it off after ionization causes the voltage across the tube to drop.

Figure 4.18 illustrates this concept. The resistor R can be added to aid turn-off of the SIDAC by providing a small idle current resulting in a voltage drop across the impedance Z. The impedance Z could be a saturable reactor and or positive temperature coefficient thermistor. These components help to insure stability of the system comprised of the negative resistance SIDAC and negative resistance tube during starting, and promote turn off of the SIDAC.

The techniques illustrated in Figure 4.13 are also possible methods for generation of the necessary high-voltage required in fluorescent starting. The circuits

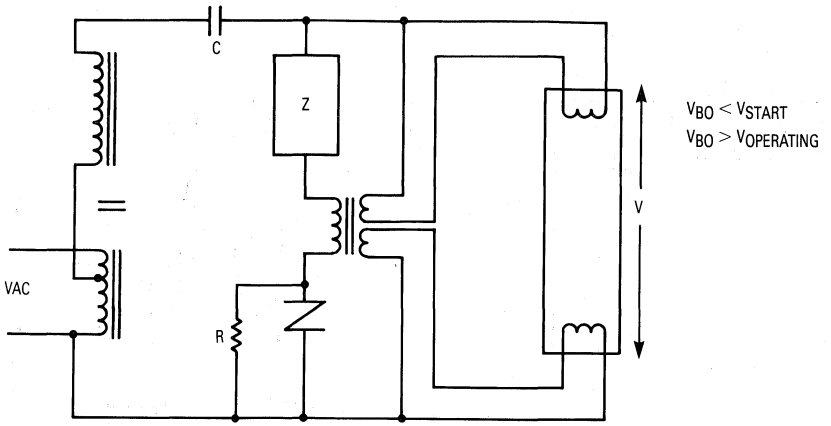


Figure 4.18. Fluorescent Starter Using SIDAC and Autotransformer Ballast

Table 4.1. Possible Sources for Thermistor Devices

Fenwal Electronics, 63 Fountain Street Framingham MA 01701
Keystone Carbon Company, Thermistor Division St. Marys, PA 15857
Thermometrics, 808 U.S. Highway 1 Edison, N.J. 08817
Therm-O-Disc, Inc. Micro Devices Product Group 1320 South Main Street, Mansfield, OH 44907
Midwest Components Inc., P.O. Box 787 1981 Port City Boulevard, Muskegon, MI 49443
Nichicon (America) Corp., Dept. G 927 E. State Pkwy, Schaumburg, IL 60195

must be modified to allow heating of the fluorescent tube cathodes if starting is to simulate the conditions existing when a glow tube is used.

Thermistors are useful in delaying the turn-on or insuring the turn-off of SIDAC devices. Table 4.1 shows possible sources of thermistor devices.

Other high voltage nominal current trigger applications are:

- Gas or oil igniters
- Electric fences
- HV electrostatic air filters
- Capacitor Discharge ignitions

Note that all these applications use similar circuits where a charged capacitor is dumped to generate a high transformer secondary voltage (Figure 4.11).

In many cases, the SIDAC current wave can be approximated by an exponential or quasi-exponential current wave (such as that resulting from a critically damped or slightly underdamped CRL discharge circuit). The question then becomes; how much "real world" surge current can the SIDAC sustain? The data sheet defines an I_{TSM} of 20 A, but this is for a 60 Hz, one cycle, peak sine wave whereas the capacitor discharge current waveform has a fast-rise time with an exponential fall time.

To generate the surge current curve of peak current versus exponential discharge pulse width, the test circuit of Figure 4.19 was implemented. It simulates the topology of many applications whereby a charged capacitor is dumped by means of a turned-on SIDAC to produce a current pulse. Timing for this circuit is derived from the nonsymmetrical CMOS astable multivibrator (M.V.) gates G1 and G2. With the component values shown, an approximate 20 second positive-going output pulse is fed to the base of the NPN small-signal high voltage transistor Q1, turning it on. The following high voltage PNP transistor is consequently turned on, allowing capacitor C1 to be charged through limiting resistor R1 in about 16 seconds. The astable M.V. then changes state for about 1.5 seconds with the positive going pulse from Gate 1 fed through integrator R2-C2 to Gate 3 and then Gate 4. The net result of about a 100 μs time delay from G4 is to ensure non-coincident timing conditions. This positive going output is then differentiated by C3-R3 to produce an approximate 1 ms, leading edge, positive going pulse which turns on NPN transistor Q3 and the following PNP transistor Q4. Thus, an approximate 15 mA, 1 ms pulse is generated for turning on SCR Q5 about 100 μs after capacitor charging transistor Q2 is turned off. The SCR now fires, discharging C1 through the current limiting resistor R4 and the SIDAC Device Under Test (D.U.T.). The peak current and its duration is set by the voltage V_C across capacitor C1 and current limiting resistor R4. The circuit has about a 240 V capability limited by C1, Q1 and Q2 (250 V, 300 V and 300 V respectively).

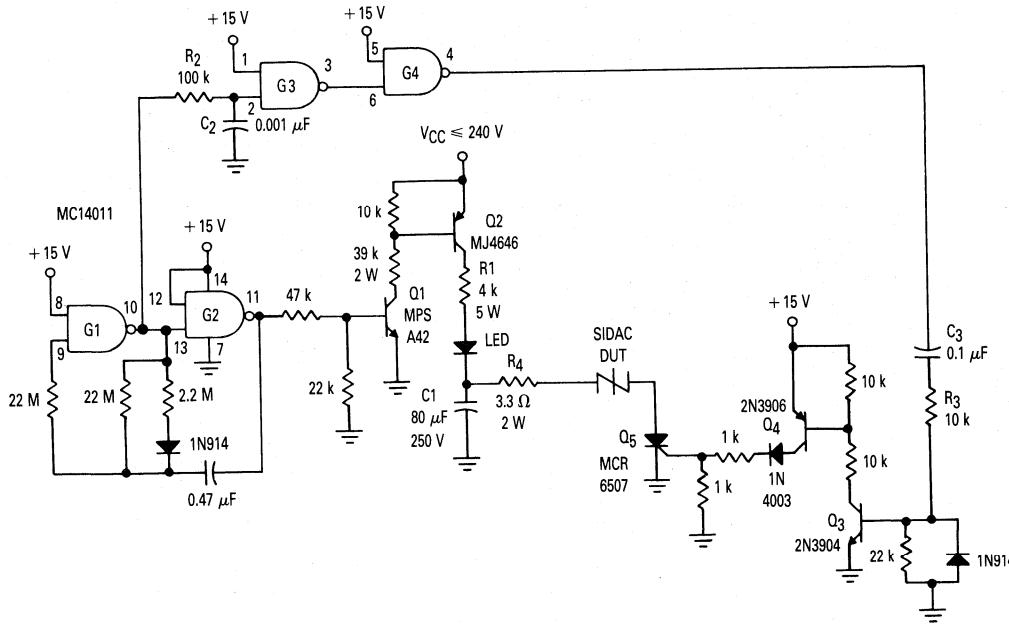


Figure 4.19. SIDAC Surge Tester

The SCR is required to fire the SIDAC, rather than the breakover voltage, so that the energy to the D.U.T. can be predictably controlled.

By varying V_C, C1 and R4, the surge current curve of Figure 15 was derived. Extensive life testing and adequate derating ensure that the SIDAC, when properly used, will reliably operate in the various applications.

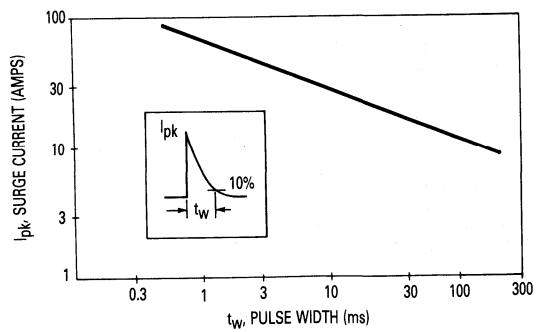


Figure 4.20. Exponential Surge Current Capability of the MK1V SIDAC. Pulse Width versus Peak Current

CHAPTER 5
SCR CHARACTERISTICS

SCR TURN-OFF CHARACTERISTICS

In addition to their traditional role of power control devices, SCRs are being used in a wide variety of other applications in which the SCR's turn-off characteristics are important. As an example — reliable high frequency inverters and converter designs (<20 kHz) require a known and controlled circuit-commutated turn-off time (t_q). Unfortunately, it is usually difficult to find the turn-off time of a particular SCR for a given set of circuit conditions.

This section discusses t_q in general and describes a circuit capable of measuring t_q. Moreover, it provides data and curves that illustrate the effect on t_q when other parameters are varied, to optimize circuit performance.

SCR TURN-OFF MECHANISM

The SCR, being a four layer device (P-N-P-N), is represented by the two interconnected transistors, as shown in Figure 5.1. This regenerative configuration allows the device to turn on and remain on when the gate trigger is removed, as long as the loop gain criteria is satisfied; i.e., when the sum of the common base current gains (α) of both the equivalent NPN transistor and PNP transistor, exceed one. To turn off the SCR, the loop gain must be brought below unity, whereby the on-state principal current (anode current i_T) limited by the external circuit impedance, is reduced below the holding current (I_H). For ac line applications, this occurs automatically during the negative going portion of the waveform. However, for dc applications (inverters, as an example), the anode current must be interrupted or diverted; (diversion of the anode current is the technique used in the t_q test fixture described later in this application note).

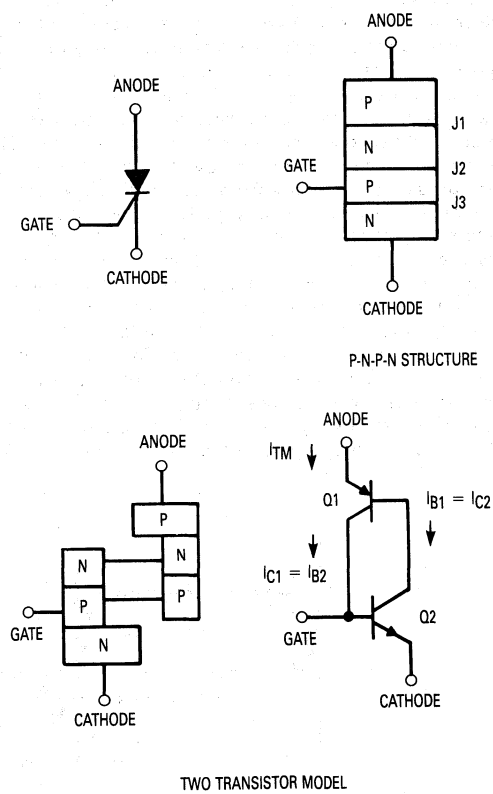


Figure 5.1. Two Transistor Analogy of an SCR

SCR TURN-OFF TIME t_q

Once the anode current in the SCR ceases, a period of time must elapse before the SCR can again block a forward voltage. This period is the SCR's turn-off time, t_q, and is dependent on temperature, forward current, and other parameters. The turn-off time phenomenon can be understood by considering the three junctions that make up the SCR. When the SCR is in the conducting state, each of the three junctions is forward biased and the N and P regions (base regions) on either side of J2 are heavily saturated with holes and electrons (stored charge). In order to turn off the SCR in a minimum amount of time, it is necessary to apply a negative (reverse) voltage to the device anode, causing the holes and electrons near the two end junctions, J1 and J3, to diffuse to these junctions. This causes a reverse current to flow through the SCR. When the holes and electrons near junc-

tions J1 and J3 have been removed, the reverse current will cease and junctions J1 and J3 will assume a blocking state. However, this does not complete the recovery of the SCR since a high concentration of holes and electrons still exist near the center junction, J2. This concentration decreases by the recombination process and is largely independent of the external circuit. When the hole and electron concentration near junction J2 has reached some low value, junction J2 will assume its blocking condition and a forward voltage can, after this time, be applied without the SCR switching back to the conduction state.

1

t_q MEASUREMENT

When measuring SCR turn-off time, t_q, it is first necessary to establish a forward current for a period of time long enough to ensure carrier equilibrium. This must be specified, since I_{TM} has a strong effect on the turn-off time of the device. Then, the SCR current is reversed at a specified di/dt rate, usually by shunting the SCR anode to some negative voltage through an inductor. The SCR will then display a "reverse recovery current," which is the charge clearing away from the junctions. A further waiting time must then elapse while charges recombine, before a forward voltage can be applied. This forward voltage is ramped up a specified dv/dt rate. The dv/dt delay time is reduced until a critical point is reached where the SCR can no longer block the forward applied voltage ramp. In effect, the SCR turns on and consequently, the ramp voltage collapses. The elapsed time between this critical point and the point at which the forward SCR current passes through zero and starts to go negative (reverse recovery phase), is the t_q of the SCR. This is illustrated by the waveforms shown in Figure 5.2.

t_q GENERAL TEST FIXTURE

The simplified circuit for generating these waveforms is schematically illustrated in Figure 5.3. This circuit is implemented with as many as eight transformers including variacs, and in addition to being very bulky, has been known to be troublesome to operate. However, the configuration is relevent and, in fact, is the basis for the design, as described in the following paragraphs.

t_q TEST FIXTURE BLOCK DIAGRAMS AND WAVEFORMS

The block diagram of the t_q Test Fixture, illustrated in Figure 5.4, consists of four basic blocks: A Line Synchronized Pulse Generator establishes system timing; a Constant Current Generator (variable in amplitude) powers the Device Under Test (DUT); a di/dt Circuit controls the rate of change of the SCR turn-off current; and the dv/dt Circuit reapplies a controlled forward blocking voltage. Note from the waveforms illustrated that the di/dt circuit, in parallel with the DUT, diverts the constant current from the DUT to produce the described anode current I_{TM}.

t_q TEST FIXTURE CHARACTERISTICS

The complete schematic of the t_q Test Fixture and the important waveforms are shown in Figures 5.5 and 5.6, respectively.

A CMOS Gate is used as the Line Synchronized Pulse Generator, configured as a wave shaping Schmitt trigger, clocking two cascaded monostable multivibrators for delay and pulse width settings (Gates 1C to 1F). The result is a pulse generated every half cycle whose width and position (where on the cycle it triggers) are adjustable by

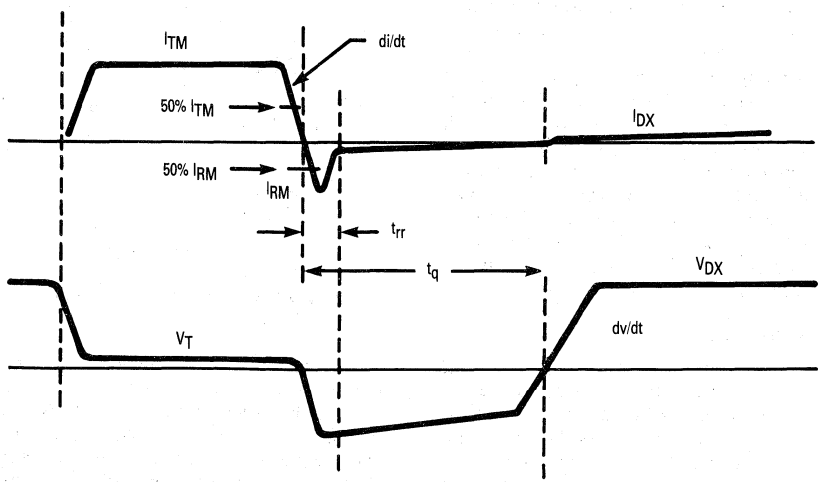


Figure 5.2. SCR Current and Voltage Waveforms During Circuit-Commutated Turn-Off

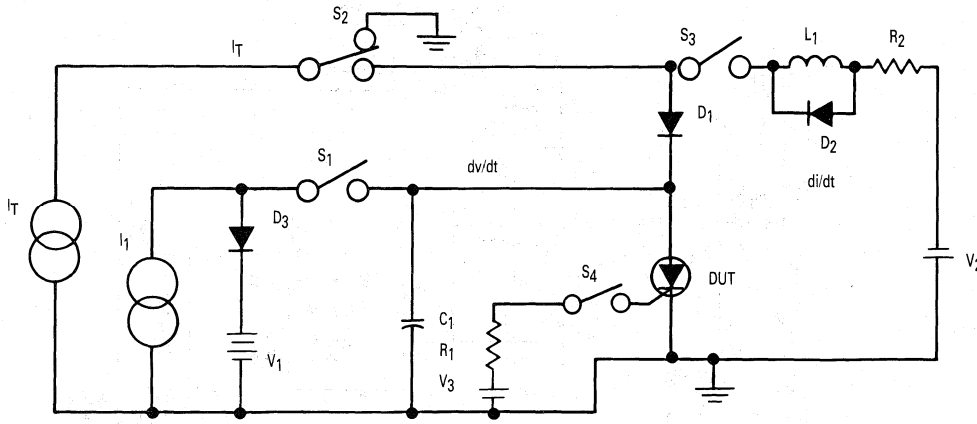


Figure 5.3. Simplified t_q Test Circuit

means of potentiometers R2 and R3, respectively. The output pulse is normally set to straddle the peak of the ac line, which not only makes the power supplies more efficient, but also allows a more consistent oscilloscope display. This pulse shown in waveform A of Figure 5.6 initiates the t_q test, which requires approximately 0.5 ms to assure the device a complete turn on. A fairly low duty cycle results, (approximately 5%) which is important in minimizing temperature effects. The repetitive nature of this test permits easy oscilloscope viewing and allows one to readily "walk in" the dv/dt ramp. This is accomplished by adjusting the appropriate potentiometer (R7) which, every 8.33 ms (every half cycle) will apply the dv/dt ramp at a controlled time delay.

To generate the appropriate system timing delays, four RC integrating network/comparators are used, consisting of op-amps U2, U5 and U6.

Op-amp U2A, along with transistor Q2, opto-coupler U4 and the following transistors Q6 and Q7, provide the gate drive pulse to the DUT (see waveforms B, C and D of Figure 5.6). The resulting gate current pulse is about 50 μs wide and can be selected, by means of switch S2, for an I_{GT} of from about 1 mA to 90 mA. Opto-coupler U4, as well as U1 in the Constant Current Circuit, provide electrical isolation between the power circuitry and the low level circuitry.

The Constant Current Circuit consists of an NPN Darlington Q3, connected as a constant current source driving a PNP tri-Darlington (Darlington Q4, Bipolar Q5). By varying the base voltage of Q3 (with Current Control potentiometer R4), the collector current of Q3 and thus the base voltage of Q4 will also vary. The PNP output transistor Q5 (MJ14003) (rated at 70 A), is also configured as a constant current source with four, parallel connected emitter resistors (approximately 0.04 ohms, 200 W), thus providing as much as 60 A test current. Very briefly, the circuit operates as follows: — CMOS Gate 1E is clocked high, turning on, in order, a) NPN transistor Q16, b) PNP transistor Q1, c) optocoupler U3, and d) transistors Q3, Q4 and Q5. The board mounted Current Set potentiometer R5, sets the maximum output current and R4, the Current Control, is a front panel, multiturn potentiometer.

Time delay for the di/dt Circuit is derived from cascaded op-amps U2B and U5 (waveforms F and G of Figure 5.6). The output gate, in turn, drives NPN transistor Q8, fol-

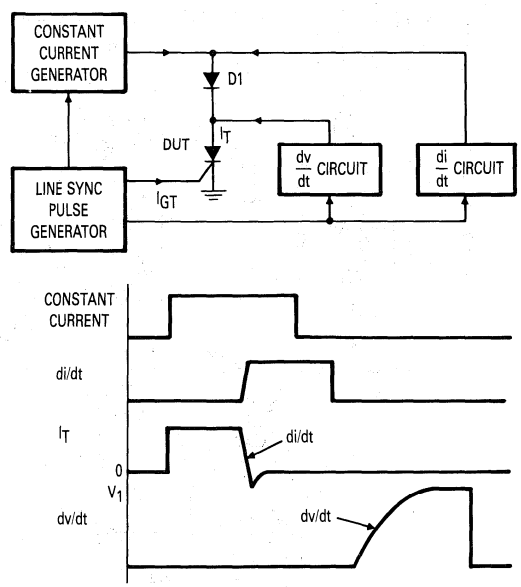


Figure 5.4. Block Diagram of the t_q Test Fixture and Waveforms

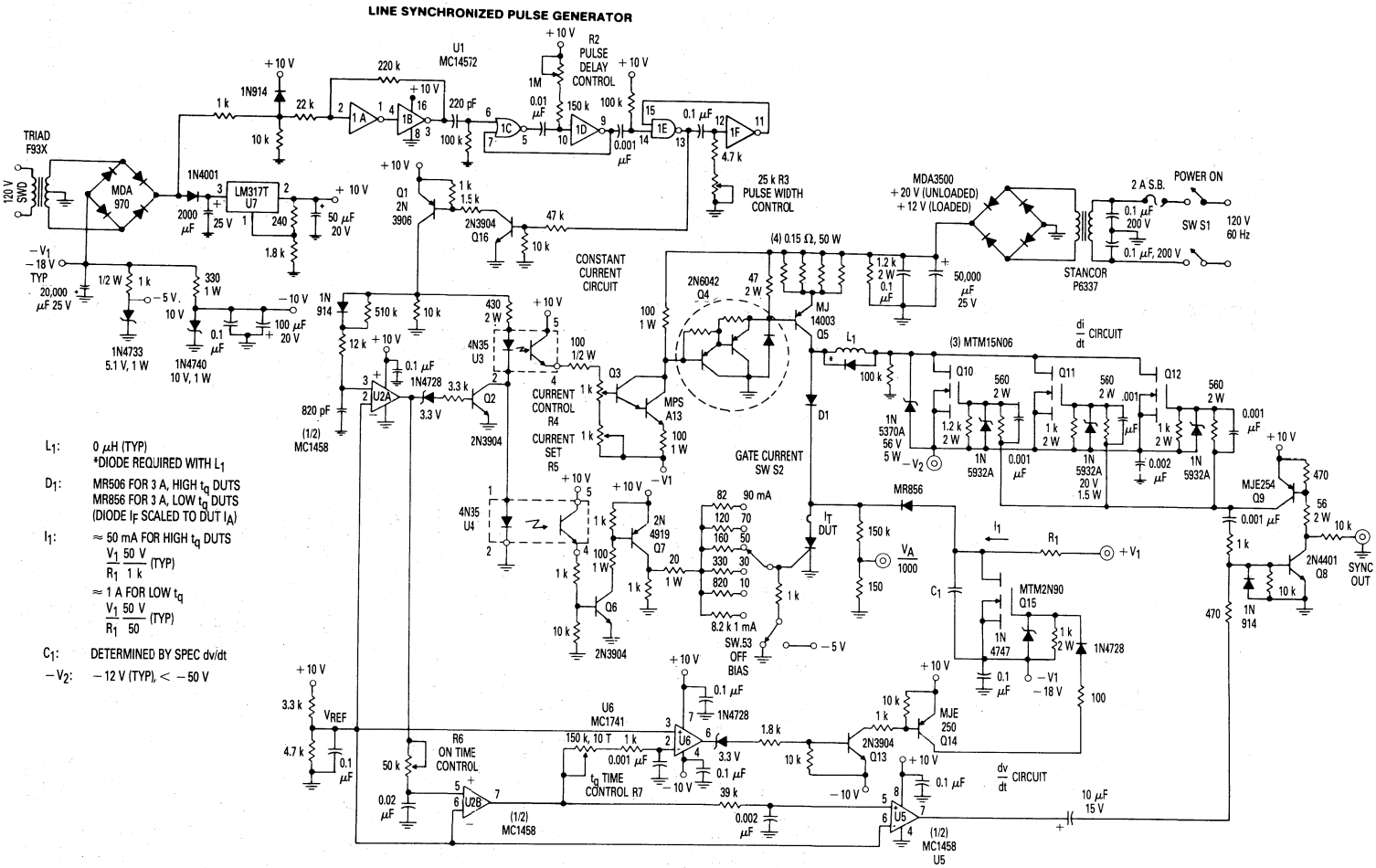


Figure 5.5. t_q Test Fixture

lowed by PNP transistor Q9, whose output provides the gate drive for the three parallel connected N-channel power MOSFET transistors Q10–Q12 (waveform H of Figure 5.6). These three FETs (MTM15N06), are rated at 15 A continuous drain current and 40 A pulsed current and thus can readily divert the maximum 60 A constant current that the Fixture can generate. The results of this diversion from the DUT is described by waveforms E, H and I of Figure 5.6, with the di/dt of I_{TM} dictated by the series inductance L1. For all subsequent testing, the inductor was a shorting bar, resulting in very little inductance and consequently, the highest di/dt (limited primarily by wiring inductance). When a physical inductor L1 is used, a clamp diode, scaled to the diverted current, should be placed across L1 to limit "inductive kicks."

dv/dt CIRCUIT

The last major portion of the Fixture, the dv/dt Circuit, is variable time delayed by the multi-turn, front panel t_q Time Control potentiometer R7, operating as part of an integrator on the input of comparator U6. Its output (waveform J of Figure 5.6) is used to turn-off, in order, a) normally on NPN transistor Q13, b) PNP transistor Q14 and c) N-channel power MOSFET Q15 (waveform L of Figure 5.6). This FET is placed across ramp generating capacitor C1, and when unclamped (turned off), the capacitor is allowed to charge through resistor R1 to the supply voltage $+V_1$. Thus, the voltage appearing on the drain will be an exponentially rising voltage with a dv/dt dictated by R1, C1, whose position in time can be advanced or delayed. This waveform is then applied through a blocking diode to the anode of the DUT for the forward blocking voltage test.

Another blocking diode, D1, also plays an important role in t_q measurements and must be properly selected. Its purpose is to prevent the dv/dt ramp from feeding back into the Current Source and di/dt Circuit and also to momentarily apply a reverse blocking voltage (a function of $-V_2$ of the di/dt circuit) to the DUT. Consequently, D1 must have a reverse recovery time t_{rr} greater than the DUT, but less than the t_q time. When measuring standard recovery SCRs, its selection — fast recovery rectifiers or standard recovery — is not that critical, however, for fast recovery, low t_q SCRs, the diode must be tailored to the DUT to produce accurate results. Also, the current rating of the diode must be compatible with the DUT test current. These effects are illustrated in the waveforms shown in Figure 5.7 where both a fast recovery rectifier and standard recovery rectifier were used in measuring t_q of a standard 2N6058 SCR. Although the di/dt's were the same, the reverse recovery current I_{RM} and t_{rr} were greater with the standard recovery rectifier, resulting in a somewhat shorter t_q (59 μs versus 63 μs). In fact, t_q is affected by the initial conditions (I_{TM}, di/dt, I_{RM}, dv/dt, etc.) and these conditions should be specified to maintain measurement repeatability. This is later described in the published curves and tables.

Finally, the resistor R1 and the resultant current I_1 in the dv/dt circuit must meet certain criteria: I_1 should be greater than the SCR holding current so that when the DUT does indicate t_q limitation, it latches up, thus suppressing the dv/dt ramp voltage; and, for fast SCRs (low t_q), I_1 should be large enough to ensure measurement repeatability. Typical values of I_1 for standard and fast SCRs may be 50 mA and 500 mA, respectively. Obviously, for high forward blocking voltage $+V_1$ tests, the power requirements must be met.

EFFECTS OF GATE BIAS ON t_q

Examples of the effects of I_1 on t_q are listed in Table 5.III whereby standard and fast SCRs were tested with about 50 mA and 1 A, respectively. Note that the low t_q SCR's required fast recovery diodes and high I_1 current.

TEST FIXTURE POWER SUPPLIES

Most of the power supplies for the system are self-contained, including the +12 V supply for the Constant Current Circuit. This simple, unregulated supply furnishes up to 60 A peak pulsed current, primarily due to the line synchronized operation of the system. Power supplies $+V_1$ and $-V_2$, for this exercise, were external supplies, since they are variable, but they can be incorporated in the system. The reverse blocking voltage to the DUT is supplied by $-V_2$ and is typically set for about

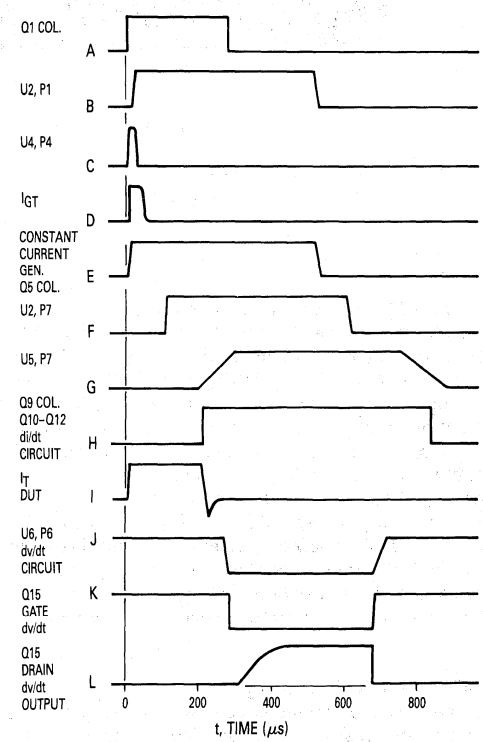
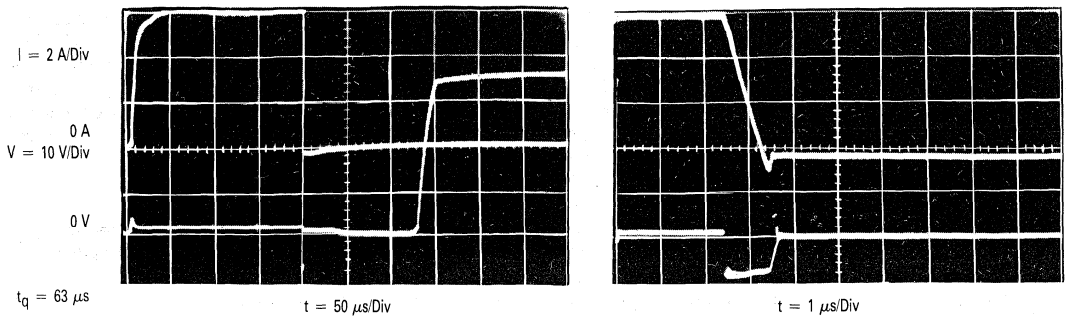
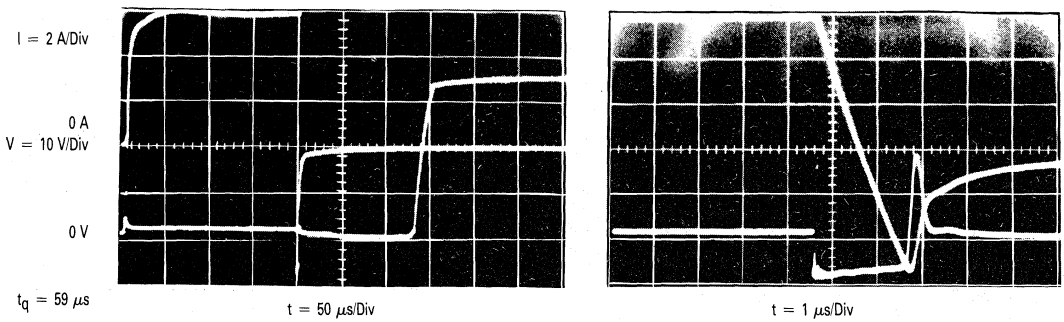


Figure 5.6. t_q Test Fixture System Waveforms



D1 = MR856, FAST RECOVERY RECTIFIER



D1 = MR502, STANDARD RECOVERY RECTIFIER

Figure 5.7. The Effects of Blocking Diode D1 on t_q of a 2N6058 SCR

−10 V to −20 V, being limited to the breakdown voltage of the diverting power MOSFETS (V_{DSS} = 60 V). The +12 V unregulated supply can be as high as +20 V when unloaded; therefore, $-V_2$ (MAX), in theory, would be −40 V but should be limited to less than −36 V due to the 56 V protective Zener across the drain-source of the FETs. Also, $-V_2$ must be capable of handling the peak 60 A, diverting current, if so required.

The reapplied forward blocking voltage power supply $+V_1$, may be as high as the DUT V_{DRM} which conceivably can be 600 V, 1,000 V or greater and, since this supply is on most of the time, must be able to supply the required I_1. Due to the sometimes high power requirements, $+V_1$ test conditions may have to be reduced for extremely fast SCRs.

PARAMETERS AFFECTING t_q

To see how the various circuit parameters can affect t_q, one condition at a time is varied while the others are held constant. The parameters to be investigated are a) forward current magnitude (I_{TM}), b) forward current duration, c) rate of change of turn-off current (di/dt), d) reverse-current magnitude (I_{RM}), e) reverse voltage (V_{RM}), f) rate of reapplied forward voltage (dv/dt), g) magnitude limit of reapplied voltage, h) gate-cathode resistance and i) gate drive magnitude (I_{GT}).

Typical data of this kind, taken for a variety of SCRs, including standard SCRs, high speed SCRs, is condensed and shown in Table 5.1. The data consists of the different conditions which the particular SCR types were subjected to; ten SCRs of each type were serialized and tested to each condition and the ten t_q's were averaged to yield a "typical t_q."

The conditions listed in Column A in Table 5.I, are typical conditions that might be found in circuit operation. Columns B through J in Table 5.I, are in order of increasing t_q; the conditions listed in these columns are only the conditions that were modified from those in Column A and if a parameter is not listed, it is the same as in Column A.

Table 5.I. Parameters Affecting t_q

Device	A	B	C	D	E	F	G	H	I
2N6508 25 A 600 V	R_{GK} = 1 k dv/dt = 15 V/μs I_{TM} = 25 A I_{RM} = 14 A di/dt = −100 A/μs I_{TM} duration = 275 μs I_{GT} = 30 mA	R_{GK} = 100 dv/dt = 2.4 V/μs I_{TM} = 1 A I_{RM} = 1.8 A di/dt = 32 A/μs	R_{GK} = 100 dv/dt = 2.4 V/μs I_{TM} = 2 A I_{RM} = 50 mA di/dt = 0.5 μs	R_{GK} = 100 dv/dt = 2.4 V/μs I_{RM} = 50 mA di/dt = 0.45 A/μs	R_{GK} = 100 dv/dt = 2.4 V/μs	R_{GK} = ∞ dv/dt = 2.4 V/μs	R_{GK} = 100 dv/dt = 2.4 V/μs I_{TM} = 37 A	R_{GK} = 100	I_{GT} = 90 mA
	typ t_q = 68 μs	typ t_q = 42 μs	typ t_q = 45 μs	typ t_q = 49 μs	typ t_q = 60 μs	typ t_q = 64 μs	typ t_q = 64 μs	typ t_q = 65 μs	typ t_q = 68 μs
2N6398 12 A	R_{GK} = 1 k dv/dt = 90 V/μs I_{TM} = 12 A I_{RM} = 11 A di/dt = −100 A/μs I_{TM} duration = 275 μs I_{GT} = 30 mA	R_{GK} = 100 dv/dt = 2.5 V/μs I_{TM} = 1 A I_{RM} = 50 mA di/dt = −0.5 A/μs	R_{GK} = 100 dv/dt = 2.5 V/μs I_{TM} = 1 A I_{RM} = 2.7 A di/dt = 56 A/μs	R_{GK} = 100 dv/dt = 2.5 V/μs I_{RM} = 50 mA di/dt = 32 A/μs	R_{GK} = 100 dv/dt = 2.5 V/μs I_{TM} = 18 A I_{RM} = 50 mA di/dt = 0.3 A/μs	R_{GK} = ∞ dv/dt = 2.5 V/μs I_{RM} = 50 mA di/dt = 0.35 A/μs	R_{GK} = 100	I_{GT} = 90 mA	
	typ t_q = 48 μs	typ t_q = 30 μs	typ t_q = 31 μs	typ t_q = 32 μs	typ t_q = 33 μs	typ t_q = 35.5 μs	typ t_q = 45 μs	typ t_q = 48 μs	

Table 5.I. Continued

Device	A	B	C	D	E	F	G	H	I	J
C106F 4 A	I_{GT} = 1 mA R_{GK} = 1 k dv/dt = 5 V/μs I_{TM} = 4 A I_{RM} = 4 A di/dt = 50 A/μs I_{TM} duration = 275 μs V_{DX} = 50 V	I_{TM} = 2 A I_{RM} = 2.5 A di/dt = −30 A/μs V_{DX} = 50 V	I_{TM} = 6 A I_{RM} = −1 A/μs di/dt = −1 A/μs V_{DX} = 150 V	I_{TM} = 6 A I_{RM} = 0.1 A di/dt = −1 A/μs V_{DX} = 50 V	dv/dt = 1.4 V/μs I_{TM} = 2 A I_{RM} = 0.2 A di/dt = −1.4 A/μs	$-V_2$ = 35 V I_{RM} = 0.2 A di/dt = −1.4 A/μs	I_{RM} = 0.15 A $-V_2$ = 4 V di/dt = −1.4 A/μs	dv/dt = 1.4 V/μs I_{RM} = 0.15 A di/dt = 1.4 A/μs	I_{GT} = 90 mA dv/dt = 1.4 V/μs I_{RM} = 2 A di/dt = −1.4 A/μs	dv/dt = 28 V/μs I_{RM} = 0.15 A di/dt = −1.4 A
	typ t_q = 28 μs	typ t_q = 25 μs	typ t_q = 26 μs	typ t_q = 26 μs	typ t_q = 26 μs	typ t_q = 27 μs	typ t_q = 27 μs	typ t_q = 27 μs	typ t_q = 27 μs	typ t_q = 29 μs
2N6240 4 A	R_{GK} = 1 k dv/dt = 40 V/μs I_{TM} = 4 A I_{RM} = 4 A di/dt = 50 A/μs I_{TM} duration = 275 μs I_{GT} = 1 mA V_{DX} = 50 V	R_{GK} = 100 dv/dt = 1.3 V/μs I_{TM} = 1 A I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA V_{DX} = 150 V	R_{GK} = 100 dv/dt = 1.75 V/μs I_{TM} = 1 A I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA	R_{GK} = 100 dv/dt = 1.75 V/μs I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA	dv/dt = 1.75 V/μs R_{GK} = 100 I_{TM} = 6 A I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA	R_{GK} = 100 I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA	R_{GK} = 100 I_{GT} = 900 mA	R_{GK} = ∞ dv/dt = 1.75 V/μs I_{TM} = 1 A I_{RM} = 50 mA di/dt = −0.5 A/μs I_{GT} = 90 mA	I_{GT} = 90 mA	
	typ t_q = 44.8 μs	typ t_q = 26 μs	typ t_q = 26.2 μs	typ t_q = 27.7 μs	typ t_q = 28.6 μs	typ t_q = 30 μs	typ t_q = 32.7 μs	typ t_q = 37.2 μs	typ t_q = 41.4 μs	
MCR100-6 0.8 A	R_{GK} = 1 k dv/dt = 160 V/μs I_{TM} = 0.8 A I_{RM} = 0.8 A di/dt = 12 A/μs V_{DX} = 50 V I_{TM} duration = 275 μs	dv/dt = 30 V/μs I_{TM} = 0.25 A I_{RM} = 40 mA di/dt = −0.6 A/μs	dv/dt = 30 V/μs Ir = 40 mA di/dt = −0.8 A/μs	$-V_2$ = 9 V I_{RM} = 20 mA di/dt = −0.4 A/μs	$-V_2$ = 1 V Ir = 40 mA di/dt = −0.8 A/μs	dv/dt = 30 V/μs I_{TM} = 1.12 A I_{RM} = 40 mA di/dt = −0.8 A/μs	dv/dt = 30 V/μs I_{TM} = 1.12 A I_{RM} = 40 mA di/dt = −0.8 A/μs V_{DX} = 100 V			
	typ t_q = 14.4 μs	typ t_q = 12.7 μs	typ t_q = 13.5 μs	typ t_q = 13.7 μs	typ t_q = 13.9 μs	typ t_q = 14.4 μs	typ t_q = 14.4 μs			
2N5063 0.8 A	R_{GK} = 1 k dv/dt = 30 V/μs I_{TM} = 0.8 A I_{RM} = 0.8 A di/dt = 12 A/μs I_{TM} duration = 275 μs V_{DX} = 50 V	dv/dt = 5 V/μs I_{TM} = 0.2 A I_{RM} = 50 mA di/dt = −0.6 A/μs	dv/dt = 5 V/μs I_{RM} = 50 mA di/dt = −0.8 A/μs	dv/dt = 5 V/μs I_{TM} = 1.12 A I_{RM} = 50 mA di/dt = −0.8 A/μs	I_{RM} = 40 mA $-V_2$ = 9 V di/dt = −0.45 A/μs	I_{RM} = 40 mA $-V_2$ = 1 V di/dt = −0.8 A/μs	V_{DX} = 100 V dv/dt = 5 V/μs I_{TM} = 1.12 A I_{RM} = 50 mA di/dt = −0.8 A			
	typ t_q = 28.9 μs	typ t_q = 27/μs	typ t_q = 30/μs	typ t_q = 31 μs	typ t_q = 31.2 μs	typ t_q = 31.4 μs	typ t_q = 31.7 μs			
2N5061 0.8 A	dv/dt = 10 V/μs I_{TM} = 0.8 A I_{RM} = 0.8 A di/dt = 18 A/μs I_{TM} duration = 275 μs R_{GK} = 1 k V_{DX} = 30 V	dv/dt = 3.5 V/μs I_{TM} = 0.25 A I_{RM} = 40 mA di/dt = −0.7 A/μs	dv/dt = −3.5 V/μs I_{RM} = 40 mA di/dt = −0.8 A/μs	dv/dt = 3.5 V/μs I_{TM} = 1.12 A I_{RM} = 40 mA di/dt = −0.8 A/μs V_{DX} = 60 V	dv/dt = 3.58/μs I_{TM} = 1.12 A I_{RM} = 40 mA di/dt = −0.7 A/μs	$-V_2$ = 4 V I_{RM} = 20 mA di/dt = −0.2 A/μs	$-V_2$ = 1 V I_{RM} = 40 mA di/dt = −0.8 A/μs			
	typ t_q = 31.7 μs	typ t_q = 19.1 μs	typ t_q = 19/μs	typ t_q = 19.8 μs	typ t_q = 20.2 μs	typ t_q = 30 μs	typ t_q = 30.2 μs			

Table 5.II is a condensed summary of Table 5.I and shows what happens to the t_q of the different devices when a parameter is varied in one direction or the other.

THE EFFECTS OF CHANGING PARAMETERS ON t_q

From Tables 5.I and 5.II, it is clear that some parameters affect t_q more than others. The following discussion describes the effect on t_q of the various parameters.

FORWARD CURRENT MAGNITUDE (I_{TM})

Of the parameters that were investigated, forward-current magnitude and the di/dt rate have the strongest effect on t_q. Varying the I_{TM} magnitude over a realistic range of I_{TM} conditions can change the measured t_q by about 30%. The change in t_q is attributed to varying current densities (stored charge) present in the SCR's junctions as the I_{TM} magnitude is changed. Thus, if a large SCR must have a short t_q when a low I_{TM} is present, a large gate trigger pulse (I_{GT} magnitude) would be advantageous. This turns on a large portion of the SCR to minimize the high current densities that exists if only a small portion of the SCR were turned on (by a weak gate pulse) and the low I_{TM} did not fully extend the turned on region.

In general, the SCR will exhibit longer t_q times with increasing I_{TM}. Increasing temperature also increases the t_q time.

di/dt RATE

Varying the turn-off rate of change of anode current di/dt does have some effect on the t_q of SCRs. Although the increase in t_q versus increasing di/dt was nominal for the SCRs illustrated, the percentage change for the fast SCRs was fairly high (about 30–40%).

REVERSE CURRENT MAGNITUDE (I_{RM})

The reverse current is actually due to the stored charge clearing out of the SCR's junctions when a negative voltage is applied to the SCR anode. I_{RM} is very closely related to the di/dt rate; an increasing di/dt rate causing an increase of I_{RM} and a decreasing di/dt rate causing a lower I_{RM}.

By using different series inductors and changing the negative anode turn-off voltage, it is possible to keep the di/dt rate constant while changing I_{RM}. It was found that I_{RM} has little or no effect on t_q when it is the only variable changed (see Table 5.I C106D, Columns F and G, for example).

Parameter Changed	Device	Columns	1st (μs)	2nd (μs)
I_{GT} Increase	2N6508	AI	68	68
	2N6398	AG	48	48
	2N6240	AI	44.8	41.4
	C106D	HI	27	27
Decrease R_{GK} 1 k to 100 ohms	2N6508	AH	68	65
	2N6398	AG	48	45
	2N6240	GI	41.4	32.7
Increase R_{GK} 1 k to ∞	2N6508	EF	60	64
	2N6398	DF	32	35.5
	2N6240	CH	26.2	37.2
V_{DX}	C106D	DC	26	26
	2N6240	BC	26.2	26
	MCR100-6	FG	14.4	14.4
	2N5063	DG	31	31.7
	2N5061	DE	20.2	19.8
Decrease dv/dt Rate	2N6508	EH	65	60
	C106D	HJ	29	27
	2N6240	DF	30	27.7
Increase I_{TM}	2N6508	EG	60	64
	2N6398	DE	32	33
	C106D	EH	26	27
	2N6240	DC	26.2	27.7
		DE	27.7	28.6
		CE	26.2	28.6
	MCR100-6	CF	13.5	14.4
	2N5063	CD	30.7	31
	2N5061	BE	19.1	20.7

Table 5.II. The Effects of Changing Parameters on t_q

REVERSE ANODE VOLTAGE (V_{RM})

Reverse anode voltage has a strong effect on the I_{RM} magnitude and the di/dt rate, but when V_{RM} alone is varied, with I_{RM} and di/dt held constant, little or no change in t_q time was noticed. V_{RM} must always be within the reverse voltage of the device.

Device	Gate Bias		Conditions		$+V_1$	RI	dv/dt (v/μs)
	0V	−5 V	$-V_2 = -10$ V, $I_F = 3$ A		50 V	1 k/50	2.5/50
	t_q[1]	t_q[2]	Diode DI	dv/dt (V/μs)	Remarks		
2N6508	40 μs	30 μs	Slow MR502	2.5	Slow diode faster than fast diode, (lower t_q)		
2N6240	16 μs	9 μs	Slow	2.5	Slow diode faster. 2.5 V/μs faster than 50 V/μs		
2N6398	30 μs	25 μs	Slow	2.5	Tested slow diode only		
C106D	13 μs	8 μs	Slow	2.5	Tested slow diode only		

Table 5.III. The Effects of Gate Bias on t_q

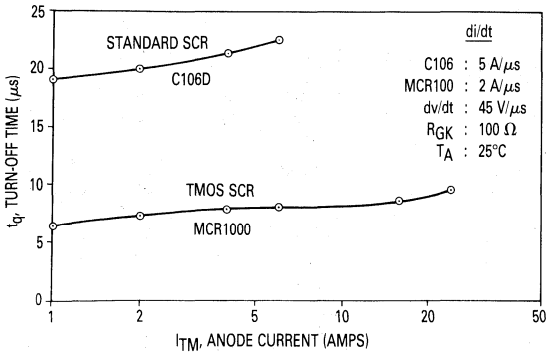


Figure 5.8. Standard SCR & TMOS SCR Turn-Off Time t_q as a Function of Anode Current I_{TM}

REAPPLIED dv/dt RATE

Varying the reapplied dv/dt rate across the range of dv/dt's commonly encountered can vary the t_q of a given SCR by more than 10%. The effect of the dv/dt rate on t_q is due to the Anode-Gate capacitance. The dv/dt applied at the SCR anode injects current into the gate through this capacitance (i_{GT} = C dv/dt). As the dv/dt rate increased, the gate current also increases and can trigger the SCR on. To complicate matters, this injected current also adds to the current due to leakage or stored charge left in the junctions just after turn-off.

The stored charge remaining in the center junction is the main reason for long t_q times and, for the most part, the charge is removed by the recombination process. If the reapplied dv/dt rate is high, more charge is injected into this junction and prevents it from returning to the blocking state, as soon as if it were a slow dv/dt rate. The higher the dv/dt rate, the longer the t_q times will be.

MAGNITUDE LIMIT OF REAPPLIED dv/dt (V_{DX})

Changing the magnitude limit of the reapplied dv/dt voltage has little or no effect on a given SCR's t_q time when the maximum applied voltage is well below the voltage breakdown of the SCR. The t_q times will lengthen if the SCR is being used near its voltage breakdown, since the leakage present near breakdown is higher than at lower voltage levels. The leakage will lengthen the time it takes for the charge to be swept out of the SCR's center junction, thus lengthening the time it takes for this junction to return to the blocking state.

GATE CATHODE RESISTANCE (R_{GK})

In general, the lower the R_{GK} is, the shorter the t_q time will be for a given SCR. This is because low R_{GK} aids in the removal of stored charge in the SCR's junctions. An approximate 15% change in the t_q time is seen by changing R_{GK} from 100 ohms to 1000 ohms for the DUTs.

GATE DRIVE MAGNITUDE (I_{GT})

Changing the gate drive magnitude has little effect on a SCR's t_q time unless it is grossly overdriven or underdriven. When it is overdriven, there is an unnecessary large amount of charge in the SCR's junction. When underdriven, it is possible that only a small portion of the chip at the gate region turns on. If the anode current is not large enough to spread the small turned on region, there is a high current and charge density in this region that consequently lengthens the t_q time.

FORWARD CURRENT DURATION

Forward current duration had no measurable effect on t_q time when varied from 100 μs to 300 μs, which were the limits of the Motorola t_q Tester. Longer I_{TM} durations heat up the SCR which causes temperature effects; very short I_{TM} durations affect the t_q time due to the lack of time for the charges in the SCR's junctions to reach equilibrium, but these effects were not seen in the range tested.

REVERSE GATE BIAS VOLTAGE

As in transistor operation, reverse biasing the gate of the SCR decreases the turn-off time, due to the rapid "sweeping out" of the stored charge. The reduction in t_q for standard SCRs is quite pronounced, approaching perhaps 50% in some cases; for fast SCRs, only nominal improvement might result. Table 5.III shows this effect on six SCRs where the gate bias was set for 0 V and −5 V, respectively (the 1 k gate resistor of the DUT was either grounded or returned to −5 V). Due to the internal, monolithic resistor of most SCRs, the actual reverse bias voltage between the gate-cathode is less than the reverse bias supply.

CHARACTERIZING SCRs FOR CROWBAR APPLICATIONS

The use of a crowbar to protect sensitive loads from power supply overvoltage is quite common and, at the first glance, the design of these crowbars seems like a straightfoward, relatively simple task. The crowbar SCR is selected so as to handle the overvoltage condition and a fuse is chosen at 125 to 250% of the supply's rated full-load line current. However, upon further investigation, other questions and problems are encountered.

How much overvoltage and for how long (energy) can the load take this overvoltage? Will the crowbar respond too slowly and thus not protect the load or too fast resulting in false, nuisance triggering? How much energy can the crowbar thyristor (SCR) take and will it survive until the fuse opens or the circuit breaker opens? How fast will the fuse open, and at what energy level? Can the fuse adequately differentiate between normal current levels — including surge currents — and crowbar short circuit conditions?

It is the attempt of this section to answer these questions — to characterize the load, crowbar, and fuse and thus to match their characteristics to each other.

The type of regulator of most concern is the low voltage, series pass regulator where the filter capacitors to

be crowbarred, due to 60 Hz operation, are relatively large and the charge and energy stored correspondingly large. On the other hand, switching regulators operating at about 20 kHz require smaller capacitors and thus have lower crowbar constraints.

These regulators are quite often line-operated using a high voltage, two-transistor inverter, half bridge or full bridge, driving an output step-down transformer. If a transistor were to fail, the regulator-transformed power would be less and the output voltage would drop, not rise, as is the case for the linear series regulator with a shorted pass transistor. Thus, the need for overvoltage protection of these types of switching regulators is minimized.

This premise, however, does not consider the case of the lower power series switching regulator where a shorted transistor would cause the output voltage to rise. Nor does it take into account overvoltage due to transients on the output bus or accidental power supply hookup. For these types of operations, the crowbar SCR should be considered.

HOW MUCH OVERVOLTAGE CAN THE LOAD TAKE?

Crowbar protection is most often needed when ICs are used, particularly those requiring a critical supply voltage such as TTL or expensive LSI memories and MPUs.

If the load is 5 V TTL, the maximum specified continuous voltage is 7 V. (CMOS, with its wide power supply range of 3 to 18 V, is quite immune to most overvoltage conditions.) But, can the TTL sustain 8 V or 10 V or 15 V and, if so, for how long and for how many power cycles? Safe Operating Area (SOA) of the TTL must be known. Unfortunately, this information is not readily available and has to be generated.

Using the test circuit illustrated in Appendix III, a quasi-SOA curve for a typical TTL gate was generated (Figure 5.9). Knowing this overvoltage-time limit, the crowbar and fuse energy ratings can be determined.

The two possible configurations are illustrated in Figure 5.10, the first case shows the crowbar SCR across the

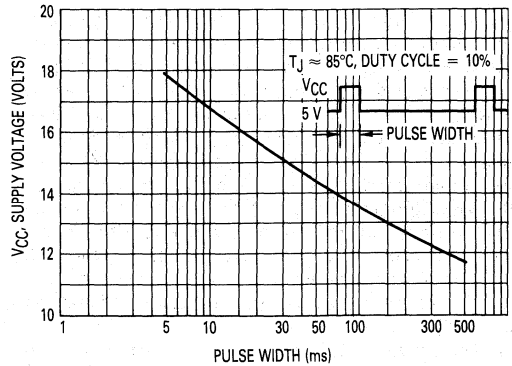
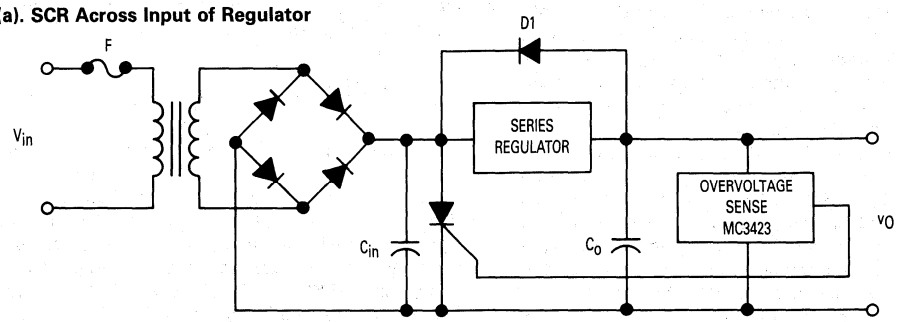
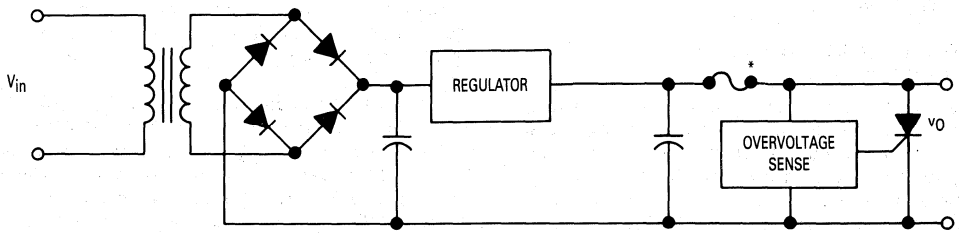


Figure 5.9. Pulsed Supply Voltage versus Pulse Width

(a). SCR Across Input of Regulator



(b). SCR Across Output of Regulator



*NEEDED IF SUPPLY NOT CURRENT LIMITED

Figure 5.10. Typical Crowbar Configurations

input of the regulator and the second, across the output. For both configurations, the overvoltage comparator senses the load voltage at the remote load terminals, particularly when the I_R drop of the supply leads can be appreciable. As long as the output voltage is less than that of the comparator reference, the crowbar SCR will be in an off state and draw no supply current. When an over-voltage condition occurs, the comparator will produce a gate trigger to the SCR, firing it, and thus clamping the regulator input, as in the first case — to the SCRs on-state drop of about 1 to 1.5 V, thereby protecting the load.

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Placing the crowbar across the input filter capacitors, although effectively clamping the output, has several disadvantages.

1. There is a stress placed on the input rectifiers during the crowbarring short circuit time before the line fuse opens, particularly under repeated operation.
2. Under low line conditions, the minimum short circuit current can be of the same magnitude as the maximum primary line current at high line, high load, making the proper fuse selection a difficult choice.
3. The capacitive energy to be crowbarred (input and output capacitor through rectifier D1) can be high.

When the SCR crowbar and the fuse are placed in the dc load circuit, the above problems are minimized. If crowbarring occurs due to an external transient on the line and the regulator's current limiting is working properly, the SCR only has to crowbar the generally smaller output filter capacitor and sustain the limited regulator current.

If the series pass devices were to fail (short), even with current limiting or foldback disabled, the crowbarred energy would generally be less than of the previous case. This is due to the higher impedance of the shorted regulator (due to emitter sharing and current sensing resistors) relative to that of rectifier D1.

Fuse selection is much easier as a fault will now give a greater percentage increase in dc load current than when measuring transformer primary or secondary rms current. The disadvantage, however, of placing the fuse in the dc load is that there is no protection for the input rectifier, capacitor, and transformer, if one of these components were to fail (short). Secondly, the one fuse must protect not only the load and regulator, but also have adequate clearing time to protect the SCR, a situation which is not always readily accomplished. The input circuitry can be protected with the addition of a primary fuse or a circuit breaker.

HOW MUCH ENERGY HAS TO BE CROWBARRED?

This is dictated by the power supply filter capacitors, which are a function of output current. A survey of several linear power supply manufacturers showed the output filter capacitor size to be from about 100 to 400 microfarads per ampere with about 200 μF/A being typical. A 30 A regulator might therefore have a 6000 μF output filter capacitor.

Additionally, the usually much larger input filter capacitor will have to be dumped if the regulator were to short, although that energy to be dissipated will be dependent on the total resistance in the circuit between that capacitor and the SCR crowbar.

The charge to be crowbarred would be

$$Q = CV = I_T,$$

the energy,

$$E = 1/2\ CV^2,$$

and the peak surge current

$$i_{pk} = \frac{V_C}{R_T}$$

When the SCR crowbars the capacitor, the current waveform will be similar to that of Figure 5.11, with the peak surge current, i_{pk}, being a function of the total impedance in the circuit (Figure 5.12) and will thus be limited by the Equivalent Series Resistance (ESR) and inductance (ESL) of the capacitor plus the dynamic impedance of the SCR, any external current limiting resistance, (and inductance) of the interconnecting wires and circuit board conductors.

The ESR of computer grade capacitors, depending on the capacitor size and working voltage, might vary from 10 to 1000 milliohms (mΩ). Those used in this study were in the 25 to 50 mΩ range.

The dynamic impedance of the SCR (the slope of the on-state voltage, on-state current curve), at high currents, might be in the 10 to 20 mΩ range. As an example, from the on-state characteristics of the MCR70, 35 A rms SCR, the dynamic impedance is

$$r_d = \frac{\Delta V_F}{\Delta I_F} = \frac{(4.5 - 3.4)V}{(300 - 200)A} = \frac{1.1\ V}{100\ A} \cong 11\ m\Omega.$$

The interconnecting wire might offer an additional 5 mΩ (#20 solid copper wire $\cong$ 20 mΩ/ft) so that the total circuit resistance, without additional current limiting, might be in the 40 to 70 mΩ range. The circuit inductance was considered low enough to ignore so far as i_{pk} is concerned for this exercise, being in hundreds of nanohenry range (ESL $\cong$ 3 nH, L wire $\cong$ 500 nH/ft). However, di/dt will be affected by the inductance.

HOW MUCH ENERGY CAN THE CROWBAR SCR SUSTAIN?

There are several factors which contribute to possible SCR failures or degradation — the peak surge current, di/dt, and a measure of the device's energy capability, I^2t.

If the peak current and/or duration of the surge is large, destruction of the device due to excessive dissipation can occur. Obviously, the i_{pk} can be reduced by inserting additional impedance in the crowbar path, at an increase in dump time. However, this time, which is a measure of how long the overvoltage is present, should be within the SOA of the load.

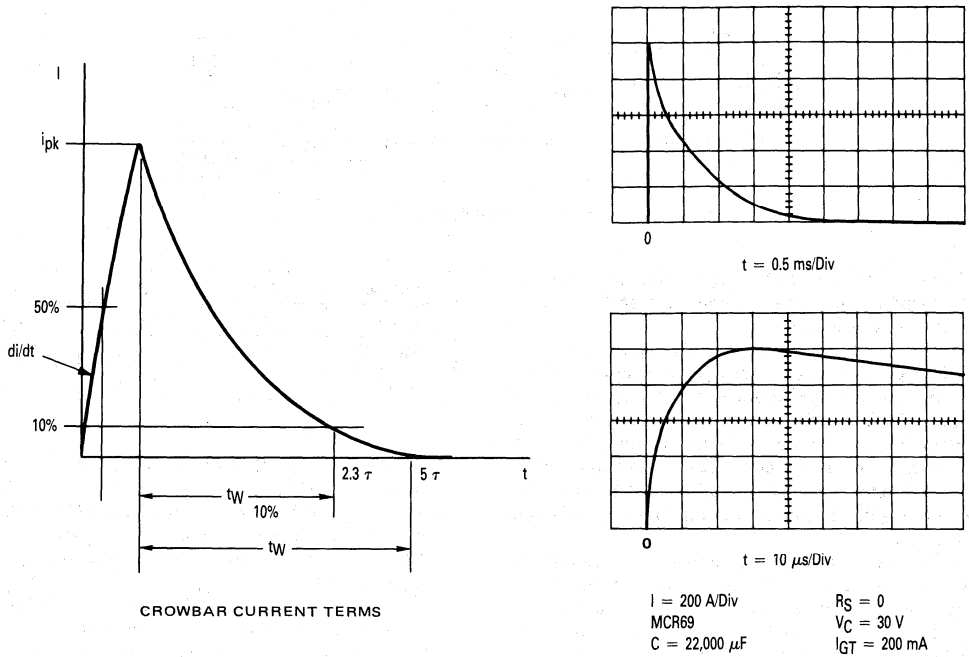

Figure 5.11. Typical SCR Crowbar Waveform

The energy stored in the capacitor being a constant for a particular voltage would suggest that the I^2t integral for any limiting resistance is also a constant. In reality, this is not the case as the thermal response of the device must be taken into consideration. It has been shown that the dissipation capability of a device varies as to the $\sqrt{t}$ for the first tens of milliseconds of the thermal response and, in effect, the measure of a device's energy capability would be closer to $i^2\sqrt{t}$. This effect is subsequently illustrated in the empirically derived i_{pk} versus time derating curves being a non-linear function. However, for comparison with fuses, which are rated in I^2t, the linear time base, "t," will be used.

The di/dt of the current surge pulse is also a critical parameter and should not exceed the device's ratings (typically about 200 A/μs for 50 A or less SCRs). The magnitude of di/dt that the SCR can sustain is controlled by the device construction and, to some extent, the gate drive conditions. When the SCR gate region is driven on, conduction across the junction starts in a small region and progressively propagates across the total junction. Anode current will initially be concentrated in this small conducting area, causing high current densities which can degrade and ultimately destroy the device. To minimize this di/dt effect, the gate should be turned on hard and fast such that the area turned on is initially maximized. This can be accomplished with a gate current pulse approaching five times the maximum specified continuous gate current, I_{gt}, and with a fast rise time (< 1 μs). The gate current pulse width should be greater than the propagation time; a figure of 10 μs minimum should satisfy most SCRs with average current ratings under 50 A or so.

The wiring inductance alone is generally large enough to limit the di/dt. Since most SCRs are good for over 100 A/μs, this effect is not too large a problem. However, if the di/dt is found excessive, it can be reduced by placing

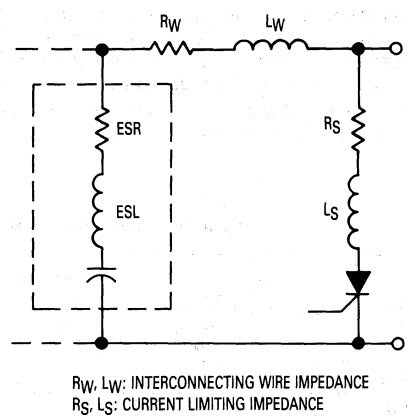

Figure 5.12. Circuit Elements Affecting SCR Surge Current

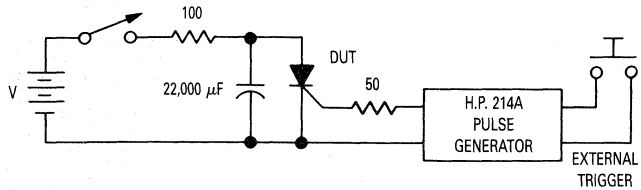

Figure 5.13

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an inductance in the loop; but, again, this increases the circuit's response time to an overvoltage and the trade-off should be considered.

Since many SCR applications are for 60 Hz line operation, the specified peak non-repetitive surge current I_{TSM} and circuit fusing I^2t are based on 1/2 cycle (8.3 ms) conditions. For some SCRs, a derating curve based on up to 60 or 100 cycles of operation is also published. This rating, however, does not relate to crowbar applications. To fully evaluate a crowbar system, the SCR must be characterized with the capacitor dump exponential surge current pulse.

A simple test circuit for deriving this pulse is shown in Figure 5.13, whereby a capacitor is charged through a limiting resistor to the supply voltage, V, and then the charge is dumped by the SCR device under test (DUT). The SCR gate pulse can be varied in magnitude, pulse width, and rise time to produce the various I_{GT} conditions. An estimate of the crowbar energy capability of the DUT is determined by first dumping the capacitor charged to a low voltage and then progressively increasing the voltage until the DUT fails. This is repeated for several devices to establish an average and minimum value of the failure points cluster.

This procedure was used to test several different SCRs of which the following Table 5.IV describes several of the pertinent energy specifications and also the measured crowbar surge current at the point of device failure.

This one-shot destruct test was run with a gate current of five $I_{GT(MAX)}$ and a 22,000 μF capacitor whose ESR produced the exponentially decaying current pulse about 1.5 ms wide at its 10% point. Based on an appropriate derating, ten devices of each line where then successfully tested under the following conditions.

Device	V_C	i_{pk}	t
MCR68	12 V	250 A	1.5 ms
MCR69	30 V	800 A	1.5 ms
MCR70	30 V	800 A	1.5 ms

To determine the effect of gate drive on the SCRs, three devices from each line were characterized at non-destruct levels using three different capacitors (200, 6,000, and 22,000 μF), three different capacitor voltages (10, 20, and 30 V), and three different gate drives ($I_{GT(MAX)}$, 5 $I_{GT(MAX)}$, and a ramp $I_{GT(MAX)}$ with a di/dt of about 1 mA/μs). Due to its energy limitations, the MCR68 was tested with only 10 V across the larger capacitors.

The slow ramp, I_{GT}, was used to simulate overvoltage sense applications where the gate trigger rise time can be slow such as with a coupling zener diode.

No difference in SCR current characteristics were noted with the different gate current drive conditions; the peak currents were a function of capacitor voltage and circuit impedance, the fall times related to R_TC, and the rise times, t_r, and di/dt, were more circuit dependent (wiring inductance) and less device dependent (SCR turn-on time, t_{on}). Since the wiring inductance limits, t_r, the effect of various I_{GT}s was masked, resulting in virtually identical waveforms.

The derated surge current, derived from a single (or low number) pulse test, does not truly reflect what a power supply crowbar SCR might have to see over the life of the supply. Life testing over many cycles have to be performed; thus, the circuit described in Appendix IV was developed. This life test fixture can simultaneously

Table 5.IV. Specified and Measured Current Characteristics of Three SCRs

Device	Case	Maximum Specified Values					Measured Crowbar Surge Current I_{pk}		
		$I_{T(rms)}$ (A)	$I_{T(AV)}$ (A)	I_{TSM}* (A)	I^2t (A^2s)	$I_{GT(Max)}$ (mA)	Min (A)	Max (A)	Ave (A)
MCR68	TO-220	12	8	100	40	30	380	750	480
MCR69	TO-220	25	16	300	375	30	1050	1250	1100
MCR70	TO-208	35	22	350	510	30	1100	1300	1200

*I_{TSM} = Peak Non-Repetitive Surge Current, 1/2 cycle sine wave, 8.3 ms.

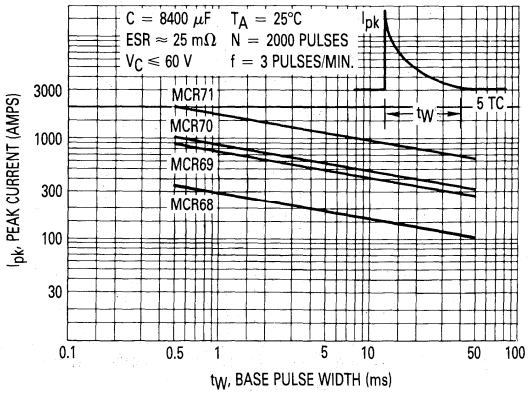
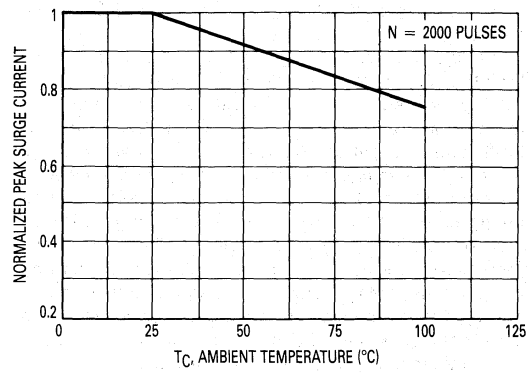


Figure 5.14(a). Peak Surge Current versus Pulse Width



(b). Peak Surge Current versus Ambient Temperature

test ten SCRs under various crowbar energy and gate drive conditions.

Each of the illustrated SCRs of Figure 5.14(a) were tested with as many as four limiting resistors (0, 50, 100, and 240 mΩ) and run for 1000 cycles at a nominal energy level. If no failures occurred, the peak current was progressively increased until a failure(s) resulted. Then the current was reduced by 10% and ten new devices were tested for 2000 cycles (about six hours at 350 cycles/hour). If this test proved successful, the data was further derated by 20% and plotted as shown on log-log paper with a slope of $-1/4$. This theoretical slope, due to the $I^2\sqrt{t}$ one-dimensional heat-flow relationship (see Appendix VI), closely follows the empirical results. Of particular interest is that although the peak current increases with decreasing time, as expected, the I^2t actually decreases.

Figure 5.14(b) shows the effect of elevated ambient temperature on the peak current capability of the illustrated SCRs.

FUSE CHARACTERISTICS

SCRs, like rectifiers, are generally rated in terms of average forward current, $I_{T(AV)}$, due to their half-wave operation. Additionally, an rms forward current, $I_{T(rms)}$, a peak forward surge current, I_{TSM}, and a circuit-fusing energy limit, I^2t, may be shown. However, these specifications, which are based one-half cycle 60 Hz operation, are not related to the crowbar current pulse and some means must be established to define their relationship. Also, fuses which must ultimately match the SCR and the load, are rated in rms currents.

The crowbar energy curves are based on an exponentially decaying surge current waveform. This can be converted* to I_{rms} by the equation.

$$I_{rms} = 0.316\ i_{pk}$$

which now allows relating the SCR to the fuse.

*See Appendix V

The logic load has its own overvoltage SOA as a function of time (Figure 5.9). The crowbar SCR must clamp the overvoltage within a specified time, and still be within its own energy rating; thus, the series-limiting resistance, R_S, in the crowbar path must satisfy both the load and SCR energy limitations. The overvoltage response time is set by the total limitations. The overvoltage response time is set by the total limiting resistance and dumped capacitor(s) time constant. Since the SOA of the TTL used in this exercise was derived by a rectangular overvoltage pulse (in effect, over-energy), the energy equivalent of the real-world exponentially falling voltage waveform must be made. An approximation can be made by using an equivalent rectangular pulse of 0.7 times the peak power and 0.7 times the base time.

Once an overvoltage is detected and the crowbar is enabled, in addition to sustaining the peak current, the SCR must handle the regulator short-circuit current for the time it takes to open the fuse.

Thus, all three elements are tied together — the load can take just so much overvoltage (over-energy) and the crowbar SCR must repeatedly sustain for the life of the equipment an rms equivalent current pulse that lasts for the fuse response time.

It would seem that the matching of the fuse to the SCR would be straightforward — simply ensure that the fuse rms current rating never exceed the SCR rms current rating (Figure 5.15), but still be sufficient to handle steady-state and normal overload currents. The more exact relationship would involve the energy dissipated in the system $\int I^2R dt$, which on a comparative basis, can be reduced to I^2t. Thus, the "let-through" I^2t of the fuse should not exceed I^2t capability of the SCR under all operating conditions. These conditions are many, consisting of "available fault current," power factor of the load, supply voltage, supply frequency, ambient temperature, and various fuse factors affecting the I^2t.

There has been much detailed information published on fuse characteristics and, rather than repeat the text which would take many pages, the reader is referred to

those sources. Instead, the fuse basics will be defined and an example of matching the fuse to the SCR will be shown.

In addition to interrupting high current, the fuse should limit the current, thermal energy, and overvoltage due to the high current. Figure 5.16 illustrates the condition of the fuse at the moment the over-current starts. The peak let-through current can be assumed triangular in shape for a first-order approximation, lasting for the clearing time of the fuse. This time consists of the melting or pre-arcing time and the arcing time. The melting time is an inverse function of over-current and, at the time that the fuse element is opened, an arc will be formed causing the peak arc voltage. This arc voltage is both fuse and circuit dependent and under certain conditions can exceed the peak line voltage, a condition the user should ensure does not overstress the electronics.

The available short-circuit current is the maximum current the circuit is capable of delivering and is generally limited by the input transformer copper loss and reactance when the crowbar SCR is placed at the input to the regulator or the regulator current limiting when placed at the output. For a fuse to safely protect the circuit, it should limit the peak let-through current and clear the fault in a short time, usually less than 10 ms.

Fuse manufacturers publish several curves for characterizing their products. The current-time plot, which describes current versus melting time (minimum time being 10 ms), is used in general industrial applications, but is not adequate for protecting semiconductors where the clearing time must be in the subcycle range. Where protection is required for normal multicycle overloads, this curve is useful.

Two other useful curves, the total clearing I^2t characteristic and the peak let-through current I_{PLT} characteristic, are illustrated in Figures 5.17 and 5.18 respectively. Some vendors also show total clearing time curves (overlayed on Figure 5.17 as dotted lines) which then allows direct comparison with the SCR energy limits. When this clearing time information is not shown, then the designer should determine the I_{PLT} and I^2t from the respective curves and then solve for the clearing time from the approximate equation relating these two parameters.

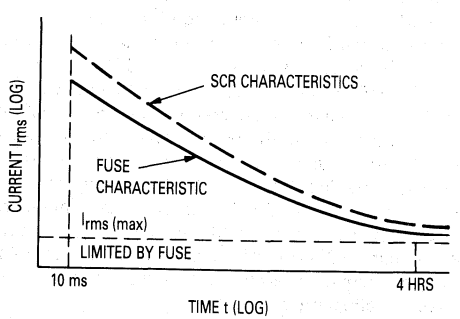


Figure 5.15. Time-Current Characteristic Curves of a Crowbar SCR and a Fuse

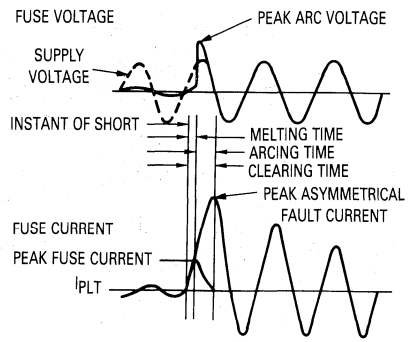


Figure 5.16. Typical Fuse Timing Waveforms During Short Circuit

Assuming a triangular waveform for I_{PLT}, the total clearing time, t_c, would then approximately be

$$t_c \approx \frac{3\ I^2t}{I_{PLT}^2}$$

Once t_c of the fuse is known, the comparison with the SCR can readily be made. As long as the I^2t of the fuse is less than the I^2t of the SCR, the SCR is protected. It should be pointed out that these calculations are predicated on a known value of available fault current. By inspection of Figure 5.18, it can be seen that I_{PLT} can vary greatly with available fault current, which could have a marked effect on the degree of protection. Also, the illustrated curves are for particular operating conditions; the curves will vary somewhat with applied voltage and frequency, initial loading, load power factor, and ambient temperature. Therefore, the reader is referred to the manufacturer's data sheet in those cases where extrapolation will be required for other operating conditions. The final proof is obtained by testing the fuse in the actual circuit under worst-case conditions.

CROWBAR EXAMPLE

To illustrate the proper matching of the crowbar SCR to the load and the fuse, consider the following example. A 50 A TTL load, powered by a 60 A current limited series regulator, has to be protected from transients on the supply bus by crowbarring the regulator output. The output filter capacitor of 10,000 μF (200 μF/A) contributes most of the energy to be crowbarred (the input capacitor is current limited by the regulator). The transients can reach 18 V for periods of 100 ms.

Referring to Figure 5.9, it is seen that this transient exceeds the empirically derived SOA. To ensure safe operation, the overvoltage transient must be crowbarred within 5 ms. Since the TTL SOA is based on a rectangular power pulse even though plotted in terms of voltage, the equivalent crowbarred energy pulse should also be derived. Thus, the exponentially decaying voltage waveform should be multiplied by the exponentially decaying current to result in an energy waveform proportional to

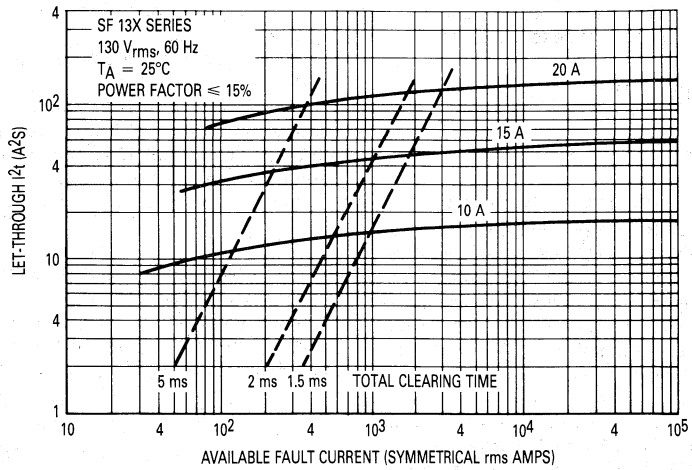


Figure 5.17. Maximum Clearing I^2t Characteristics for 10 to 20 A Fuses

e^{-2x}. The rectangular equivalent will have to be determined and then compared with the TTL SOA. However, for simplicity, by using the crowbarred exponential waveform, a conservative rating will result.

To protect the SCR, a fuse must be chosen that will open before the SCR's I^2t is exceeded, the current being the regulator limiting current which will also be the available fault current to the fuse.

The fuse could be eliminated by using a 60 A SCR, but the cost versus convenience trade-off of not replacing the fuse is not warranted for this example. A second fuse or circuit breaker will protect the rectifiers and regulator for internal faults (shorts), but its selection, which is based on the respective energy limits of those components, is not part of this exercise.

If a crowbar discharge time of 3 ms were chosen, it would not only be within the rectangular pulsed SOA, but also be well within the derived equivalent rectangular model of the exponential waveform. It would also require about 1.3 time constants for the overvoltage to decay from 18 V to 5 V; thus, the RC time constant would be 3 ms/1.3 or 2.3 ms.

The limiting resistance, R_S would simply be

$$R_S = \frac{2.3\ \text{ms}}{10{,}000\ \mu\text{F}} = 0.23\ \Omega \cong 0.2\ \Omega$$

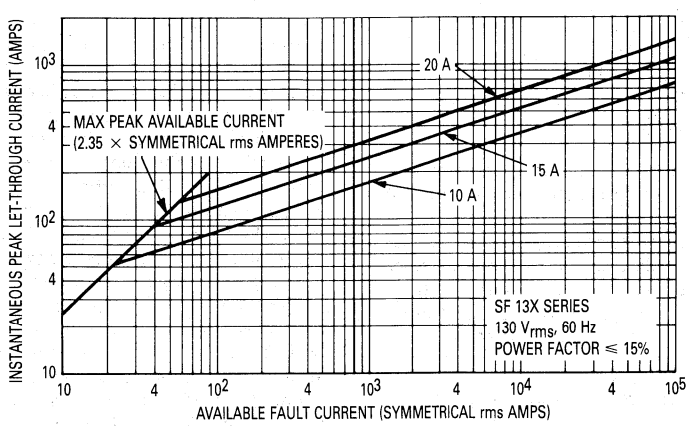


Figure 5.18. Peak Let-Through Current versus Fault Current for 10 to 20 A Fuses

Since the capacitor quickly charges up to the overvoltages V_{CC1} of 18 V, the peak capacitor discharge current would be

$$I_{pk} = \frac{V_{CC1}}{R_S} = \frac{18\ V}{0.2\ \Omega} = 90\ A$$

The rms current equivalent for this exponentially decaying pulse would be

$$I_{rms} = 0.316\ I_{pk} = 0.316(90) = 28.4\ A\ rms$$

Now referring to the SCR peak current energy curves (Figure 5.14), it is seen that the MCR68 can sustain 210 A peak for a base time of 3 ms. This 12 A SCR must also sustain the 60 A regulator limited current for the time required to open the fuse. The MCR68 has a specified peak forward surge current rating of 100 A (1/2 cycle, sine wave, 60 Hz, non-repetitive) and a circuit fusing rating of 40 A^2s.

The non-repetitive rating implies that the device can sustian 100 occurrences of this 1/2 cycle surge over the life of the device; the SCR crowbar surge current curves were based on 2000 cycles.

For the 3 ms time frame, the $I_1{}^2t_1$ for the exponential waveform is

$$I_1{}^2t_1 = (28.4\ A)^2(3\ ms) = 2.4\ A^2s$$

Assuming that the fuse will open within 6 ms, the approximate energy that the SCR must sustain would be 60 A for an additional 3 ms. By superposition, this would amount to

$$I_2{}^2t_2 = (60\ A)^2(6\ ms) = 21.6\ A^2s$$

which, when added to the exponential energy, would result in 24 A^2.

The MCR68 has a 40 A^2s rating based on a 1/2 cycle of 8.3 ms. Due to the one-dimensional heat flow in the device, the energy capability is not linearly related to time, but varies as to the $\sqrt{t}$. Therefore, with a 6 ms 1/2-cycle sine wave, the 40 A^2t rating would now decrease to approximately (see Appendix VI for derivation).

$$I_2{}^2t_2 = I_1{}^2t_1\left(\frac{t_2}{t_1}\right)^{1/2}$$
$$= 40\ A^2s\left(\frac{6\ ms}{8.3\ ms}\right)^{1/2}$$
$$= 34\ A^2s$$

Although the 1/2 cycle extrapolated rating is greater than the actual crowbar energy, it is only characterized for 100 cycles of operation.

To ensure 2000 cycles of operation, at a somewhat higher cost, the 25 A MCR69 could be chosen. Its exponential peak current capability, at 3 ms, is about 560 A and has a specified I_{TSM} of 300 A for 8.3 ms. The I^2t rating is not specified, but can be calculated from the equation

$$I^2t = \frac{(I_{TSM})^2}{2}t = \frac{(300\ A)^2}{2}(8.3\ ms) = 375\ A^2s$$

Extrapolating to 6 ms results in about 318 A^2s, an I^2t rating much greater than the circuit 24 A^2s value.

The circuit designer can then make the cost/performance trade-offs.

All of these ratings are predicated on the fuse opening within 6 ms.

With an available fault current of 60 A, Figure 5.17 shows that a 10 A (SF13X series) fuse will have a let-through I^2t of about 10 A^2 s and a total clearing time of about 6 ms, satisfying the SCR requirements, that is,

$$I^2t\ fuse < I^2t\ SCR$$
$$t_c \leq 6\ ms$$

Figure 5.18 illustrates that for the same conditions, instantaneous peak let-through current of about 70 A would result. For fuse manufacturers that don't show the clearing time information, the approximate time can be calculated from the triangular model, as follows

$$t_c = \frac{3\ I^2t}{I_{PLT}{}^2} = \frac{3(10)}{(70)^2} = 6.1\ ms$$

The fuse is now matched to the SCR which is matched to the logic load. Other types of loads can be similarly matched, if the load energy characteristics are known.

CHARACTERIZING SWITCHES AS LINE-TYPE MODULATORS

In the past, hydrogen thyratrons have been used extensively as discharge switches for line type modulators. In general, such devices have been highly satisfactory from an electrical performance standpoint, but they have some major drawbacks including relatively large size and weight, low efficiency (due to filament power requirements), and short life expectancy compared with semiconductor devices, now can be eliminated through the use of silicon controlled rectifiers.

A line type modulator is a modulator whose output-pulse characteristics are determined by a lumped-constant transmission line (pulse forming network) and by the proper match of the line impedance (PFN) to the load impedance.

A switch for this type modulator should only initiate conduction and should have no effect on pulse characteristics. This is in contrast to a hard switch modulator where output pulse characteristics are determined by the "hard" relationship of grid (base) control of conduction through a vacuum tube (transistor) switch.

Referring to the schematic (Figure 5.25), when the power supply is first turned on, no charge exists in the PFN, and energy is transferred from the power supply to the PFN via the resonant circuit comprising the charging choke and PFN capacitors. At the time that the voltage across the PFN capacitors reaches twice the power supply voltage, current through the charging choke tries to reverse and the power supply is disconnected due to the back biased impedance of the hold-off diode. If we assume this diode to be perfect, the energy remains stored in the PFN until the discharge switch is triggered to its on state. When this occurs, assuming that the pulse transformer has been designed to match the load impedance to the PFN impedance, all energy stored in the PFN

reactance will be transferred to the load if we neglect switch losses. Upon completion of the transfer of energy the switch must return to its off condition before allowing transfer of energy once again from the power supply to the PFN storage element.

OPTIMUM SWITCH CHARACTERISTICS

FORWARD BREAKOVER VOLTAGE

Device manufacturers normally apply the variable-amplitude output of a half-wave rectifier across the SCR. Thus, forward voltage is applied to the device for only a half cycle and the rated voltage is applied only as an ac peak. While this produces a satisfactory rating for ac applications, it does not hold for dc.

An estimated 90% of devices tested for minimum breakover voltage (V_{BO}) in a dc circuit will not meet the data sheet performance specifications. A switch designed for the pulse modulator application should therefore specify a minimum continuous forward breakover voltage at rated maximum leakage current for maximum device temperatures.

THE OFF SWITCH

The maximum forward leakage current of the SCR must be limited to a low value at maximum device temperature. During the period of device nonconduction it is desired that the switch offer an off impedance in the range of megohms to hundreds of megohms. This is required for two reasons: (1) to prevent diminishing the efficiency of recharge by an effective shunt path across the PFN, and (2) to prevent the bleeding off of PFN charge during the interpulse period. This second factor is especially important in the design of radar tansponders wherein the period between interrogations is variable. Change of the PFN voltage during the interpulse period could result in frequency shift, pulse instabilities, and loss of power from the transmitter being modulated.

THE ON SWITCH

At present, SCR design is more limited in the achievable maximum forward sustaining voltage than in the current that the device will conduct. For this reason modulators utilizing SCRs can be operated at lower impedance levels than comparable thyratron circuits of yesterday. It is not uncommon for the characteristic impedance of the pulse forming network to be in the order of 5 to 10 ohms or less. Operating the SCR at higher current to switch the same equivalent pulse power as a thyratron requires the SCR on impedance to be much lower so that the I^2R loss is a reasonable value, in order to maintain circuit efficiency. Low switch loss, moreover, is mandatory because internal power dissipation can be directly translated into junction-temperature-rise and associated leakage current increase which, if excessive, could result in thermal runaway.

TURN-ON TIME

In radar circuits the *pulse-power* handling capability of an SCR, rather than the normally specified *average-power* capability, is of primary importance.

For short pulses at high PRFs the major portion of semiconductor dissipation occurs during the initial turn-on during the time that the anode rises from its forward leakage value to its maximum value. It is necessary, therefore, that turn-on time be as short as possible to prevent excessive power dissipation.

The function of radar is to provide distance information measured as a function of time. It is important, therefore, that any delay introduced by a component be fixed in relation to some variable parameter such as signal strength or temperature. For radar pulse modulator applications, a minimal delay variation versus temperature is required and any such variation must be repetitive from SCR to SCR, in production lots, so that adequate circuit compensation may be provided.

PULSE GATE CURRENT TO FIRE

The time of delay, the time of rise, and the delay variation versus temperature associated with SCR turn-on are functions of the gate triggering current available and the trigger pulse duration. In order to predict pulse circuit operation of the SCR, the pulse gate current required to turn the device on when switching the low-impedance modulator should be specified and the limits of turn-on-time variation for the specified pulse trigger current and collector load should be given at the high and low operating temperature extremes.

RECOVERY TIME

After the cessation of forward conducting current in the on device, a time of SCR circuit isolation must be provided to allow the semiconductor to return to its off state. Recovery time cannot be given as an independent parameter of device operation, but must include factors as determined by the external circuit, such as: (1) pulse current and rate of decay; (2) availability of an inverse voltage immediately following pulse-current conduction; (3) level of base bias following pulse current conduction; (4) rate of rise of reapplied positive voltage and its amplitude in relation to SCR breakover voltage; and (5) maximum circuit ambient temperature.

In the reverse direction the controlled rectifier behaves like a conventional silicon diode. Under worst circuit conditions, if an inverse voltage is generated through the existence of a load short circuit, the current available will be limited only by the impedance of the pulse forming network and SCR inverse characteristics. The reverse current is able to sweep out some of the carriers from the SCR junctions. Intentional design of the load impedance to something less than the network impedance allows development of an inverse voltage across the SCR imme-

diately after pulse conduction, enhancing switch turn-off time. Careful use of a fast clamp diode in series with a fast zener diode, the two in shunt across the SCR, allows application of a safe value of circuit-inverse-voltage without preventing the initial useful reverse current. Availability of a negative base-bias following pulse current conduction provides a similar enhancement of switch turn-off time.

If removal of carriers from the SCR junction enables a faster switch recovery time, then, conversely, operation of the SCR at high temperatures with large forward currents and with slow rate of current decay all increase device recovery time.

1

HOLDING CURRENT

One of the anomalies that exist in the design of a pulse SCR is the requirement for a high holding current. This need can be determined by examining the isolation component that disconnects the power supply from the discharge circuit during the time that PFN energy is being transferred to the transmitter and during the recovery time of the discharge switch. An inductance resonating with the PFN capacitance at twice the time of recharge is normally used for power supply isolation. Resonant charging restricts the initial flow of current from the power supply, thereby maximizing the time at which power supply current flow will exceed the holding current of the SCR. If the PFN recharge current from the power supply exceeds the holding current of the SCR before it has recovered, the SCR will again conduct without the application of a trigger pulse. As a result continuous conduction occurs from the power supply through the low impedance path of the charging choke and on switch. This lock-on condition can completely disable the equipment employing the SCR switch.

The charging current passed by the inductance is given as (the PFN inductance is considered negligible):

$$i_c(t) = \frac{E_{bb} - V_n(0)}{\sqrt{L_c/C_n}} \left(\frac{\cos \dfrac{T_r - 2t}{2\sqrt{L_c C_n}}}{\sin \dfrac{T_r}{2\sqrt{L_c C_n}}} \right)$$

Where

E_{bb} = power supply voltage
$V_n(0)$ = 0 volts if the PFN employs a clamp diode or is matched to the load
T_r = time of resonant recharge and is usually equal to $\frac{1}{PRF}$
L_c = value of charging inductance
C_n = value of total PFN capacity

For a given radar pulse modulator design, the values of power supply voltage, time of resonant recharge, charging choke inductance, and PFN capacitance are established. If the time (t) represents the recovery time of the SCR being used as the discharge switch, i_c then represents the minimum value of holding current required by the SCR to prevent power supply lock-on. Conversely, if the modulator design is about an existing SCR where holding current, recovery time, and forward breakover voltage are known, the charge parameters can be derived by rewriting the above formula as follows:

$$I_H = \frac{V_{BO} - V_n(0)}{\sqrt{L_c/C_n}} \left(\frac{\cos \dfrac{T_r - 2(\text{recovery time})}{2\sqrt{L_c C_n}}}{\sin \dfrac{T_r}{2\sqrt{L_c C_n}}} \right)$$

The designer may find that for the chosen SCR the desired characteristics of modulator pulse width and pulse repetition frequency are not obtainable.

One means of increasing the effective holding current of an SCR is for the semiconductor to exhibit some turn-off gain characteristic for the residual current flow at the end of the modulator pulse. The circuit designer then can provide turn-off base current, making the SCR more effective as a pulse circuit element.

THE SCR AS A UNIDIRECTIONAL SWITCH

When tirggered to its on state, the SCR, like the hydrogen thyratron, is capable of conducting current in one direction. A load short circuit could result in an inverse voltage across the SCR due to the reflection of voltage from the pulse forming network. The circuit designer may wish to provide an intentional load-to-PFN mismatch such that some inverse voltage is generated across the SCR to enhance its turn-off characteristics. Nevertheless, since the normal circuit application is unidirectional, the semiconductor device designer could take advantage of this fact in restricting the inverse-voltage rating that the SCR must withstand. The circuit designer, in turn, can accommodate this lack of peak-inverse-voltage rating by use of a suitable diode clamp across the PFN or across the SCR.

SCRs TESTS FOR PULSE CIRCUIT APPLICATION

The suitability for pulse circuit applications of SCRs not specifically characterized for such purposes can be determined from measurements carried out with relatively simple test circuits under controlled conditions. Applicable test circuits and procedures are outlined in the following section.

FORWARD BLOCKING VOLTAGE AND LEAKAGE CURRENT

Mount the SCRs to a heat sink and connect the units to be tested as shown in Figure 5.21. Place the assembly in an oven and stabilize at maximum SCR rated temperature. Turn on the power supply and raise the voltage to rated V_{BO}. Allow units to remain with the voltage applied for a minimum of four hours. At the end of the temperature soak, determine if any units exhibit thermal runaway by checking for blown fuses (without removing the power). Reject any units which have blown circuit fuses. The forward leakage current, I_{LF}, of the remaining units may be calculated after measuring the voltage V_L, across resistor R2. Any units with a leakage current greater than manufacturer's rating should be rejected.

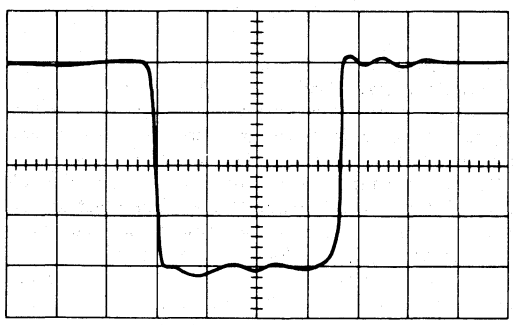


Figure 5.20. Vertical Set to 4 cm, Horizontal 0.2 μs/cm. Detected RF Magnetron Pulse

TURN-ON TIME, VARIATION AND ON IMPEDANCE

This circuit assumes that the pulse gate current required to switch a given modulator load current is specified by the manufacturer or that the designer is able to specify the operating conditions. Typical operating values might be:

Time of trigger pulse t = 1 μs
Pulse gate current I_G = 200 mA
Forward blocking voltage V_{BO} = 400 V
Load current I_{Load} = 30 A

To measure turn-on time using a Tektronix 545 oscilloscope (or equivalent) with a dual trace type CA plug-in, connect probes of Channels A and B to Test Points A and B. Place the Mode selector switch in the Added Algebraically position and the Channel B Polarity switch in the Inverted position. Adjust the HR212A pulse generator to give a positive pulse 1 μs wide (100 pps) as viewed at Test Point A. Adjust the amplitude of the "added" voltage across the 100-ohm base resistor for the specified pulse gate current (200 mA in this example).

Switch the Mode selector knob to the alternate position. Connect Channel A to Test Point D. Leave the oscilloscope probe, Channel B, at Test Point B, thereby displaying the input trigger waveform. Measure the time between the 50 percent voltage amplitudes of the two waveforms. This is the Turn-On Time ($t_D + t_R$).

To measure turn-on time versus temperature, place the device to be tested on a suitable heat sink and place the assembly in a temperature chamber. Stabilize the chamber at minimum rated (cold) temperature. Repeat the above measurements. Raise the chamber temperature to maximum rated (hot) temperature and stabilize. Repeat the measurements above.

To measure the turn-on impedance for the specified current load, the on impedance can be measured as an SCR forward voltage drop. The point in time of mea-

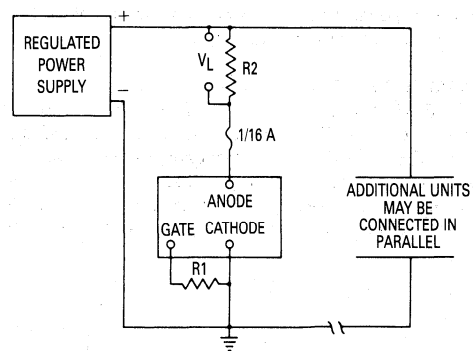


Figure 5.21. Test Setup for SCR Forward Blocking Voltage and Leakage Current Measurements

RESISTOR R1 IS USED ONLY IF MANUFACTURER CALLS FOR BIAS RESISTOR BETWEEN GATE AND CATHODE. RESISTOR R2 CAN HAVE ANY SMALL VALUE WHICH, WHEN MULTIPLIED BY MAXIMUM ALLOWABLE LEAKAGE CURRENT, WILL PROVIDE A CONVENIENT READING OF VOLTAGE V_L.

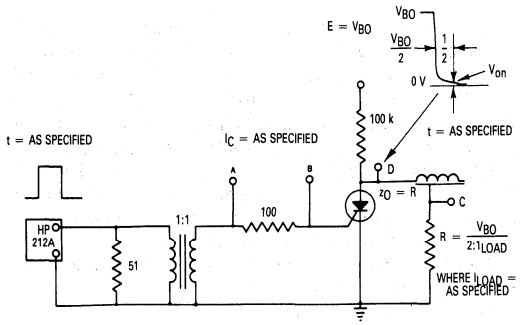


Figure 5.22. Suggested Test Circuit for SCR "On" Measurements

surement shall be half the output pulse width. For a 1 μs output pulse, the measurement procedure would be:

Connect the oscilloscope probe, Channel B, to Point D shown in Figure 5.22. Use the oscilloscope controls Time/CM and Multiplier to a setting of 0.5 μs per centimeter or faster. With the Amplitude Control set to view 100 volts per centimeter (to prevent amplifier overloading) measure the amplitude of the voltage drop, V_F, across the SCR 0.5 μs after the PFN voltage waveform has dropped to half amplitude. It may be necessary to check ground reference several times during this test to provide the needed accuracy of measurement.

HOLDING CURRENT

The SCR holding current can be measured with or without a gate turn-off current, according to the position of switch S2. The Motorola Trigger Pulse Generator is a transistor circuit capable of generating a 1.5 μs turn-on pulse followed by a variable-duration turn-off pulse. Measurements should be made at the maximum expected temperature of operation. Resistor R1 should be chosen to allow an initial magnitude of current flow at the device pulse current rating.

To measure holding current, connect the SCRs under test as illustrated in Figure 5.23. Place SCRs in oven and stabilize at maximum expected operating temperature. View the waveform across R1 by connecting the oscilloscope probe (Tektronix 2465) Channel A to Point A, and Channel B to Point B. Place the Mode Selector switch in the Added Algebraically position. Place the Polarity swich of Channel B in the Inverted position. Adjust both Volts/CM switches to the same scale factor, making sure that each Variable knob is in its Calibrated position. Adjust pulse generator for a positive pulse, 1 μs wide, and 1,000 pps pulse repetition frequency. Adjust power supply voltage to rated V_{BO}. Adjust input pulse amplitude until unit fully triggers. Measure amplitude of voltage drop across R1, V(A − B), and calculate holding current in mA from the equation

$$mA = \frac{V(A - B)}{R1} + \frac{V_{BO}}{100\ k\ \Omega}$$

Any unit which turns on but does not turn off has a holding current of less than

$$\frac{V_{BO}\ V}{100\ k\Omega}$$

The approximate voltage setting to view the amplitude of the holding current will be 10 or 20 volts per centimeter. The approximate sweep speed will be 2 to 5 μs per centimeter. These settings will, of course, vary, depending upon the holding current of the unit under test.

SCR recovery time is greatly dependent upon the circuit in which the device is used. However, any test of SCR recovery time should suffice to compare devices of various manufacturers, as long as the test procedure is standardized. Further evaluation of the selected devices could be made in an actual modulator circuit tester wherein techniques conducive to SCR turn-off are used.

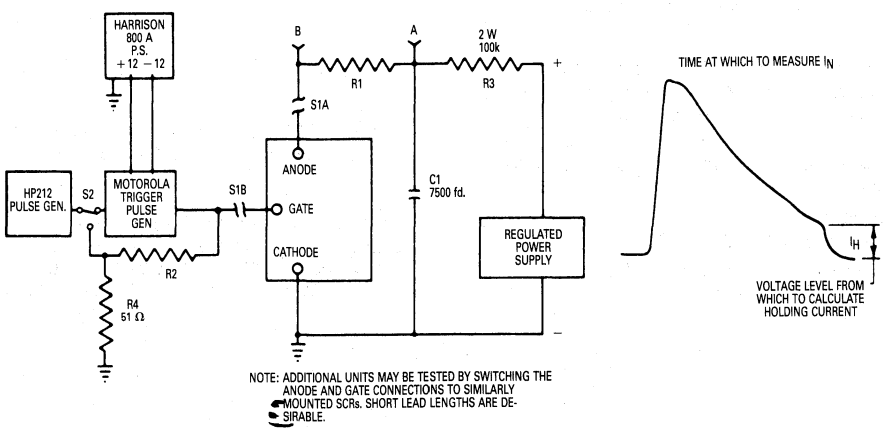


Figure 5.23. Test Setup for Measuring Holding Current

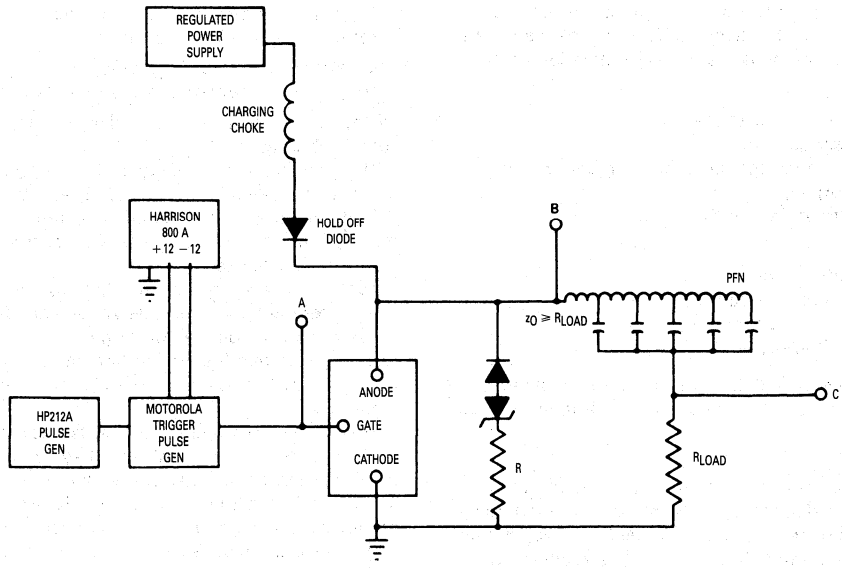


Figure 5.24. Modulator Circuit for SCR Tests

The above circuit setup shown in Figures 5.24 and 5.25 can be employed for such tests. A slight load to PFN mismatch is called for to generate an inverse voltage across the SCR at the termination of the output pulse. An SCR gate turn-off pulse is used. The recharge component is a charging choke, providing optimized conditions of reapplied voltage to the PFN (and across the SCR). Adequate heat sinking of the SCR should be provided.

PARALLEL CONNECTED SCRs

When an application requires current capability in excess of a single economical SCR, it can be worthwhile to consider paralleling two or more devices. To help deter-

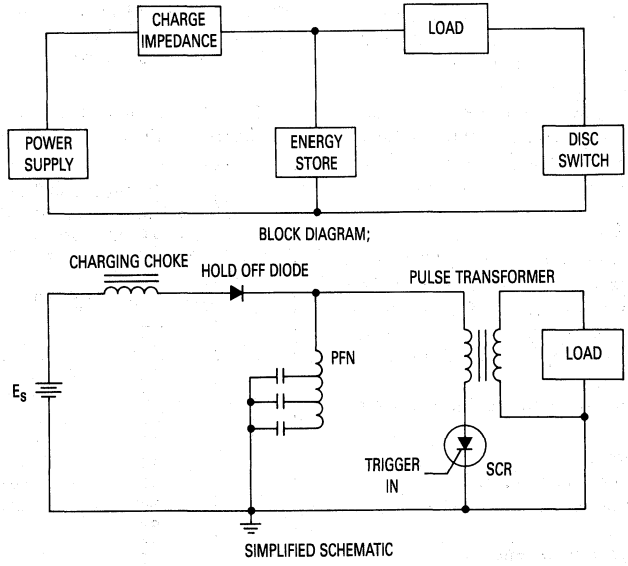


Figure 5.25. Radar Modulator, Resonant Line Type

mine if two or more SCRs in parallel are more cost effective than one high current SCR, some of the advantages and disadvantages are listed for parallel devices.

Advantages
1. Less expensive to purchase
2. Less expensive to mount
3. Less expensive to replace, in case of failure
4. Ease of mounting
5. Ease of isolation from sink

Disadvantages
1. Increased SCR count
2. Selected or matched devices
3. Increased component count
4. Greater R & D effort

There are several factors to keep in mind in paralleling and many are pertinent for single SCR operations as well.

GATE DRIVE

The required gate current (I_{GT}) amplitude can vary greatly and can depend upon SCR type and load being switched. As a general rule for parallel SCRs, I_{GT} should be at least two or three times the $I_{GT(MAX)}$ specification on the data sheet and ideally close to, but never exceeding, the maximum specified gate power dissipation or peak current. Adequate gate current is necessary for rapid turn-on of all the parallel SCRs and to ensure simultaneous turn-on without excessive current crowding across any of the individual die. The rise time of the gate drive pulse should be fast, ideally $\leq$ 100 ns. Each gate should be driven from a good current source and through its own resistor, even if transformer drive is used. Gate pulse width requirements vary but should be of sufficient width to ensure simultaneous turn-on and last well beyond the turn-on delay of the slowest device, as well as beyond the time required for latching of all devices. Ideally, gate current would flow for the entire conduction period to ensure latching under all operating conditions.

With low voltage switching, which includes conduction angles near 180° and near zero degrees, the gate drive requirements can be more critical and special emphasis may be required of gate pulse amplitude and width.

PARAMETER MATCHING

For reliable current sharing with parallel SCRs, there are certain device parameters that should be matched or held within close tolerances. The degree of matching required varies and can be affected by type of load (resistive, inductive, incandescent lamp or phase controlled loads) being switched.

The most common device parameters that can effect current sharing are:

1) t_d — turn-on delay time
2) t_r — turn-on rise time of anode current
3) $V_{A(MIN)}$ — minimum anode voltage at which device will turn on
4) Static on-state voltage and current
5) I_L — Latching current

The four parameters shown in Table 5.VI were measured with a curve tracer and are:

I_L, latching current; V_{TM}, on-state voltage; I_{GT} and V_{GT}, minimum gate current and voltage for turn on.

Of the four parameters, I_L and V_{TM} can greatly affect current sharing.

The latching current of each SCR is important at turn-on to ensure each device turns on and will stay on for the entire conduction period. On-state voltage determines how well the SCRs share current when cathode ballasting is not used.

Table 5.V gives turn-on delay time (t_d) and turn-on rise time (t_r) of the anode-cathode voltage and the minimum forward anode voltage for turn-on. These parameters were measured in the circuits shown in Figures 5.28 and 5.29. One SCR at a time was used in the circuit shown in Figure 5.28.

Turn-on delay on twenty-five SCRs was measured (only ten are shown in Table 5.V) and they could be from one or more production lots. The variation in t_d was slight and ranged from 35 to 44 ns but could vary considerably on other production lots and this possible variation in t_d would have to be considered in a parallel application.

Waveforms for minimum forward anode voltage for turn-on are shown in Figure 5.26. The trailing edge of the gate current pulse is phase delayed (R_3) so that the SCR is not turned on. The width of the gate current pulse is now increased (R_5) until the SCR turns on and the forward anode voltage switches to the on-state at about 0.73 V. This is the minimum voltage at which this SCR will turn on with the circuit conditions shown in Figure 5.28.

For dynamic turn-on current sharing, t_d, t_r and $V_{A(MIN)}$ are very important. As an example, with a high wattage incandescent lamp load, it is very important that the inrush current of the cold filament be equally shared by the parallel SCRs. The minimum anode voltage at which a device turns on is also very important. If one of the

Table 5.V. 2N6394 Turn-On Delay, Rise Time and Minimum Forward Anode Voltage For Turn-On

Device	Turn-On Delay and Rise Time Off-State Voltage = 8 V Peak R_L = 10 Ohms, $I_A \cong$ 6.5 A Peak I_G = 100 mA (PW = 100 μs) Conduction Angle 90 Degrees		Minimum Anode Voltage For Turn-On Off-State Voltage = 4 V Peak R_L = 0.5 Ohm I_A = 5A I_G = 100 mA
	t_d(ns)	t_r(μs)	(Volts)
1	35	0.80	0.70
2	38	0.95	0.81
3	45	1	0.75
4	44	1	0.75
5	44	0.90	0.75
6	43	0.85	0.75
7	38	1.30	0.75
8	38	1.25	0.70
9	38	1	0.75
10	37	0.82	0.70

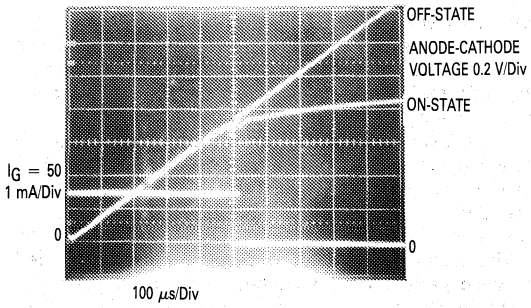


Figure 5.26. Minimum Anode Voltage For Turn-On Off-State Voltage = 4 V Peak, R_L = 0.5 Ohm, $I_A \approx$ 5 A, I_G = 75 mA

parallel devices turns on before the other devices and its on-state voltage is lower than the required minimum anode voltage for turn-on of the unfired devices, they therefore cannot turn on. This would overload the device which turned on, probably causing failure from overcurrent and excessive junction temperature.

Turn-off time — t_q is important in higher frequency applications which require the SCR to recover from the forward conduction period and be able to block the next cycle of forward voltage. Thus, t_q matching for high frequency operation can be as important as t_d, t_r and $V_{A(MIN)}$ matching for equal turn-on current sharing.

Due to the variable in t_q measurement, no further attempt will be made here to discuss this parameter and the reader is referred to Application Note AN914.

The need for on-state matching of current and voltage is important, especially in unforced current sharing circuits.

UNFORCED CURRENT SHARING

When operating parallel SCRs without forced current sharing, such as without cathode ballasting using resistors or inductors, it is very important that the device parameters be closely matched. This includes t_d, t_r, minimum forward anode voltage for turn-on and on-state voltage matching. The degree of matching determines the success of the circuit.

In circuits without ballasting, it is especially important that physical layout, mounting of devices and resistance paths be identical for good current sharing, even with on-state matched devices.

Figure 5.27 shows how anode current can vary on devices closely matched for on-state voltage (1, 3 and 4) and a mismatched device (2). Without resistance ballasting, the matched devices share peak current within one ampere and device 2 is passing only nine amps, seven amps lower than device 1. Table 5.VI shows the degree of match or mismatch of V_{TM} of the four SCRs.

With unforced current sharing (R_K = 0), there was a greater tendency for one device (1) to turn-on, preventing the others from turning on when low anode switching voltage (≤ 10 V rms) was tried. Table 5.V shows that the minimum anode voltage for turn-on is from 7 to 14% lower for device 1 than on 2, 3 and 4. Also, device 1 turn-on delay is 35 ns versus 38, 45 and 44 ns for devices 2, 3 and 4.

The tendency for device 1 to turn on, preventing the other three from turning on, is most probably due to its lower minimum anode voltage requirement and shorter turn on delay. The remedy would be closer matching of the minimum anode voltage for turn-on and driving the gates hard (but less than the gate power specifications) and increasing the width of the gate current pulse.

FORCED CURRENT SHARING

Cathode ballast elements can be used to help ensure good static on-state current sharing. Either inductors or resistors can be used and each has advantages and disadvantages. This section discusses resistive ballasting, but it should be kept in mind that the inductor method is usually better suited for the higher current levels. Although they are more expensive and difficult to design, there is less power loss with inductor ballasting as well as other benefits.

The degree of peak current sharing is shown in Figure 5.27 for four parallel 2N6394 SCRs using cathode resistor ballasting with an inductive anode load. With devices 1, 3 and 4, on-state voltage is matched within 10 mV at an anode current of 15 A (See Table 5.VI) and are within 1 A of each other in Figure 5.27, with cathode resistance (R_K) equal to zero. As R_K increases, the current sharing becomes even closer. The unmatched device 2, with a V_{TM} of 1.41 V (Table 5.VI), is not carrying its share of current (Figure 5.29) with R_K equal zero. As R_K increases, device 2 takes a greater share of the total current and with R_K around 0.25 ohm, the four SCRs are sharing peak current quite well. The value of R_K depends on how close the on-state voltage is matched on the SCRs and the degree of current sharing desired, as well as the permissible power dissipation in R_K.

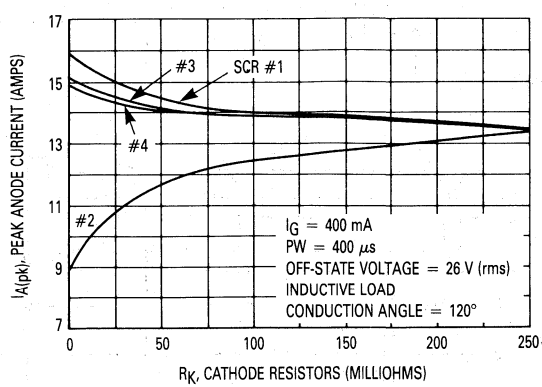


Figure 5.27. Effects Of Cathode Resistor On Anode Current Sharing

Table 5.VI. 2N6394 Parameters Measured On Curve Tracer, T_C = 25°C

Device #	I_L, Latching Current V_D=12 Vdc I_G=100 mA	V_{TM}, On-State Voltage I_A=15 A PW=300 μs	Minimum Gate Current & Voltage for Turn-On V_D=12 Vdc, R_L=140 Ω	
			I_{GT}	V_{GT}
1	13 mA	1.25 V	5.6 mA	0.615 V
2	27	1.41	8.8	0.679
3	28	1.26	12	0.658
4	23	1.26	9.6	0.649
5	23	1.28	9.4	0.659
6	23	1.26	9.6	0.645
7	18	1.25	7.1	0.690
8	19	1.25	7	0.687
9	19	1.25	8.4	0.694
10	16	1.25	6.9	0.679

LINE SYNCHRONIZED DRIVE CIRCUIT

Gate drive for phase control of the four parallel SCRs is accomplished with one complementary MOS hex gate, MC14572, and two bipolar transistors (Figure 5.28). This adjustable line-synchronized driver permits SCR conduction from near zero to 180 degrees. A Schmitt trigger clocks a delay monostable multivibrator that is followed by a pulse-width monostable multi-vibrator.

Line synchronization is achieved through the half-wave section of the secondary winding of the full-wave, center-tapped transformer (A). This winding also supplies power to the circuit through rectifiers D_1 and D_2.

The full-wave signal is clipped by diode D_5, referenced to a +15 volt supply, so that the input limit of the CMOS chip is not exceeded. The waveform is then shaped by the Schmitt trigger, which is composed of inverters U_{1-a} and U_{1-b}. A fast switching output signal B results.

The positive-going edge of this pulse is differentiated by the capacitive-resistive network of C_1 and R_2 and triggers the delay multivibrator that is composed of U_{1-c} and U_{1-d}. As a result, the normally high output is switched low. The trailing edge of this pulse (C) then triggers the following multivibrator, which is composed of NAND gate U_{1-e} and inverter U_{1-f}. The positive going output pulse (waveform D) of this multivibrator, whose width is set by potentiometer R_6, turns on transistors Q_1 and Q_2, which drives the gates of the four SCRs. Transistor Q_2 supplies about 400 mA drive current to each gate through 100 ohm resistors and has a rise time of ≤ 100 ns.

PARALLEL SCR CIRCUIT

The four SCRs are 2N6394s, housed in the TO-220 package, rated at 12 A rms, 50 V and are shown schematically

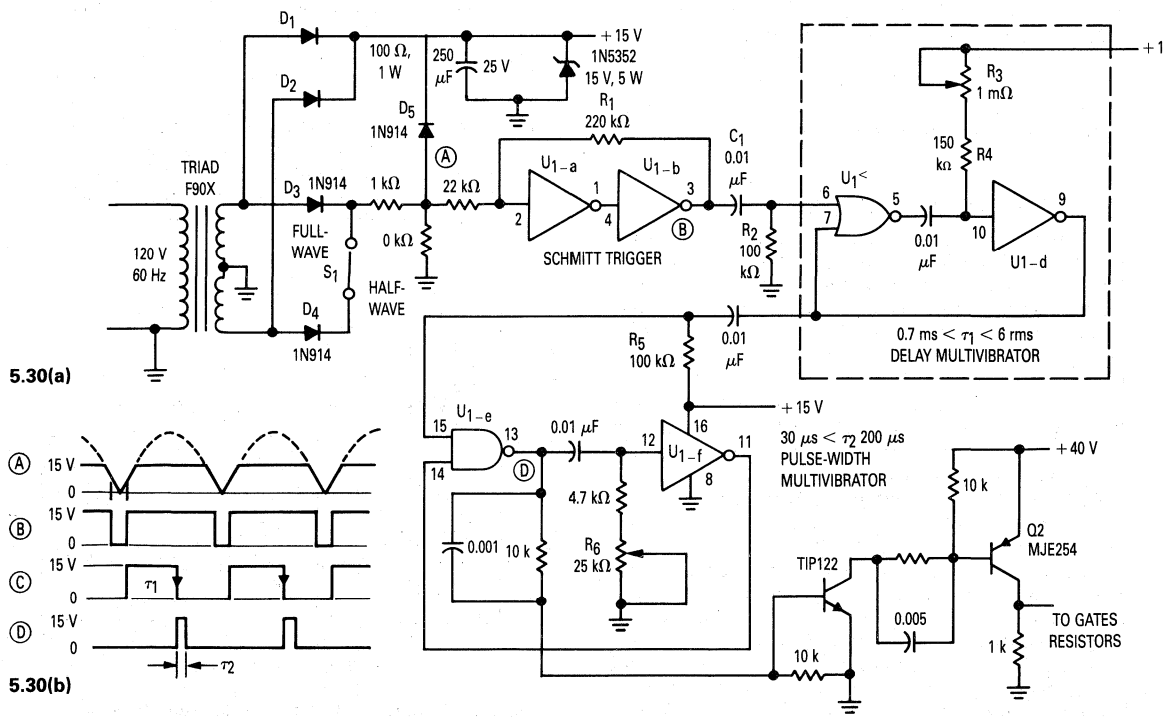


Figure 5.28. Line-Synchronized Gate Driver

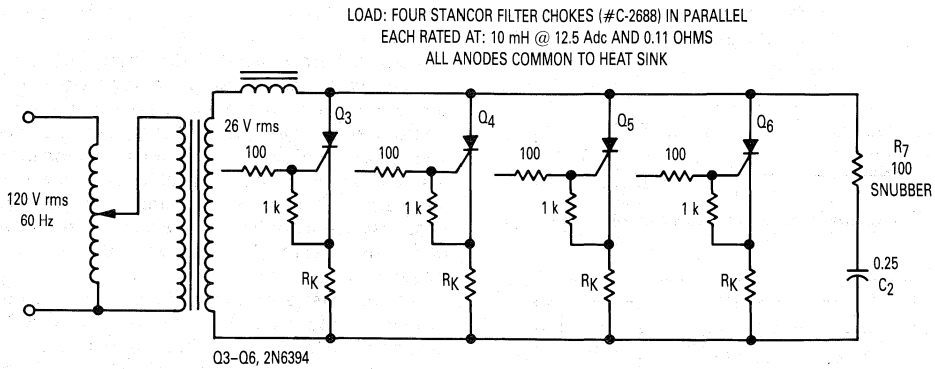


Figure 5.29. Parallel Thyristors

in Figure 5.29. Due to line power limitations, it was decided to use a voltage step down transformer and not try working directly from the 120 V line. Also, line isolation was desirable in an experiment of this type.

The step down transformer ratings were 120 V rms primary, 26 V rms secondary, rated at 100 A, and was used with a variable transformer for anode voltage adjustment. The inductive load consisted of four filter chokes in parallel (Stancor #C-2688 with each rated at 10 mH, 12.5 Adc and 0.11 ohm).

For good current sharing with parallel SCRs, symmetry in layout and mounting is of primary importance. The four SCRs were mounted on a natural finish aluminum heat sink and torqued to specification which is 8 inch pounds. Cathode leads and wiring were identical, and when used, the cathode resistors R_K were matched within 1%. An RC snubber network (R_7 and C_2) was connected across the anodes-cathodes to slow down the rate-of-rise of the off-state voltage, preventing unwanted turn-on.

CHARACTERIZING RFI SUPPRESSION IN THYRISTOR CIRCUITS

In order to understand the measures for suppression of EMI, characteristics of the interference must be explored first. To have interference at all, we must have a transmitter, or creator of interference, and a receiver, a device affected by the interference. Neither the transmitter nor the receiver need be related in any way to those circuits commonly referred to as radio-frequency circuits. Common transmitters are opening and closing of a switch or relay contacts, electric motors with commutators, all forms of electric arcs, and electronic circuits with rapidly changing voltages and currents. Receivers are generally electronic circuits, both low and high impedance which are sensitive to pulse or high frequency energy. Often the very circuits creating the interference are sensitive to similar interference from other circuits nearby or on the same power line.

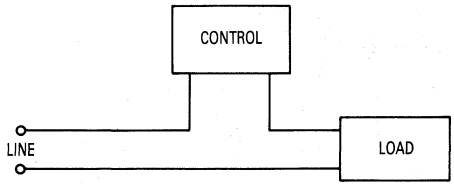
EMI can generally be separated into two categories — radiated and conducted. Radiated interference travels by way of electro-magnetic waves just as desirable RF energy does. Conducted interference travels on power, communications, or control wires. Although this separation and nomenclature might seem to indicate two neat little packages, independently controllable, such is not the case. The two are very often interdependent such that in some cases control of one form may completely eliminate the other. In any case, both interference forms must be considered when interference elimination steps are taken.

Phase control circuits using thyristors (SCRs, triacs, etc.) for controlling motor speed or resistive lighting and heating loads are particularly offensive in creating interference. They can completely obliterate most stations on any AM radio nearby and will play havoc with another control on the same power line. These controls are generally connected in one of the two ways shown in the block diagrams of Figure 5.30.

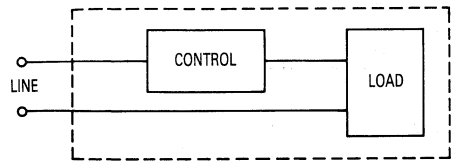
A common example of the connection of 5.30(a) is the wall mounted light dimmer controlling a ceiling mounted lamp. A motorized appliance with a built-in control such as a food mixer is an example of the connection shown in 5.30(b).

Figure 5.30(a) may be re-drawn as shown in Figure 5.31, illustrating the complete circuit for RF energy. The switch in the control box represents the thyristor, shown in its blocking state. In phase control operation, this switch is open at the beginning of each half cycle of the power line alternations. After a delay determined by the remainder of the control circuitry, the switch is closed and remains that way until the instantaneous current drops to zero. This switch is the source from which the RF energy flows down the power lines and through the various capacitors to ground.

If the load is passive, such as a lamp or a motor which does not generate interference, it may be considered as an impedance bypassed with the wire-to-wire capacitance of its leads. If it is another RF energy source, however, such as a motor with a commutator, it must be treated separately to reduce interference from that



(a). Separately Mounted Control



(b). Control and Load in the Same Enclosure

Figure 5.30. Block Diagrams of Control Connections

source. The power supply may be considered as dc since the interference pulse is extremely short (10 μs) compared to the period of the power line frequency (16 ms for 60 Hz). The inductance associated with the power source comes from two separate phenomena. First is the leakage impedance of the supply transformer, and second is the self-inductance of the wires between the power line transformer and the load.

One of the most difficult parameters to pin down in the system is the effect of grounding. Most industrial and commercial wiring and many homes use a grounded conduit system which provides excellent shielding of radiated energy emanating from the wiring. However, a large number of homes are being wired with two to three wire insulated cable without conduit. In three-wire systems, one wire is grounded independently of the power system even though one of the power lines is already grounded. The capacitances to ground shown in Figure 5.31 will be greatly affected by the type of grounding used. Of course, in any home appliance, filtering must be provided suitable for all three different systems.

Before the switch in the control is closed, the system is in a steady-state condition with the upper line of the power line at the system voltage and the bottom line and the load at ground potential. When the switch is closed, the upper line potential instantaneously falls due to the line and source inductance, then it rises back to its original value as the line inductance is charged. While the upper line is rising, the line from the control to the load also rises in potential. The effect of both of these lines increasing in potential together causes an electro-static field change which radiates energy. In addition, any other loads connected across the power lines at point A, for example, would be affected by a temporary loss of voltage created by the closing of the switch and by the line and source inductance. This is a form of conducted interference.

A second form of radiated interference is inductive coupling in which the power line and ground form a one-turn primary of an air core transformer. In this mode, an unbalanced transient current flows down the power lines with the difference current flowing to ground through the various capacitive paths available. The secondary is the radio antenna or the circuit being affected. This type of interference is a problem only when the receiver is within about one wavelength of the transmitter at the offending frequency.

Radiated interference from the control circuit proper is of little consequence due to several factors. The lead lengths in general are so short compared to the wavelengths in question that they make extremely poor antenna. In addition, most of these control circuits are mounted in metal enclosures which provide shielding for radiated energy generated within the control circuitry.

A steel box will absorb radiated energy at 150 kHz such that any signal inside the box is reduced 12.9 dB per mil of thickness of the box. In other words, a 1/16 inch thick steel box will attenuate radiated interference by over 800 dB! A similar aluminum box will attenuate 1 dB per mil or 62.5 dB total. Thus, even in an aluminum box, the control circuitry will radiate very little energy.

Both forms of radiated interference which are a problem are a result of conducted interference on the power lines which is in turn caused by a rapid rise in current. Thus, if this current rise is slowed, all forms of interference will be reduced.

RFI SOLUTIONS

Since the switch in Figure 5.31, when it closes, provides a very low impedance path, a capacitor in parallel with it will show little benefit in slowing down the rise of current. The capacitor will be charged to a voltage deter-

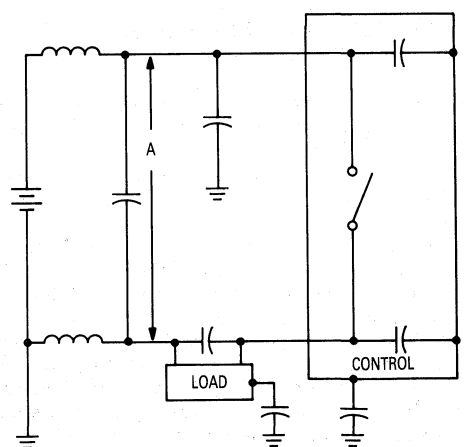


Figure 5.31. RF Circuit for Figure 5.30(a)

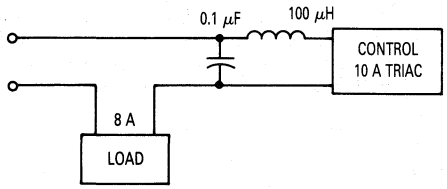


Figure 5.32. One Possible EMI Reduction Circuit

mined by the circuit constants and the phase angle of the line voltage just before the switch closes. When the switch closes, the capacitor will discharge quickly, its current limited only by its own resistance and the resistance of the switch. However, a series inductor will slow down the current rise in the load and thus reduce the voltage transient on all lines. A capacitor connected as shown in Figure 5.32 will also help slow down the current rise since the inductor will now limit the current out of the capacitor. Thus, the capacitor voltage will drop slowly and correspondingly the load voltage will increase slowly.

Although this circuit will be effective in many cases, the filter is unbalanced, providing an RF current path through the capacitances to ground. It has, therefore, been found advantageous to divide the inductor into two parts and to put half in each line to the control. Figure 5.33 illustrates this circuit showing the polarity marks of two coils which are wound on the same core.

A capacitor at point A will help reduce interference further. This circuit is particularly effective when used with the connection of Figure 5.30(b) where the load is not always on the grounded side of the power line. In this case, the two halves of the inductor would be located in the power line leads, between the controlled circuit and the power source.

Where the control circuit is sensitive to fast rising line transients, a capacitor at point B will do much to eliminate this problem. The capacitor must charge through the impedance of the inductor, thus limiting the rate of voltage change (dv/dt) applied to the thyristor while it is in the blocking state.

DESIGN CRITERIA

Design equations for the split inductor have been developed based on parameters which should be known before attempting a design. The most difficult to determine is t_r, the minimum allowable current rise time which will not cause objectionable interference. The value of this parameter must be determined empirically in each situation if complete interference reduction is needed. Motorola has conducted extensive tests using an AM radio as a receiver and a 600 Watt thyristor lamp dimmer as a transmitter. A rate of about 0.35 Amp per μs seems to be effective in eliminating objectionable interference as well as materially reducing false triggering of the thyristor due to line transients. The value of t_r may be calculated by dividing the peak current anticipated by the allowable rate of current rise.

Ferrite core inductors have proved to be the most practical physical configuration. Most ferrites are effective; those with highest permeability and saturation flux density are preferred. Those specifically designed as high frequency types are not necessarily desirable.

Laminated iron cores may also be used; however, they require a capacitor at point A in Figure 5.33 to be at all effective. At these switching speeds, the iron requires considerable current in the windings before any flux change can take place. We have found currents rising to half their peak value in less than one μs before the inductance begins to slow down the rise. The capacitor supplies this current for the short period without dropping in voltage, thus eliminating the pulse on the power line.

Once a core material has been selected, wire size is the next decision in the design problems. Due to the small number of turns involved (generally a single layer) smaller sizes than normally used in transformers may be chosen safely. Generally, 500 to 800 circular mills per ampere is acceptable, depending on the enclosure of the filter and the maximum ambient temperature expected.

An idea of the size of the core needed may be determined from the equation:

$$(1) \quad A_c A_w = \frac{26\, A_{wire}\, E_{rms}\, t_r}{B_{MAX}}$$

where:

A_c = the effective cross-sectional area of the core in in^2
A_w = available core window area in in^2
A_{wire} = wire cross section in circular mils
B_{MAX} = core saturation flux density in gauss
t_r = allowable current rise time in seconds
E_{rms} = line voltage

(A factor of 3 has been included in this equation to allow for winding space factor.) Once a tentative core selection has been made, the number of turns required may be found from the equation:

$$(2) \quad N = \frac{11\, E_{rms}\, t_r \times 10^6}{B_{MAX} A_c}$$

where:

N = the total number of turns on the core

The next step is to check how well the required number of turns will fit onto the core. If the fit is satisfactory, the core design is complete; if not, some trade-offs will have to be made.

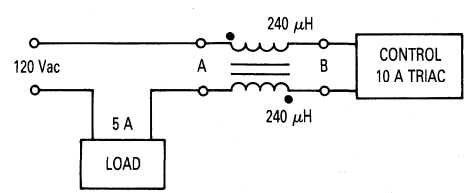


Figure 5.33. Split Inductor Circuit

In most cases, the inductor as designed at this point will have far too much inductance. It will support the entire peak line voltage for the time selected as t_r and will then saturate quickly, giving much too fast a current rise. The required inductance should be calculated from the allowable rise time and load resistance, making the rise time equal to two time constants. Thus:

$$\frac{2L}{R} = t_r \quad \text{or} \quad L = \frac{R\,t_r}{2} \tag{3}$$

Paper or other insulating material should be inserted between the core halves to obtain the required inductance by the equation:

$$l_g = \frac{3.19\,N^2\,A_c \times 10^{-8}}{L} - \frac{l_c}{\mu} \tag{4}$$

where:

l_g = total length of air gap in inches
μ = effective ac permeability of the core material at the power line frequency
l_c = effective magnetic path length of the core in inches
A_c = effective cross sectional area of the core in square inches
L = inductance in henries

DESIGN EXAMPLE

Consider a 600 watt, 120 Volt lamp dimmer using a Motorola 2N6148 triac. Line current is $\frac{600}{120} = 5$ amperes. #16 wire will provide about 516 circular mils per ampere.

For core material, type 3C5 of Ferroxcube Corporation of America, Saugerties, New York, has a high B_{max} and μ. The company specifies B_{MAX} = 3800 gauss and μ = 1900 for material.

As was previously mentioned, a current rise rate of about 0.35 ampere per μs has been found to be acceptable for interference problems with ac-dc radios in most wiring situations. With 5 amperes rms, 7 amperes peak,

$$t_r = \frac{7}{0.35} = 20\ \mu s$$

Then by equation (1):

$$A_cA_w = \frac{26x2580x120x20x10^{-6}}{3800\ \text{gauss}} = 0.044$$

Core part number 1F30 of the same company in a U-1 configuration has an A_cA_w product of 0.0386, which should be close enough.

$$N = \frac{10.93x120x20x10^{-6}x10^{6}}{3800 \times 0.137} = 42\ \text{turns}$$

Two coils of 21 turns each should be wound on either one or two legs and be connected as shown in Figure 5.35.

The required inductance of the coil is found from equation (3).

$$L = \frac{R\,t_r}{2} = \frac{E_{rated}}{I_{rated}} \times \frac{t_r}{2} = \frac{120}{5} \times \frac{20}{2} \times 10^{-6} = 240 \times 10^{-6}$$

$L = 240\ \mu H$

To obtain this inductance, the air gap should be

$$l_g = \frac{3.19x42^2x0.137x10^{-8}}{240x10^{-6}} - \frac{3.33}{1900} = 0.0321 - 0.00175$$

$l_g = 0.03035$

Thus, 15 mils of insulating material in each leg will provide the necessary inductance.

If a problem still exists with false triggering of the thyristor due to conducted interference, a capacitor at point B in Figure 5.33 will probably remedy the situation.

CHAPTER 6
APPLICATIONS

Because they are reliable solid state switches, thyristors have many applications, especially as controls.

One of the most common uses for thyristors is to control ac loads such as electric motors. This can be done either by controlling the part of each ac cycle when the circuit conducts current (phase control) or by controlling the number of cycles per time period when current is conducted (cycle control).

In addition, thyristors can serve as the basis of relaxation oscillators for timers and other applications. Most of the devices covered in this book have control applications.

PHASE CONTROL WITH THYRISTORS

The most common method of electronic ac power control is called *phase control.* Figure 6.1 illustrates this concept. During the first portion of each half-cycle of the ac sine wave, an electronic switch is opened to prevent the current flow. At some specific phase angle, α, this switch is closed to allow the full line voltage to be applied to the load for the remainder of that half-cycle. Varying α will control the portion of the total sine wave that is applied to the load (shaded area), and thereby regulate the power flow to the load.

The simplest circuit for accomplishing phase control is shown in Figure 6.2. The electronic switch in this case is a triac (Q) which can be turned on by a small current pulse to its gate. The TRIAC turns off automatically when the current through it passes through zero. In the circuit shown, capacitor C_T is charged during each half-cycle by the current flowing through resistor R_T and the load. The fact that the load is in series with R_T during this portion of the cycle is of little consequence since the resistance of R_T is many times greater than that of the load. When the voltage across C_T reaches the breakdown voltage of the DIAC bilateral trigger (D), the energy stored in capacitor C_T is released. This energy produces a current pulse in the DIAC, which flows through the gate of the TRIAC and turns it on. Since both the DIAC and the TRIAC are bidirectional devices, the values of R_T and C_T will determine the phase angle at which the TRIAC will be triggered in both the positive and negative half-cycles of the ac sine wave.

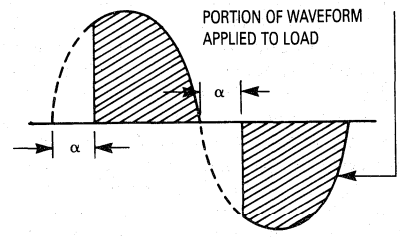


Figure 6.1. Phase Control of AC Waveform

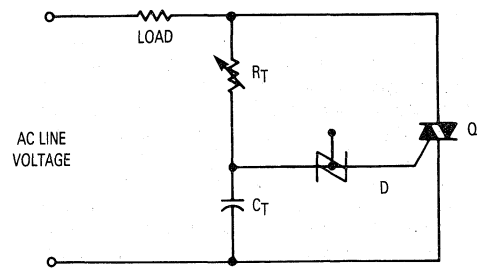


Figure 6.2. Simplest Circuit for Phase Control

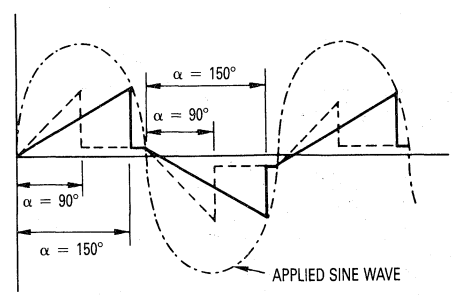


Figure 6.3. Waveforms of Capacitor Voltage at Two Phase Angles

The waveform of the voltage across the capacitor for two typical control conditions (α = 90° and 150°) is shown in Figure 6.3. If a silicon controlled rectifier is used in this

circuit in place of the TRIAC, only one half-cycle of the waveform will be controlled. The other half-cycle will be blocked, resulting in a pulsing dc output whose average value can be varied by adjusting R_T.

CONTROL OF INDUCTION MOTORS

Shaded-pole motors driving low-starting-torque loads such as fans and blowers may readily be controlled using any of the previously described full-wave circuits. One needs only to substitute the winding of the shaded-pole motor for the load resistor shown in the circuit diagrams.

Constant-torque loads or high-starting-torque loads are difficult, if not impossible, to control using the voltage controls described here. Figure 6.4 shows the effect of varying voltage on the speed-torque curve of a typical shaded-pole motor. A typical fan-load curve and a constant-torque-load curve have been superimposed upon this graph. It is not difficult to see that the torque developed by the motor is equal to the load torque at two different points on the constant-torque-load curve, giving two points of equilibrium and thus an ambiguity to the speed control. The equilibrium point at the lower speed is a condition of high motor current because of low counter EMF and would result in burnout of the motor winding if the motor were left in this condition for any length of time. By contrast, the fan speed-torque curve crosses each of the motor speed-torque curves at only one point, therefore causing no ambiguities. In addition, the low-speed point is one of low voltage well within the motor winding's current-carrying capabilities.

Permanent-split-capacitor motors can also be controlled by any of these circuits, but more effective control is achieved if the motor is connected as shown in Figure 6.5. Here only the main winding is controlled and the capacitor winding is continuously connected to the entire ac line voltage. This connection maintains the phase shift between the windings, which is lost if the capacitor phase is also controlled. Figure 6.6(a) shows the effect of voltage on the speed-torque characteristics of this motor and a superimposed fan-load curve.

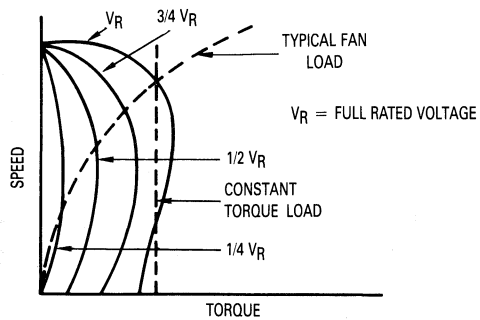


Figure 6.4. Characteristics of Shaded-Pole Motors at Several Voltages

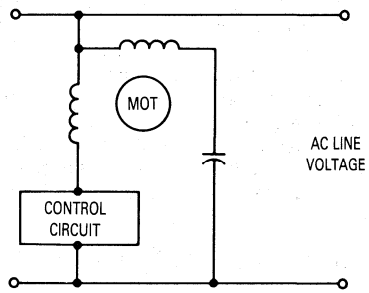
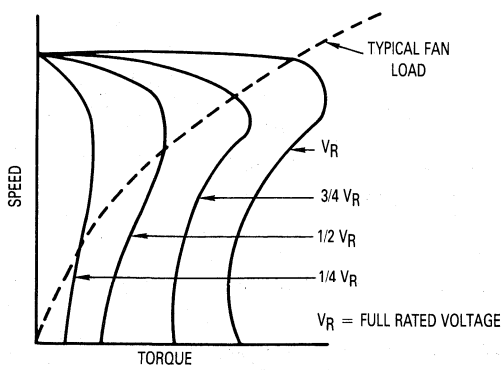
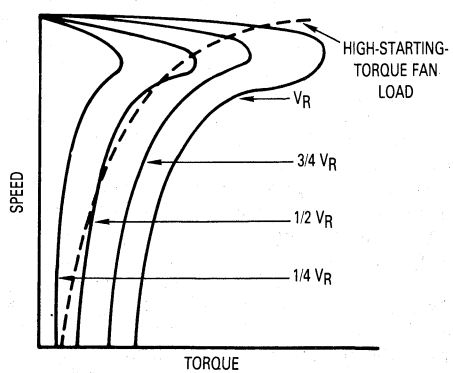


Figure 6.5. Connection Diagram for Permanent-Split-Capacitor Motors

Not all induction motors of either the shaded-pole or the permanent-split-capacitor types can be controlled effectively using these techniques, even with the proper loads. Motors designed for the highest efficiencies and, therefore, low slip also have a very low starting torque and may, under certain conditions, have a speed-torque



(a). High-Starting-Torque Motor



(b). High-Efficiency Motor

Figure 6.6. Speed-Torque Curves for Permanent-Split-Capacitor Motors at Various Applied Voltages

characteristic that could be crossed twice by a specific fan-load speed-torque characteristic. Figure 6.6(b) shows motor torque-speed characteristic curves upon which has been superimposed the curve of a fan with high starting torque. It is therefore desirable to use a motor whose squirrel-cage rotor is designed for medium-to-high impedance levels and, therefore, has a high starting torque. The slight loss in efficiency of such a motor at full rated speed and load is a small price to pay for the advantage of speed control.

A unique circuit for use with *capacitor-start motors* in explosive or highly corrosive atmospheres, in which the arcing or the corrosion of switch contacts is severe and undesirable, is shown in Figure 6.7. Resistor R1 is connected in series with the main running winding and is of such a resistance that the voltage drop under normal full-load conditions is approximately 0.2 V peak. Since starting currents on these motors are quite high, this peak voltage drop will exceed 1 V during starting conditions, triggering the TRIAC, which will cause current to flow in the capacitor winding. When full speed is reached, the current through the main winding will decrease to about 0.2 V, which is insufficient to trigger the TRIAC — thus the capacitor winding will no longer be energized. Resistor R2 and capacitor C2 form a dv/dt suppression network; this prevents the TRIAC from turning on due to line transients and inductive switching transients.

CONTROL OF UNIVERSAL MOTORS

Any of the half-wave or full-wave controls described previously can be used to control universal motors. *Non-feedback,* manual controls, such as those shown in Figure 6.2, are simple and inexpensive, but they provide very little torque at low speeds. A comparison of typical speed-torque curves using a control of this type with those of *feedback* control is shown in Figure 6.8.

These motors have some unique characteristics which allow their speed to be controlled very easily and efficiently with a feedback circuit such as that shown in Figure 6.9. This circuit provides phase-controlled half-wave power to the motor; that is, on the negative half-cycle, the SCR blocks current flow in the negative direction causing the motor to be driven by a pulsating direct current whose amplitude is dependent on the phase control of the SCR.

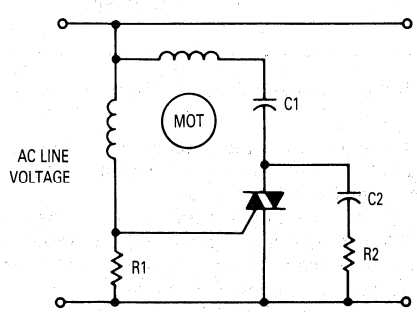


Figure 6.7. Circuit Diagram for Capacitor-Start Motor

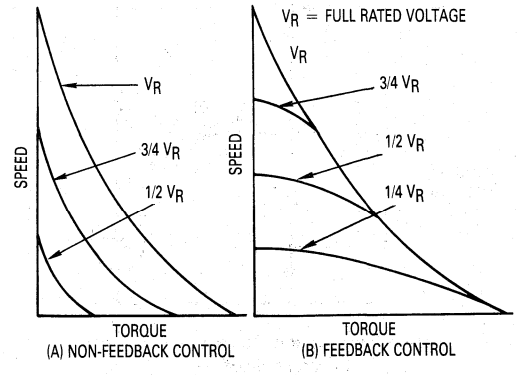


Figure 6.8. Comparison of Feedback Control with Non-Feedback Control

The theory of operation of this control circuit is not at all difficult to understand. Assuming that the motor has been running, the voltage at point A in the circuit diagram must be larger than the forward drop of Diode D1, the gate-to-cathode drop of the SCR, and the EMF generated by the residual MMF in the motor, to get sufficient current flow to trigger the SCR.

The waveform at point A (V_A) for one positive half-cycle is shown in 6.9(b), along with the voltage levels of the SCR gate (V_{SCR}), the diode drop (V_D), and the motor-generated EMF (V_M). The phase angle (α) at which the SCR would trigger is shown by the vertical dotted line. Should the motor for any reason speed up so that the generated motor voltage would increase, the trigger point would move upward and to the right along the curve so that the SCR would trigger later in the half-cycle and thus provide less power to the motor, causing it to slow down again.

Similarly, if the motor speed decreased, the trigger point would move to the left and down the curve, causing the TRIAC to trigger earlier in the half-cycle providing more power to the motor, thereby speeding it up.

Resistors R1, R2, and R3, along with diode D2 and capacitor C1 form the ramp-generator section of the circuit. Capacitor C1 is charged by the voltage divider R1, R2, and R3 during the positive half-cycle. Diode D2 prevents negative current flow during the negative half-cycle, therefore C1 discharges through only R2 and R3 during that half-cycle. Adjustment of R3 controls the amount by which C1 discharges during the negative half-cycle. Because the resistance of R1 is very much larger than the ac impedance of capacitor C1, the voltage waveform on C1 approaches that of a perfect cosine wave with a dc component. As potentiometer R2 is varied, both the dc and the ac voltages are divided, giving a family of curves as shown in 6.9(c).

The gain of the system, that is, the ratio of the change of effective SCR output voltage to the change in generator

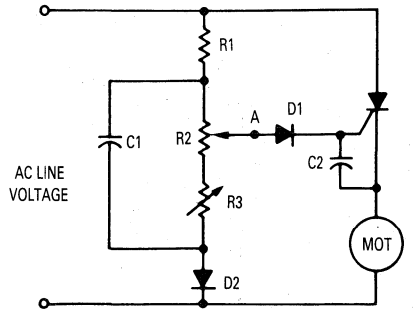
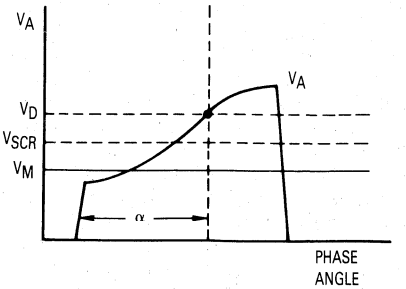
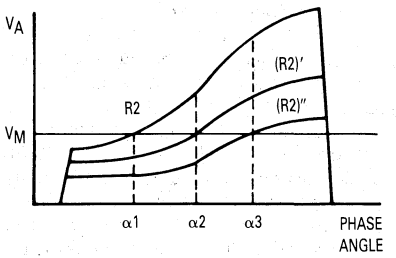


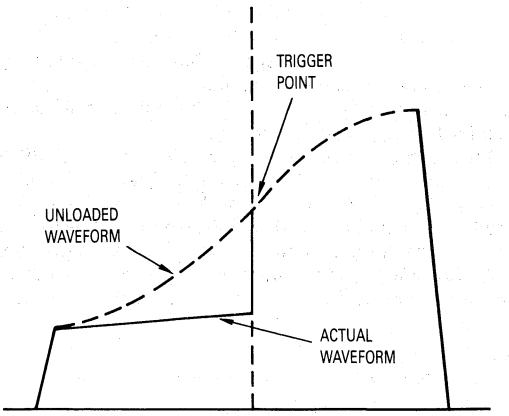
Figure 6.9 (a). Speed-Control Scheme for Universal Motors



(b). Waveform for One Positive Half-Cycle of Circuit



(c). Voltage Waveform at Point "A" for Three Settings of Potentiometer R2



(d). Point "A" Voltage with Excessive Resistance R1

EMF, is considerably greater at low speed settings than it is at high speed settings. This high gain coupled with a motor with a very low residual EMF will cause a condition sometimes known as cycle skipping. In this mode of operation, the motor speed is controlled by skipping entire cycles or groups of cycles, then triggering one or two cycles early in the period to compensate for the loss in speed. Loading the motor would eliminate this condition; however, the undesirable sound and vibration of the motor necessitate that this condition be eliminated. This can be done in two ways.

The first method is used if the motor design is fixed and cannot be changed. In this case, the impedance level of the voltage divider R1, R2, and R3 can be lowered so that C1 will charge more rapidly, thus increasing the slope of the ramp and lowering the system gain. The second method, which will provide an overall benefit in improved circuit performance, involves a redesign of the motor so that the residual EMF becomes greater. In general, this means using a lower grade of magnetic steel for the laminations. As a matter of fact, some people have found that ordinary cold-rolled steel used as rotor laminations makes a motor ideally suited for this type of electronic control.

Another common problem encountered with this circuit is that of thermal runaway. With the speed control set at low or medium speed, at high ambient temperatures the speed may increase uncontrollably to its maximum value. This phenomenon is caused by an excessive impedance in the voltage-divider string for the SCR being triggered. If the voltage-divider current is too low, current will flow into the gate of the SCR without turning it on, causing the waveform at point A to be as shown in 6.9(d). The flat portion of the waveform in the early part of the half-cycle is caused by the SCR gate current loading the voltage divider before the SCR is triggered. After the SCR is triggered, diode D1 is back-biased and a load is no longer on the voltage divider so that it jumps up to its unloaded voltage. As the ambient temperature increases, the SCR becomes more sensitive, thereby requiring less gate current to trigger, and is triggered earlier in the half-cycle. This early triggering causes increased current in the SCR, thereby heating the junction still further and increasing still further the sensitivity of the SCR until maximum speed has been reached.

The solutions to this problem are the use of the most sensitive SCR practical and a voltage divider network of sufficiently low impedance. As a rough rule of thumb, the average current through the voltage divider during the positive half-cycle should be approximately three times the current necessary to trigger the lowest-

sensitivity (highest gate current) SCR being used.

In addition to the type of steel used in the motor laminations, consideration should also be given to the design of motors used in this half-wave speed control. Since the maximum rms voltage available to the motor under half-wave conditions is 85 V, the motor should be designed for use at that voltage to obtain maximum speed. However, U.L. requirements state that semiconductor devices used in appliance control systems must be able to be short-circuited without causing danger. Many designers have found it advantageous, therefore, to use 115 V motors with this system and provide a switch to apply full-wave voltage to the motor for high-speed operation. Figure 6.10 shows the proper connection for this switch. If one were to simply short-circuit the SCR for full-speed operation, a problem could arise. If the motor were operating at full speed with the switch closed, and the switch were then opened during the negative half-cycle, the current flowing in the inductive field of the motor could then break down the SCR in the negative direction and destroy the control. With the circuit as shown, the energy stored in the field of the motor is dissipated in the arc of the switch before the SCR is connected into the circuit.

CONTROL OF PERMANENT-MAGNET MOTORS

As a result of recent developments in ceramic permanent-magnet materials that can be easily molded into complex shapes at low cost, the permanent-magnet motor has become increasingly attractive as an appliance component. Electronic control of this type of motor can be easily achieved using techniques similar to those just described for the universal motor. Figure 6.11 is a circuit diagram of a control system that we have developed and tested successfully to control permanent-magnet motors presently being used in blenders. Potentiometer R3 and diode D1 form a dc charging path for capacitor C1; variable resistor R1 and resistor R2 form an ac charging path which creates the ramp voltage on the capacitor. Resistor R4 and diode D2 serve to isolate the motor control circuit from the ramp generator during the positive and negative half-cycles, respectively.

A small amount of cycle skipping can be experienced at low speeds using this control, but not enough to necessitate further development work. Since the voltage generated during off time is very high, the thermal runaway problem does not appear at all. Typical speed-torque curves for motors of this type are shown in Figure 6.12.

MOTOR SPEED CONTROL WITH FEEDBACK

While many motor speed control circuits have used SCRs, the TRIAC has not been very popular in this application. At first glance, it would appear that the TRIAC would be perfect for speed control because of its bilateral characteristics. There are a couple of reasons why this is not true. The major difficulty is the TRIAC's dv/dt characteristic. Another reason is the difficulty of obtaining a feedback signal because of the TRIAC's bilateral nature.

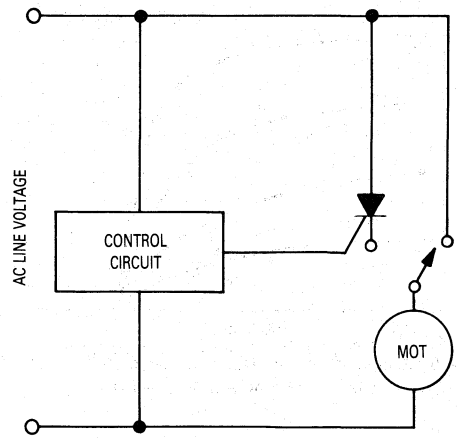


Figure 6.10. Switching Scheme for Full-Wave Operation

While the TRIAC has its disadvantages, it does offer some advantages. In a SCR speed control either two SCRs must be used, or the line voltage must be full-wave rectified using relatively high current rectifiers, or the control must be limited to half-wave. The TRIAC eliminates all these difficulties. By using a TRIAC the part count, package size, and cost can be reduced. Figure 6.13 shows a TRIAC motor speed control circuit that derives its feedback from the load current and does not require separate connections to the motor field and armature windings. Therefore, this circuit can be conveniently built into an appliance or used as a separate control.

The circuit operates as follows: When the TRIAC conducts, the normal line voltage, less the drop across the TRIAC and resistor R5, is applied to the motor. By delaying the firing of the TRIAC until a later portion of the cycle, the rms voltage applied to the motor is reduced and its speed is reduced proportionally. The use of feedback maintains torque at reduced speeds.

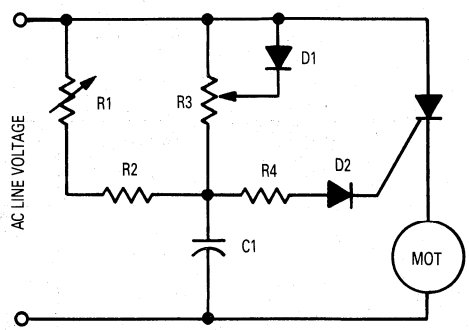


Figure 6.11. Circuit Diagram for Controlling Permanent-Magnet Motors

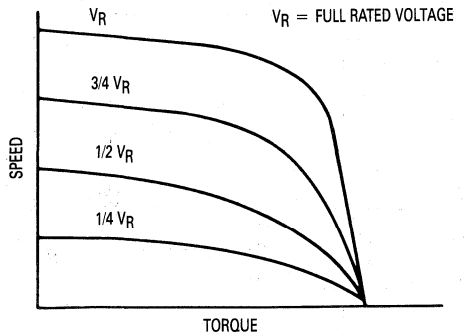


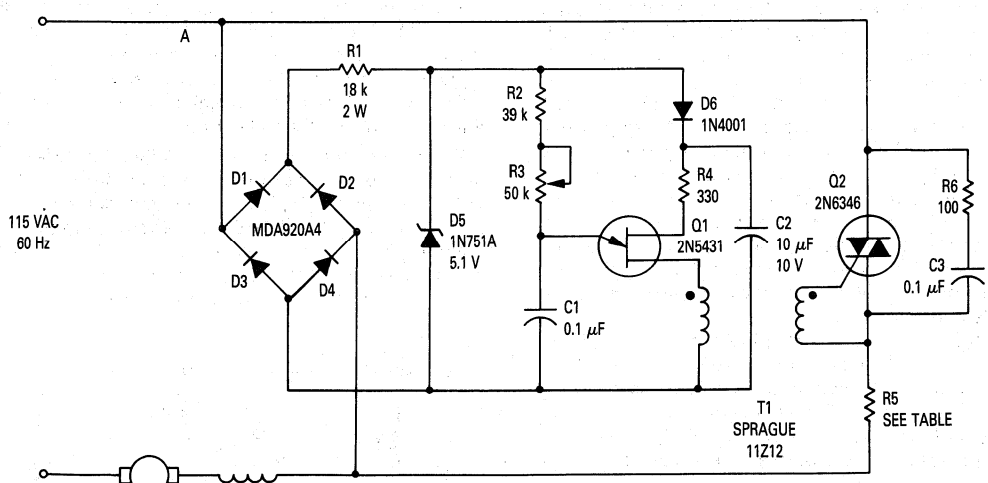
Figure 6.12. Speed-Torque Characteristic of Permanent-Magnet Motors at Various Applied Voltages

Diodes D1 through D4 form a bridge which applies full-wave rectified voltage to the phase-control circuit. Phase control of the TRIAC is obtained by the charging of capacitor C1 through resistors R2 and R3 from the voltage level established by zener diode D5. When C1 charges to the firing voltage of unijunction transistor Q1, the TRIAC is triggered through transformer T1. C1 discharges through the emitter of Q1. While the TRIAC is conducting, the voltage drop between points A and B falls below the breakdown voltage of D5. Therefore, during the conduction period, the voltage on C1 is determined by the voltage drop from A to B and by resistors R1, R2, and R3. Since the voltage between A and B is a function of motor current due to resistor R5, C1 is charged during the conduction period to a value which is proportional to the motor current. The value of R5 is chosen so that C1 cannot charge to a high enough voltage to fire Q1 during the conduction period. However, the amount of charging required to fire Q1 has been decreased by an amount proportional to the motor current. Therefore, the firing angle at which Q1 will fire has been advanced in proportion to the motor current. As the motor is loaded and draws more current, the firing angle of Q1 is advanced even more, causing a proportionate increase in the rms voltage applied to the motor, and a consequent increase in its available torque.

Since the firing voltage of Q1 depends on the voltage from base one to base two, it is necessary to support the base two voltage during the conduction portion of the cycle to prevent the feedback voltage from firing Q1. D6 and C2 perform this function.

Because the motor is an inductive load, it is necessary to limit the commutation dv/dt for reliable circuit operation. R6 and C3 perform this function.

Nominal values for R5 can be obtained from the table or they can be calculated from the equation given. Exact values for R5 depend somewhat on the motor characteristics. Therefore, it is suggested that R5 be an adjust-



NOMINAL R5 VALUES			
Motor Rating (Amperes)	R5		$R5 = \frac{2}{I_M}$
	OHMS	Watts	
2	1	5	
3	0.67	10	I_M = Max. Rated Motor Current (rms)
6.5	0.32	15	

Figure 6.13. Motor Speed Control with Feedback

able wirewound resistor which can be calibrated in terms of motor current, and the speed control can be adapted to many different motors. If the value of R5 is too high, feedback will be excessive and surging or loss of control will result. If the value is too low, a loss of torque will result. The maximum motor current flows through R5, and its wattage must be determined accordingly.

This circuit has been operated successfully with 2 and 3 ampere 1/4-inch drills and has satisfactorily controlled motor speeds down to 1/3 or less of maximum speed with good torque characteristics.

CONSTANT SPEED MOTOR CONTROL USING TACHOMETER FEEDBACK

Tachometer feedback sensing rotor speed provides excellent performance with electric motors. The principal advantages to be gained from tachometer feedback are the ability to apply feedback control to shaded-pole motors, and better brush life in universal motors used in feedback circuits. This latter advantage results from the use of full-wave rather than half-wave control, reducing the peak currents for similar power levels.

THE TACHOMETER

The heart of this system is, of course, the speed-sensing tachometer itself. Economy being one of the principal goals of the design, it was decided to use a simple magnetic tachometer incorporating the existing motor fan as an integral part of the magnetic circuit. The generator consists of a coil wound on a permanent magnet which is placed so that the moving fan blades provide a magnetic path of varying reluctance as they move past the poles of the magnet. Several possible configurations of the magnetic system are shown in Figure 6.14.

Flux in a magnetic circuit can be found from the "magnetic Ohm's law":

$$\phi = \frac{MMF}{R},$$

where ϕ = the flux,
MMF = the magnetomotive force (strength of the magnet), and
R = the reluctance of the magnetic path.

Assuming the MMF of the permanent magnet to be constant, it is readily apparent that variations in reluctance will directly affect the flux. The steel fan blades provide a low-reluctance path for the flux once it crosses the air gap between them and the poles of the magnet. If the magnet used has a horseshoe or U shape, and is placed so that adjacent fan blades are directly opposite each pole in one position of the motor armature, the magnetic path will be of relatively low reluctance; then as the motor turns the reluctance will increase until one fan blade is precisely centered between the poles of the magnet. As rotation continues, the reluctance will then alternately increase and decrease as the fan blades pass the poles of the magnet. If a bar- or L-shaped magnet is used so that one pole is close to the shaft or the frame of the motor and the other is near the fan blades, the magnetic path reluctance will vary as each blade passes the magnet pole near the fan. In either case the varying reluctance causes variations in the circuit flux and a voltage is generated in the coil wound around the magnet. The voltage is given by the equation:

$$e = -N \frac{d\phi}{dt} \times 10^{-8},$$

where e = the coil voltage in volts,
N = the number of turns in the coil, and
$\frac{d\phi}{dt}$ = the rate of change of flux in lines per second.

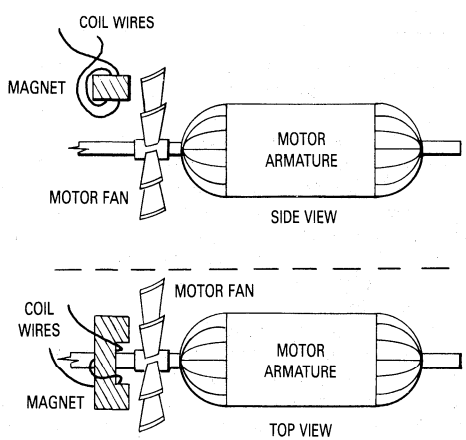
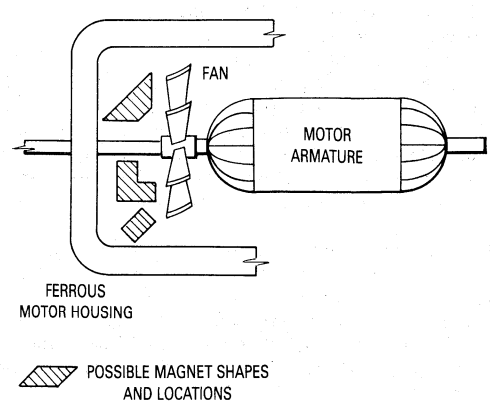


Figure 6.14(a). Locations for Magnetic Sensing Tachometer Generator Using a Horseshoe Magnet



(b). Locations for Magnetic Sensing Tachometer Generator Using an "L" or Bar Magnet

In a practical case, a typical small horseshoe magnet wound with 1000 turns of wire generated a voltage of about 0.5 volts/1000 rpm when mounted in a blender.

Since both generated voltage and frequency are directly proportional to the motor speed, either parameter can be used as the feedback signal. However, circuits using voltage sensing are less complex and therefore less expensive. Only that system will be discussed here.

THE ELECTRONICS

In one basic circuit, which is shown in Figure 6.15, the generator output is rectified by rectifier D1, then filtered and applied between the positive supply voltage and the base of the detector transistor Q1. This provides a negative voltage which reduces the base-voltage on Q1 when the speed increases.

The emitter of the detector transistor is connected to a voltage divider which is adjusted to the desired tachometer output voltage. In normal operation, if the tachometer voltage is less than desired, the detector transistor, Q1, is turned on by current through R1 into its base. Q1 then turns on Q2 which causes the timing capacitor for unijunction transistor Q3 to charge quickly. Standard unijunction transistor circuitry is used to trigger the thyristor.

As the tachometer output approaches the voltage desired, the base-emitter voltage of Q1 is reduced to the point at which Q1 is almost cut off. Thereby, the collector current of Q2, which charges the unijunction timing capacitor, is reduced, causing that capacitor to charge slowly and trigger the thyristor later in the half cycle. In this manner, the average power to the motor is reduced until just enough power to maintain the desired motor speed is allowed to flow.

Input circuit variations are used when the tachometer output voltage is too low to give a usable signal with a silicon rectifier. In the variation shown in Figure 6.15(b), the tachometer is connected between a voltage divider and the base of the amplifier transistor. The voltage divider is set so that with no tachometer output the transistor is just barely in conduction. As the tachometer output increases, Q_T is cut off on negative half cycles and conducts on positive half cycles. Resistors R9 and R10 provide a fixed gain for this amplifier stage, providing the h_{FE} of Q_T is much greater than the ratio of R9 to R10. Thus the output of the amplifier is a fixed multiple of the positive values of the tachometer waveform. The rectifier diode D1 prevents C1 from discharging through R9 on negative half cycles of the tachometer. The remainder of the filter and control circuitry is the same as the basic circuit.

In the second variation, shown in 6.15(c), R8 has been replaced by a semiconductor diode, D2. Since the voltage and temperature characteristics more closely match those of the transistor base-to-emitter junction, this circuit is easier to design and needs no initial adjustments as does the circuit in 6.15(b). The remainder of this circuit is identical to that of Figure 6.14.

In the second basic circuit, which is shown in Figure 6.16, the rectified and filtered tachometer voltage is added to the output voltage of the voltage divider formed by R1 and R2. If the sum of the two voltages is less than $V1 - V_{BE\ Q1}$ (where $V_{BE\ Q1}$ is the base-emitter voltage of Q1), Q1 will conduct a current proportional to $V1 - V_{BE\ Q1}$, charging capacitor C. If the sum of the two voltages is greater than $V1 - V_{BE\ Q1}$, Q1 will be cut off and no current will flow into the capacitor. The operation of the remainder of the circuit is the same as the previously described circuits.

All of the circuits that have been described show a unijunction transistor as the trigger device. A three-layer bilateral trigger diode may also be used as shown in Figure 6.17. The rectifier diode which is connected to the pulsing dc voltage, V2, discharges the capacitor at the end of each half cycle of the line voltage alternations, providing synchronization to the line voltage.

Complete circuit diagrams using the two basic circuits are shown in Figure 6.18.

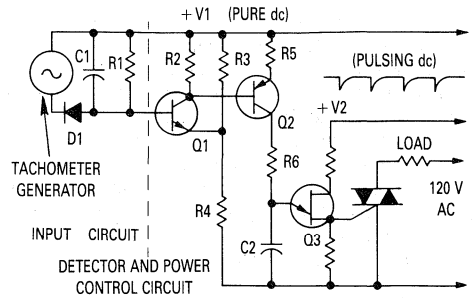
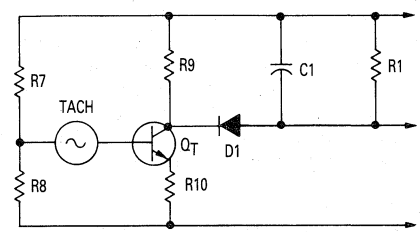
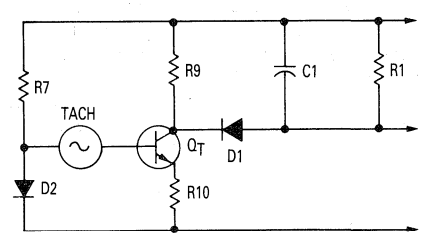


Figure 6.15(a). Basic Tachometer Control Circuit



(b). Variation Used when the Tachometer Output is Too Low for Adequate Control



(c). Variation Providing Better Temperature Tracking and Easier Initial Adjustment

PHASE CONTROL WITH TRIGGER DEVICES

Phase control using thyristors is one of the most common means of controlling the flow of power to electric motors, lamps, and heaters. With an ac voltage applied to the circuit, the gated thyristor (SCR, TRIAC, etc.) remains in its off-state for the first portion of each half cycle of the power line, then, at a time (phase angle) determined by the control circuit, the thyristor switches on for the remainder of the half cycle. By controlling the phase angle at which the thyristor is switched on, the relative power in the load may be controlled.

PHASE CONTROL WITH UNIJUNCTION TRANSISTORS

Unijunction transistors provide a simple, convenient

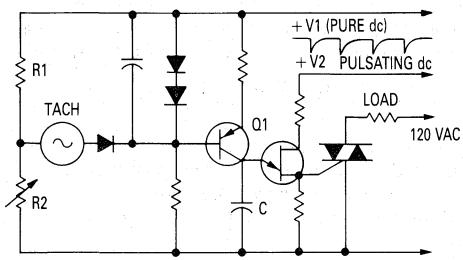


Figure 6.16. Another Basic Tachometer Circuit

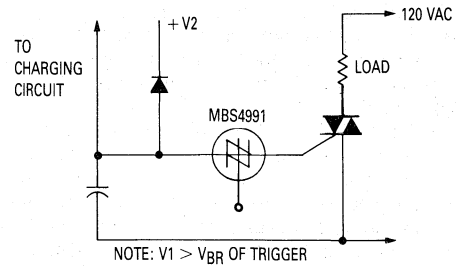


Figure 6.17. Use of a Bidirectional Switch as the Triggering Device Instead of the Unijunction Transistors Shown in Other Figures

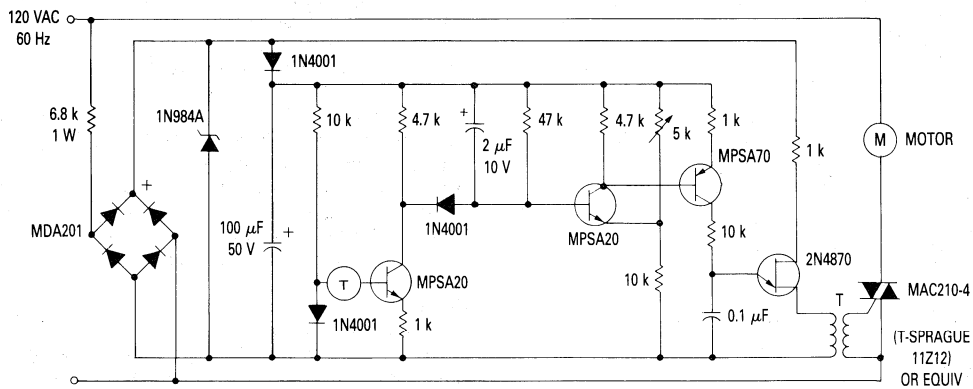


Figure 6.18(a). Complete Diagram of Tachometer Speed Control Shown in 6.15

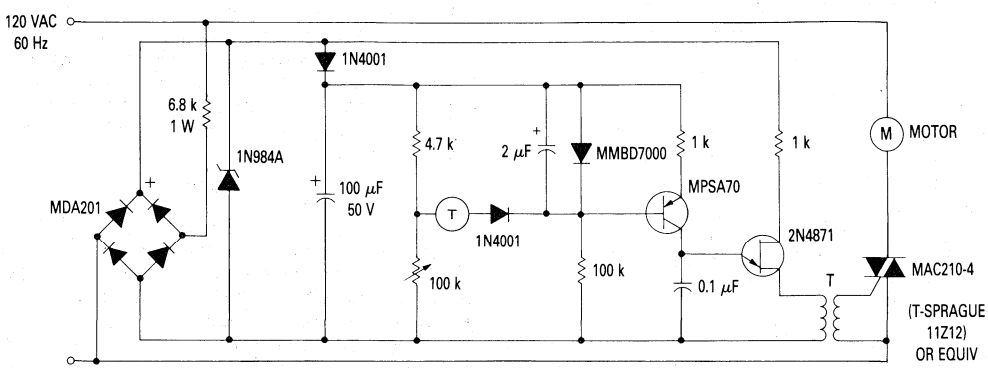


Figure 6.18(b). Complete Diagram of Tachometer Speed Control Shown in 6.16

means for obtaining the thyristor trigger pulse synchronized to the ac line at a controlled phase angle.

These circuits are all based on the simple relaxation oscillator circuit of Figure 6.19. R_T and C_T in the figure form the timing network which determines the time between the application of voltage to the circuit (represented by the closing of S1) and the initiation of the pulse. In the case of the circuit shown, with V_S pure dc, the oscillator is free running, R_T and C_T determine the frequency of oscillation. The peak of the output pulse voltage is clipped by the forward conduction voltage of the gate to cathode diode in the thyristor. The principal waveforms associated with the circuit are shown in Figure 6.19(b).

Operation of the circuit may best be described by referring to the capacitor voltage waveform. When the power is applied, C_T charges at the rate determined by its own capacitance and the value of R_T until its voltage reaches the peak point voltage of the UJT emitter. At that time the UJT switches into the conduction condition, discharging C_T through R_{B1} and the gate of the thyristor. With V_S pure dc, the cycle then repeats immediately; however, in many cases V_S is derived from the anode voltage of the thyristor so that the timing cycle cannot start again until the thyristor is blocking forward voltage and once again provides V_S.

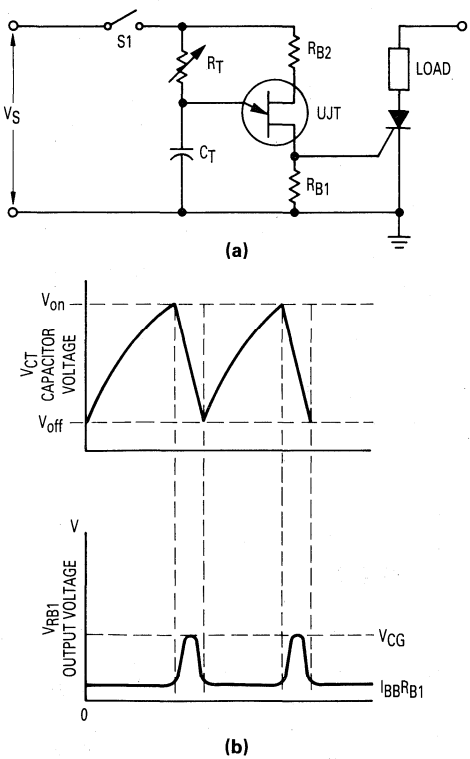


Figure 6.19. Basic Relaxation Oscillator Circuit (a) and Waveforms (b)

During the time in which the capacitor is being charged, current flows through the interbase resistance (r_{BB}) of the unijunction. R_{B1} is included in the circuit to provide a path for this current so that it does not flow through the gate of the thyristor and cause an undesirable turn on. Its value is selected so that a maximum voltage developed across it will be less than 0.2 volt. For a typical unijunction with r_{BB} = 4 to 9 kohms, and a typical operating voltage of 20 volts, the value of R_{B1} would be:

$$R_{B1} = \frac{0.2\ R_{BB(MIN)}}{V_S} = \frac{0.2 \times 4\ k}{20} = 40 \text{ ohms}$$

R_{B2} is necessary only if some degree of temperature compensation is necessary. In most cases, particularly in the feedback systems described later, it may be left out and base 2 of the unijunction may be connected to the positive side of V_S.

It is often necessary to synchronize the timing of the output pulses to the power line voltage zero-crossing points. One simple method of accomplishing synchronization is shown in Figure 6.20. Zener diode D1 clips the rectified supply voltage resulting in a V_S as shown in 6.20(b). Since V_{BB}, and therefore the peak point voltage of the unijunction drops to zero each time the line voltage corsses zero, C_T is discharged at the end of every half cycle and begins each half cycle in the discharged state. Thus, even if the UJT has not triggered during one half cycle, the capacitor begins the next half cycle discharged so that the phase angle at which the pulse occurs is directly controlled for each cycle by the values of R_T and C_T. The zener diode also provides voltage stabilization for the timing circuit giving the same pulse phase angle regardless of normal line voltage fluctuations.

APPLICATIONS

The most elementary application of the UJT trigger circuit, shown in Figure 6.21, is a half-wave control circuit. In this circuit, R_D is selected to limit the current through D_1 so that the diode dissipation capability is not exceeded. Dividing the allowable diode dissipation by one-half the zener voltage will give the allowable positive current in the diode since it is conducting in the voltage regulating mode only during positive half cycles. Once the positive half-cycle current is found, the resistor value may be calculated by subtracting 0.7 times the zener voltage from the rms line voltage and dividing the result by the positive current:

$$R_D = \frac{E_{rms} - 0.7\ V_z}{I_{positive}}$$

The power rating of R_D must be calculated on the basis of full wave conduction as D1 is conducting on the negative half cycle acting as a shunt rectifier as well as providing V_S on the positive half cycle.

The thyristor is acting both as a power control device and a rectifier, providing variable power to the load during the positive half cycle and no power to the load during

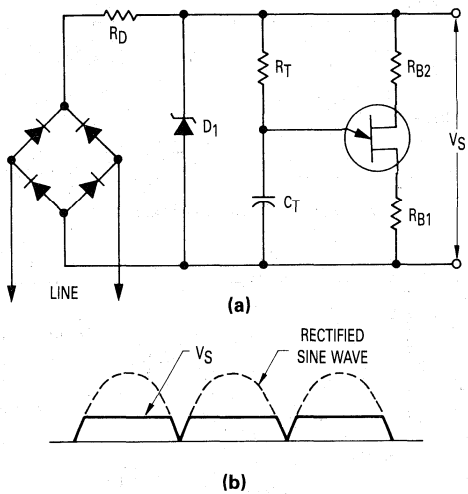


Figure 6.20. Control Circuit (a) with Zener Clipped, Rectified Voltage (b)

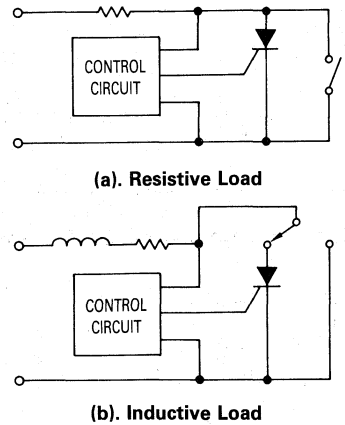


Figure 6.22. Half Wave Controls with Switching for Full Wave Operation

the negative half cycle. The circuit is designed to be a two terminal control which can be inserted in place of a switch. If full wave power is desired as the upper extreme of this control, a switch can be added which will short circuit the SCR when R_T is turned to its maximum power position. The switch may be placed in parallel with the SCR if the load is resistive; however, if the load is inductive, the load must be transferred from the SCR to the direct line as shown in Figure 6.22.

Full wave control may be realized by the addition of a bridge rectifier, a pulse transformer, and by changing the thyristor from an SCR to a TRIAC, shown in Figure 6.23. In this circuit R_{B1} is not necessary since the pulse transformer isolates the thyristor gate from the steady-state UJT current.

Occasionally a circuit is required which will provide constant output voltage regardless of line voltage changes. Adding potentiometer P1 to the circuits of Figures 6.21 and 6.23 will provide an approximate solution to this problem. The potentiometer is adjusted to provide reasonably constant output over the desired range of line voltage. As the line voltage increases, so does the voltage on the wiper of P1 increasing V_{BB} and thus the peak point voltage of the UJT. The increased peak point voltage results in C_T charging to a higher voltage and thus taking more time to trigger. The additional delay reduces the thyristor conduction angle and maintains the average voltage at a reasonably constant value.

FEEDBACK CIRCUITS

The circuits described so far have been manual control circuits; i.e., the power output is controlled by a potentiometer turned by hand. Simple feedback circuits may be constructed by replacing R_T with heat or light-dependent sensing resistors; however, these circuits have no means of adjusting the operating levels. The addition of a transistor to the circuits of Figures 6.21 and 6.23 allows complete control.

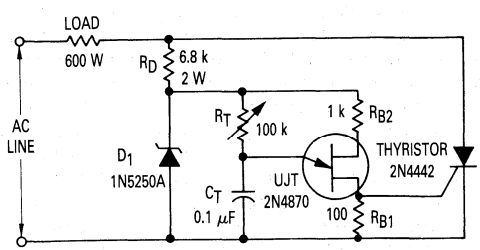


Figure 6.21. Half Wave Control Circuit with Typical Values for a 600 Watt Resistive Load

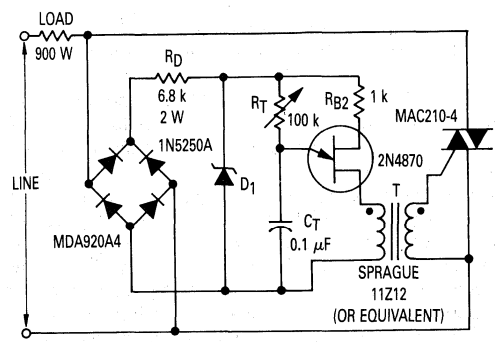


Figure 6.23. A Simple Full Wave Trigger Circuit with Typical Values for a 900 Watt Resistive Load

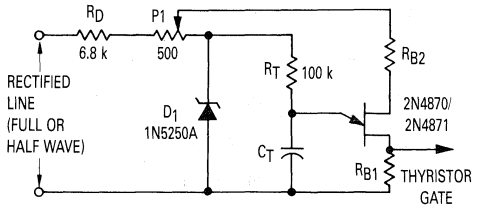


Figure 6.24. Circuit for Line Voltage Compensation

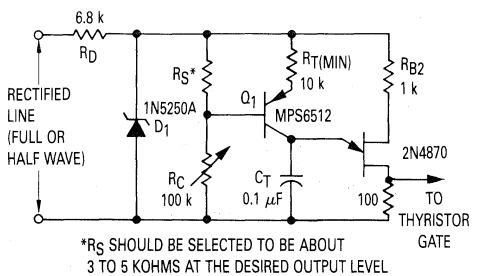


Figure 6.25. Feedback Control Circuit

Figure 6.25 shows a feedback control using a sensing resistor for feedback. The sensing resistor may respond to any one of many stimuli such as heat, light, moisture, pressure, or magnetic field. R_S is the sensing resistor and R_C is the control resistor that establishes the desired operating point. Transistor Q_1 is connected as an emitter follower such that an increase in the resistance of R_S decreases the voltage on the base of Q_1, causing more current to flow. Current through Q_1 causes voltage to charge C_T, triggering the UJT at some phase angle. As R_S becomes larger, more current flows into the capacitor, the voltage builds up faster, causing the UJT to trigger at a smaller phase angle and more power is applied to the load. When R_S decreases, less power is applied to the load. Thus, this circuit is for a sensing resistor which decreases in response to too much power in the load. If the sensing resistor increases with load power, then R_S and R_C should be interchanged.

If the quantity to be sensed can be fed back to the circuit in the form of an isolated, varying dc voltage such as the output of a tachometer, it may be inserted between the voltage divider and the base of Q1 with the proper polarity. In this case, the voltage divider would be a potentiometer to adjust the operating point. Such a circuit is shown in Figure 6.26.

In some cases, average load voltage is the desired feedback variable. In a half wave circuit this type of feedback usually requires the addition of a pulse transformer, shown in Figure 6.27. The RC network, R_1, R_2, C_1, averages load voltage so that it may be compared with the set point on R_S by Q_1. Full wave operation of this type of circuit requires dc in the load as well as the control circuit. Figure 6.28 is one method of obtaining this full wave control.

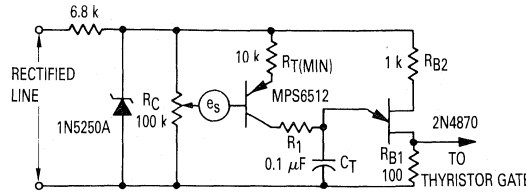


Figure 6.26. Voltage Feedback Circuit

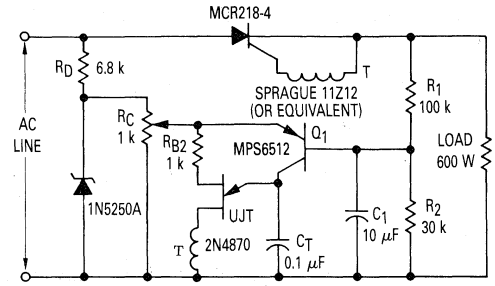


Figure 6.27. Half Wave, Average Voltage Feedback

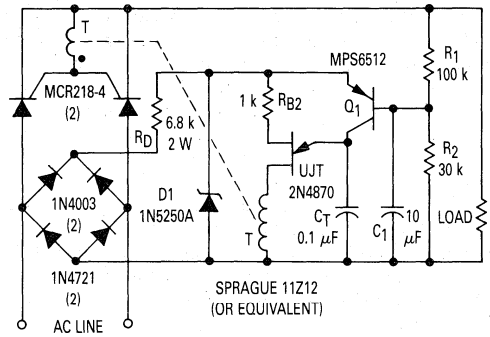


Figure 6.28. Full Wave, Average Voltage Feedback Control

There are, of course, many more sophisticated circuits which can be derived from the basic circuits discussed here. If, for example, very close temperature control is desired, the circuit of Figure 6.25 might not have sufficient gain. To solve this problem a dc amplifier could be inserted between the voltage divider and the control transistor gate to provide as close a control as desired. Other modifications to add multiple inputs, switched gains, ramp and pedestal control, etc., are all simple additions to add sophistication. Basically, however, it is the UJT itself which provides the fast rising, high current pulse, which is desirable for reliable thyristor operation. The ease of adding feedback and relative insensitivity to line voltage changes are additional benefits gained from using this trigger device.

CYCLE CONTROL WITH OPTICALLY ISOLATED TRIAC DRIVERS

In addition to the phase control circuits, TRIAC drivers can also be used for ac power control by on-off or burst control, of a number of ac cycles. This form of power control allows logic circuits and microprocessors to easily control ac power with TRIAC drivers of both the zero-crossing and non zero-crossing varieties.

USING NON-ZERO CROSSING OPTICALLY ISOLATED TRIAC DRIVERS

USING THE MOC3011 ON 240 VAC LINES

The rated voltage of a MOC3011 is not sufficiently high for it to be used directly on 240 V line; however, the designer may stack two of them in series. When used this way, two resistors are required to equalize the voltage dropped across them as shown in Figure 6.29.

REMOTE CONTROL OF AC VOLTAGE

Local building codes frequently require all 115 V light switch wiring to be enclosed in conduit. By using a MOC3011, a TRIAC, and a low voltage source, it is possible to control a large lighting load from a long distance through low voltage signal wiring which is completely isolated from the ac line. Such wiring usually is not required to be put in conduit, so the cost savings in installing a lighting system in commercial or residential buildings can be considerable. An example is shown in Figure 6.29. Naturally, the load could also be a motor, fan, pool pump, etc.

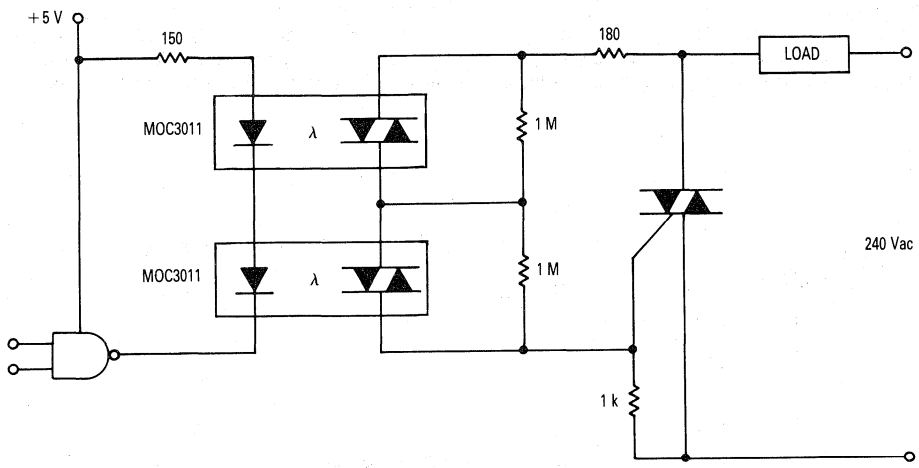


Figure 6.29. Two MOC3011 TRIAC Drivers in Series to Drive 240 V TRIAC

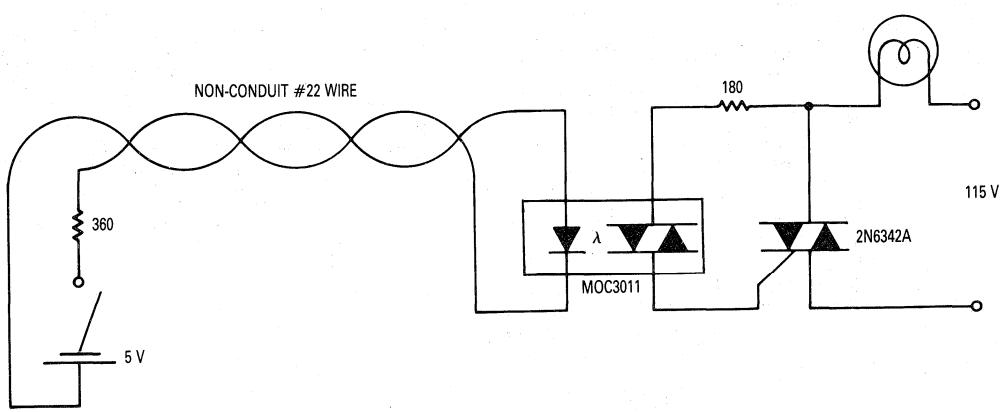


Figure 6.30. Remote Control of AC Loads Through Low Voltage Non-Conduit Cable

SOLID STATE RELAY

Figure 6.30 shows a complete general purpose, solid state relay snubbed for inductive loads with input protection. When the designer has more control of the input and output conditions, he can eliminate those components which are not needed for his particular application to make the circuit more cost effective.

INTERFACING MICROPROCESSORS TO 115 VAC PERIPHERALS

The output of a typical microcomputer input-output (I/O) port is a TTL-compatible terminal capable of driving one or two TTL loads. This is not quite enough to drive the MOC3011, nor can it be connected directly to an SCR or TRIAC, because computer common is not normally referenced to one side of the ac supply. Standard 7400 series gates can provide an input compatible with the output of an MC6821, MC6846 or similar peripheral interface adaptor and can directly drive the MOC3011. If the second input of a 2 input gate is tied to a simple timing circuit, it will also provide energization of the TRIAC only at the zero crossing of the ac line voltage as shown in Figure 6.32. This technique extends the life of incandescent lamps, reduces the surge current strains on the TRIAC, and reduces EMI generated by load switching. Of course, zero crossing can be generated within the microcomputer itself, but this requires considerable software overhead and usually just as much hardware to generate the zero-crossing timing signals.

APPLICATIONS USING THE ZERO CROSSING TRIAC DRIVER

For applications where EMI induced, non-zero crossing-load switching is a problem, the zero crossing TRIAC driver is the answer. This TRIAC driver can greatly simplify the suppression of EMI for only a nominal increased cost. Examples of several applications using the MOC3031, 41 follows.

MATRIX SWITCHING

Matrix, or point-to-point switching, represents a method of controlling many loads using a minimum number of components. On the 115 V line, the MOC3031 is ideal for this application; refer to Figure 6.33. The large static dv/dt rating of the MOC3031 prevents unwanted

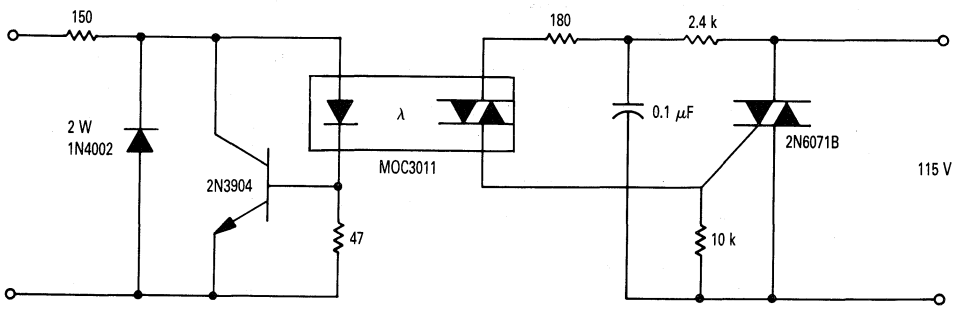


Figure 6.31. Solid-State Relay

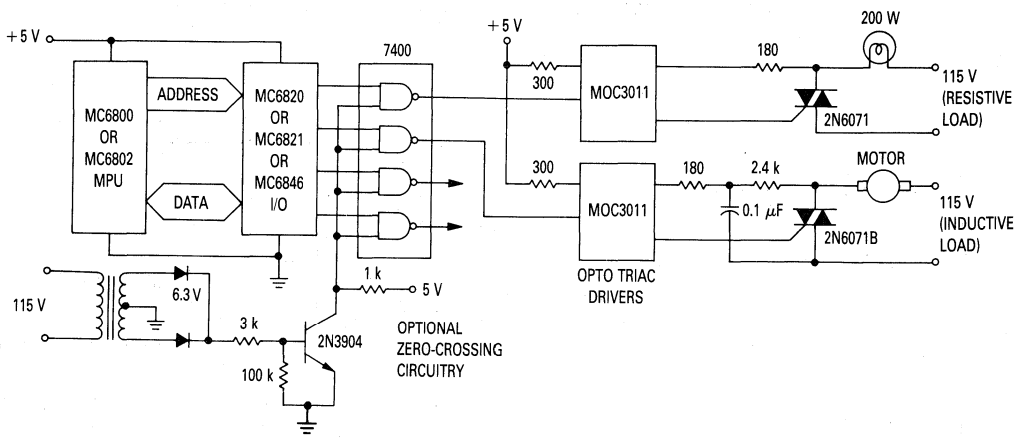


Figure 6.32. Interfacing an M6800 Microcomputer System to 115 Vac Loads

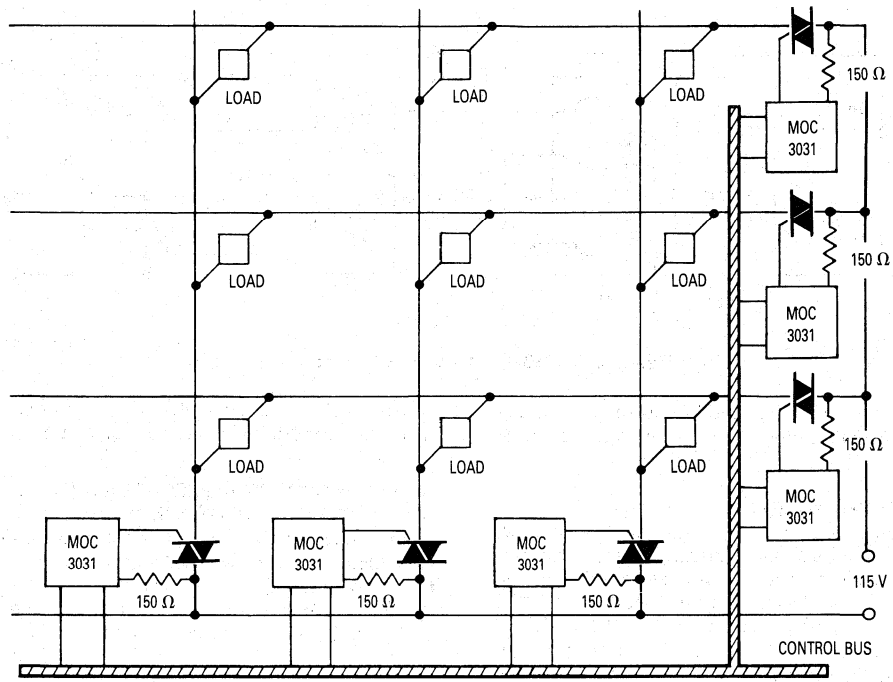


Figure 6.33. Matrix Switching

loads from being triggered on. This might occur, in the case of non-zero crossing TRIAC drivers, when a TRIAC driver on a vertical line was subjected to a large voltage ramp due to a TRIAC on a horizontal line being switched on. Since non-zero crossing TRIAC drivers have lower static dv/dt ratings, this ramp would be sufficiently large to trigger the device on.

R is determined as before:

$$R_{(min)} = \frac{V_{in(pk)}}{I_{TSM}}$$

$$= \frac{170\ V}{1.2\ A} = 150\ \text{ohms}$$

POWER RELAYS

The use of high-power relays to control the application of ac power to various loads is a very widespread practice. Their low contact resistance causes very little power loss and many options in power control are possible due to their multipole-multithrow capability. The MOC3041 is well suited to the use of power relays on the 230 Vac line; refer to Figure 6.34. The large static dv/dt of this device makes a snubber network unnecessary, thus reducing component count and the amount of printed circuit board space required. A non-zero crossing TRIAC

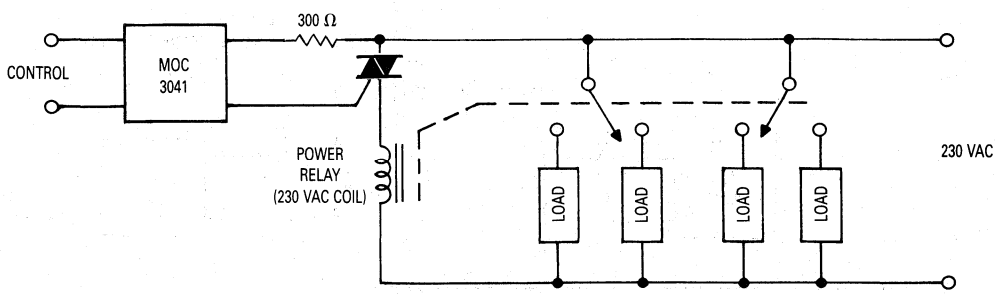


Figure 6.34. Power Relay Control

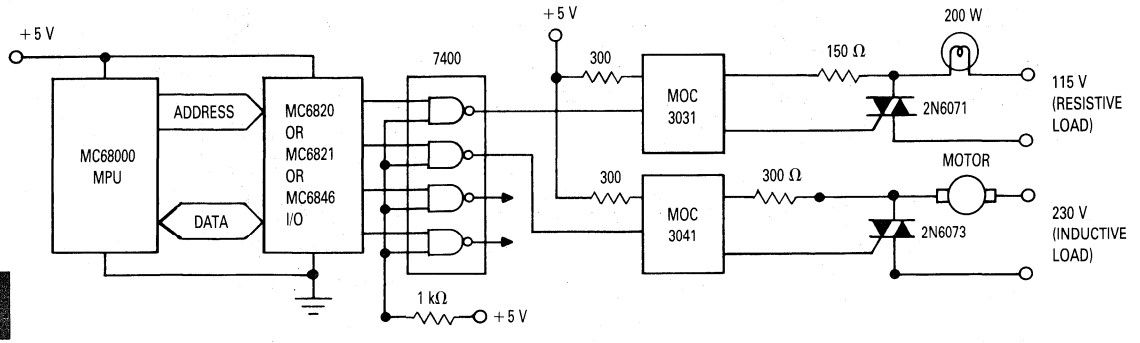


Figure 6.35. M68000 Microcomputer Interface

driver (MOC3021) could be used in this application, but its lower static dv/dt rating would necessitate a snubber network.

MICROCOMPUTER INTERFACE

The output of most microcomputer input/output (I/O) ports is a TTL signal capable of driving several TTL gates. This is insufficient to drive a zero-crossing TRIAC driver. In addition, it cannot be used to drive an SCR or TRIAC directly, because computer common is not usually referenced to one side of the ac supply. However, standard 7400 NAND gates can be used as buffers to accept the output of the I/O port and in turn, drive the MOC3031 and/or MOC3041; refer to Figure 6.35.

The zero-crossing feature of these devices extends the life of incandescent lamps, reduces inrush currents and minimizes EMI generated by load switching.

AC MOTORS

The large static dv/dt rating of the zero-crossing TRIAC drivers make them ideal when controlling ac motors. Figure 6.36 shows a circuit for reversing a two phase motor using the MOC3041. The higher voltage MOC3041 is required, even on the 115 Vac line, due to the mutual and self-inductance of each of the motor windings, which may cause a voltage much higher than 115 Vac to appear across the winding which is not conducting current.

DETERMINING LIMITING RESISTOR R FOR A HIGH-WATTAGE INCANDESCENT LAMP

Many high-wattage incandescent lamps suffer shortened lifetimes when switched on at ac line voltages other than zero. This is due to a large inrush current destroying the filament. A simple solution to this problem is the use of the MOC3041 as shown in Figure 6.37. The MOC3041 may be controlled from a switch or some form of digital logic.

The minimum value of R is determined by the maximum surge current rating of the MOC3041 (I_{TSM}):

$$R_{(min)} = \frac{V_{in(pk)}}{I_{TSM}} = \frac{V_{in(pk)}}{1.2\ A} \quad (10)$$

On a 230 Vac Line:

$$R_{(min)} = \frac{340\ V}{1.2\ A} = 283\ \text{ohms} \quad (11)$$

In reality, this would be a 300 ohm resistor.

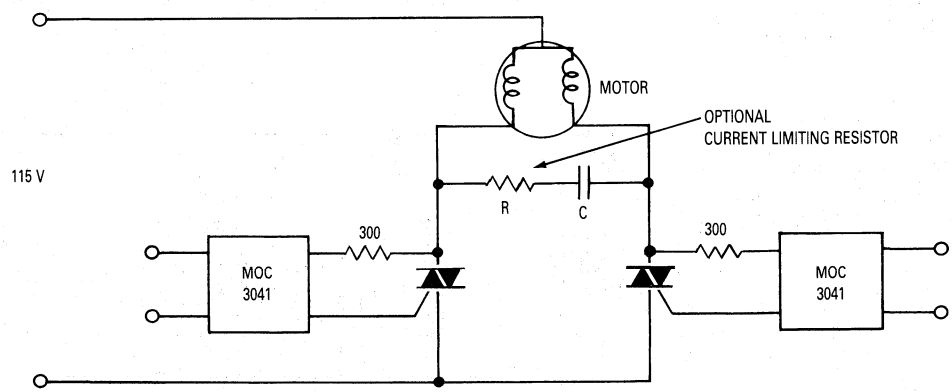


Figure 6.36. Reversing Motor Circuit

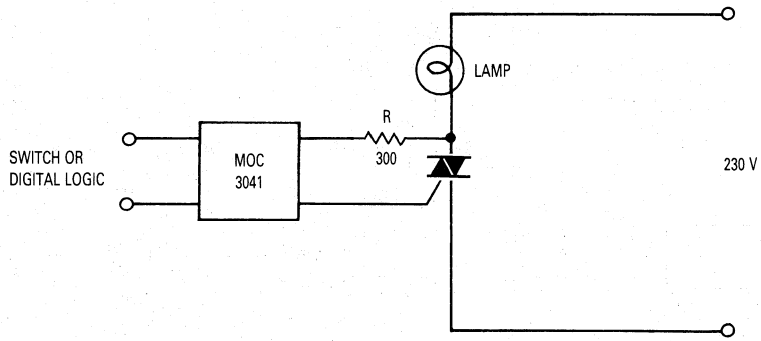


Figure 6.37. High-Wattage Lamp Control

AC POWER CONTROL WITH SOLID-STATE RELAYS

The Solid-State Relay (SSR) as described below, is a relay function with:

a. Four Terminals (Two Input, Two Output)
b. DC or AC Input
c. Optical Isolation Between Input and Output
d. Thyristor (SCR or TRIAC) Output
e. Zero Voltage Switching Output (Will Only Turn On Close to Zero Volts)
f. AC Output (50 or 60 Hz)

Figure 6.38 shows the general format and waveforms of the SSR. The input on/off signal is conditioned (perhaps only by a resistor) and fed to the Light-Emitting-Diode (LED) of an optoelectronic-coupler. This is ANDed with a go signal that is generated close to the zero-crossing of the line, typically ≤ 10 Volts. Thus, the output is not gated on via the amplifier except at the zero-crossing of the line voltage. The SSR output is then re-gated on at the beginning of every half-cycle until the input on signal is removed. When this happens, the thyristor output stays on until the load current reaches zero, and then turns off.

ADVANTAGES AND DISADVANTAGES OF SSRs

The SSR has several advantages that make it an attractive choice over its progenitor, the Electromechanical Relay (EMR) although the SSR generally costs more than its electromechanical counterpart. These advantages are:

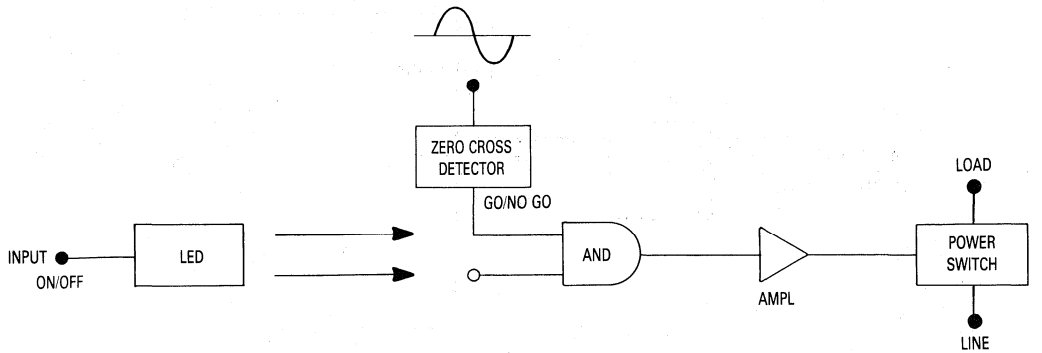
1) No Moving Parts — the SSR is all solid-state. There are no bearing surfaces to wear, springs to fatigue, assemblies to pick up dust and rust. This leads to several other advantages.
2) No Contact Bounce — this in turn means no contact wear, arcing, or Electromagnetic Interference (EMI) associated with contact bounce.
3) Fast Operation — usually less than 10 μs. Fast turn-on time allows the SSR to be easily synchronized with line zero-crossing. This also minimizes EMI and can greatly increase the lifetime of tungsten lamps, of considerable value in applications such as traffic signals.
4) Shock and Vibration Resistance — the solid-state contact cannot be "shaken open" as easily as the EMR contact.
5) Absence of Audible Noise — this devolves from the lack of moving mechanical parts.
6) Output Contact Latching — the thyristor is a latching device, and turns off only at the load current zero-crossing, minimizing EMI.
7) High Sensitivity — the SSR can readily be designed to interface directly with TTL and CMOS logic, simplifying circuit design.
8) Very Low Coupling Capacitance Between Input and Output. This is a characteristic inherent in the optoelectronic-coupler used in the SSR, and can be useful in areas such as medical electronics where the reduction of stray leakage paths is important.

This list of advantages is impressive, but of course, the designer has to consider the following disadvantages:

1) Voltage Transient Resistance — the ac line is not the clean sine wave obtainable from a signal generator. Superimposed on the line are voltage spikes from motors, solenoids, EMRs (ironical), lightning, etc. The solid-state components in the SSR have a finite voltage rating and must be protected from such spikes, either with RC networks (snubbing), zener diodes, MOVs or selenium voltage clippers. If not done, the thyristors will turn on for part of a half cycle, and at worst, they will be permanently damaged, and fail to block voltage. For critical applications a safety margin on voltage of 2 to 1 or better should be sought.

   The voltage transient has at least two facets — the first is the sheer amplitude, already discussed. The second is its frequency, or rate-of-rise of voltage (dv/dt). All thyristors are sensitive to dv/dt to some extent, and the transient must be snubbed, or "soaked up," to below this level with an RC network.(1) Typically this rating ("critical" or "static"

(1) For a more thorough discussion of snubbers, see page 1-3-9.



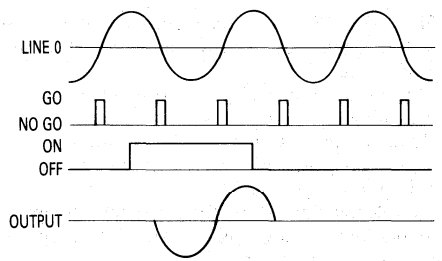


Figure 6.38. SSR Block Diagram

dv/dt) is 50 to 100 V/μs at maximum temperature. Again the failure mode is to let through, to a half-cycle of the line, though a high energy transient can cause permanent damage. Table 6.I gives some starting points for snubbing circuit values. The component values required depend on the characteristics of the transient, which are usually difficult to quantify. Snubbing across the line as well as across the SSR will also help.

Load Current A rms	Resistance Ω	Capacitance μF
5	47	0.047
10	33	0.1
25	10	0.22
40	22	0.47

Table 6.I. Typical Snubbing Values

2) Voltage Drop — The SSR output contact has some offset voltage — approximately 1 V, depending on current, causing dissipation. As the thyristor has an operating temperature limit of +125°C, this heat must be removed, usually by conduction to air via a heat sink or the chassis.
3) Leakage Current — When an EMR is open, no current can flow. When an SSR is open however, it does not have as definite an off condition. There is always some current leakage through the output power switching thyristor, the control circuitry, and the snubbing network. The total of this leakage is usually 1 to 10 mA rms — three or four orders of magnitude less than the on-state current rating.
4) Multiple Poles — are costly to obtain in SSRs, and three phase applications may be difficult to implement.
5) Nuclear Radiation — SSRs will be damaged by nuclear radiation.

TRIAC SSR CIRCUIT

Many SSR circuits use a TRIAC as the output switching device. Figure 6.39(a) shows a typical TRIAC SSR circuit. The control circuit is used in the SCR relay as well, and is defined separately. The input circuit is TTL compatible. Output snubbing for inductive loads will be described later.

A sensitive-gate SCR (SCR1) is used to gate the power TRIAC, and a transistor amplifier is used as an interface between the optoelectronic-coupler and SCR1. (A sensitive-gate SCR and a diode bridge are used in preference to a sensitive gate TRIAC because of the higher sensitivity of the SCR.)

CONTROL CIRCUIT OPERATION

The operation of the control circuit is straightforward. The AND function of Figure 6.38 is performed by the wired-NOR collector configuration of the small-signal transistors Q1 and Q2. Q1 clamps the gate of SCR1 if optoelectronic-coupler OC1 is off. Q2 clamps the gate if there is sufficient voltage at the junction of the potential divider R4, R5 to overcome the V_{BE} of Q2. By judicious selection of R4 and R5, Q2 will clamp SCR1's gate if more than approximately 5 Volts appear at the anode of SCR1; i.e., Q2 is the zero-crossing detector.

If OC1 is on, Q1 is clamped off, and SCR1 can be turned on by current flowing down R6, only if Q2 is also off — which it is only at zero crossing.

The capacitors are added to eliminate circuit race conditions and spurious firing, time ambiguities in operation. Figure 6.39(b) shows the full-wave rectified line that appears across the control circuit. The zero voltage firing

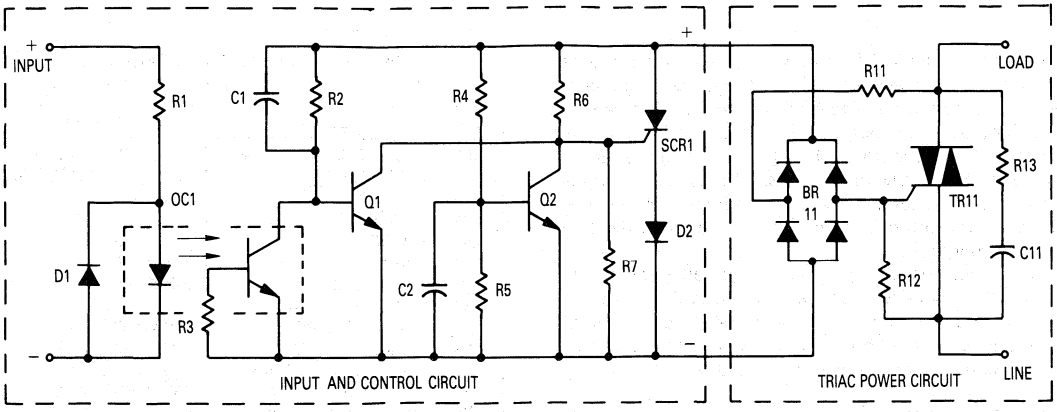


Figure 6.39(a). TRIAC SSR Circuit

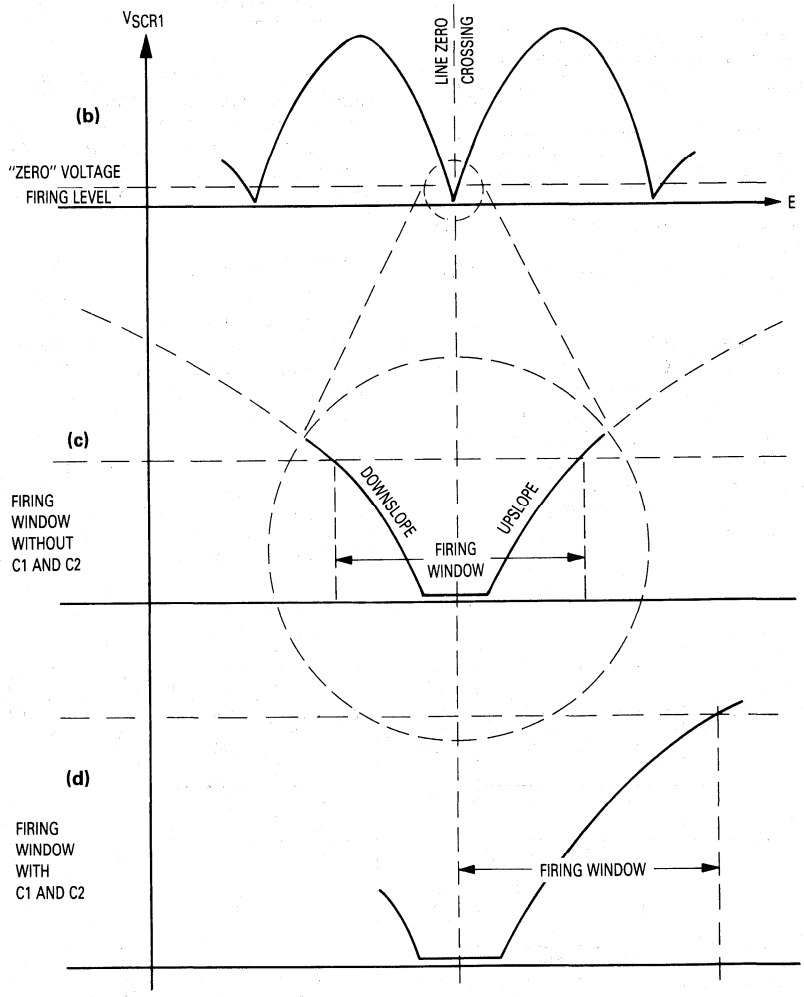


Figure 6.39. Firing Windows

level is shown in 6.39(b) and 6.39(c), expanded in time and voltage. A race condition exists on the up-slope of the second half-cycle in that SCR1 may be triggered via R6 before Q1 has enough base current via R2 to clamp SCR1's gate. C1 provides current by virtue of the rate of change of the supply voltage, and Q1 is turned on firmly as the supply voltage starts to rise, eliminating any possibility of unwanted firing of the SSR; thus eliminating the race condition.

This leaves the possibility of unwanted firing of the SSR on the down-slope of the first half cycle shown. C2 provides a phase shift to the zero voltage potential divider, and Q2 is held on through the real zero-crossing. The resultant window is shown in 6.39(d).

1

CONTROL CIRCUIT COMPONENTS

The parts list for the control circuit at two line voltages is shown in Table 6.II.

R1 limits the current in the input LED of OC1. The input circuit will function over the range of 3 to 33 Vdc.

D1 provides reverse voltage protection for the input of OC1.

D2 allows the gate of SCR1 to be reverse biased, providing better noise immunity and dv/dt performance.

R7 eliminates pickup on SCR1's gate through the zero-crossing interval.

SCR1 is a sensitive gate SCR; the 2N5064 is a TO-92 device, the 2N6240 is a Case 77 device.

Alternatives to the simple series resistor (R1) input circuit will be described later.

POWER CIRCUIT COMPONENTS

The parts list for the TRIAC power circuit in Figure 6.39(a) is shown in Table 6.III for several rms current ratings, and two line voltages. The metal TRIACs are in the half-inch pressfit package in the isolated stud configuration; the plastic TRIACs are in the TO-220 Thermowatt package. R12 is chosen by calculating the peak control circuit off-state leakage current and ensuring that the voltage drop across R12 is less than the $V_{GT(MIN)}$ of the TRIAC.

Table 6.II. Control Circuit Parts List

	Line Voltage	
Part	**120 V rms**	**240 V rms**
C1	220 pF, 20%, 200 Vdc	100 pF, 20%, 400 Vdc
C2	0.022 μF, 20%, 50 Vdc	0.022 μF, 20%, 50 Vdc
D1	1N4001	1N4001
D2	1N4001	1N4001
OC1	MOC1005	MOC1005
Q1	MPS5172	MPS5172
Q2	MPS5172	MPS5172
R1	1 kΩ, 10%, 1 W	1 kΩ, 10%, 1 W
R2	47 kΩ, 5%, 1/2 W	100 kΩ, 5%, 1 W
R3	1 MΩ, 10%, 1/4 W	1 MΩ, 10%, 1/4 W
R4	110 kΩ, 5%, 1/2 W	220 kΩ, 5%, 1/2 W
R5	15 kΩ, 5%, 1/4 W	15 kΩ, 5%, 1/4 W
R6	33 kΩ, 10%, 1/2 W	68 kΩ, 10%, 1 W
R7	10 kΩ, 10%, 1/4 W	10 kΩ, 10%, 1/4 W
SCR1	2N5064	2N6240

C11 must be an ac rated capacitor, and with R13 provides some snubbing for the TRIAC. The values shown for this network are intended more for inductive load commutating dv/dt snubbing than for voltage transient suppression. Consult the individual data sheets for the dissipation, temperature, and surge current limits of the TRIACs.

TRIACs AND INDUCTIVE LOADS

The TRIAC is a single device which to some extent is the equivalent of two SCRs inverse parallel connected; certainly this is so for resistive loads. Inductive loads however, can cause problems for TRIACs, especially at turn-off.

A TRIAC turns off every line half-cycle when the line current goes through zero. With a resistive load, this coincides with the line voltage also going through zero. The TRIAC must regain blocking-state before there are more than 1 or 2 Volts of the reverse polarity across it — at 120 V rms, 60 Hz line this is approximately 30 μs. The TRIAC has not completely regained its off-state charac-

Table 6.III. TRIAC Power Circuit Parts List

Voltage		**120 V rms**				**240 V rms**			
rms Current Amperes		**8**	**12**	**25**	**40**	**8**	**12**	**25**	**40**
BR11		MDA102A	MDA102A	MDA102A	MDA102A	MDA104A	MDA104A	MDA104A	MDA104A
C11, μF (10%, line voltage ac rated)		0.047	0.047	0.1	0.1	0.047	0.047	0.1	0.1
R11 (10%, 1 W)		39	39	39	39	39	39	39	39
R12 (10%, 1/2 W)		18	18	18	18	18	18	18	18
R13 (10%, 1/2 W)		620	620	330	330	620	620	330	330
TR11	Plastic	2N6342	2N6342A	—	—	2N6343	2N6343A	—	—
	Metal	—	T4121B	2N6163	T6420B	—	T4121D	T4121D	T6420D

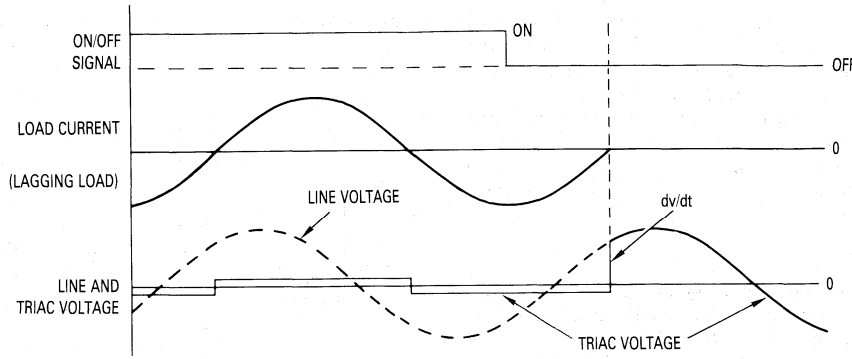


Figure 6.40. Commutating dv/dt

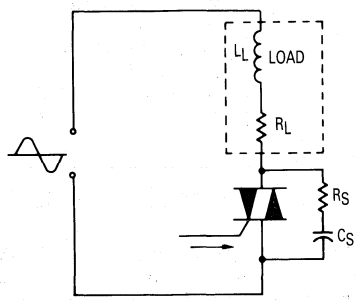


Figure 6.41. TRIAC with Snubber Network

teristics, but does so as the line voltage increases at the 60 Hz rate.

Figure 6.40 indicates what happens with an inductive or lagging load. The on signal is removed asynchronously and the TRIAC, a latching device, stays on until the next current zero. As the current is lagging the applied voltage, the line voltage at that instant appears across the TRIAC. It is this rate-of-rise of voltage, the commutating dv/dt, that must be limited in TRIAC circuits, usually to a few volts per microsecond. This is normally done by use of a snubber network R_S and C_S as shown in Figure 6.41.

SCRs have less trouble as each device has a full half-cycle to turn off and, once off, can resist dv/dt to the critical value of 50 to 100 V/μs.

CHOOSING THE SNUBBING COMPONENTS(1)

There are no easy methods for selecting the values of R_S and C_S in Figure 6.41 required to limit commutating dv/dt. The circuit is a damped tuned circuit comprised by R_S, C_S, R_L and L_L, and to a minor extent the junction capacitance of the TRIAC. At turn-off this circuit receives a step impulse of line voltage which depends on the power factor of the load. Assuming the load is fixed, which is normally the case, the designer can vary R_S and

(1) For a more thorough discussion of snubbers, see page 1-3-9.

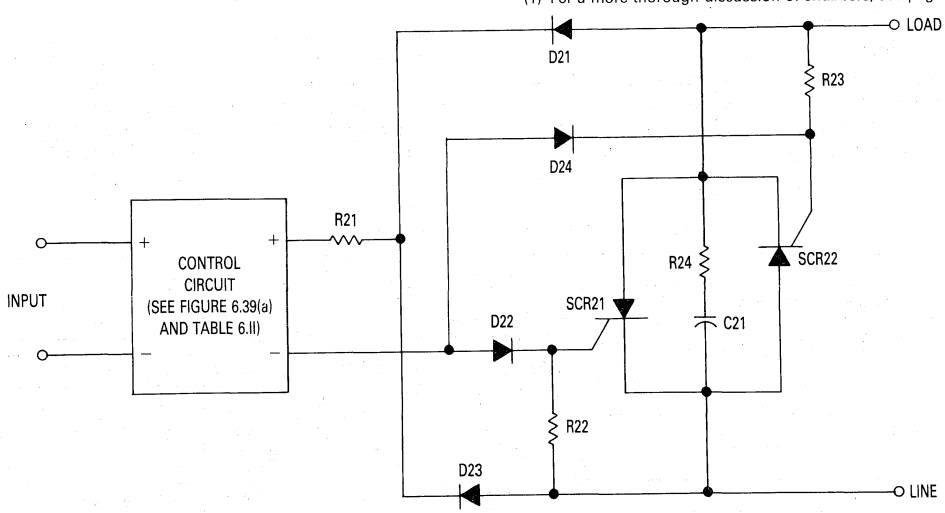


Figure 6.42. SCR SSR Circuit

$_S$. C_S can be increased to decrease the commutating dv/dt; R_S can be increased to decrease the resonant overring of the tuned circuit — to increase damping. This can be done empirically, beginning with the values for C11 and R13 given in Table 6.III, and aiming at close to critical damping and the data sheet value for commutating dv/dt. Reduced temperatures, voltages, and off-going di/dt (rate-of-change of current at turn-off) will give some safety margin.

SCR SSR CIRCUIT

The inverse parallel connected Silicon Controlled Rectifier (SCR) pair (shown in Figure 6.42) is less sensitive to commutating dv/dt. Other advantages are the improved thermal and surge characteristics of having two devices; the disadvantage is increased cost.

The SCR power circuit can use the same control circuit as the TRIAC Circuit shown in Figure 6.39(a). In Figure 6.42, for positive load terminal and when the control circuit is gated on, current flows through the load, D21, R21, SCR1, D22, the gate of SCR21 and back to the line, thus turning on SCR21. Operation is similar for the other line polarity. R22 and R23 provide a path for the off-state leakage of the control circuit and are chosen so that the voltage dropped across them is less than the $V_{GT(MIN)}$ of the particular SCR. R24 and C21 provide snubbing and line transient suppression, and may be chosen from Table 6.IV or from the C11, R13 rows of Table 6.III. The latter values will provide less transient protection but also less off-state current, with the capacitor being smaller. Other circuit values are shown in Table 6.IV.

Consult the individual data sheets for packages and dissipation, temperature, and surge current limits.

While the SCRs have much higher dv/dt commutation ability, with inductive loads, attention should be paid to maintaining the dv/dt below data sheet levels.

ALTERNATE INPUT CIRCUITS

CMOS COMPATIBLE

The 1 kΩ resistor, R1, shown in Figure 6.39(a) and Table 6.II, provide an input that is compatible with the current that a TTL gate output can sink. The resistor R1 must be changed for CMOS compatibility, aiming at 2 mA in the LED for adequate performance to 100°C. At 2 mA do not use the CMOS output for any other function, as a LOGIC 0 or 1 may not be guaranteed. Assume a forward voltage drop of 1.1 V for the LED, and then make the Ohm's Law calculation for the system dc supply voltage, thus defining a new value for R1.

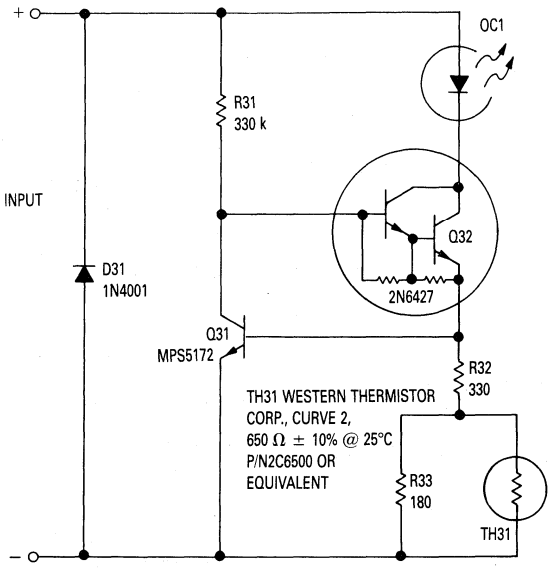


Figure 6.43. TTL/CMOS Compatible Input

TTL/CMOS COMPATIBLE

To be TTL compatible at 5 Volts and CMOS compatible over 3 to 15 Volts, a constant current circuit is required, such as the one in Figure 6.43. The current is set by the V_{BE} of Q31 and the resistance of the R32, R33, and thermistor TH31 network, and is between 1 and 2 mA, higher at high temperatures to compensate for the reduced transmission efficiency of optoelectronic-couplers at higher temperature. The circuit of Figure 6.43 gives an equivalent impedance of approximately 50 kΩ. The circuit

Table 6.IV. SCR Power Circuit Parts List

Voltage		120 V rms				240 V rms			
rms Current Amperes		5	11	22	49	5	11	22	49
C21 (10%, line voltage ac rated)		← SEE TEXT →							
D21–24		1N4003	1N4003	1N4003	1N4003	1N4004	1N4004	1N4004	1N4004
R21 (10%, 1 W)		39	39	39	39	39	39	39	39
R22, 23 (10%, 1/2 W)		18	18	18	18	18	18	18	18
R24		← SEE TEXT →							
SCR21,22	Plastic	2N6239	MCR218-4	2N6402	—	2N6240	MCR218-6	2N6403	—
	Metal	—	2N4170	2N6168	2N6172	—	2N4172	2N6169	2N6173

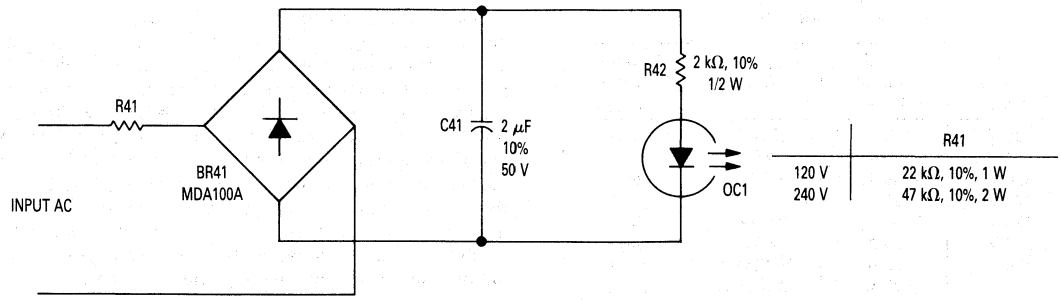


Figure 6.44. AC Compatible Input

performs adequately over 3 to 33 Vdc and −40 to +100°C. Note that though the SSR is protected against damage from improperly connected inputs, the external circuit is not, as D31 acts as a bypass for a wrongly connected input driver.

AC LINE COMPATIBLE

To use SSRs as logic switching elements is inefficient, considering the availability and versatility of logic families such as CMOS. When it is convenient to trigger from ac, a circuit such as shown in Figure 6.44 may be used. The capacitor C41 is required to provide current to the LED of OC1 through the zero-crossing time. An in-phase input voltage gives the worst case condition. The circuit gives 2 mA minimum LED current at 75% of nominal line voltage.

INVERSE PARALLEL SCRs FOR POWER CONTROL

TRIACs are very useful devices. They end up in solid state relays, lamp drivers, motor controls, sensing and detection circuits; just about any industrial full-wave application. But in high-frequency applications or those requiring high voltage or current, their role is limited by their present physical characteristics, and they become very expensive at current levels above 40 amperes rms.

SCRs can be used in an inverse-parallel connection to bypass the limitations of a TRIAC. A simple scheme for doing this is shown in Figure 6.45. Table 6.V. lists suggested SCR's for this circuit configuration. The control device can take any of many forms, shown is the reed relay (Figure 6.45). TRIACs and Opto couplers can be inserted at point A-A to replace the reed relay.

Compared to a TRIAC, an inverse-parallel configuration has distinct advantages. Voltage and current capabilities are dependent solely on SCR characteristics with ratings today of over a thousand volts and several hundred amps. Because each SCR operates only on a half-wave basis, the system's rms current rating is $\sqrt{2}$ times the SCR's rms current rating (see Suggested SCR chart). The system has the same surge current rating as the SCRs do. Operation at 400 Hz is also no problem. While turn-off time and dv/dt limits control TRIAC operating speed, the recovery characteristics of an SCR need only be better than the appropriate half-wave period.

With inductive loads you no longer need to worry about commutating dv/dt, either. SCRs only need to withstand static dv/dt, for which they are typically rated an order of magnitude greater than TRIACs are for commutating dv/dt.

Better reliability can be achieved by replacing the reed relay with a low current TRIAC to drive the SCRs, although some of its limitations come with it. In the preferred circuit of Figure 6.45(b), the main requirements of

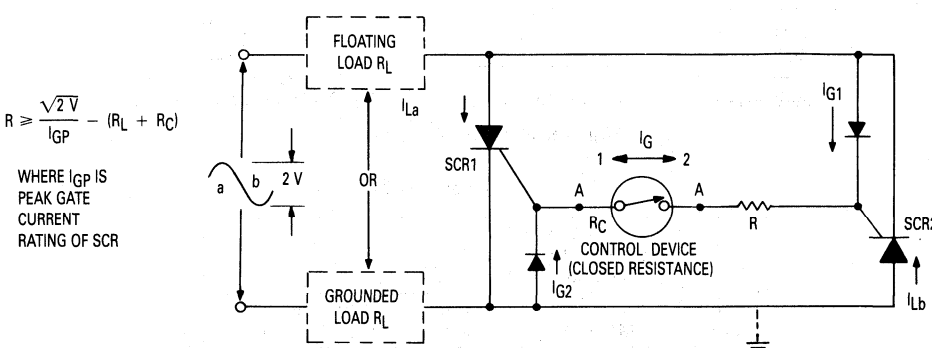


Figure 6.45. Use of Inverse Parallel SCRs

Table 6.V. Suggested SCRs

Line Voltage (rms)	Load Current (rms)	SCR	SCR Current Rating (rms)
120 V	25 A	2N5169	20 A
220 V	25 A	2N5170	20 A
120 V	35 A	2N6506	25 A
220 V	35 A	2N6507	25 A
120 V	50 A	2N3897	35 A
220 V	50 A	2N3898	35 A
120 V	75 A	MCR64-4	55 A
220 V	75 A	MCR64-6	55 A

1

the TRIAC are that it be able to block the peak system voltage and that it have a surge current rating compatible with the gate current requirements of the SCRs. This is normally so small that a TO-92 cased device is adequate to drive the largest SCRs.

In circuits like Figure 6.45, the control devices alternately pass the gate currents I_{G1} and I_{G2} during the "a" and "b" half cycles, respectively. I_{La} and I_{Lb} are the load currents during the corresponding half cycles. Each SCR then gets the other half cycle for recovery time. Heat sinking can also be done more efficiently, since power is being dissipated in two packages, rather than all in one. The load can either be floated or grounded. If the SCRs are not of the shunted-gate variety, a gate-cathode resistance should be added to shunt the leakage current at higher temperatures. The diodes act as steering diodes so the gate-cathode junctions are not avalanched. The blocking capability of the diodes need only be as high as the V_{GT} of the SCRs. A snubber can also be used if conditions dictate.

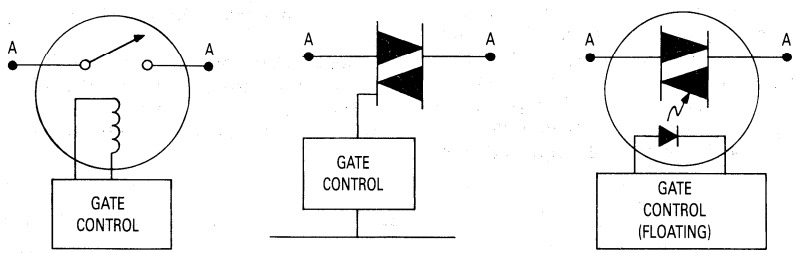
This circuit offers several benefits. One is a considerable increase in gain. This permits driving the TRIAC with almost any other semiconductors such as linear ICs, photosensitive devices and logic, including MOS. If necessary, it can use an optically coupled TRIAC driver to isolate (up to 7500 V isolation) delicate logic circuits from the power circuit (see Figure 6.46(c)). Table 6.VI. lists suggested components. Another benefit is being able to gate the TRIAC with a supply of either polarity. Probably the most important benefit of the TRIAC/SCR combination is its ability to handle variable-phase applications — nearly impossible for non solid-state control devices.

INTERFACING DIGITAL CIRCUITS TO THYRISTOR CONTROLLED AC LOADS

Because they are bidirectional devices, TRIACs are the most common thyristor for controlling ac loads. A TRIAC can be triggered by either a positive or negative gate signal on either the positive or negative half-cycle of applied MT2 voltage, producing four quadrants of operation. However, the TRIAC's trigger sensitivity varies with the quadrant, with quadrants II and III (gate signal negative and MT2 either positive or negative) being the most sensitive and quadrant IV (gate positive, MT2 negative) the least sensitive.

For driving a TRIAC with IC logic, quadrants II and III are particularly desirable, not only because less gate trigger current is required, but also because IC power dissipation is reduced since the TRIAC can be triggered by an "active low" output from the IC.

There are other advantages to operating in quadrants II and III. Since the rate of rise of on-state current of a TRIAC (di/dt) is a function of how hard the TRIAC's gate



(a). Reed Relay **(b). Low-Current TRIAC** **(c). Optically Coupled TRIAC Driver**

Figure 6.46. Control Devices

Table 6.VI. Driver TRIACs

Line Voltage	Gate Negative Or In Phase With Line Voltage	Gate Positive	Optically Coupled
120	MAC97A4	MAC97A4	MOC3030*, 3011
220	MAC97A6	MAC97A6	MOC3020, MOC3021

*Includes inhibit circuit for zero crossover firing.

is turned on, a given IC output in quadrants II and III will produce a greater di/dt capability than in the less sensitive quadrant IV. Moreover, harder gate turn-on could reduce di/dt failure. One additional advantage of quadrant II and III operation is that devices specified in all four quadrants are generally more expensive than devices specified in quadrants I, II and III, due to the additional testing involved and the resulting lower yields.

USING TRIACs

Once the TRIAC load requirements are defined, an appropriate device selection can be made by referring to the TRIAC current ratings of Table 6.VII.

Two important thyristor parameters are gate trigger current (I_{GT}) and gate trigger voltage (V_{GT}).

I_{GT} (Gate Trigger Current) is the amount of gate trigger current required to turn the device on. I_{GT} has a negative temperature coefficient — that is, the trigger current required to turn the device on increases with decreasing temperature. If the TRIAC must operate over a wide temperature range, its I_{GT} requirement could double at the low temperature extreme from that of its 25°C rating.

It is good practice, if possible, to trigger the thyristor with three to ten times the I_{GT} rating for the device. This increases its di/dt capability and ensures adequate gate trigger current at low temperatures.

V_{GT} (Gate Trigger Voltage) is the voltage the thyristor gate needs to ensure triggering the device on. This voltage is needed to overcome the input threshold voltage of the device. To prevent thyristor triggering, gate voltage should be kept to approximately 0.4 V or less.

Like I_{GT}, V_{GT} increases with decreasing temperature.

INDUCTIVE LOAD SWITCHING

Switching of inductive loads, using TRIACs, may require special consideration in order to avoid false triggering. This false-trigger mechanism is illustrated in Figure 6.47 which shows an inductive circuit together with the accompanying waveforms.

Table 6.VII. TRIACs with Various Current Ratings

Sensitive Gate TRIACs	
TRIAC	**$I_{T(rms)}$**
MAC97, 97A, 97B	0.6 A
2N6068A	4 A
MAC228, A Series	8 A
Non-sensitive Gate TRIACs	
TRIAC	**$T_{T(rms)}$**
MAC97, A Series	0.6
2N6071-75	4
2N6342-49	8
2N6342A-49A	12
MAC15A, 15AFP Series	15
MAC223A, 223AFP Series	25
2N6160-65	30
2N5444-46	40

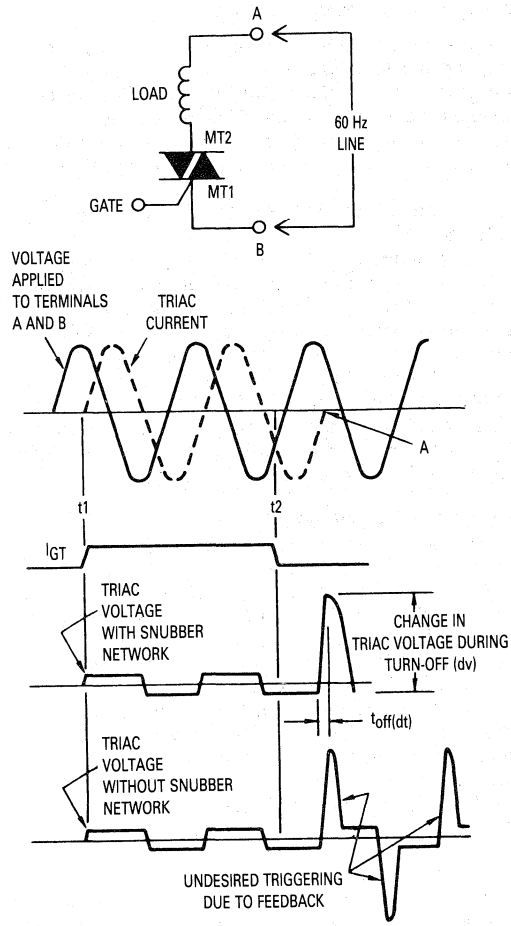


Figure 6.47. Inductive Load TRIAC Circuit and Equivalent Waveforms

As shown, the TRIAC is triggered on, at t1, by the positive gate current (I_{GT}). At that point, TRIAC current flows and the voltage across the TRIAC is quite low since the TRIAC resistance, during conduction, is very low.

From point t1 to t2 the applied I_{GT} keeps the TRIAC in a conductive condition, resulting in a continuous sinusoidal current flow that lags the applied voltage by 90° for this pure inductive load.

At t2, I_{GT} is turned off, but TRIAC current continues to flow until it reaches a value that is less than the sustaining current (I_H), at point A. At that point, TRIAC current is cut off and TRIAC voltage is at a maximum. Some of that voltage is fed back to the gate via the internal capacitance (from MT2 to gate) of the TRIAC.

TTL-TO-THYRISTOR INTERFACE

The subject of interfacing requires a knowledge of the output characteristics of the driving stages as well as the input requirements of the load. This section describes the driving capabilities of some of the more popular TTL circuits and matches these to the input demands of thyristors under various practical operating conditions.

TTL CIRCUITS WITH TOTEM-POLE OUTPUTS (e.g. 5400 SERIES)

The configuration of a typical totem-pole connected TTL output stage is illustrated in Figure 6.48(a). This stage is capable of "sourcing" current to a load, when the load is connected from V_{out} to ground, and of "sinking" current from the load when the latter is connected from V_{out} to V_{CC}. If the load happens to be the input circuit of a TRIAC (gate to MT_1), the TRIAC will be operating in quadrants I and IV (gate goes positive) when connected from V_{out} to ground, and of "sinking" II and III (gate goes negative) when connected from V_{out} to V_{CC}.

QUADRANT I-IV OPERATION

Considering first the gate-positive condition, Figure 6.48(b), the operation of the circuit is as follows:

When V_{in} to the TTL output stage is low (logical "zero"), transistors Q1 and Q3 of that stage are cut off, and Q2 is conducting. Therefore, Q2 sources current to the thyristor, and the thyristor would be triggered on during the $V_{in} = 0$ condition.

When V_{in} goes high (logical "one"), transistors Q1 and Q3 are on and Q2 is off. In this condition depicted by the equivalent circuit transistor Q3 is turned on and its collector voltage is, essentially, $V_{CE(sat)}$. As a result, the TRIAC is clamped off by the low internal resistance of Q3.

QUADRANT II-III OPERATION

When the TRIAC is to be operated in the more sensitive quadrants II and III (negative-gate turn-on), the circuit in Figure 6.49(a) may be employed.

With Q3 in saturation, as shown in the equivalent circuit of 6.49(b), its saturation voltage is quite small, leaving virtually the entire $-V_{EE}$ voltage available for thyristor turn-on. This could result in a TRIAC gate current that exceeds the current limit of Q3, requiring a current-limiting series resistor, ($R_{(lim)}$).

When the V_{out} level goes high, Q3 is turned off and Q2 becomes conductive. Under those conditions, the TRIAC gate voltage is below V_{GT} and the TRIAC is turned off.

DIRECT-DRIVE LIMITATIONS

With sensitive-gate TRIACs, the direct connection of a TRIAC to a TTL circuit may sometimes be practical. However, the limitations of such circuits must be recognized.

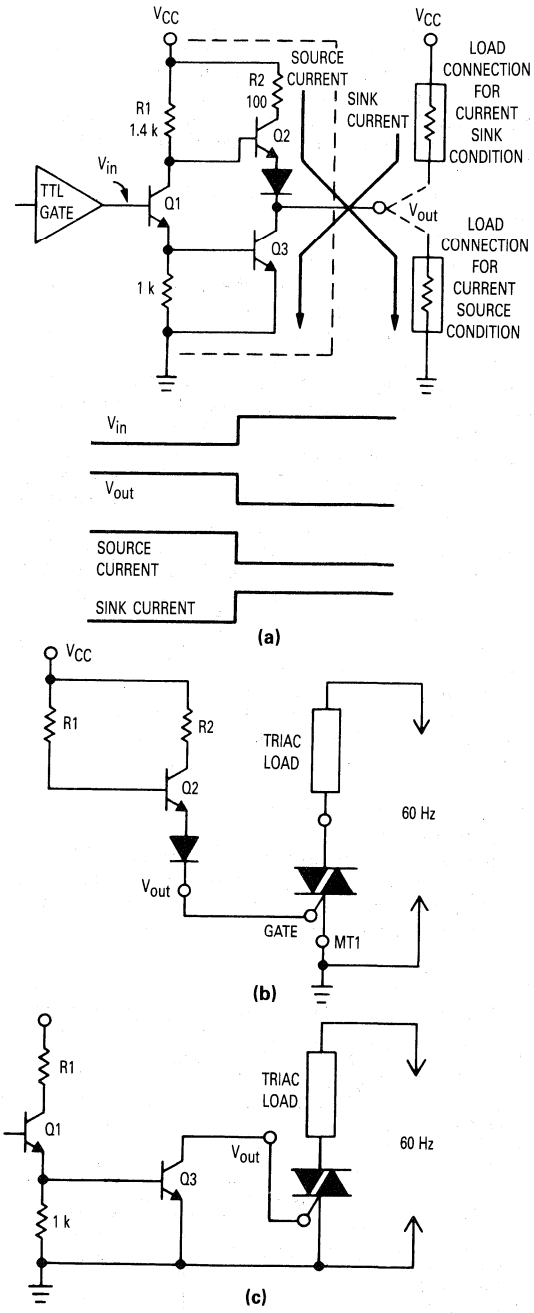


Figure 6.48. Totem-Pole Output Circuit TTL Logic, Together with Voltage and Current Waveforms, (b) Equivalent Circuit for Triggering TRIAC with a Positive Voltage — TRIAC-On Condition, (c) TRIAC-Off Condition

For example:

For TTL circuits, the "high" logic level is specified as 2.4 volts. In the circuit of Figure 6.48(a), transistor Q2 is capable of supplying a short-circuit output current (I_{SC}) of 20 to 55 mA (depending on the tolerances of R1 and R2, and on the h_{FE} of Q2). Although this is adequate to turn a sensitive-gate TRIAC on, the specified 2.4 volt (high) logic level can only be maintained if the sourcing current is held to a maximum of 0.4 mA — far less than the current required to turn on any thyristor. Thus, the direct conneciton is useful only if the driver need not activate other logic circuits in addition to a TRIAC.

A similar limiting condition exists in the Logic "0" condition of the output, when the thyristor is to be clamped off. In this condition, Q3 is conducting and V_{out} equals the saturation voltage ($V_{CE(sat)}$) of Q3. TTL specifications indicate that the low logic level (logic "0") may not exceed 0.4 volts, and that the sink current must be limited to 16 mA in order not to exceed this value. A higher value of sink current would cause ($V_{CE(sat)}$) to rise, and could trigger the thyristor on.

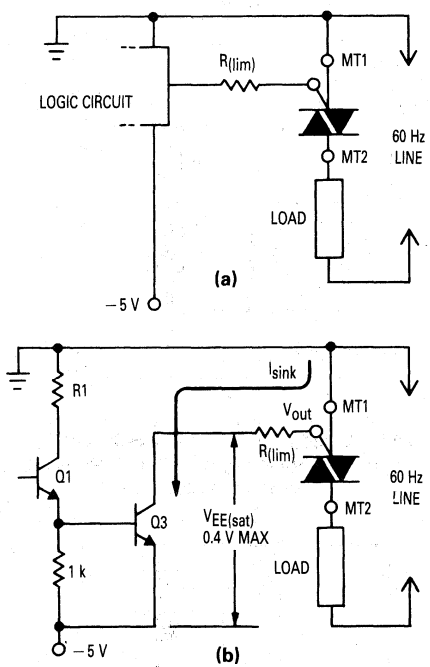


Figure 6.49. TTL Circuit for Quadrant II and III TRIAC Operation Requiring Negative V_{GT}, (b) Schematic Illustrates TRIAC Turn-On Condition, V_{out} = Logical "0"

CIRCUIT DESIGN CONSIDERATIONS

Where a 5400-type TTL circuit is used solely for controlling a TRIAC, with positive-gate turn-on (quadrants I-IV), a sensitive gate TRIAC may be directly coupled to the logic output, as in Figure 6.48. If the correct logic levels must be maintained, however, a couple of resistors must be added to the circuit, as in Figure 6.50(a). In this diagram, R1 is a pull-up which allows the circuit to source more current during a high logical output. Its value must be large enough, however, to limit the sinking current below the 16 mA maximum when V_{out} goes low so that the logical zero level of 0.4 volts is not exceeded.

Resistor R2, a voltage divider in conjunction with R1, insures V_{OH} (the "high" output voltage) to be 2.4 V or greater.

The resistor values may be calculated as follows:

For a supply voltage of 5 V and a maximum sinking current of 16 mA

$$R_1 \geqslant V_{CC}/16\text{ mA} \geqslant 5/0.016 \geqslant 312\ \Omega$$

Thus, 330 Ω, 1/4 W resistor may be used. Assuming R_1 to be 330 Ω and a thyristor gate on voltage (V_{GT}) of 1 V, the equivalent circuit of Figure 6.50(b) exists during the logical "1" output level. Since the logical "1" level must be maintaned at 2.4 volts, the voltage drop across R_2 must be 1.4 V. Therefore,

$$R_2 = 1.4/I_R = 1.4/V_{R_1/R_2} = 1.4/(2.6/330) \cong 175\ \Omega$$

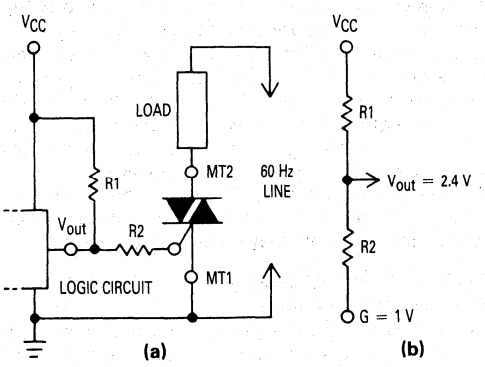


Figure 6.50. Practical Direct-Coupled TTL TRIAC Circuit; (b) Equivalent Circuit Used for Calculation of Resistor Values

A 180 Ω resistor may be used for R_2. If the V_{GT} is less than 1 volt, R_2 may need to be larger.

The MAC97A and 2N6071A TRIACs are compatible devices for this circuit arrangement, since they are guaranteed to be triggered on by 5 mA, whereas the current through the circuit of Figure 6.50(b) is approximately 8 mA, (V_{R_1}/R_1).

When the TRIAC is to be turned on by a negative gate voltage, as in Figure 6.49(b), the purpose of the limiting resistor $R_{(lim)}$ is to hold the current through transistor Q3 to 16 mA. With a 5 V supply, a TRIAC V_{GT} of 1 V and a maximum sink current of 16 mA

$$R_{(lim)} = (V_{CC} - V_{GT})/I_{sink} = (5 - 1)/0.016 \geq 250\ \Omega$$

In practice, a 270 Ω, 1/4 W resistor may be used.

1

OPEN COLLECTOR TTL CIRCUIT

The output section of an open-collector TTL gate is shown in Figure 6.51(a).

A typical logic gate of this kind is the 5401 type Q2-input NAND gate circuit. This logic gate also has a maximum sink current of 16 mA (V_{OL} = 0.4 V max.) because of the Q1 (sat) limitations. If this logic gate is to source any current, a pull-up-collector resistor, R_1 (6.51b) is needed. When this TTL gate is used to trigger a thyristor, R_1 should be chosen to supply the maximum trigger current available from the TTL circuit (≈ 16 mA, in this case). The value of R_1 is calculated in the same way and for the same reasons as in Figure 6.50. If a logical "1" level must be maintained at the TTL output (2.4 V min.), the entire circuit of Figure 6.50 should be used.

For direct drive (logical "0") quadrants II and III triggering, the open collector, negative supplied (−5 V) TTL circuit of Figure 6.52 can be used. Resistor R_1 can have a value of 270 Ω, as in Figure 6.49. Resistor R_2 ensures that the TRIAC gate is referenced to MT1 when the TTL gate goes high (off), thus preventing unwanted turn-on. An R_2 value of about 1 k should be adequate for sensitive gate TRIACs and still draw minimal current.

Circuits utilizing Schottky TTL are generally designed in the same way as TTL circuits, although the current source/sink capabilities may be slightly different.

TRIGGERING THYRISTORS FROM LOGIC GATES USING INTERFACE TRANSISTORS

For applications requiring thyristors that demand more gate current than a direct-coupled logic circuit can supply, an interface device is needed. This device can be a small-signal transistor or an opto coupler.

The transistor circuits can take several different configurations, depending on whether a series or shunt switch design is chosen, and whether gate-current sourcing (quadrants I and IV) or sinking (quadrants II and III) is selected. An example of a series switch, high output (logic 1) activation, is shown in Figure 6.53. Any logic family can be used as long as the output characteristics are known. The NPN interface transistor, Q1, is configured in the common-emitter mode — the simplest approach — with the emitter connected directly to the gate of the thyristor.

Depending on the logic family used, resistor R_1 (pull-up resistor) and R_3 (base-emitter leakage resistor) may or may not be required. If, for example, the logic is a typical TTL totem-pole output gate that must supply 5 mA to the base of the NPN transistor and still maintain a "high" (2.4 V) logic output, then R_1 and R_2 are required. If the "high" logic level is not required, then the TTL circuit can directly source the base current, limited by resistor R_2.

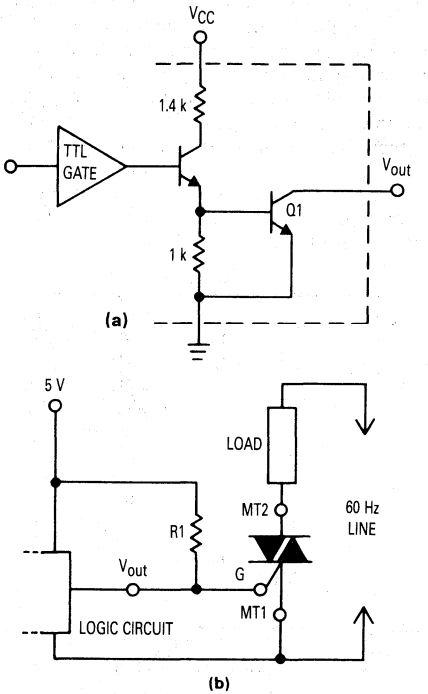


Figure 6.51. Output Section of Open-Collector TTL, (b) For Current Sourcing, A Pull-up Resistor, R1, Must Be Added

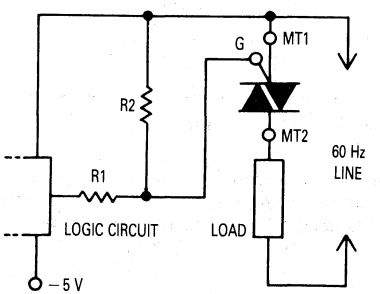


Figure 6.52. Negative-Supplied (−5 V) TTL Gate Permits TRIAC Operation in Quadrants II and III

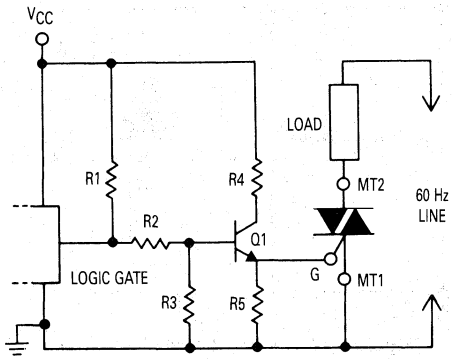


Figure 6.53. Series Switch, High Output (Logic "1")

To illustrate this circuit, consider the case where the selected TRIAC requires a positive-gate current of 100 mA. The interface transistor, a popular 2N4401, has a specified minimum h_{FE} (at a collector current of 150 mA) of 100. To ensure that this transistor is driven hard into saturation, under "worse case" (low temperature) conditions, a forced h_{FE} of 20 is chosen — thus, 5 mA of base current. For this example, the collector supply is chosen to be the same as the logic supply (+ 5 V); but for the circuit configuration, it could be a different supply, if required. The collector-resistor, R_4, is simply

$$R_4 = (V_{CC} - V_{CE(sat)} - V_{GT(typ)})/I_{GT}$$
$$= (5 - 1 - 0.9)/100 \text{ mA} = 40\ \Omega$$

A 39 ohm, 1 W resistor is then chosen, since its actual dissipation is about 0.4 W.

If the "logic 1" output level is not important, then the base limiting resistor R_2 is required, and the pull-up resistor R_1 is not. Since the collector resistor of the TTL upper totem-pole transistor, Q2, is about 100 Ω, this resistor plus R_2 should limit the base current to 5 mA.

Thus R_2 calculates to

$$R_2 = [(V_{CC} - V_{BE} - V_{GT})/5 \text{ mA}] - 100\ \Omega$$
$$= [(5 - 0.7 - 0.9)/0.005]\ 100\ \Omega$$
$$\approx 560\ \Omega \text{ (specified)}$$

When the TTL output is low, the lower transistor of the totem-pole, Q3, is a clamp, through the 560 Ω resistor, across the 2N4401; and, since the 560 Ω resistor is relatively low, no leakage-current shunting resistor, R_3, is required.

In a similar manner, if the TTL output must remain at "logic 1" level, the resistor R_1 can be calculated as described earlier (R_3 may or may not be required).

For low-logic activation (logic "0"), the circuit of Figure 6.54 can be used. In this example, the PNP-interface transistor 2N4403, when turned on, will supply positive-gate current to the thyristor. To ensure that the high logic level will keep the thyristor off, the logic gate and the transistor emitter must be supplied with the same power supply. The base resistors, as in the previous example, are dictated by the output characteristics of the logic family used. Thus if a TTL gate circuit is used, it must be able to sink the base current of the PNP transistor ($I_{OL(MAX)}$ = 16 mA).

When thyristor operation in quadrants II and III is desired, the circuits of Figures 6.55 and 6.56 can be used; Figure 6.55 is for high logic output activation and Figure 6.56 is for low. Both circuits are similar to those on Figures 6.53 and 6.54, but with the transistor polarity and power supplies reversed.

Figure 6.55 sinks current from the thyristor gate through a switched NPN transistor whose emitter is referenced to a negative supply. The logic circuit must also be referenced to this negative supply to ensure that transistor Q1 is turned off when required; thus, for TTL gates, V_{EE} would be −5 V.

In Figure 6.56, the logic-high bus, which is now ground, is the common ground for both the logic, and the thyristor and the load. As in the first example (Figure 6.53), the negative supply for the logic circuit ($-V_{EE}$) and the collector supply for the PNP transistor need not be the same supply. If, for power-supply current limitations, the collector supply is chosen to be another supply ($-V_{CC}$), it must be within the V_{CEO} ratings of the PNP transistor.

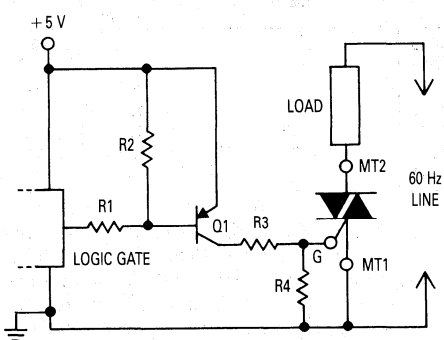


Figure 6.54. Low-Logic Activation with Interface Transistor

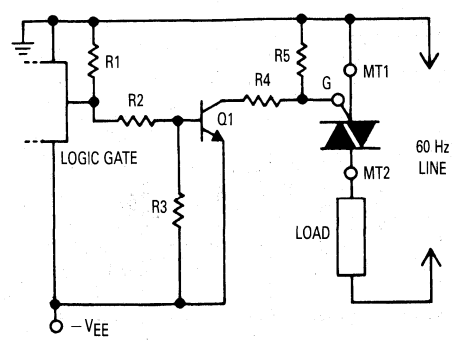


Figure 6.55. High-Logic Output Activation

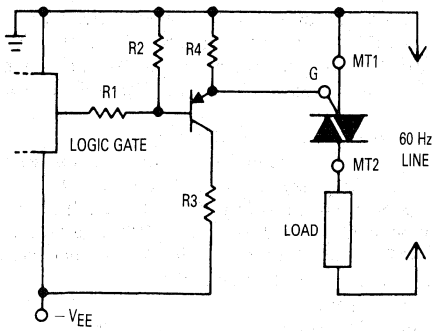


Figure 6.56. Low-Logic Output Activation

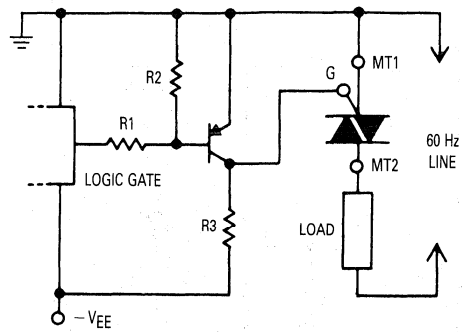


Figure 6.58. Shunt-Interface Circuit (Quadrants I and III Operation)

Also, the power dissipation of collector resistor, R3, is a function of $-V_{CC}$ — the lower $-V_{CC}$, the lower the power rating.

The four examples shown use gate-series switching to activate the thyristor and load (when the interface transistor is off, the load is off). Shunt-switching can also be used if the converse is required, as shown in Figures 6.57 and 6.58. In Figure 6.57, when the logic output is high, NPN transistor, Q1, is turned on, thus clamping the gate of the thyristor off. To activate the load, the logic output goes low, turning off Q1 and allowing positive gate current, as set by resistor R3, to turn on the thyristor.

In a similar manner, quadrant's II and III operation is derived from the shunt interface circuit of Figure 6.58.

OPTICAL ISOLATORS/COUPLERS

An Optoelectronic isolator combines a light-emitting device and a photo detector in the same opaque package that provides ambient light protection. Since there is no electrical connection between input and output, and the emitter and detector cannot reverse their roles, a signal can pass through the coupler in one direction only.

Since the opto-coupler provides input circuitry protection and isolation from output-circuit conditions, ground-loop prevention, dc level shifting, and logic control of high voltage power circuitry are typical areas where opto-couplers are useful.

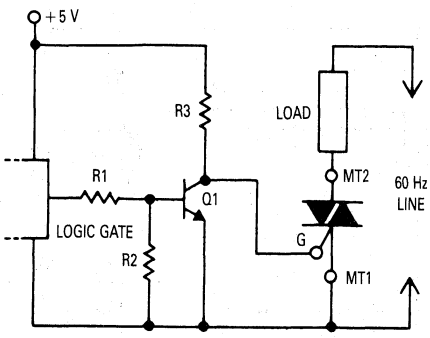


Figure 6.57. Shunt-Interface Circuit (High-Logic Output)

Figure 6.59 shows a photo-TRIAC used as a driver for a higher-power TRIAC. The photo-TRIAC is light sensitive and is turned on by a certain specified light density (H), which is a function of the LED current. With dark conditions (LED current = 0) the photo-TRIAC is not turned on, so that the only output current from the coupler is leakage current, called peak-blocking current (I_{DRM}). The coupler is bilateral and designed to switch ac signals.

The photo-TRIAC output current capability is, typically, 100 mA, continuous, or 1 A peak.

Any Motorola TRIAC can be used in the circuit of Figure 6.59 by using Table 6.VII. The value of R is based on the photo-TRIAC's current-handling capability. For example, when the MOC3011 operates with a 120 V line voltage (approximately 175 V peak), a peak I_{GT} current of 175 V/180 ohm (approximately 1 A) flows when the line voltage is at its maximum. If less than 1 A of I_{GT} is needed, R can be increased. Circuit operation is as follows:

Table 6.VII. Specifications for Typical Optically Coupled TRIAC Drivers

Device Type	Maximum Required LED Trigger Current (mA)	Peak Blocking Voltage	R(Ohms)
MOC3009	30	250	180
MOC3011	15	250	180
MOC3011	10	250	180
MOC3020	30	400	260
MOC3021	15	400	360
MOC3030	30	250	51
MOC3031	15	250	51

When an op-amp, logic gate, transistor or any other appropriate device turns on the LED, the emitted light triggers the photo-TRIAC. Since, at this time, the main TRIAC is not on, MT2-to-gate is an open circuit. The 60 Hz line can now cause a current flow via R, the photo-TRIAC, Gate-MT1 junction and load. This Gate-MT1 current triggers the main TRIAC, which then shorts and turns

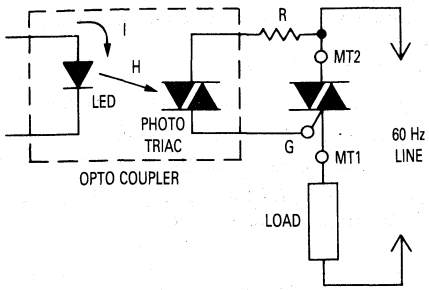


Figure 6.59. Optically-Coupled TRIAC Driver is Used to Drive a Higher-Power TRIAC

off the photo-TRIAC. The process repeats itself every half cycle until the LED is turned off.

Triggering the main TRIAC is thus accomplished by turning on the LED with the required LED-trigger current indicated in Table 6.VII.

MICROPROCESSORS

Microprocessor systems are also capable of controlling ac power loads when interfaced with thyristors. Commonly, the output of the MPU drives a PIA (peripheral interface adaptor) which then drives the next stage. The PIA Output Port generally has a TTL compatible output with significantly less current source and sink capability than standard TTL. (MPUs and PIAs are sometimes constructed together on the same chip and called microcontrolllers.)

When switching ac loads from microcomputers, it is good practice to optically isolate them from unexpected load or ac line phenomena to protect the computer system from possible damage. In addition, optical isolation will make UL recognition possible.

A typical TTL-compatible microcontroller, such as the MC3870P offers the following specifications:

I_{OH} = 300 μA (V_{OH} = 2.4 V)
I_{OL} = 1.8 mA (V_{OL} = 0.4 V)
V_{CC} = 5 V

Since this is not adequate for driving the optocoupler directly (10 mA for the MOC3011), an interface transistor is necessary.

The circuit of Figure 6.60 may be used for thyristor triggering from the 3870 logical "1."

The interface transistor, again, can be the 2N4401. With 10 mA of collector current (for the MOC3011) and a base current of 0.75 mA, the $V_{CE(sat)}$ will be approximately 0.1 V.

R_1 can be calculated as in a previous example. Specifically:

1.8 mA (maximum I_{OL} for the 3870)
$> 5\ V/R_1$; $R_1 > 2.77$ k

R_1 can be 3 k, 1/4 W

With a base current of 0.75 mA, R_1 will drop (0.75 mA) (3 k) or 2.25 V. This causes a V_{OH} of 2.75 V, which is within the logical "1" range.

$R_2 = [2.75\ V - V_{BE(on)}]/I_B = (2.75 - 0.75)/0.75 = 2.66$ k

R_2 can be a 2.7 k, 1/4 W resistor.

R_3 must limit I_C to 10 mA:

$$R_3 = [5\ V - V_{CE(sat)} - V_F(\text{diode})/10\ \text{mA}]$$
$$= (5 - 0.1 - 1.2)/10\ \text{mA} = 370\ \Omega$$

Since R_3 is relatively small, no base-emitter leakage resistor is required.

Figure 6.61 shows logical "0" activation. Resistor values are calculated in a similar way.

THE CMOS INTERFACE

Another popular logic family, CMOS, can also be used to drive thyristors.

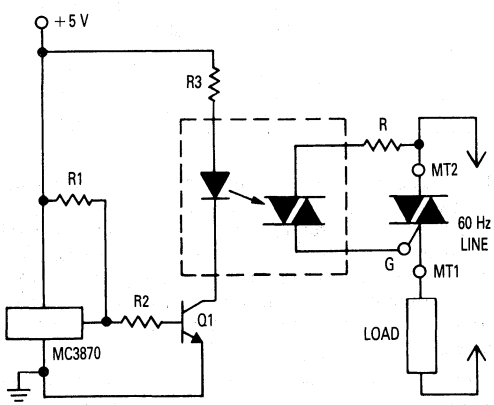


Figure 6.60. Logical "1" Activation from MC3870P Microcomputer

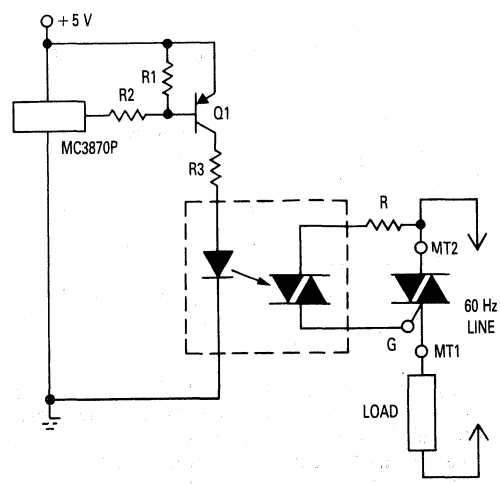


Figure 6.61. Logical "0" Activation

As shown in Figure 6.62(a), the output stage of a typical CMOS Gate consists of a P-channel MOS device connected in series with an N-channel device (drain-to-drain), with the gates tied together and driven from a common input signal. When the input signal goes high, logical 1, the P-channel device is essentially off and conducts only leakage current (I_{DSS}), on the order of picoamps. The N-channel unit is forward-biased and, although it has a relatively high on resistance ($r_{DS(on)}$), the drain-to-source voltage of the N-channel device (V_{DS}) is very low (essentially zero) because of the very low drain current (V_{DSS}) flowing through the device. Conversely, when the input goes low (zero), the P-channel device is turned fully on, the N-channel device is off and the output voltage will be very near V_{DD}.

When interfacing with transistors or thyristors, the CMOS Gate is current-limited mainly by its relatively high on resistance, the dc resistance between drain and source, when the device is turned on.

The equivalent circuits for sourcing and sinking current into an external load is shown in Figures 6.62(b) and 6.62(c). Normally, when interfacing CMOS to CMOS, the logic outputs will be very near their absolute maximum states (V_{DD} or 0 V) because of the extremely small load currents. With other types of loads (e.g. TRIACs), the current, and the resulting output voltage, is dictated by the simple voltage divider of $r_{DS(on)}$ and the load resistor R_L, where $r_{DS(on)}$ is the total series and/or parallel resistance of the devices comprising the NOR and NAND function.

Interfacing CMOS gates with thyristors requires a knowledge of the on resistance of the gate in the source and sink conditions. The on-resistance of CMOS devices is not normally specified on data sheets.

It can easily be calculated, however, from the output drive currents, which are specified. The drive (source/sink) currents of typical CMOS gates at various supply voltages are shown in Table 6.VIII. From this information,

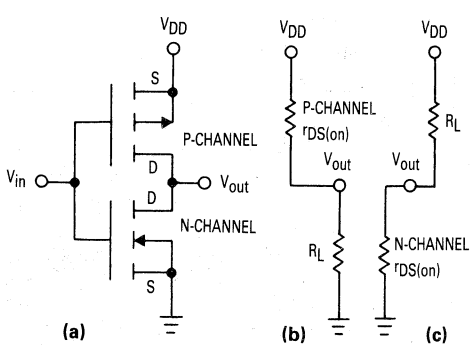


Figure 6.62. Output Section of a Typical CMOS Gate, (b) Equivalent Current-Sourcing Circuit is Activated when V_{in} goes Low, Turning the P-Channel Device Fully On, (c) Equivalent Current Sinking Circuit is Activated when the Input Goes High and Turns the N-Channel Device On

the on resistance for worst case design is calculated as follows:

For the source condition

$$r_{DS(on)(MAX)} = (V_{DD} - V_{OH})/I_{OH(MIN)}$$

Similarly, for the sink current condition

$$r_{DS(on)(MAX)} = V_{OL}/I_{OL(MIN)}$$

Values of $r_{DS(on)}$ for the various condition shown in Table 6.VIII are tabulated in Table 6.IX.

Table 6.VIII. CMOS Characteristics

Specified source/sink currents to maintain logical "1" and logical "0" levels for various power-supply (V_{DD}) voltages. The I_{OH} and I_{OL} values are used to calculate the "on" resistance of the CMOS output.

Output Drive Current	CMOS AL Series mA, dc		CMOSCL/CP Series mA, dc	
	Min	Typ	Min	Typ
$I_{(source)} - I_{OH}$				
V_{DD} = 5 V; V_{OH} = 2.5 V	−0.5	−1.7	−0.2	−1.7
V_{DD} = 10 V; V_{OH} = 9.5 V	−0.5	−0.9	−0.2	−0.9
V_{DD} = 15 V; V_{OH} = 13.5 V		−3.5		−3.5
$I_{(sink)} - I_{OL}$				
V_{DD} = 5 V; V_{OL} = 0.4 V	0.4	7.8	0.2	7.8
V_{DD} = 10 V; V_{OL} = 0.5 V	0.9	2	0.5	2
V_{DD} = 15 V; V_{OL} = 1.5 V		7.8		7.8

Table 6.IX
Calculated CMOS On Resistance Values For Current Sourcing and Sinking at Various V_{DD} Options

Operating Conditions	Output Resistance, $r_{DS(on)}$ Ohms	
	Typical	Maximum
Source Condition		
V_{DD} = 5 V	1.7 k	12.5 k
10 V	500	2.5 k
15 V	430	—
Sink Condition		
V_{DD} = 5 V	500	2 k
10 V	420	1 k
15 V	190	—

It is apparent from this table that the on resistance decreases with increasing supply voltage.

Although the minimum currents are now shown on the data sheet for the 15 V case, the maximum on resistance can be no greater than the 10 V example and, therefore, can be assumed for worst case approximation to be 1 and 2.5 kohms for sink-and-source current cases, respectively.

The sourcing on resistance is greater than the sinking case because the difference in carrier mobilities of the two channel types.

Since $r_{DS(on)}$ for both source and sink conditions var-

ies with supply voltage (V_{DD}), there are certain drive limitations. The relative high $r_{DS(on)}$ of the P-channel transistor could possibly limit the direct thyristor drive capability; and, in a like manner, the N-channel $r_{DS(on)}$ might limit its clamping capability. With a 10 or 15 V supply, the device may be capable of supplying more than 10 mA, but should be limited to that current, with an external limiting resistor, to avoid exceeding the reliable limits of the unit metalization.

DC MOTOR CONTROL WITH THYRISTORS

In order to control the speed of a dc series field motor at different required torque levels, it is necessary to adjust the voltage applied to the motor. For any particular applied voltage the motor speed is determined solely by the torque requirements and top speed is reached under minimum torque conditions. When a series motor is used as a traction drive for vehicles, it is desirable to control the voltage to the motor to fit the various torque requirements of grades, speed and load. The common method of varying the speed of the motor is by inserting resistance in series with the motor to reduce the supplied voltage. This type of motor speed control is very inefficient due to the I^2R loss, especially under high current and torque conditions.

A much more efficient method of controlling the voltage applied to the motor is the pulse width modulation method shown in Figure 6.63. In this method, a variable width pulse of voltage is applied to the motor at the same rate to proportionally vary the average voltage applied to the motor. A diode is placed in parallel with the inductive motor path to provide a circuit for the inductive motor current and prevent abrupt motor current change. Abrupt current changes would cause high induced voltage across the switching device.

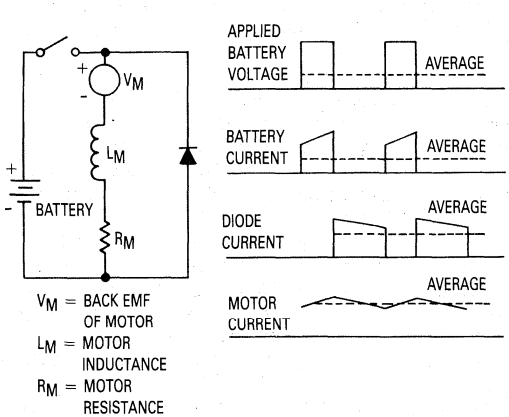


Figure 6.63. Basic Pulse Width Modulated Motor Speed Control

The circulating current through the diode decreases only in response to motor and diode loss. With reference to Figure 6.63, it can be seen that the circulating diode current causes more average current to flow through the motor than is taken from the battery. However, the power taken from the battery is approximately equal to the power delivered to the motor, indicating that energy is stored in the motor inductance at the battery voltage level and is delivered to the motor at the approximate current level when the battery is disconnected.

To provide smooth and quiet motor operation, the current variations through the motor should be kept to a minimum during the switching cycle. There are limitations on the amount of energy that can be stored in the motor inductance, which, in turn, limits the power delivered to the motor during the off time; thus the off time must be short. To operate the motor at low speeds, the on time must be approximately 10 percent of the off time and therefore, a rapid switching rate is required that is generally beyond the capabilities of mechanical switches. Practical solutions can be found by the use of semiconductor devices for fast, reliable and efficient switching operations.

SCR DC MOTOR CONTROL

SCRs offer several advantages over power transistors as semiconductor switches. They require less driver power, are less susceptible to damage by overload currents and can handle more voltage and current. Their disadvantages are that they have a higher power dissipation due to higher voltage drops and the difficulty in commutating to the off condition.

The SCR must be turned off by either interrupting the current through the anode-cathode circuit or by forcing current through the SCR in the reverse direction so that the net flow of forward current is below the holding current long enough for the SCR to recover blocking ability. Commutation of the SCR in high current motor control circuits is generally accomplished by discharging a capacitor through the SCR in the reverse direction. The value of this capacitor is determined approximately from the following equation:

$$C_c = \frac{T_q I_A}{V_c}$$

Where:

C_c = value of necessary commutating capacitance
T_q = turn-off time of the SCR
I_A = value of anode current before commutation
V_c = voltage of C_c before commutation

This relationship shows that to reduce the size of C_c, the capacitor should be charged to as high a voltage as possible and the SCR should be selected with as low a turn-off time as possible.

If a 20 microsecond turn-off time SCR is commutated by a capacitor charged to 36 volts, it would take over 110 μF to turn off 200 amperes in the RC commutating circuit of Figure 6.64. If a 50 cycle switching frequency is desired,

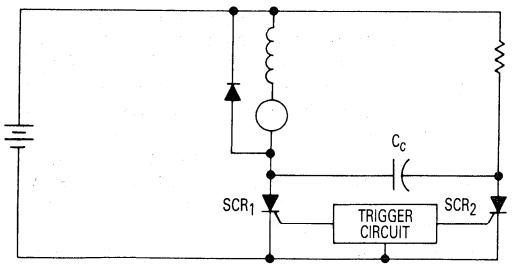


Figure 6.64. Speed Control with Resistive Charging

the value of R_1 would be approximately 5 ohms to allow charging time with an on duty cycle of 10 percent. The value of this resistor would give approximately 260 watts dissipation in the charging circuit with 90 percent off duty cycle.

If the resonant charging commutating circuitry of Figure 6.65 is used, the capacitor is reduced to approximately 55 μF. In this circuit, SCR_3 is gated on at the same time as SCR_1 and allows the resonant charging of C_C through L_C to twice the supply voltage. SCR_3 is then turned off by the reversal of voltage in the resonant circuit before SCR_2 is gated on. It is apparent that there is very little power loss in the charge circuit depending upon the voltage drop across SCR_3 and the resistance in L_C.

If the commutating capacitor is to be reduced further, it is necessary to use a transformer to charge the capacitor to more than twice the supply voltage. This type of circuit is illustrated by the transformer charge circuit shown in Figure 6.66. In this circuit the capacitor can be charged to several times the supply voltage by transformer action through diode D_1 before commutating SCR_1. The disadvantage of this circuit is in the high motor current that flows through the transformer primary winding.

HEAVY DUTY MOTOR CONTROL WITH SCRs

Another advantage of SCRs is their high surge current capabilities, demonstrated in the motor drive portion of the golf cart controller shown in Figure 6.67. Germanium power transistors were used because of the low saturation voltages and resulting low static power loss. However, since switching speeds are slow and leakage currents are high, additional circuit techniques are required to ensure reliable operation:

1) The faster turn-on time of the SCR (Q9) over that of the germanium transistors shapes the turn-on load line.
2) The parallelled output transistors (Q3–Q8) require a 6 V reverse bias.
3) The driver transistor Q2 obtains reverse bias by means of diode D4.

To obtain the 6 V bias, the 36 V string of 6 V batteries are tapped, as shown in the schematic. Thus, the motor is powered from 30 V and the collector supply for Q2 is 24 V, minimizing the dissipation in colllector load resistor R1.

Total switching loss in switchmode applications is the result of the static (on-state) loss, dynamic (switching) loss and leakage current (off-state) loss. The low saturation voltage of germanium transistors produces low static loss. However, switching speeds of the germanium transistors are low and leakage currents are high. Loss due to leakage current can be reduced with off bias, and load line shaping can minimize switching loss. The turn-off switching loss was reduced with a standard snubber network (D5, C1, R2) see Figure 6.67.

Turn-on loss was uniquely and substantially reduced by using a parallel connected SCR (across the germanium transistors) the MCR265-4 (55 A rms, 550 A surge). This faster switching device diverts the initial turn-on motor load current from the germanium output transistors, reducing both system turn-on loss and transistor SOA stress.

The main point of interest is the power switching portion of the PWM motor controller. Most of the readily available PWM ICs can be used (MC3420, MC34060, TL494, SG1525A, UA78S40, etc.), as they can source at

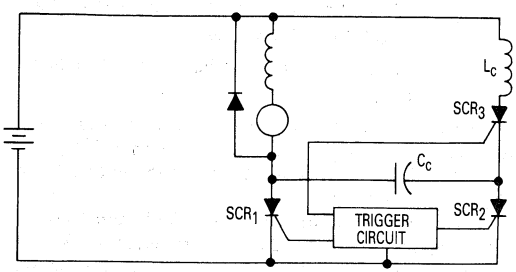


Figure 6.65. Speed Control with Inductive Charging

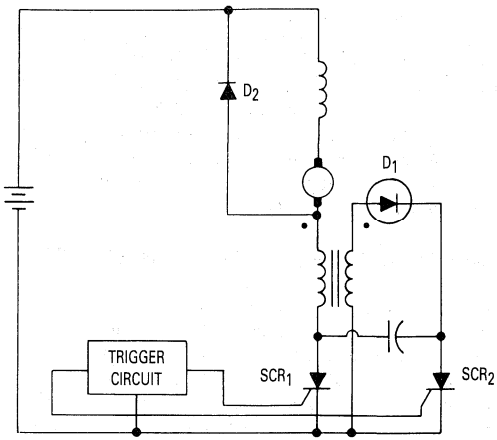


Figure 6.66. SCR Motor Control with Transformer Charging

least a 10 mA, +15 V pulse for driving the following power MOSFET.

Due to the extremely high input impedance of the power MOSFET, the PWM output can be directly connected to the FET gate, requiring no active interface circuitry. The positive going output of the PWM is power gained and inverted by the TMOS FET Q1 to supply the negative going base drive to PNP transistor Q2. Diode D1 provides off-bias to this paraphase amplifier, the negative going pulse from the emitter furnishing base drive to the six parallel connected output transistors and the positive going collector output pulse supplying the SCR gate trigger coupled through transformer T1.

Since the faster turn-on SCR is triggered on first, it will carry the high, initial turn-on motor current. Then the slower turn-on germanium transistors will conduct clamping off the SCR, and carry the full motor current. For the illustrated 2HP motor and semiconductors, a peak exponentially rising and falling SCR current pulse of 120 A lasting for about 60 μs was measured. This current is well within the rating of the SCR. Thus, the high turn-on stresses are removed from the transistors providing a much more reliable and efficient motor controller while using only a few additional components.

DIRECTION AND SPEED CONTROL FOR MOTORS

For a shunt motor, a constant voltage should be applied to the shunt field to maintain constant field flux so that the armature reaction has negligible effect. When constant voltage is applied to the shunt field, the speed is a direct function of the armature voltage and the armature current. If the field is weak, then the armature reaction may counterbalance the voltage drop due to the brushes, windings and armature resistances, with the net result of a rising speed-load characteristic.

The speed of a shunt-wound motor can be controlled with a variable resistance in series with the field or the armature. Varying the field current for small motor provides a wide range of speeds with good speed regulation. However, if the field becomes extremely weak, a rising speed-load characteristic results. This method cannot provide control below the design motor speed. Varying the resistance in series with the armature results in speeds less than the designed motor speed; however, this method yields poor speed regulation, especially at low speed settings. This method of control also increases power dissipation and reduces efficiency and the torque since the maximum armature current is reduced. Neither type of resistive speed control is very satisfactory. Thyristor drive controls, on the other hand, provide continuous control through the range of speed desired, do not have the power losses inherent in resistive circuits, and do not compromise the torque characteristics of motors.

Although a series-wound motor can be used with either dc or ac excitation, dc operation provides superior performance. A universal motor is a small series-wound mo-

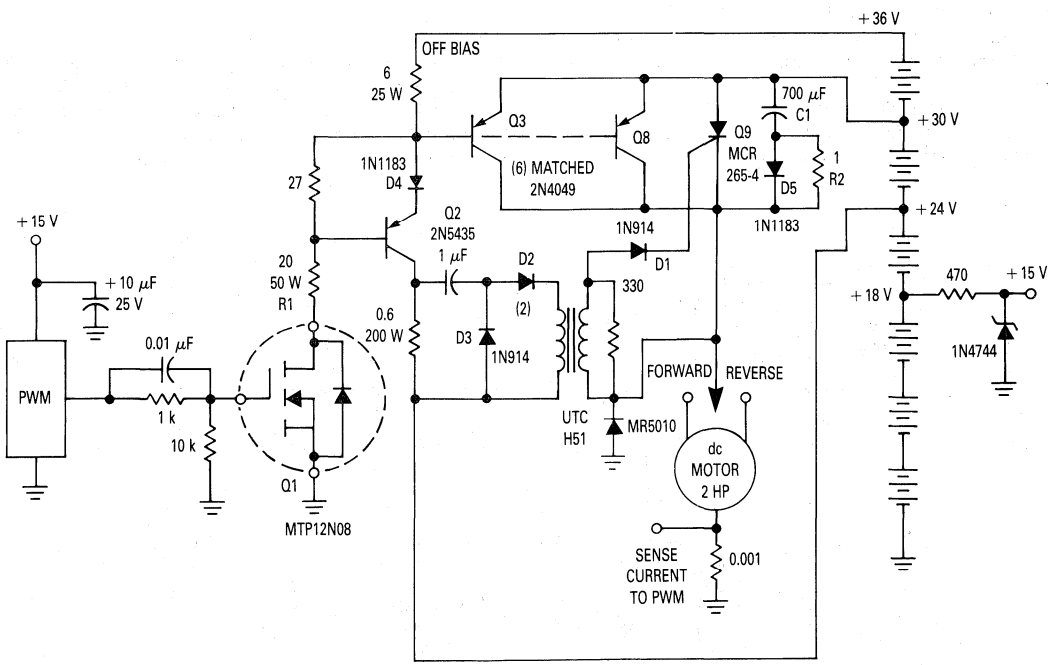


Figure 6.67. PWM DC Motor Controller Using SCR Turn-On Feature

tor designed to operate from either a dc or an ac supply of the same voltage. In the small motors used as universal motors, the winding inductance is not large enough to produce sufficient current through transformer action to create excessive commutation problems. Also, high-resistance brushes are used to aid commutation. The characteristics of a universal motor operated from alternating current closely approximate those obtained for a dc power source up to full load; however, above full load the ac and dc characteristics differ. For a series motor that was not designed as a universal motor, the speed-torque characteristic with ac rather than dc is not as good as that for the universal motor. At eight loads, the speed for ac operation may be greater than for dc since the effective ac field strength is smaller than that obtained on direct current. At any rate, a series motor should not be operated in a no-load condition unless precaution is are taken to limit the maximum speed.

1

SERIES-WOUND MOTORS

The circuit shown in Figure 6.68 can be used to control the speed and direction of rotation of a series-wound dc motor. Silicon controlled rectifiers Q1–Q4, which are connected in a bridge arrangement, are triggered in diagonal pairs. Which pair is turned on is controlled by switch S1 since it connects either coupling transformer T1 or coupling transformer T2 to a pulsing circuit. The current in the field can be reversed by selecting either SCRs Q2 and Q3 for conduction, or SCRs Q1 and Q4 for conduction. Since the armature current is always in the same direction, the field current reverses in relation to the armature current, thus reversing the direction of rotation of the motor.

A pulse circuit is used to drive the SCRs through either transformer T1 or T2. The pulse required to fire the SCR is obtained from the energy stored in capacitor C1. This capacitor charges to the breakdown voltage of zener diode D5 through potentiometer R1 and resistor R2. As the capacitor voltage exceeds the zener voltage, the zener conducts, delivering current to the gate of SCR Q5. This turns Q5 on, which discharges C1 through either T1 or T2 depending on the position of S1. This creates the desired triggering pulse. Once Q5 is on, it remains on for the duration of the half cycle. This clamps the voltage across C1 to the forward voltage drop of Q5. When the supply voltage drops to zero, Q5 turns off, permitting C1 to begin charging when the supply voltage begins to increase.

The speed of the motor can be controlled by potentiometer R1. The larger the resistance in the circuit, the longer required to charge C1 to the breakdown voltage of zener D5. This determines the conduction angle of either Q1 and Q4, or Q2 and Q3, thus setting the average motor voltage and thereby the speed.

SHUNT-WOUND MOTORS

If a shunt-wound motor is to be used, then the circuit in Figure 6.69 is required. This circuit operates like the one shown in Figure 6.68. The only differences are that the field is placed across the rectified supply and the armature is placed in the SCR bridge. Thus the field current is unidirectional but armature current is reversible; consequently the motor's direction of rotation is reversible. Potentiometer R1 controls the speed as explained previously.

RESULTS

Excellent results were obtained when these circuits were used to control 1/15 hp, 115 V, 5,000 r/min motors. This circuit will control larger, fractional-horsepower motors provided the motor current requirements are within the semiconductor ratings. Higher current devices will

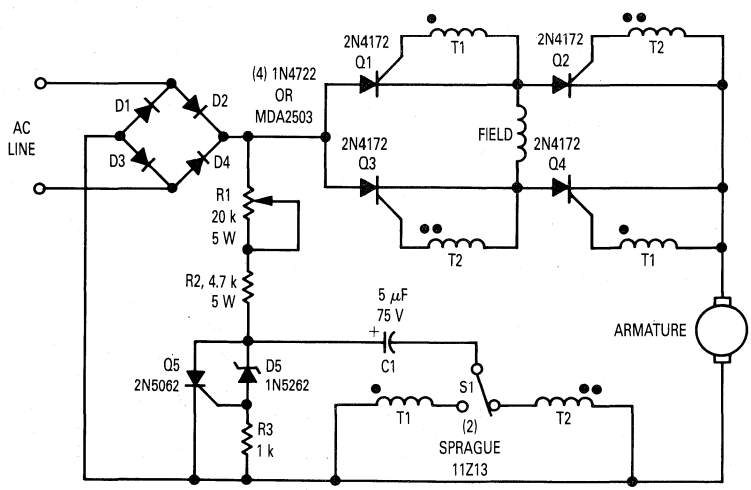


Figure 6.68. Direction and Speed Control for Series-Wound or Universal Motor

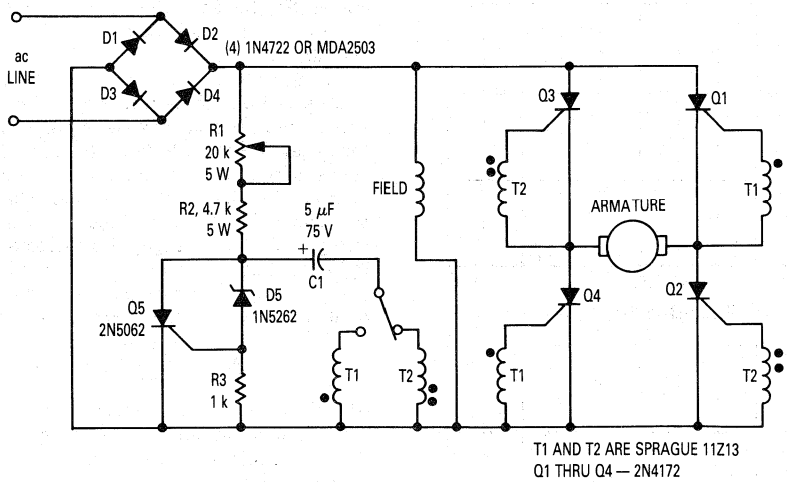


Figure 6.69. Direction and Speed Control for Shunt-Wound Motor

permit control of even larger motors, but the operation of the motor under worst case must not cause anode currents to exceed the ratings of the semiconductor.

UNIJUNCTION TRANSISTOR APPLICATIONS USING RELAXATION OSCILLATORS

Most UJT oscillator circuits employ the basic relaxation oscillator circuit in some way or another. As mentioned previously, either the output at base-one, base-two, or the emitter can be utilized in order to fulfill a specified requirement.

THE BASIC UJT RELAXATION OSCILLATOR CIRCUIT

The UJT relaxation oscillator, the basic building block in most UJT timer and oscillator circuits (Figure 6.70), operates as follows:

When power is applied, the capacitor C_E charges exponentially through the resistor R_E until the voltage on the capacitor equals the emitter firing voltage V_p. At this voltage, the emitter base-one junction becomes forward biased and the emitter characteristic goes into the negative resistance region. The capacitor C_E discharges through the emitter and a positive going pulse will be available at base-one. This pulse is shown in Figure 6.71(a) for a circuit having

R_E = 10 kΩ, C_E = 0.01 μF, R2 = 200 Ω, R1 = 47 Ω, and V1 = 20 V.

Prior to firing, a current I_{B2} is flowing from base-two to base-one. When emitter current starts to flow, this current will increase to $I_{B2(mod)}$ since the resistance from base-two to ground is decreasing. A negative going voltage pulse will therefore appear at base-two, and this waveform is shown in Figure 6.71(b).

When the voltage at the emitter has decreased to V_0, a voltage approximately equal to the valley voltage when R1 is purely resistive, the UJT will turn off if R_E meets certain conditions. C_E will start to charge up again, and the cycle repeats. The waveform that appears at the emitter is shown in Figure 6.71(c).

In order for the above sequence of events to take place, R_E has to meet certain conditions. What these conditions are can best be explained by means of the emitter characteristic curve in Figure 6.72. (This curve is not drawn to scale in order to show more detail.)

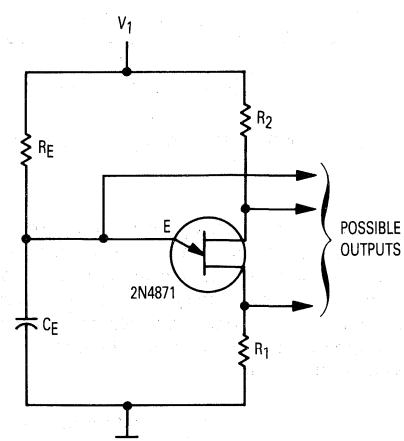
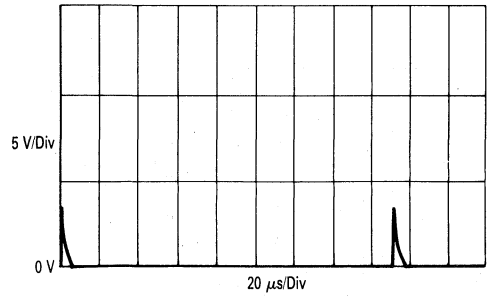
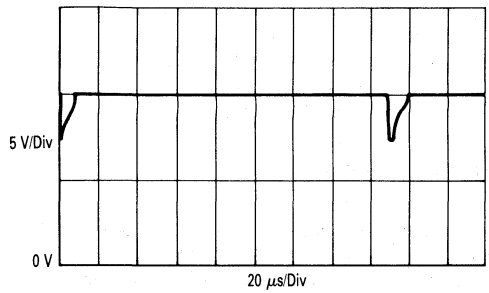


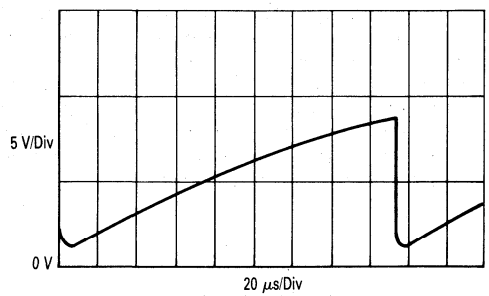
Figure 6.70. Basic UJT Relaxation Oscillator



(a). Base-One Waveform



(b). Base-Two Waveform



(c). Emitter Waveform

Figure 6.71. UJT Oscillator Waveforms

The emitter capacitor C_E, will charge until the emitter voltage is equal to V_p. At this point on the characteristic curve, peak point emitter current I_p will be flowing, and in order to fire the unijunction, the value of emitter resistor R_E must be small enough to allow a current somewhat larger than I_p to flow. R_E must, therefore, meet the following requirement:

$$R_E < \frac{V_1 - V_p}{I_p} = R_{E(max)} \qquad (1)$$

where V1 is the applied bias voltage.

Referring to Figure 6.72, this means that a load line intersecting the characteristic curve in the cutoff region, as illustrated by load line one, would keep the UJT from ever firing.

Having selected $R_E < R_{E(max)}$, the UJT will turn on, and C_E will discharge through the emitter. If R_E is too small, however, and allows an emitter current larger than the valley current I_V to flow, the UJT will not turn off. A stable state in the saturation region will result, and the load line will intersect the emitter characteristic curve somewhere to the right of the valley point. Load line 2 in Figure 6.76 intersecting the characteristic curve at P2 illustrates this condition.

The minimum R_E that can be used in order to assure oscillation can be defined by this formula:

$$R_E > \frac{V_1 - V_V}{I_V} = R_{E(min)}^{**} \qquad (2)$$

where V_V is the valley voltage.

An emitter resistance selected to meet the requirements in equations (1) and (2) will result in a load line which intersects the characteristic curve somewhere in the negative resistance region. An example is given by load line 3, intersecting the curve at P3. The theory explaining the turn-off will be given in the appendix.

The time required for a complete period can be calculated. The voltage on C_E at any time is given by the equation:

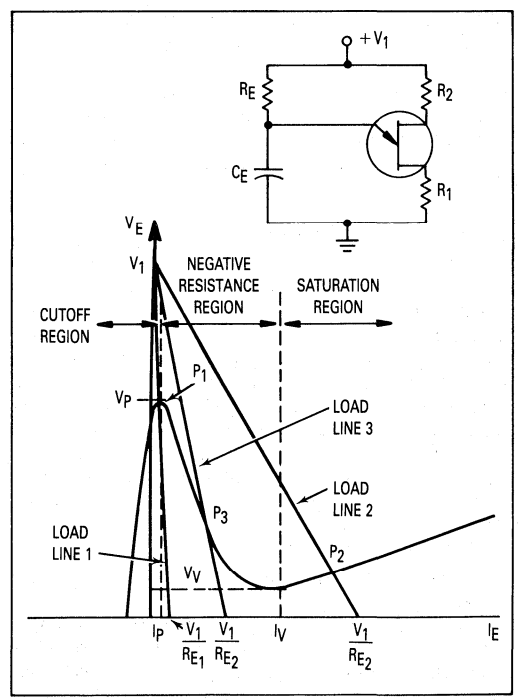


Figure 6.72. UJT Emitter Characteristic Load Lines

$$V_{CE} = V_V + (V_1 - V_V)(1 - e^{-t/R_E C_E}) \quad (3)$$

Substituting $V_{CE} = V_p = V_D + \eta V_{B2B1}$ *

$$V_D + \eta V_{B2B1} = V_V + (V_1 - V_V)(1 - e^{-t/R_E C_E}) \quad (4)$$

Solving this equation for t will give the time to charge C_E from V_V to V_p.

$$t = R_E C_E \ln \frac{V_1 - V_V}{V_1 - V_D - \eta V_{B2B1}} \quad (5)$$

A complete period T also includes the switching time of the UJT and the formula for T becomes:

$$T = R_E C_E \ln \frac{V_1 - V_V}{V_1 - V_D - \eta V_{B2B1}} + t_{on} + t_{off} \quad (6)$$

The following simplifications can be made in this formula:

The turn-on time is generally much smaller than t_{off} and can be omitted. V1 is also usually an order of magnitude larger than V_D or V_V, and when R1 and R2 are small, $V_{B2B1} \approx V1$. Equation (6) can therefore be written:

$$T \approx R_E C_E \ln \frac{1}{1 - \eta} + t_{off} \quad (7)$$

To summarize: The condition for stable operation of the relaxation oscillator is that R_E must be chosen such that the load line intersects the emitter characteristic in the negative resistance region. The approximate period can be found from Equation (7).

A SIMPLE TIME DELAY

The basic building block can be used without modification in simple time delay circuits. One such circuit is shown in Figure 6.70. The circuit operation is as follows:

The circuit values are determined by means of the equations developed previously. The value of the emitter resistor R_E is chosen by means of Equations (1) and (2) to meet the requirements.

$$R_{E(min)} < R_E < R_{E(max)}$$

When $R_E = 10\ M\Omega$, $C_E = 10\ \mu F$, and $\eta = 0.8$, the time required for one complete period can be found from Equation (7):

$$T \approx R_E C_E \ln \frac{1}{1 - \eta} = 10 \cdot 10^6 x 10 \cdot 10^{-6} x \ln \frac{1}{1 - 0.8}$$

$T \approx 160$ seconds

R2 is selected as 1 kΩ to provide maximum temperature compensation with the UJT used. The theory of temperature compensation is given in the appendix.

After the first cycle, the relay will normally be energized. When push-button S1, being normally closed is activated, the SCR turns off, the relay is de-energized, and power is applied to the relaxation oscillator and the load. After a time delay varying from less than a second to approximately 2.5 minutes, as determined by the setting of the 10 MΩ pot, the unijunction will fire and turn on the SCR. The relay will energize and power is removed from the oscillator and the load. The relay K1 will stay energized until the push button S1 is pushed again.

As shown in the example, the UJT trigger output from Base 1 directly drives the gate of the SCR. However, where isolation between the UJT trigger, or any other type of trigger, and the thyristor power circuit is required, then a simple pulse transformer interfacing the two elements will suffice.

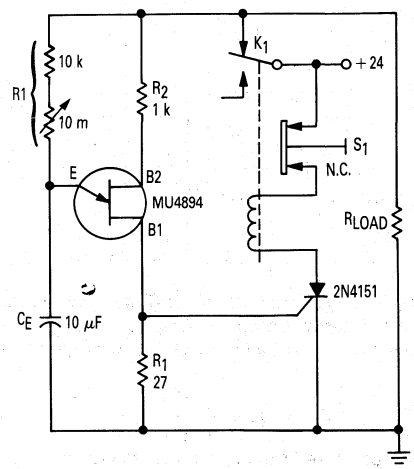


Figure 6.73. A Simple Time Delay Circuit

TEMPERATURE STABILIZATION OF THE PEAK POINT VOLTAGE

Practically all UJT characteristics are temperature dependent. The interbase resistance r_{BB} and emitter reverse current I_{EO} increase, while the peak, valley voltage, current, intrinsic standoff ratio η, and the junction diode drop decrease with increasing temperature.

The peak point voltage is given by the equation:

$$V_p = V_D + \eta V_{B2B1} \quad (10)$$

Since both V_D and η decrease with temperature, V_p will also decrease. This is, of course, a very undesirable condition in many applications, and particularly in timers and oscillator circuits. It has been found, however, that the change in V_p can be compensated for by adding a resistor R2 in series with the base two terminal.

If this resistor is selected properly, V_p can be made to vary less than 1% over a 50°C temperature variation. An equivalent circuit for the UJT in the cutoff region, including the external resistor R2, is shown in Figure 6.74. The peak point voltage is now given by:

*In practice, the variation of the emitter voltage in the neighborhood of the valley point is so small that in order to assure turn-off, R_E should be selected two to three times larger than $R_{E(min)}$.

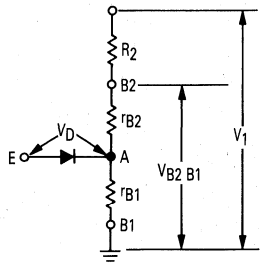


Figure 6.74. Electrical Equivalent Circuit for the UJT with the External Compensating Resistor R2. The Equivalent Circuit is Valid in the Cutoff Region Only.

$$V_p = V_D + \frac{\eta V_1 r_{BB}}{(r_{BB} + R2)} \qquad (11)$$

When temperature is increased, V_D decreases, and for V_p to remain unchanged, the second term in Equation (6), representing the voltage at point A, must increase. Since r_{BB} will increase and R2 will remain unchanged, the second term in the equation will indeed increase.

It would seem like a relatively simple task to calculate R2 from Equation (11) by taking the derivative of V_p with respect to temperature, setting $\frac{dv}{dt}$ p equal to zero, and solve for R2. This procedure would result in the following equation for R2.

$$R2 = 1/2 \left[-(2\, r_{BB} + V_1\, \eta \frac{K_2}{K_1} + V_1\, r_{BB} \frac{K_3}{K_1}) \pm (2\, r_{BB} + V_1\, \eta \frac{K_2}{K_1} + V_1\, r_{BB} \frac{K_3}{K_1} - 4\, r_{BB}\, (1 + V1 \frac{K_3}{K_1}) \right] \qquad (12)$$

$$\text{where } K_1 = \frac{dv_D}{dt} = -2.7 \text{ mV/°C}$$

$$K_2 = \frac{d\, r_{BB}}{dt} = \frac{r_{BBN} \cdot 0.77}{100} \text{ /°C}$$

$$K_3 = \frac{d\eta}{dt} = \frac{\eta N \cdot 0.06}{100} \text{ /°C}$$

and the subscript N denotes the value at 25°C.

When solving this equation, it would also have to be taken into account that r_{BB} is voltage dependent and when V_{B2B1} is increased by one volt, r_{BB} increases 1.2% (based on the value of r_{BB} at V_{B2B1} = 3 V). Furthermore, the temperature dependency of r_{BB} affects not only terms containing K2 but also terms containing r_{BB} itself.

In equation (12), r_{BB}, K1, K2, and η are voltage dependent and V_{B2B1} is dependent on R2. This interdependency results in a very complex equation that is difficult to solve, which in turn greatly reduces the usefulness of Equation (12).

In most applications, the variation predominates and can be compensated for by choosing R2 as follows:

$$R_2 = 15\%\ r_{BB} \qquad (13)$$

If a high degree of accuracy is required, however, a final adjustment of R2 should be made in the actual operating circuit.

Frequency variation as a function of temperature for a typical annular UJT is shown in Figure 6.75. Temperature curves for several values of R2 ranging from 250 ohms to 3 kohms are shown, and an R2 of approximately 1.5 kohms can be seen to compensate very well from −5°C to +85°C. A smaller resistor should be used for operation below −5°C.

DYNAMIC OPERATING PATHS

In order to determine the dynamic operating path of the relaxation oscillator, the circuit in Figure 6.76 can be used. Figure 6.80(b) shows the emitter characteristic curve for V_{B2B1} = 20 volts with the dynamic operating path of the oscillator shown with dotted lines. (V_1 is adjusted for V_{B2B1} = 20 volts.)

At the beginning of a cycle, C_E will start to charge from point A on the characteristic curve. At point B, where the voltage of C_E equals V_p, the UJT will fire and the characteristic curve goes into the negative resistance region. The voltage on the capacitor cannot change instantaneously, however, and the dynamic operating path will move from point B to point C. The time required to move from B to C is approximately 1 μs. From point C, the operating path follows an essentially straight line to point D, which is approximately equal to the valley point. This straight line has a slope of approximately 32 volts/ampere or 32 ohms, and is composed of R1 and the UJT emitter base one saturation resistance. There is not enough emitter current available to sustain operation at point D, and the operating path tries to follow the characteristic curve to the point where it is intersected by the load line determined by R_E. If the emitter circuit of the UJT were purely resistive (i.e. no capacitor C_E), this intersection point would be a stable operating point. To reach this point, however, the emitter voltage must increase. The resistance from emitter to base-one will also increase, and the emitter current will decrease somewhat. When the emitter voltage increases, however, current starts to

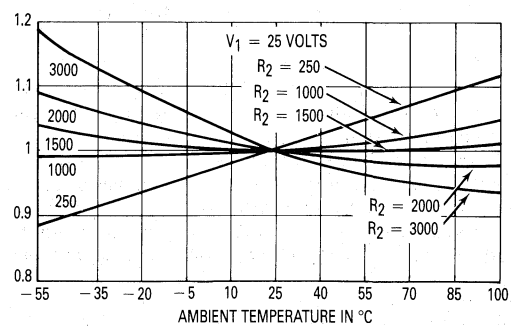
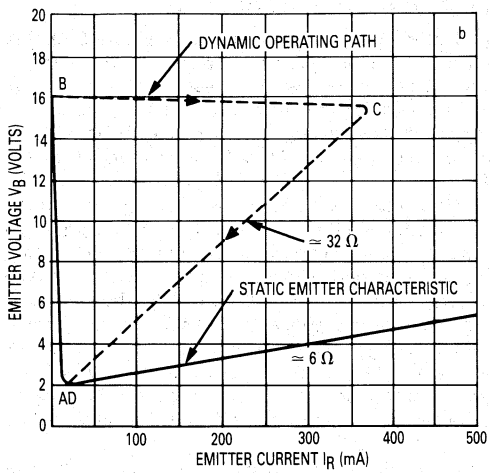


Figure 6.75. Frequency versus Temperature for a UJT Relaxation Oscillator Circuit. (Frequency is Normalized to 25°C and R2 is a Variable Parameter)



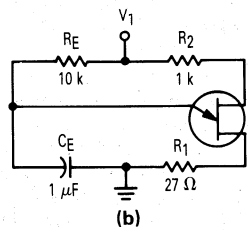


Figure 6.76. Determining the Dynamic Operating Path

flow into the capacitor and the emitter current is reduced more than that required by the characteristic curve. So, with a capacitor C_E in the emitter circuit, there are no stable operating points in the negative resistance region, and from point D, therefore, the operating path goes to point A again and the cycle repeats. It takes about 180 μs to traverse the path from point C to point A in the circuit shown.

The shape of the dynamic operating path is determined by the capacitor C_E, the bias voltage, and the resistor R1. Figure 6.77 shows operating paths for different values of C_E while 6.77(b) shows operating paths for fixed C_E but varying interbase voltage.

When an inductive load, like a relay coil, for example, is substituted for R1, the dynamic operating path will be somewhat different. Figure 6.78(a) shows a relaxation oscillator having a pulse transformer instead of R1, and Figure 6.78(b) shows the resulting dynamic operating path. An important difference here is that the emitter no longer ceases to conduct when valley voltage is reached but goes down to less than 0.5 volts before turning off. Figure 6.78(c) shows how the turn-off voltage is dependent on the bias voltage.

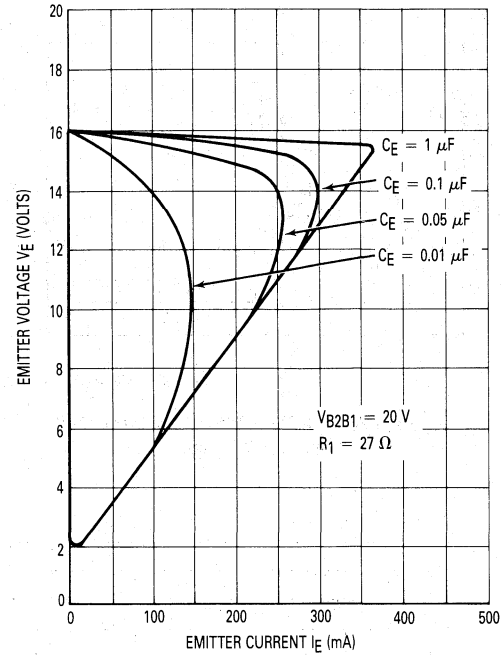


Figure 6.77(a). Dynamic Operating Path versus Emitter Capacitance (Circuit Figure 6.80)

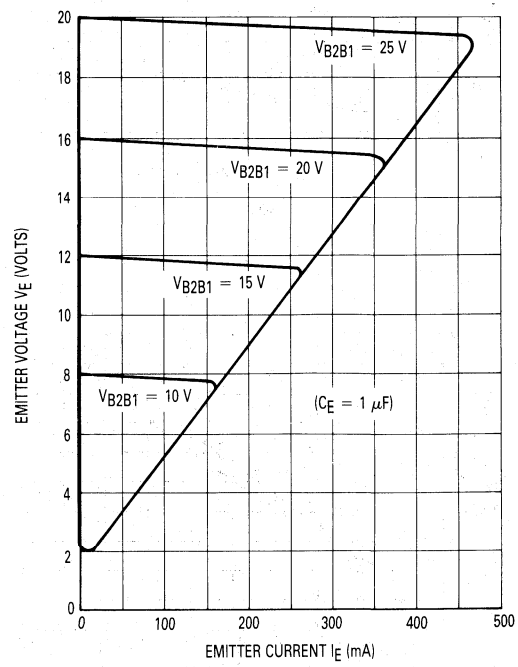


Figure 6.77(b). Dynamic Operation Path versus Interbase Voltage (Circuit Figure 6.80)

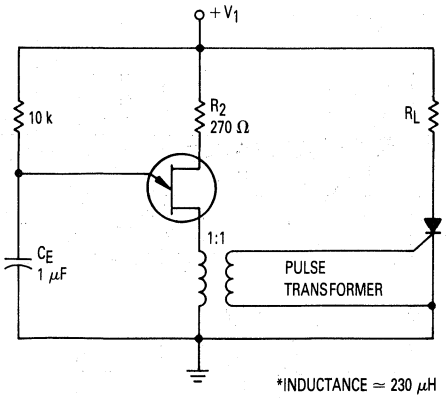
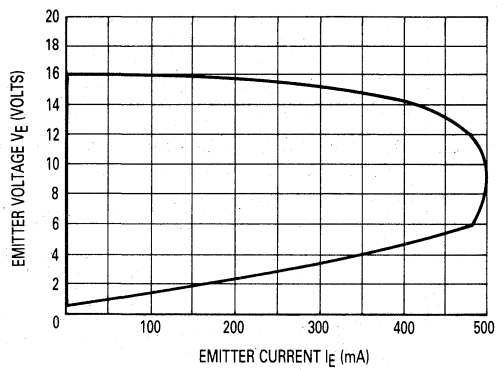
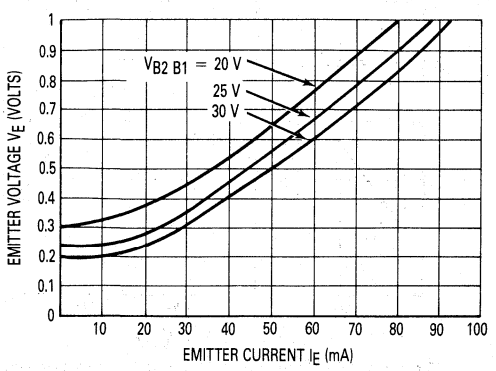


Figure 6.78(a). A Relaxation Oscillator with Inductive* Load



(b)



(c)

BATTERY CHARGER USING A UJT

The battery charger circuit shown in Figure 6.79(a) is a very simple circuit utilizing the relaxation oscillator.

The circuit will not work unless the battery to be charged is connected with proper polarity. The battery voltage controls the charger and when the battery is fully charged, the charger will not supply current to the battery.

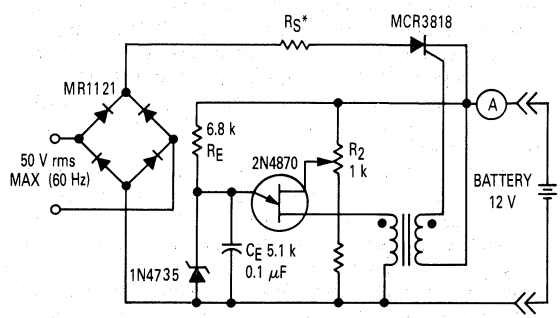
The circuit operation is as follows: The battery charging current is obtained through the SCR when it is triggered into the conducting state by the UJT relaxation oscillator. The oscillator is only activated when the battery voltage is low. V_{B2B1} of the UJT is derived from the voltage of the battery to be charged, and since $V_p = V_D + \eta V_{B2B1}$, the higher V_{B2B1}, the higher V_p. When V_p exceeds the breakdown voltage of the zener diode Z1, the UJT will cease to fire and the SCR will not conduct. This indicates that the battery has attained its desired charge as set by R2.

The relaxation oscillator itself and the waveforms associated with the operation is shown in Figure 6.79(b). (The voltage increase will tend to change the pulse repetition rate, but this is not important, since the battery will tend to average the output.)

FEEDBACK CIRCUITS

The circuits described so far have been manual control circuits; i.e., the power output is controlled by a potentiometer turned by hand. Simple feedback circuits may be constructed by replacing R_T with heat or light-dependent sensing resistors; however, these circuits have no means of adjusting the operating levels. The addition of a transistor to the circuits allows complete control.

Figure 6.81 shows a feedback control using a sensing resistor for feedback. The sensing resistor may respond



T1 — PRIMARY = 30 TURNS #22
SECONDARY = 45 TURNS #22
CORE = FERROXCUBE 203 F 181-3C3
* RS — SERIES RESISTANCE TO LIMIT CURRENT THROUGH SCR.
MCR 2818-3 IS RATED AT 20 AMPS rms.

Figure 6.79(a). A 12 Volt Battery Charger Control (20 Amps rms Max)

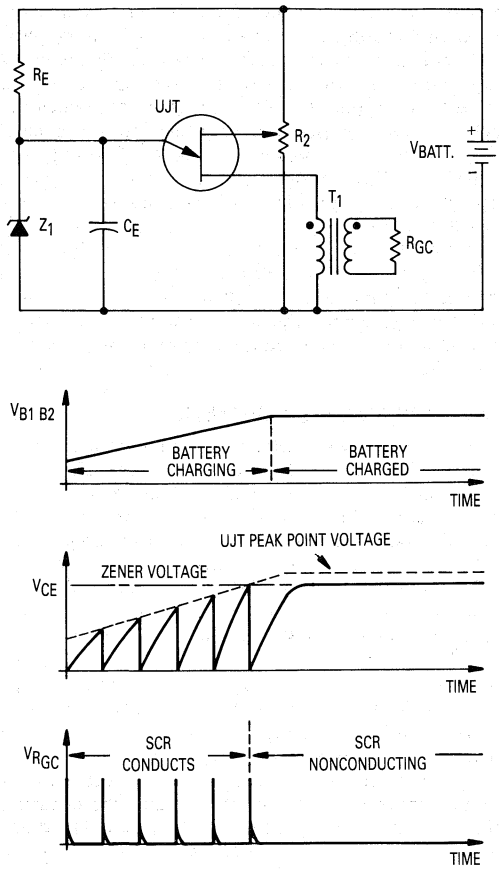


Figure 6.79(b). Waveforms Associated with Battery Charger Operation

to any one of many stimuli such as heat, light, moisture, pressure, or magnetic field. R_S is the sensing resistor and R_C is the control resistor that establishes the desired operating point. Transistor Q_1 is connected as an emitter follower such that an increase in the resistance of R_S decreases the voltage on the base of Q_1, causing more current to flow. Current through Q_1 causes voltage to charge C_T, triggering the UJT at some phase angle. As R_S becomes larger, more current flows into the capacitor, the voltage builds up faster, causing the UJT to trigger at a smaller phase angle and more power is applied to the load. When R_S decreases, less power is applied to the load. Thus, this circuit is for a sensing resistor which decreases in response to too much power in the load. If the sensing resistor increases with load power, then R_S and R_C should be interchanged.

If the quantity to be sensed can be fed back to the circuit in the form of an isolated, varying dc voltage such as the output of a tachometer, it may be inserted between the voltage divider and the base of Q1 with the proper polarity. In this case, the voltage divider would be a potentiometer to adjust the operating point. Such a circuit is shown in Figure 6.82.

In some cases, average load voltage is the desired feedback variable. In a half wave circuit this type of feedback usually requires the addition of a pulse transformer, shown in Figure 6.83. The RC network, R_1, R_2, C_1 averages load voltage so that it may be compared with the set point on R_S by Q_1. Full wave operation of this type of circuit requires dc in the load as well as the control circuit. Figure 6.84 is one method of obtaining this full wave control.

There are, of course, many more sophisticated circuits which can be derived from the basic circuits discussed here. If, for example, very close temperature control is desired, the circuit of Figure 6.81 might not have sufficient gain. To solve this problem a dc amplifier could be inserted between the voltage divider and the control transistor gate to provide as close a control as desired. Other modifications to add multiple inputs, switched gains, ramp and pedestal control, etc., are all simple additions to add sophistication. Basically, however, it is the UJT itself which provides the fast rising, high current pulse,

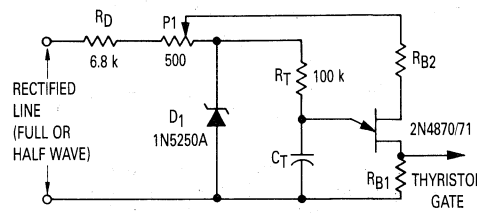


Figure 6.80. Circuit for Line Voltage Compensation

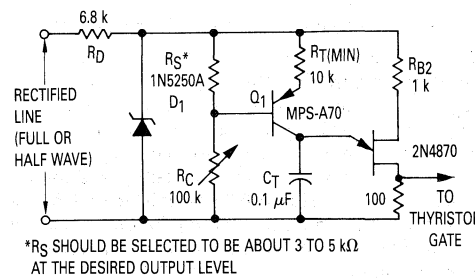


Figure 6.81. Feedback Control Circuit

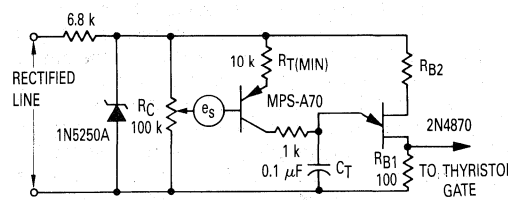


Figure 6.82. Voltage Feedback Circuit

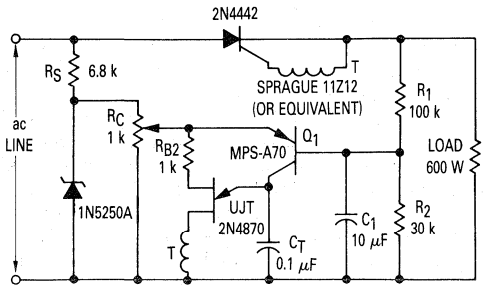


Figure 6.83. Half Wave, Average Voltage Feedback

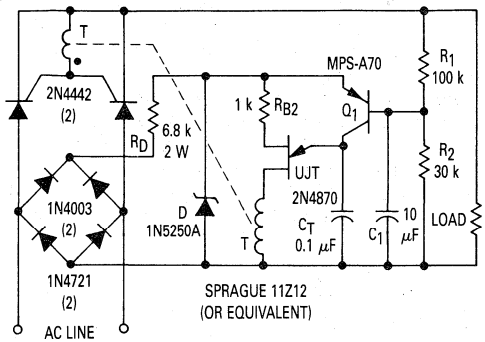


Figure 6.84. Full Wave, Average Voltage Feedback Control

which is desirable for reliable thyristor operation. The ease of adding feedback and relative insensitivity to line voltage changes are additional benefits gained from using this trigger device.

TEMPERATURE-SENSITIVE HEATER CONTROL

Figure 6.85 shows a heater circuit which is controlled by the room temperature. This circuit eliminates several of the disadvantages of the mechanical control: large size, high price, unreliability, and poor power regulation. The mechanical control is an on-off control and is not capable of regulating the power. By using phase control, the circuit shown in Figure 6.85 is able to reduce the power to the load as the desired temperature is reached, thus eliminating much of the overshoot inherent in mechanical controls.

The line voltage is full-wave rectified by the bridge consisting of D1 through D4. The output of the bridge is applied to the control circuit through resistor R1 and clamped to 20 volts by zener diode D5. The thermistor R_T and variable resistor R2 control the base current for transistor Q1. R2 is adjusted so Q1 is off at the desired temperature. When Q1 is off, no current can flow to capacitor C1, and C1 cannot charge to the firing voltage of unijunction transistor Q2. Therefore, Q2 cannot fire the TRIAC. If the temperature decreases, the resistance of R_T increases, Q1 is turned on and current flows to C1. C1 charges to the firing voltage of Q2, Q2 fires and turns the TRIAC on through pulse transformer T1. If the temperature continues to decrease, the resistance of R_T increases more and Q1 is turned on more. C1 charges faster and the TRIAC is triggered earlier, delivering more power to the load. As the temperature increases, the resistance of R_T decreases and Q1 will conduct less. C1 takes longer to charge and the TRIAC is triggered later in the cycle. When the desired temperature is reached, Q1 is off and the TRIAC is off.

The circuit shown is for a heater load, but the circuit could also be used to control a motor with a constant load such as a blower motor, as indicated by the dotted portion of Figure 6.85. The circuit is shown for a heating application but can be used for cooling by interchanging R_T and R2.

800 W LIGHT-DIMMER CIRCUITS

Figure 6.86 shows a wide-range light-dimmer circuit using a unijunction transistor and a pulse transformer to provide phase control for the TRIAC. The circuit operates from a 115 volt, 60 Hz source and can control up to 800 watts of power to incandescent lights. The power to the lights is controlled by varying the conduction angle of the TRIAC from 0° to about 170°. The power available at 170° conduction is better than 97% of that at the full 180°.

Operation begins when ac voltage is applied to the diode bridge consisting of diodes D1 through D4. The bridge rectifies the input voltage and applies a dc voltage to resistor R1 and zener diode D5. The zener applies a constant voltage of 22 volts to unijunction transistor Q1 except at the end of each alternation when the line voltage drops to zero. During each half cycle, capacitor C1 charges through variable resistor R2 until the capacitor voltage equals the emitter firing voltage V_P of Q1. When V_P is reached, Q1 fires and C1 discharges through the emitter of Q1 and a pulse is applied to the TRIAC through pulse transformer T1. Once the TRIAC turns on, voltage to the timing circuit is removed and no further pulses can occur during that half cycle. Since the line voltage is full-wave rectified and applied to the phase control circuit, the operation is the same for both positive and negative half cycles. Variable resistor R2 varies the time constant of the timing circuit thus providing phase control of the TRIAC.

800 W SOFT-START LIGHT DIMMER

The circuit shown in Figure 6.87 is a light dimmer with soft-start operation. Soft starting is desirable because of the very low resistance of a cold filament compared to its hot resistance. This low resistance causes very high inrush currents when a lamp is first turned on, and this leads to short lamp life. Soft starting also allows the use of TRIACs with lower current ratings. Failures caused by high inrush currents are eliminated by the soft-start feature, which applies current to the bulb slowly enough to eliminate high surges. Accidental turn on, which could nullify this advantage, is prevented by a special dv/dt network (R6, C3) across the TRIAC.

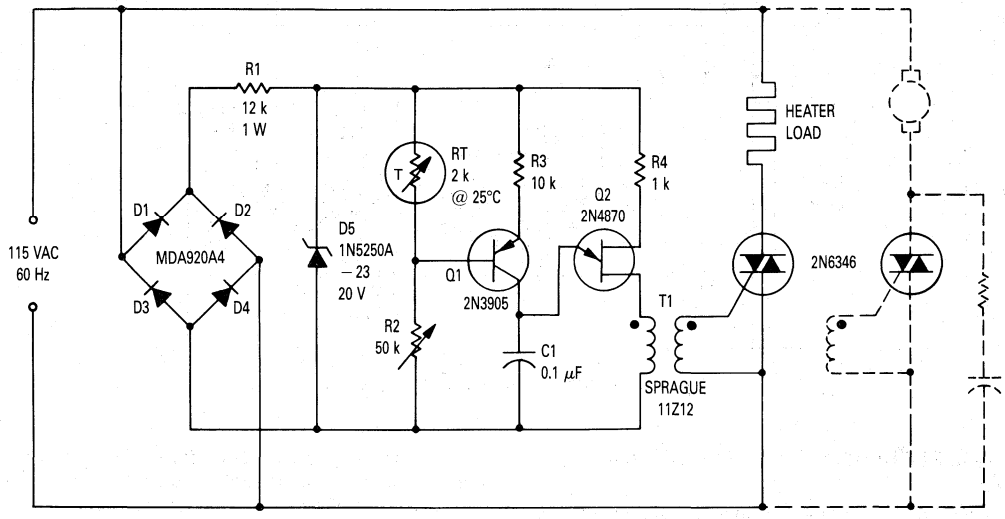


Figure 6.85. Temperature-Sensitive Heater Control

Operation of this circuit begins when 115 Vac input voltage is applied to the diode bridge consisting of D1 through D4. The bridge rectifies the input and applies a dc voltage to resistor R1 and zener diode D5. The zener provides a constant voltage of 20 volts to unijunction transistor Q1, except at the end of each half-cycle of the input when the line voltage drops to zero. Initially the voltage across capacitor C1 is zero and capacitor C2 cannot charge to trigger Q1. C1 will begin to charge, but because the voltage is low, C2 will be charged to a voltage adequate to trigger C1 only near the end of the half cycle. Although the lamp resistance is low at this time, the voltage applied to the lamp is low and the inrush current is small. Then the voltage on C1 rises, allowing C2 to trigger Q1 earlier in the cycle. At the same time the lamp is being heated by the slowly increasing applied voltage, and by the time the peak voltage applied to the lamp has reached its maximum value, the bulb has been heated sufficiently so that the peak inrush current is kept to a reasonable value. Resistor R4 controls the charging rate of C2 and provides the means to dim the lamp. Diode D6 and resistor R7 improve operation at low conduction angles.

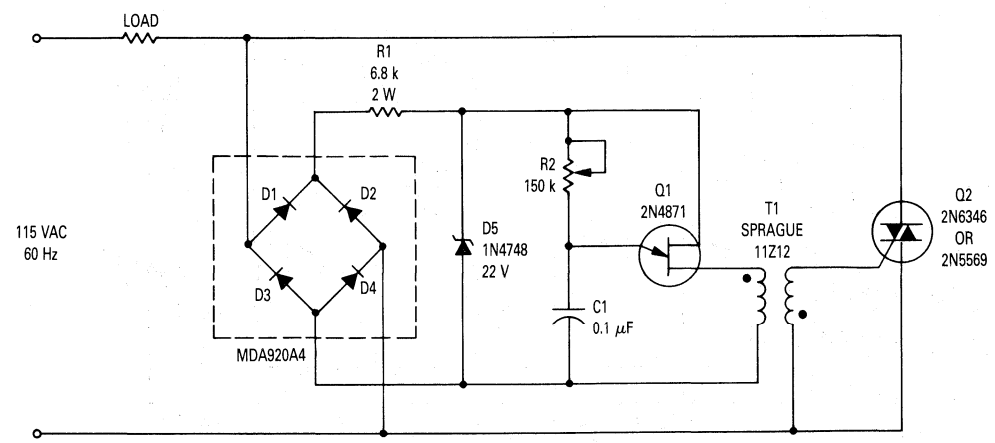


Figure 6.86. 800 W TRIAC Light Dimmer

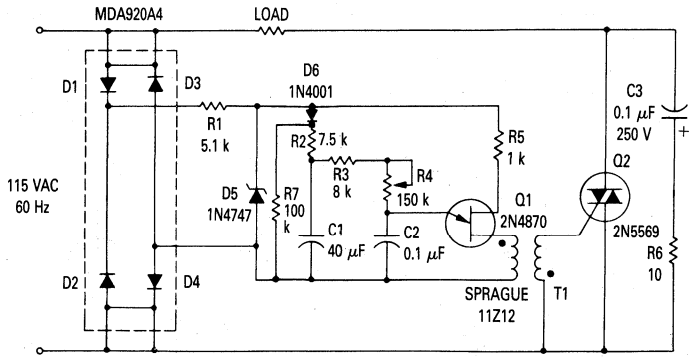


Figure 6.87. 800 W Soft-Start Light Dimmer

VOLTAGE REGULATOR FOR A PROJECTION LAMP

The circuit shown in Figure 6.88 will regulate the rms output voltage across the load (a projection lamp) to 100 volts ±2% for an input voltage between 105 and 250 volts ac. This is accomplished by indirectly sensing the light output of lamp L1 and applying this feedback signal to the firing circuit (Q1 and Q2) which controls the conduction angle of TRIAC Q3. The load is a 150-watt projection lamp which has a reflector mirror included inside the glass envelope. If the light output of the lamp were sensed directly by the photocell, it would respond to the 60 Hz variations of the supply voltage unless additional filter components were used. Therefore, another approach was used to generate the feedback signal. The reflector inside the lamp's envelope glows red due to the heat of the filament. Since the reflector has a relatively large mass it cannot respond to the supply frequency, and its light output provides a form of integration. This light is then used as a feedback signal. To eliminate 60 Hz modulation of the photocell, it is mounted at one end of a black tube with the other end of the tube directed at the back side of the reflector in the lamp. The lamp voltage is provided by TRIAC Q3, whose conduction angle is set by the firing circuit for unijunction transistor Q2. The circuit is synchronized with the line through the full-wave bridge rectifier. The voltage to the firing circuit is limited by zener diode D5. Phase control of the supply voltage is set by the charging rate of capacitor C1. Q2 will fire when the voltage on C1 reaches approximately 0.65 times the zener voltage. The charging rate of C1 is set by the conduction of Q1, which is controlled by the resistance of photocell R2. Potentiometers R3 and R4 are used to set the lamp voltage to 100 volts when the line voltage is 105 volts and 250 volts, respectively. This assures that

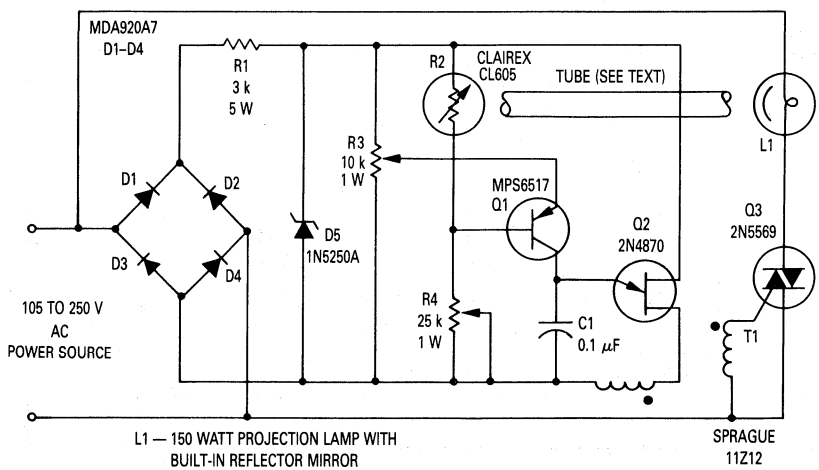
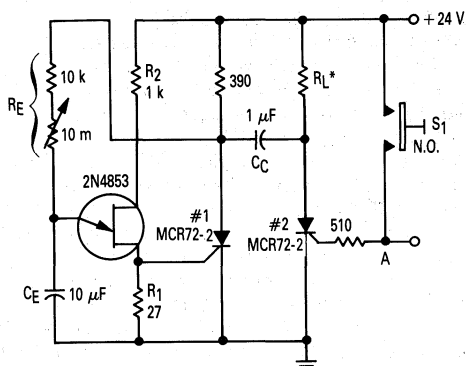


Figure 6.88. Voltage Regulator for a Projection Lamp

the lamp voltage will be within the desired tolerance over the operating range of input voltage. Some interaction will occur between R3 and R4 and the adjustment of each potentiometer may have to be made several times. Since this is an rms voltage regulator, a true rms meter must be used to adjust the load voltage.

TIMER CIRCUITS

A variation of the UJT-relay time delay circuit is shown in Figure 6.89. Here the relay is replaced by an SCR which generally reduces circuit cost. After one cycle of operation, SCR #1 will be on, and a low value of voltage is applied to the UJT emitter circuit, thus interrupting the timing function. When push button S1 is pushed, or a positive going pulse is applied at point A, SCR #2 will turn on, and SCR #1 will be turned off by commutating capacitor C_C. With SCR #1 off, the supply voltage will be applied to R_E and the circuit will begin timing again. After a period of time determined by the setting of R_E, the UJT will fire and turn SCR #1 on and commutate SCR #2 off.



*VALUE OF R_L MUST BE LOW ENOUGH TO ALLOW HOLD CURRENT TO FLOW IN THE SCR.

Figure 6.89. A Simple Time Delay Circuit Using Two SCRs

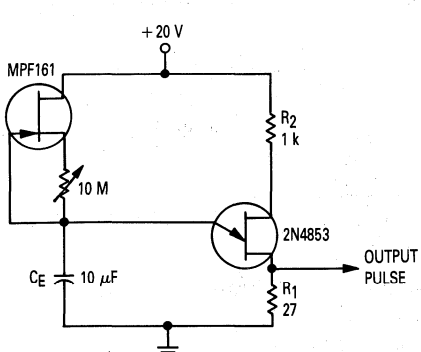


Figure 6.90. A Time Delay Circuit Featuring Constant Current Charging

The time delay is determined by the charge time of the capacitor. In order to achieve long time delays, R_E, C_E, or both will have to be large. For good accuracy and repeatability, the capacitor must have a leakage current that is much smaller than the charge current. A Mylar type capacitor has been found to be good for this purpose, but since this type of capacitor is fairly expensive for large values of capacitance, it is preferable to increase R_E in order to obtain long time delays. Large values of R_E, however, creates a problem due to the UJT peak point emitter current I_p. When the capacitor is charged almost to the peak point, only a small voltage will appear across R_E, and if R_E is very large, only a small current will be flowing. If the peak current of the UJT is appreciable, the device will never fire if the current through R_E is not sufficient to supply I_p. The annular device, having a typical I_p of 0.2 μA @ V_{B2B1} = 25 Vdc, offers an advantage in this area and large values of R_E can be used. However, when charging through a resistor, the charge current will initially be relatively large while the charge current when the voltage on the capacitor is close to V_p will be small. It would, for this reason, be advantageous to charge with a constant small current. This can be accomplished by simply replacing R_E by a junction field effect transistor as shown in Figure 6.90. Since the JFET is fully on when there is no voltage from gate to source, the 10 MΩ resistor will determine the amount of off bias applied to the FET. A constant current of less than 1 μA can easily be obtained which results in time delays up to 10 minutes.

When the capacitor is charged with a linear current, the charge time can be found from the equation:

$$t_{charge} = \frac{(V_p - V_V) \cdot C_E}{I_{charge}}$$

When C_E is in microfarads and I_{charge} is in microamperes, t will be in seconds.

However, even an emitter peak current as low as 0.2 μA is objectionable if longer time delays are desired. In the circuit shown in Figure 6.91 the peak current is supplied separately from the charging current and extremely long time delays are possible. Transistor Q1 and resistors R1, R2, and R3 form a constant current source and the charge current might be adjusted to be as low as a few nanoamperes. This current would, of course, not be sufficient to fire the UJT where I_p = 0.2 μA unless the peak current was supplied from another source. Field effect transistor Q2, acting as a source follower, supplies the current flowing into the emitter lead prior to firing and diode D1 provides a low impedance discharge path for C_E. D1 must be selected to have a leakage much lower than the charge current.

The charge current to C_E is given by the formula:

$$I_{charge} = \frac{E - V_{BE}}{R3} - I_B$$

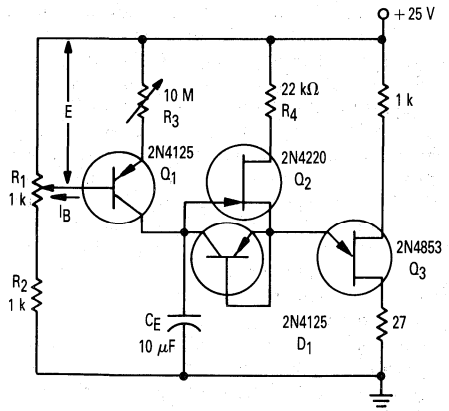


Figure 6.91. Long Duration Time Delay

Since I_B is small, the delay time will vary linearly with R3. The voltage E, applied across R3 and the base emitter junction of Q1, is set by the variable resistor R1. Time delays up to 10 hours are possible with this circuit. Resistor R4 in series with the FET drain terminal must be large enough not to allow currents in excess of I_V to flow when the UJT is on, otherwise the UJT will not turn off and the circuit will latch up.

PUT APPLICATIONS

PUTs are negative resistance devices and are often used in relaxation oscillator applications and as triggers for controlling thyristors. Due to their low leakage current, they are useful for high-impedance circuits such as long-duration timers and comparators.

TYPICAL CIRCUITS

The following circuits show a few of the many ways in which the PUT can be used. The circuits are not optimized even though performance data is shown.

In several of the circuit examples, the versatility of the PUT has been hidden in the design. By this it is meant that in designing the circuit, the circuit designer was able to select a particular intrinsic standoff ratio or he could select a particular R_G (gate resistance) that would provide a maximum or minimum valley and peak current. This makes the PUT very versatile and very easy to design with.

LOW VOLTAGE LAMP FLASHER

One advantage of the PUT over a conventional unijucntion transistor is that the PUT operates very well for low supply voltages. This is due to the low forward voltage drop of the PUT, 1.5 volts maximum for the MPU131-33, compared to the emitter saturation voltage of the UJT of 3 volts maximum for the 2N5431.

A circuit using the PUT in a low voltage application is shown in Figure 6.92 where a supply voltage of 3 volts is used. The circuit is a low voltage lamp flasher composed of a relaxation oscillator formed by Q1 and an SCR flip flop formed by Q2 and Q3.

With the supply voltage applied to the circuit, the timing capacitor C1 charges to the firing point of the PUT, 2 volts plus a diode drop. The output of the PUT is coupled through two 0.01 μF capacitors to the gate of Q2 and Q3. To clarify operation, assume that Q3 is on and capacitor C4 is charged plus to minus as shown in the figure. The next pulse from the PUT oscillator turns Q2 on. This places the voltage on C4 across Q3 which momentarily reverse biases Q3. This reverse voltage turns Q3 off. After discharging, C4 then charges with its polarity reversed to that shown. The next pulse from Q1 turns Q3 on and Q2 off. Note that C4 is a non-polarized capacitor.

For the component values shown, the lamp is on for about 1/2 second and off the same amount of time.

VOLTAGE CONTROLLED RAMP GENERATOR

The PUT provides a simple approach to a voltage controlled ramp generator, VCRG, as shown in Figure 6.93(a).

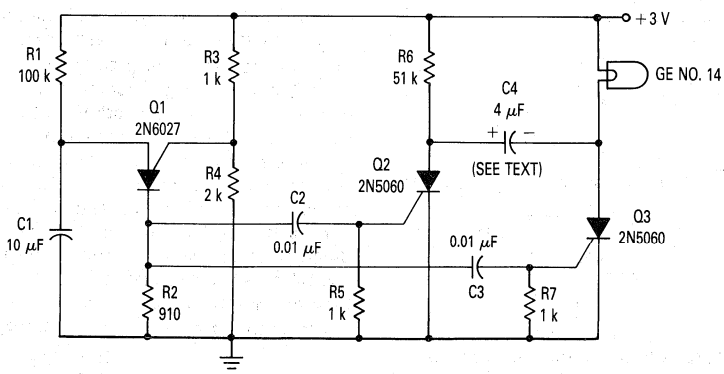


Figure 6.92. Low Voltage Lamp Flasher

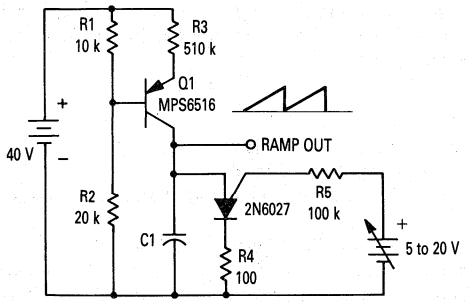
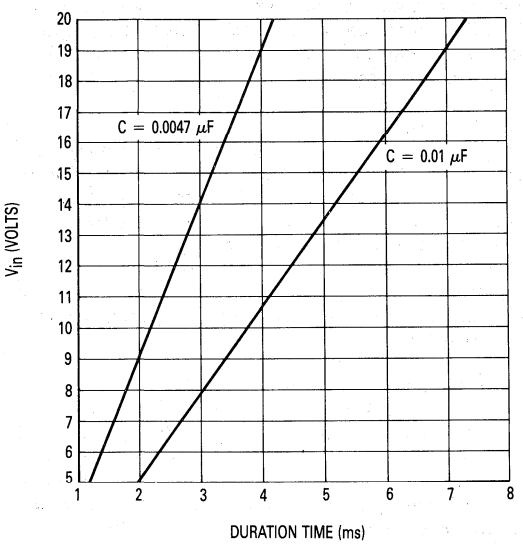


Figure 6.93(a). Voltage Controlled Ramp Generator (VCRG)



(b). Voltage versus Ramp Duration Time of VCRG

The current source formed by Q1 in conjuction with capacitor C1 set the duration time of the ramp. As the positive dc voltage at the gate is changed, the peak point firing voltage of the PUT is changed which changes the duration time, i.e., increasing the supply voltage increases the peak point firing voltage causing the duration time to increase.

Figure 6.93(b) shows a plot of voltage-versus-ramp duration time for a 0.0047 μF and a 0.01 μF timing capacitor. The figure indicates that it is possible to have a change in frequency of 3 ms and 5.4 ms for the 0.0047 μF and the 0.01 μF capacitor respectively as the control voltage is varied from 5 to 20 volts.

LOW FREQUENCY DIVIDER

The circuit shown in Figure 6.94 is a frequency divider with the ratio of capacitors C1 and C2 determining division. With a positive pulse applied to the base of Q1, assume that C1 = C2 and that C1 and C2 are discharged. When Q1 turns off, both C1 and C2 charge to 10 volts each through R3. On the next pulse to the base of Q1, C1 is again discharged but C2 remains charged to 10 volts. As Q1 turns off this time, C1 and C2 again charge. This time C2 charges to the peak point firing voltage of the PUT causing it to fire. This discharges capacitor C2 and allows capacitor C1 to charge to the line voltage. As soon as C2 discharges and C1 charges, the PUT turns off. The next cycle begins with another positive pulse on the base of Q1 which again discharges C1.

The input and output frequency can be approximated by the equation

$$f_{in} \approx \frac{(C1 + C2)}{C1} f_{out}$$

For a 10 kHz input frequency with an amplitude of 3 volts, Table 6.X shows the values for C1 and C2 needed to divide by 2 to 11.

This division range can be changed by utilizing the programmable aspect of the PUT and changing the voltage on the gate by changing the ratio R6/(R6+R5). Decreasing the ratio with a given C1 and C2 decreases the division range and increasing the ratio increases the division range.

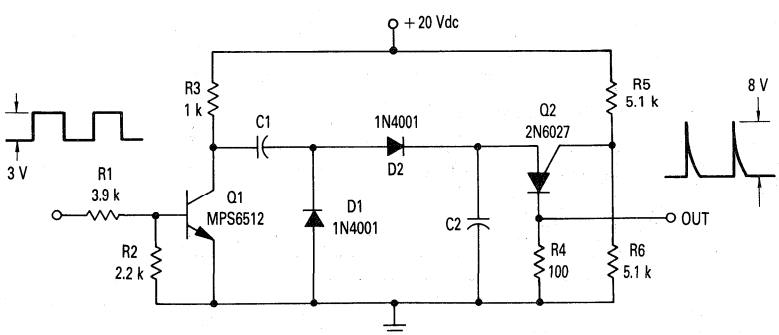


TABLE 6.X

C_1	C_2	Division
0.01 μF	0.01 μF	2
0.01 μF	0.02 μF	3
0.01 μF	0.03 μF	4
0.01 μF	0.04 μF	5
0.01 μF	0.05 μF	6
0.01 μF	0.06 μF	7
0.01 μF	0.07 μF	8
0.01 μF	0.08 μF	9
0.01 μF	0.09 μF	10
0.01 μF	0.1 μF	11

Figure 6.94. Low Frequency Divider

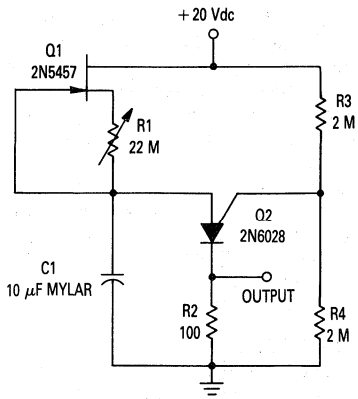


Figure 6.95. 20-Minute, Long Duration Timer

The circuit works very well and is fairly insensitive to the amplitude, pulse width, rise and fall times of the incoming pulses.

PUT LONG DURATION TIMER

A long duration timer circuit that can provide a time delay of up to 20 minutes is shown in Figure 6.95. The circuit is a standard relaxation oscillator with a FET current source in which resistor R1 is used to provide reverse bias on the gate-to-source of the JFET. This turns the JFET off and increases the charging time of C1. C1 should be a low leakage capacitor such as a mylar type.

The source resistor of the current source can be computed using the following equation:

$$V_{GS} = V_P\,(1 - \sqrt{I_O/I_{DSS}})$$

$$\therefore R1 = \frac{V_{GS}}{I_O}$$

where I_O is the current out of the current source.
V_P is the pinch off voltage,
V_{GS} is the voltage gate-to-source and,
I_{DSS} is the current, drain-to-source, with the gate shorted to the source.

The time needed to charge C1 to the peak point firing voltage of Q2 can be approximated by the following equation:

$$t = \frac{C\Delta V}{I},$$

where t is time in seconds
C is capacitance in μF,
ΔV is the change in voltage across capacitor C1, and
I is the constant current used to charge C1.

Maximum time delay of the circuit is limited by the peak point firing current, I_P, needed to fire Q2. For charging currents below I_P, there is not enough current available from the current source to fire Q2, causing the circuit to lock up. Thus a PUT works better than a unijunction in a long duration timer because it has a lower peak point firing current. Also, because of the programmable aspect of the PUT, I_P can be made very small by making R_G (the equivalent parallel resistance of R3 and R4) large, (1 MΩ) as shown in Figure 6.95.

PHASE CONTROL

Figure 6.96 shows a circuit using a PUT for phase control of an SCR. The relaxation oscillator formed by Q2 provides conduction control of Q1 from 1 to 7.8 milliseconds or 21.6° to 168.5°. This constitutes control of over 97% of the power available to the load.

Only one SCR is needed to provide phase control of both the positive and negative portion of the sine wave by putting the SCR across the bridge composed of diodes D1 through D4.

BATTERY CHARGER USING A PUT

A short circuit proof battery charger is shown in Figure 6.97 which will provide an average charging current of about 8 amperes to a 12 volt lead acid storage battery. The charger circuit has an additional advantage in that it will not function nor will it be damaged by improperly connecting the battery to the circuit.

With 115 volts at the input, the circuit commences to function when the battery is properly attached. The battery provides the current to charge the timing capacitor C1 used in the PUT relaxation oscillator. When C1 charges

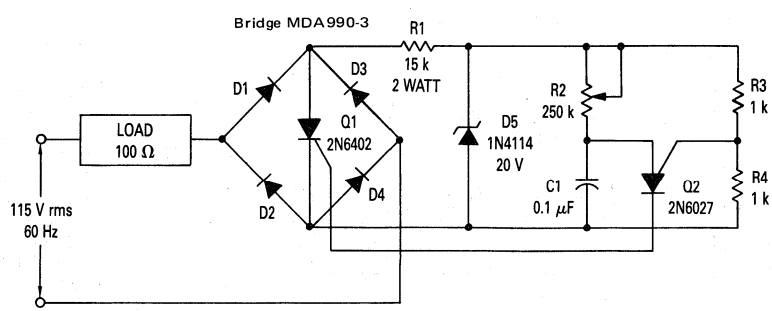


Figure 6.96. SCR Phase Control

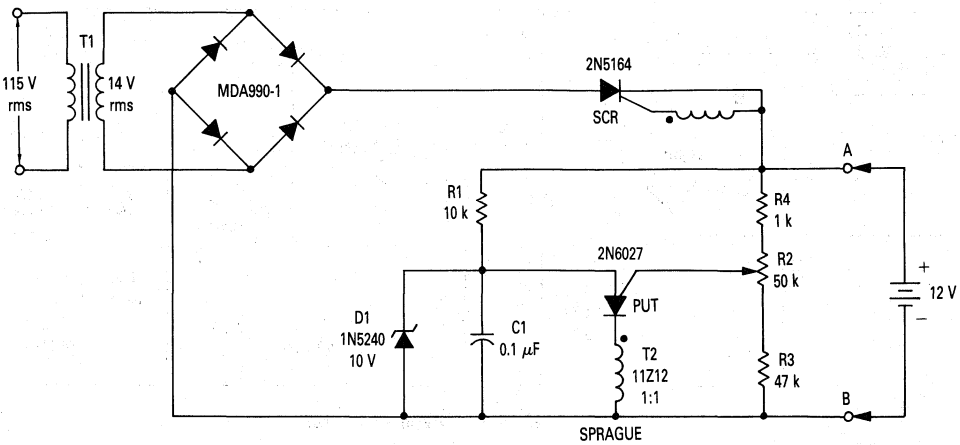


Figure 6.97(a). 12-Volt Battery Charger

to the peak point voltage of the PUT, the PUT fires turning the SCR on, which in turn applies charging current to the battery. As the battery charges, the battery voltage increases slightly which increases the peak point voltage of the PUT. This means that C1 has to charge to a slightly higher voltage to fire the PUT. The voltage on C1 increases until the zener voltage of D1 is reached which clamps the voltage on C1 and thus prevents the PUT oscillator from oscillating and charging ceases. The maximum battery voltage is set by potentiometer R2 which sets the peak point firing voltage of the PUT.

In the circuit shown, the charging voltage can be set from 10 V to 14 V, the lower limit being set by D1 and the upper limit by T1. Lower charging voltages can be obtained by reducing the reference voltage (reducing the value of zener diode D1) and limiting the charging current (using either a lower voltage transformer, T1, or adding resistance in series with the SCR).

Resistor R4 is used to prevent the PUT from being destroyed if R2 were turned all the way up.

Figure 6.97(b) shows a plot of the charging characteristics of the battery charger.

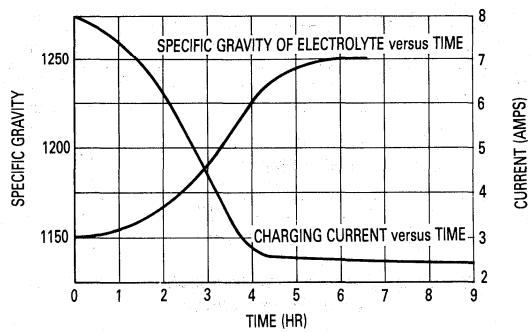


Figure 6.97(b). Charging Characteristics of Battery Charger

90 V rms VOLTAGE REGULATOR USING A PUT

The circuit of Figure 6.98 is an open loop rms voltage regulator that will provide 500 watts of power at 90 V rms with good regulation for an input voltage range of 110–130 V rms.

With the input voltage applied, capacitor C1 charges until the firing point of Q3 is reached causing it to fire. This turns Q5 on which allows current to flow through the load. As the input voltage increases, the voltage across R10 increases which increases the firing point of Q3. This delays the firing of Q3 because C1 now has to charge to a higher voltage before the peak-point voltage is reached. Thus the output voltage is held fairly constant by delaying the firing of Q5 as the input voltage increases. For a decrease in the input voltage, the reverse occurs.

Another means of providing compensation for increased input voltage is achieved by Q2 and the resistive divider formed by R6 and R7. As input voltage increases, the voltage at the base of Q2 increases causing Q2 to turn on harder which decreases the charging rate of C1 and further delays the firing of Q5.

To prevent the circuit from latching up at the beginning of each charging cycle, a delay network consisting of Q1 and its associated circuitry is used to prevent the current source from turning on until the trigger voltage has reached a sufficiently high level. This is achieved in the following way: Prior to the conduction of D2, the voltage on the base of Q1 is set by the voltage divider (R4 + R5)/(R1 + R3 + R4 + R5). This causes the base of Q1 to be more positive than the emitter and thus prevents Q1 from conducting until the voltage across R3 is sufficient to forward bias the base-emitter junction of Q1. This occurs when the line voltage has increased to about 15 volts.

The circuit can be operated over a different voltage range by changing resistors R6 and/or R4 which change the charging rate of C1.

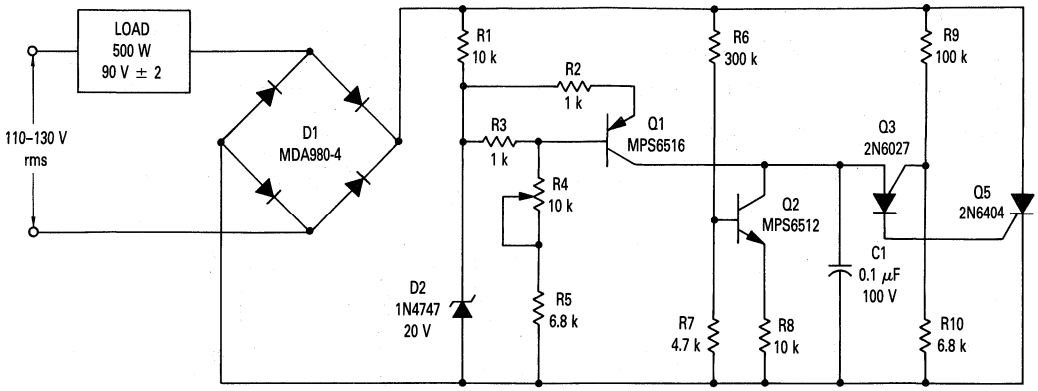
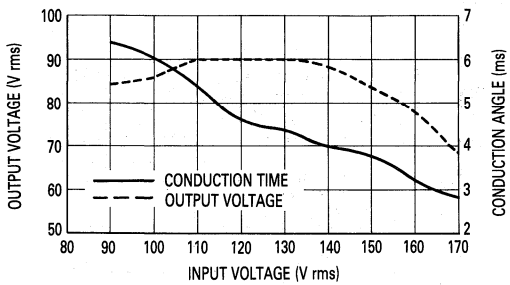


Figure 6.98(a). rms Voltage Regulator



(b). Output Voltage and Conduction Angle versus Input Voltage

Figure 6.98(b) provides a plot of output voltage and conduction angle versus input voltage for the regulator. As the figure indicates, good regulation can be obtained between the input voltage range of 110 to 130 volts.

SILICON BILATERAL SWITCH (SBS) APPLICATIONS

It is important that thyristor trigger circuitry be capable of supplying a fast rising, high current gate pulse to the power thyristors in order to prevent di/dt failure, especially when they are subjected to high inrush load currents. Because of the regenerative switching action and low dynamic on resistance of the SBS, it is ideally suited for this use.

These circuits indicate the uses for the SBS. In some applications the device switches on at V_S while in others it is turned on by drawing a small current out of the gate lead.

LAMP DIMMER

Figure 6.99 is the schematic diagram of a low cost full range lamp dimmer. Shunting the SBS with two 20 kΩ resistors minimizes the "flash-on" or hysteresis effect. V_S of the SBS is reduced to about 4 volts, and since this is below the operating voltage of the internal zener diodes, the temperature sensitivity of the device is increased.

An improved full range power controller suitable for lamp dimming and similar applications is shown in Figure 6.100.

It operates from a 120 volt, 60 Hz ac source and can control up to 1000 watts of power to incandescent bulbs. The power to the bulbs is varied by controlling the conduction angle of TRIAC Q1. Many circuits can be used for phase control, but the single RC circuit used is the simplest by far and was consequently chosen for this particular application. For settings such that no power is delivered to the load, the timing capacitor would never discharge through the SBS. The result is an abnormal amount of apparent phase shift caused by the capacitor starting to charge toward a source of voltage with a residual charge of the opposite sign. This is the cause of the hysteresis effect and is eliminated in this circuit by the addition of the two diodes and 5.1 kΩ resistor connected to the SBS gate. At the end of each positive half

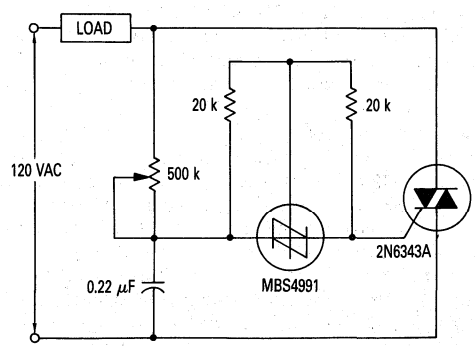


Figure 6.99. Low Cost Lamp Dimmer

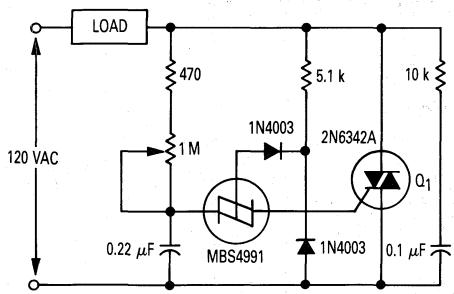


Figure 6.100. 1000 W TRIAC Light Dimmer

cycle when the applied voltage drops below that of the capacitor, gate current flows out of the SBS and it switches on, discharging the capacitor to near zero volts.

The RC network shown across the TRIAC represents a typical snubber circuit that is normally adequate to prevent line transients from accidentally firing the TRIAC.

ELECTRONIC CROWBAR

Occasionally the need arises for positive protection of expensive electrical or electronic equipment against excessive supply voltage. Such overvoltage conditions can occur due to improper switching, wiring, short circuits or failure of regulators. Where it is economically desirable to shut down equipment rather than allow it to operate on excessive supply voltage, an electronic "crowbar" circuit such as the one shown in Figure 6.101 can be employed to quickly place a short-circuit across the power lines, thereby dropping the voltage across the protected device to near zero and blowing a fuse. Since the TRIAC and SBS are both bilateral devices, the circuit is equally useful on ac or dc supply lines. With the values shown for R1, R2 and R3, the crowbar operating point can be adjusted over the range of 60 to 120 volts dc or 42 to 84 volts ac. The resistor values can be changed to cover a different range of supply voltages. The voltage rating of the TRIAC must be greater than the highest operating point as set by R2. I_1 is a low power incandescent lamp with a voltage rating equal to the supply voltage. It may be used to check the set point and operation of the unit by opening the test switch and adjusting the input or set point to fire the SBS. An alarm unit such as the Mallory Sonalert may be connected across the fuse to provide an audible indication of crowbar action. Note that this circuit may not act on short, infrequent power line transients.

SBS APPLICATIONS IN POWER CONTROL

The incandescent-lamp dimmer was one of the first circuits to use thyristors after their invention and has remained one of the most important applications of these devices. Triggering circuits for the lamp dimmer have taken many forms, from the relatively complex unijunction transistor oscillator with an RC slaving network to simpler circuits, which use a SBS trigger.

Figure 6.102 shows the basic control circuit. In the positive half-cycle, the 0.1 μF capacitor charges through the dual-section phase-shift circuit until its voltage reaches the break-over potential of the MBS4991 SBS. The MBS4991 potential then drops to about 1 volt, forcing charge from the 0.1 μF capacitor through the gate of the MAC210-4 TRIAC. This current turns the TRIAC on. When the TRIAC turns on, it removes the voltage from the timing circuit. C1 then discharges through the latched on TRIAC and the SBS if it also holds. What happens depends on the setting of R1, and the switching current, switching voltage, latching current and holding current of the SBS. Analysis of the circuit behavior is difficult and cannot be treated on a half-cycle basis because the previous half-cycle must be considered to establish the initial conditions, and because large residual voltages remain on C1. Several cycles of the ac line are needed to establish steady state conduction angles.

Hysteresis in the single RC phase controller is a result of the initial voltage on the capacitor before triggering. If the control is set to completely turn-off the lamp, triggering does not occur, and the capacitor voltage alternates up and down to some value less than V_S. If the control is advanced, the SBS fires and latches, causing the capacitor to charge from the previous polarity on-

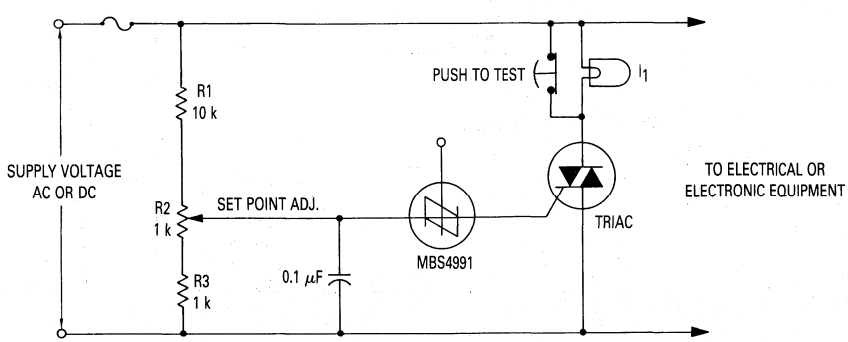


Figure 6.101. Electronic Crowbar

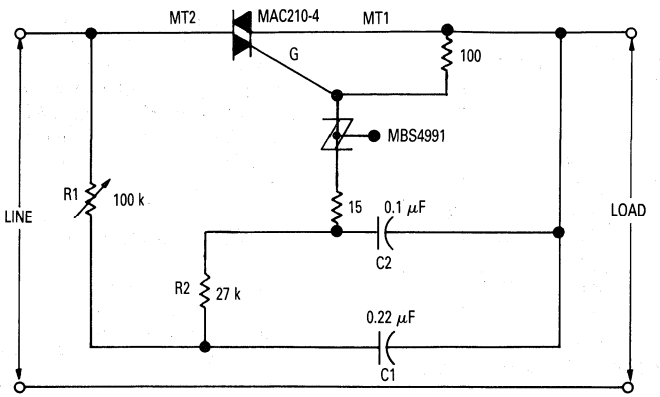


Figure 6.102. Basic Control Circuit for TRIAC Using a SBS Trigger

state voltage of about 1 V. Consequently, the control settings for dim illumination depend on whether the potentiometer is being advanced from an off state or retarded from a state with the lamp on, because the timing capacitor must charge through a different voltage gradient in the two cases.

The dual-section phase-shift network prevents hysteresis and allows reliable and stable triggering at all conduction angles. The 100 kΩ variable resistor R1 and the 0.22 μF capacitor C1 perform the basic phase shifting and serve as a charging supply for the 0.1 μF charge-storage capacitor C2. The 27 kΩ resistor isolates the trigger circuit from the phase shifter so that the voltage on C1 is only minimally affected by the triggering action. It is this isolation that reduces the hysteresis, prevalent in single-section phase-shift systems, to an unnoticeable level.

Although the same timing circuit is used for both halves of the cycle, the MBS4991 trigger is not exactly symmetrical. Therefore, the conduction angles of the MAC210-4 are not identical in the two half-cycles. As a result, a small dc component is introduced in the load.

In cases where high power must be handled or rapidly rising voltages may be encountered (high dv/dt), it is preferable to use two SCRs instead of a TRIAC for full-wave power control. Figure 6.103 illustrates a variation of the basic circuit to allow this. In the positive half-cycle, SCR-1 is triggered, through the primary of the pulse transformer (T), in exactly the same manner as in the basic circuit. In the negative half-cycle, the 0.1 μF triggering capacitor discharges through the shunt G-K diode of SCR-1 and the primary of the pulse transformer, inducing a pulse in the secondary, which triggers SCR-2.

The circuits described above were designed for incandescent-lamp dimmers and are ideally suited for this purpose. However, they may have many other uses, which are perhaps not immediately obvious. For example, uni-

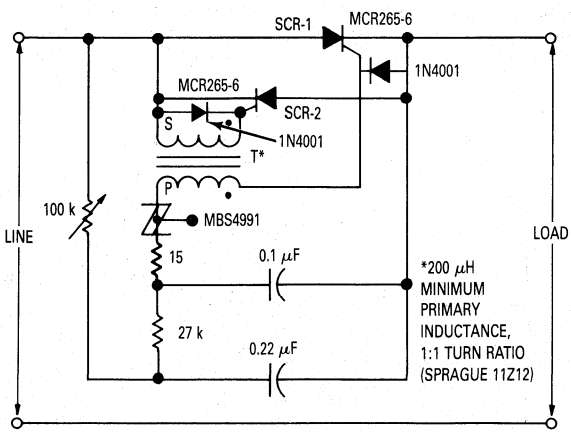


Figure 6.103. Basic Full-Wave Control Circuit for SCRs

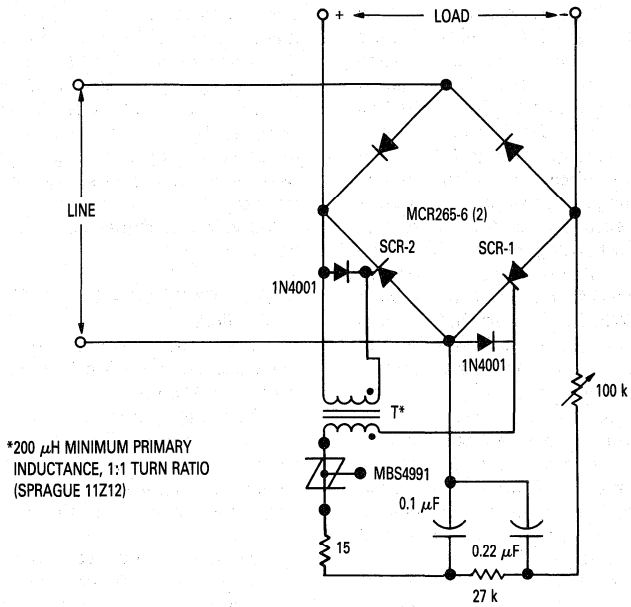


Figure 6.104. Variation of Basic Control Circuit to Provide Controlled DC Output

versal and shaded-pole motors are easily and conveniently controlled with these circuits. These motors have higher torque at low speeds when open-loop controlled in this manner rather than with rheostats or variable transformers, owing to the higher voltage pulses applied. In another application, a slight modification of the basic control circuit allows control of the dc output of a full-wave rectifier bridge using pulse-transformer coupling (Figure 6.104).

The power rating of these circuits is limited only by the thyristors employed. The control circuit will give sufficient drive for any thyristor that can be triggered with 50 mA or less gate current. For example, with MCR3818-4 controlled rectifiers mounted on a suitable heat sink, these circuits will control up to 3 kW power from a 120 V line.

TRIAC ZERO-POINT SWITCH APPLICATIONS

BASIC TRIAC ZERO-POINT SWITCH

Figure 6.105 shows a manually controlled zero-point switch useful in power control for resistive loads. Operation of the circuit is as follows. On the initial part of the positive half cycle, the voltage is changing rapidly from

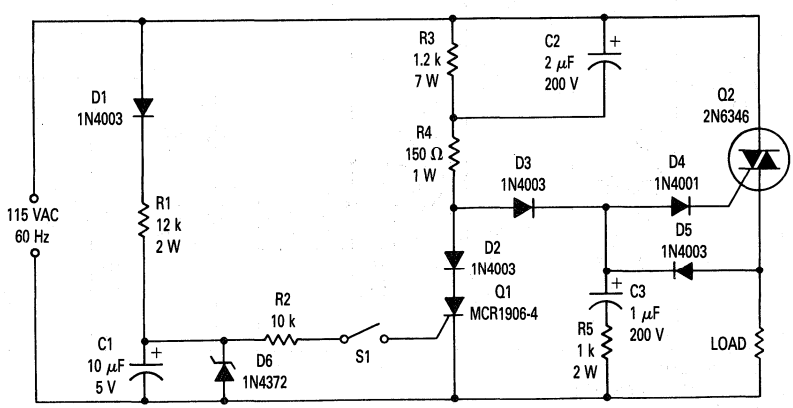


Figure 6.105. Zero-Point Switch

zero causing a large current flow into capacitor C2. The current through C2 flows through R4, D3, and D4 into the gate of the TRIAC Q2 causing it to turn on very close to zero voltage. Once Q2 turns on, capacitor C3 charges to the peak of the line voltage through D5. When the line voltage passes through the peak, D5 becomes reverse-biased and C3 begins to discharge through D4 and the gate of Q2. At this time the voltage on C3 lags the line voltage. When the line voltage goes through zero there is still some charge on C3 so that when the line voltage starts negative C3 is still discharging into the gate of Q2. Thus Q2 is also turned on near zero on the negative half cycle. This operation continues for each cycle until switch S1 is closed, at which time SCR Q1 is turned on. Q1 shunts the gate current away from Q2 during each positive half cycle keeping Q2 from turning on. Q2 cannot turn on during the negative cycle because C3 cannot charge unless Q2 is on during the positive half cycle.

If S1 is initially closed during a positive half cycle, SCR Q1 turns on but circuit operation continues for the rest of the complete cycle and then turns off. If S1 is closed during a negative half cycle, Q1 does not turn on because it is reverse biased. Q1 then turns on at the beginning of the positive half cycle and Q2 turns off.

Zero-point switching when S1 is opened is ensured by the characteristic of SCR Q1. If S1 is opened during the positive half cycle, Q1 continues to conduct for the entire half cycle and TRIAC Q2 cannot turn on in the middle of the positive half cycle. Q2 does not turn on during the negative half cycle because C3 was unable to charge during the positive half cycle. Q2 starts to conduct at the first complete positive half cycle. If S1 is opened during the negative half cycle, Q2 again cannot turn on until the beginning of the positive half cycle because C3 is uncharged.

A 3-volt gate signal for SCR Q1 is obtained from D1, R1, C1, and D6.

TEMPERATURE CONTROL WITH ZERO-POINT SWITCHING

Figure 6.106 shows a modulated TRIAC zero-point switching circuit designed to control heater loads operating from 115 Vac.

Circuit operation is best described by splitting the circuit into two parts. The circuit at the right is the zero-point switch (previously described in the section on zero-point switching) and its operation is unchanged. The circuit to the left is the proportional control for the zero-point switch.

The operation of the control circuit is as follows: diode D1, resistor R1, capacitor C1 and the load on C1 establish the dc supply voltage for the control circuit. Temperature

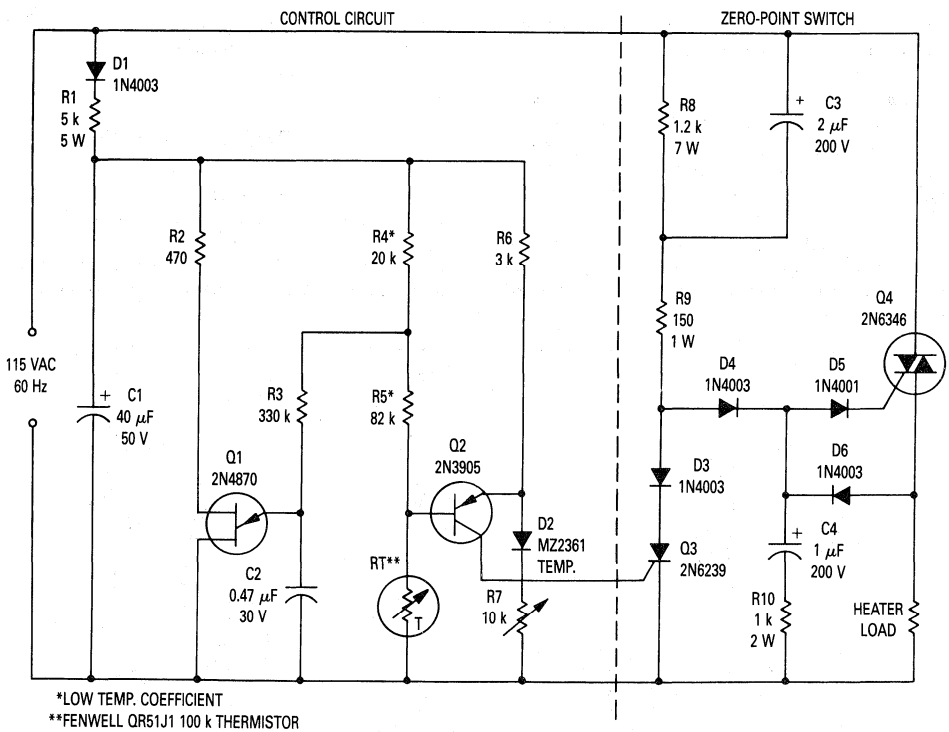


Figure 6.106. Temperature Controller Using Zero-Point Switching

is sensed by thermistor R_T, which is part of the bridge circuit consisting of R4, R5, R6, R7, D2 and R_T. The detector for the bridge is transistor Q2. R7 is set so the bridge is in balance at the desired temperature. As the temperature increases, R_T decreases, Q2 turns on and provides gate drive to SCR Q3. Q3 turns on and shunts the gate signal away from the TRIAC Q4. Q4 shuts off and removes power to the load. Now, as the temperature drops, R_T increases and Q2 turns off, SCR Q3 turns off, and full-wave power is applied to the load. Normally, the circuit would continue to cycle randomly, providing groups of full power to the heater load. However, modulation is applied to proportion the load power in response to small changes in R_T. The modulation is achieved by superimposing a sawtooth voltage on one arm of the bridge through R3. The period of the sawtooth is set to equal 12 cycles of the line frequency. From one to all 12 cycles can be applied to the load, thus allowing the load power to modulate in 8% steps from 0% to 100% duty cycle. The sawtooth voltage is generated by the unijunction transistor relaxation oscillator consisting of R2, R3, R4, C2 and Q1. The sawtooth wave modulates the bridge voltage so that over a portion of the twelve-cycle group the bridge voltage will be above the null point, and over the other portion it will be below the null point. This action divides each twelve-cycle group into an on portion and an off portion, the proportioning depending upon the amount R_T has varied from the nominal value. This circuit provides excellent control of a resistance heater as it will tend to stabilize and apply the correct amount of power on a continuous basis at a steady-state duty cycle depending on the load requirements. The temperature is therefore controlled over a very narrow range and no EMI is generated.

TRIAC RELAY-CONTACT PROTECTION

A common problem in contact switching high current is arcing which causes erosion of the contacts. A solution to this problem is illustrated in Figure 6.107. This circuit can be used to prevent relay contact arcing for loads up to 50 amperes.

There is some delay between the time a relay coil is energized and the time the contacts close. There is also a delay between the time the coil is de-energized and the time the contacts open. For the relay used in this circuit both times are about 15 ms. The TRIAC across the relay contacts will turn on as soon as sufficient gate current is present to fire it. This occurs after switch S1 is closed but before the relay contacts close. When the contacts close, the load current passes through them, rather than through the TRIAC, even though the TRIAC is receiving gate current. If S1 should be closed during the negative half cycle of the ac line, the TRIAC will not turn on immediately but will wait until the voltage begins to go positive, at which time diode D1 conducts providing gate

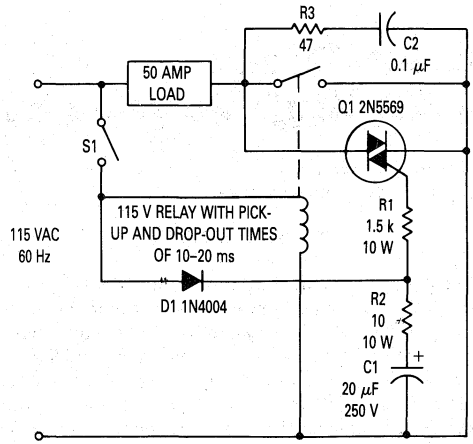


Figure 6.107. TRIAC Prevents Relay Contact Arcing

current through R1. The maximum time that could elapse before the TRIAC turns on is 8-1/3 ms for the 60 Hz supply. This is adequate to ensure that the TRIAC will be on before the relay contact closes. During the positive half cycle, capacitor C1 is charged through D1 and R2. This stores energy in the capacitor so that it can be used to keep the TRIAC on after switch S1 has been opened. The time constant of R1 plus R2 and C1 is set so that sufficient gate current is present at the time of relay drop-out after the opening of S1, to assure that the TRIAC will still be on. For the relay used, this time is 15 ms. The TRIAC therefore limits the maximum voltage, across the relay contacts upon dropout to the TRIAC's voltage drop of about 1 volt. The TRIAC will conduct until its gate current falls below the threshold level, after which it will turn off when the anode current goes to zero. The TRIAC will conduct for several cycles after the relay contacts open.

This circuit not only reduces contact bounce and arcing but also reduces the physical size of the relay. Since the relay is not required to interrupt the load current, its rating can be based on two factors: the first is the rms rating of the current-carrying metal, and the second is the contact area. This means that many well-designed 5 ampere relays can be used in a 50 ampere load circuit. Because the size of the relay has been reduced, so will the noise on closing. Another advantage of this circuit is that the life of the relay will be increased since it will not be subjected to contact burning, welding, etc.

The RC circuit shown across the contact and TRIAC (R3 and C2) is to reduce dv/dt if any other switching element is used in the line.

Series Triacs In AC High Voltage Switching Circuits

By George Templeton
Thyristor Applications Engineer

1

INTRODUCTION

This paper describes the series connection of triacs to create a high voltage switch suitable for operation at voltages up to 2000 Volts. They can replace electromechanical contactors or extend their current rating and lifetime. Motor starters and controllers operating at line voltages of 240 Volts or more require high-voltage switches. Transformer action and resonant snubber charging result in voltages much greater than the peak of the line. Triacs can be subjected to both commutating and static dV/dt when multiple switching devices are present in the circuit. Snubber designs to prevent static dV/dt turn-on result in higher voltages at turn-off. Variable load impedances also raise voltage requirements.

The benefits of series operation include: higher blocking voltage, reduced leakage, better thermal stability, higher dV/dt capability, reduced snubber costs, possible snubberless operation, and greater latitude in snubber design. The advantages of triacs as replacements for relays include:

- Small size and light weight.
- Safety — freedom from arcing and spark initiated explosions.
- Long lifespan — contact bounce and burning eliminated.
- Fast operation — turn-on in microseconds and turn-off in milliseconds.
- Quiet operation.

Triacs can be used to replace the centrifugal switch in capacitor start motors. The blocking voltage required of the triac can be much greater than the line voltage would suggest. It must block the vector sum of the line, auxiliary winding, and start capacitor voltage. This voltage increases when triac turn-off occurs at higher rpm.

TRIGGERING

Figure 1 illustrates a series thyristor switching circuit. In this circuit, the top triac triggers in Quadrant 1 when the bottom triac triggers in Quadrant 3. When the optocoupler turns on, gate current flows until the triacs latch. At that time, the voltage between the gate terminals drops to about 0.6 Volts stopping the gate current. This process repeats each half cycle. The power rating of the gate resistor can be small because of the short duration of the gate current. Optocoupler surge or triac gate ratings determine the minimum resistance value. For example, when the maximum optocoupler I_{TSM} rating is 1 A:

$$R_g >= V_{peak}/I_{max} \quad (1.0)$$

$$R_g = 750\ V/1\ A = 750\ Ohm$$

The triacs retrigger every half cycle as soon as the line voltage rises to the value necessary to force the trigger current. The instantaneous line voltage V is

$$V = I_{GT} R_g + 2 V_{GT} + 2 V_{TM} \quad (1.1)$$

where V_{GT}, I_{GT} are data book specifications for the triac and V_{TM} is the on-voltage specification for the optocoupler.

The phase delay angle is

$$\theta_d = SIN^{-1}\left[\frac{V}{\sqrt{2}\ V_{LINE}}\right] \quad (1.2)$$

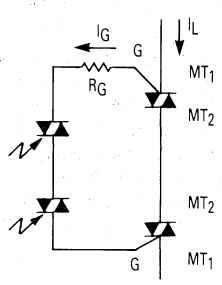


Figure 1. Series Switch

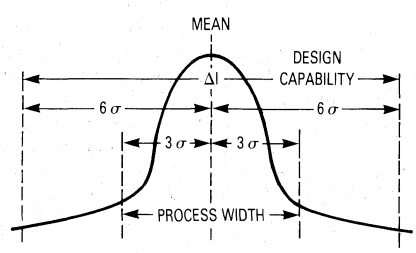


Figure 2. Designing for Probable Leakage

STATIC VOLTAGE SHARING

Maximum blocking voltage capability results when the triacs share voltage equally. The blocking voltage can be dc or ac. A combination of both results when the triac switches the start winding in capacitor start motors. In the simple series connection, both triacs operate with an identical leakage current which is less than that of either part operated alone at the same voltage. The voltages across the devices are the same only when their leakage resistances are identical. Dividing the voltage by the leakage current gives the leakage resistance. It can range from 200 kohm to 2000 megohm depending on device characteristics, temperature, and applied voltage.

Drawing a line corresponding to the measured series leakage on each device's characteristic curve locates its operating point. Figure 3a shows the highest and lowest leakage units from a sample of 100 units. At room temperature, a leakage of 350 nA results at 920 Volts. The lowest leakage unit blocks at the maximum specified value of 600 Volts, while the highest blocks 320 Volts. A 50 percent boost results.

Figure 3b shows the same two triacs at rated $T_J$max. The magnitude of their leakage increased by a factor of about 1000. Matching between the devices improved, allowing operation to 1100 Volts without exceeding the 600 Volt rating of either device.

Identical case temperatures are necessary to achieve good matching. Mounting the devices closely together on a common heatsink helps.

A stable blocking condition for operation of a single triac with no other components on the heatsink results when

$$\frac{dI_{MT}}{dT_J} \cdot \frac{dT_J}{dP_J} \cdot \frac{dP_J}{dI_{MT}} < 1 \qquad (2.0)$$

Thermal run-away is a regenerative process which occurs whenever the loop gain in the thermal feedback circuit reaches unity. An increase in junction temperature causes increased leakage current and higher power dissipation. Higher power causes higher junction temperatures which in turn leads to greater leakage. If the rate of heat release at the junction exceeds the rate of removal as temperature increases, this process repeats until the leakage current is sufficient to trigger the thyristor on.

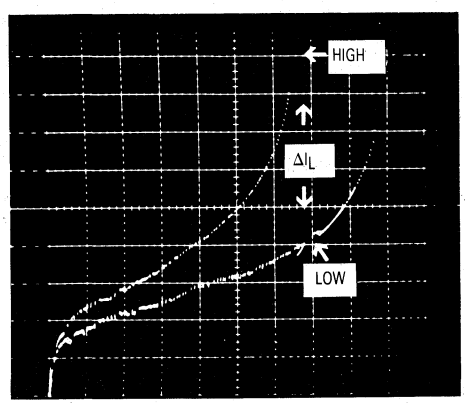
DC blocking simplifies analysis. A design providing stable dc operation guarantees ac performance. AC operation allows smaller heatsinks.

The last term in the stability equation is the applied voltage when the load resistance is low and the leakage causes negligible voltage drop across it. The second term is the thermal resistance from junction to ambient. The first term describes the behavior of leakage at the operating conditions. For example, if leakage doubles every 10°C, a triac operating with 2 mA of leakage at 800 Vdc with a 6°C/W thermal resistance is stable because

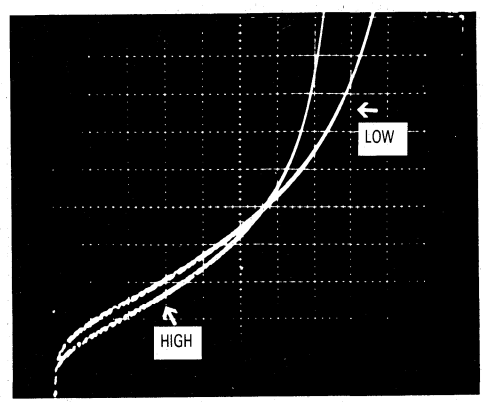
$$\frac{2\text{ mA}}{10°\text{C}} \cdot \frac{6°\text{C}}{\text{W}} \cdot 800\text{ V} = 0.96$$

Operating two triacs in series improves thermal stability. When two devices have matched leakages, each device sees half the voltage and current or 1/4 of the power in a single triac. The total leakage dissipation will approach half that of a single device operated at the same voltage. The additional voltage margin resulting from the higher total blocking voltage reduces the chance that either device will operate near its breakdown voltage where the leakage current increases rapidly with small increments in voltage. Higher voltage devices have lower leakage currents when operated near breakdown. Consequently, the highest breakover voltage unit in the pair will carry the greatest proportion of the burden. If the leakage current is large enough to cause significant changes in junction temperature, ($\Delta T_J = \phi_{JC} P_D$), the effect will tend to balance the voltage division between the two by lowering the leakage resistance of the hotter unit. If the leakage mismatch between the two is large, nearly all the voltage will drop across one device. As a result there will be little benefit connecting two in series.

Series blocking voltage depends on leakage matching. Blocking stability depends on predictable changes in leakage with temperature. Leakage has three components.



(a) 100 V/ 100 nA/ 25°C



(b) 100 V/ 100 μA/ 125°C

Figure 3. Leakage Matching versus Temperature

Surface Leakage

Passivation technique, junction design, and cleanliness determine the size of this component. It tends to be small and not very dependent on temperature.

Diffusion Leakage

Measurements with 1 volt reverse bias show that this component is less than 10 percent of the total leakage for allowed junction temperatures. It follows an equation of the form:

$$I \propto e^{-(qv/kT)} \qquad (2.1)$$

and doubles about every 10°C. Its value can be estimated by extrapolating backward from high temperature data points.

Depletion Layer Charge Generation

This component is a result of carriers liberated from within the blocking junction depletion layer. It grows with the square root of the applied voltage. The slope of the leakage versus applied voltage is the mechanism allowing for series operation with less than perfect leakage matching. Predictable diffusion processes determine this leakage. At temperatures beween 70 and 150°C it is given by:

$$i \propto e - \frac{E}{kT} \qquad (2.2)$$

where E = 1.1 eV, k = 8.62E − 5 eV/k, T = degrees Kelvin, and k = 8.62×10^{-5} eV/k.

It is useful to calculate the percentage change in leakage current with temperature:

$$A = \frac{1}{i}\frac{di}{dT_J} = \frac{E}{kT^2} = 0.08 = \frac{8\%}{°C}$$

The coefficient A was evaluated on 3 different die size triacs by curve fitting to leakage measurements every 10° from 70 to 150°C. Actual values measured 0.064 at 125° and 0.057 at 150°.

Deviations from this behavior will result at voltages and temperatures where leakage magnitude, current gain, and avalanche multiplication aid unwanted turn-on. Sensitive gate triacs are not recommended for this reason.

DERATING AND LEAKAGE MATCHING

Operation near breakdown increases leakage mismatch because of the effects of avalanche multiplication.

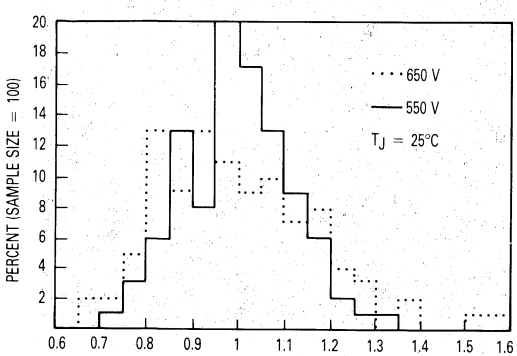


Figure 4. Normalized Leakage (Mean = 1.0)

For series operation, devices should be operated at least 100 Volts below their rating.

Figure 4 shows the leakage histogram for a triac sample operated at two different voltages. The skewedness in the high-voltage distribution is a consequence of some of the sample operating near breakdown.

HEATSINK SELECTION

Solving equations (2.0) and (2.3) for the thermal resistance required to prevent runaway gives:

$$\theta_{JA} < \frac{1}{A \cdot V \cdot i} \qquad (3.0)$$

where θ_{JA} is thermal resistance, junction to ambient, in °C/W, A = 0.08 at T_J = 125°C, V = rated V_{DRM}, and i = rated I_{DRM}.

θ_{JA} must be low enough to remove the heat resulting from conduction losses and insure blocking stability. The latter can be the limiting factor when circuit voltages are high. For example, consider a triac operated at 8 amps (rms) and 8 Watts. The allowed case temperature rise at 25° ambient is 85°C giving a required θ_{CA} (thermal resistance, case to ambient) of 10.6°C/W. Allowing 1°C/W for θ_{CHS} (thermal resistance, case to heatsink) leaves 9.6°C/W for θ_{SA} (thermal resistance, heatsink to ambient). However, thermal stability at 600 V and 2 mA I_{DRM} requires θ_{JA} = 10.4°C/W. A heatsink with θ_{SA} less than 7.4°C/W is needed, given a junction to case thermal resistance of 2°C/W.

The operation of devices in series does not change the coefficient A. When matching and thermal tracking is perfect, both devices block half the voltage. The leakage current and power divide by half and the allowed θ_{JA} for blocking stability increases by 4.

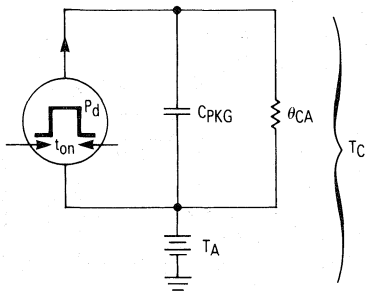
Low duty cycles allow the reduction of the heatsink size. The thermal capacitance of the heatsink keeps the junction temperature within specification. The package time constant (C_{pkg} $R_{\theta JA}$) is long in comparison with the thermal response time of the die, causing the instantaneous T_J to rise above the case as it would were the semiconductor mounted on an infinite heatsink. Heatsink design requires estimation of the peak case temperature and the use of the thermal derating curves on the data sheet. The simplest model applies to a very small heatsink which could be the semiconductor package itself. When θ_{SA} is large in comparision with θ_{CHS}, it is sufficient to lump both the package and heatsink capacitances together and treat them as a single quantity. The models provide good results when the heatsink is small and the thermal paths are short.

Model C, Figure 5 is a useful simplification for low duty cycle applications. Increasing heatsink mass adds thermal capacitance and reduces peak junction temperature. Heatsink thermal resistance is proportional to surface area and determines the average temperature.

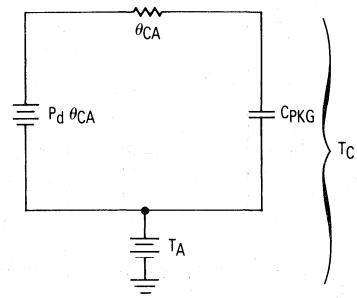
$$\theta_{SA} = 32.6\ A^{(-0.47)} \qquad (3.1)$$

where A = total surface area in square inches, θ_{SA} = thermal resistance sink to ambient in °C/W.

Analysis of heatsink thermal response to a train of periodic pulses can be treated using the methods in Motorola application note AN569 and Figure 6. For example:



(a.) Standard Thermal Analogue For a Thyristor in Free Air



(b.) Equivalent Circuit For (a)

In Circuit (B):

The steady state case temperature is given by

(5.0) $$T_{CSS} = P_d\,\theta_{CA} + T_A \text{ in } ^\circ C$$

where P_d = Applied average power, watts
θ_{CA} = Case to ambient thermal resistance, °C/W
T_A = ambient temperature, °C

The package rises toward the steady state temperature exponentially with time constant

(5.1) $$\tau = \theta_{CA}\,C_{PKG}, \text{ seconds}$$

where C_{pkg} = HM, Joules/°C
H = Specific heat, calories/(gm • °C)
M = Mass in grams

and 1 Calorie = 4.184 Joule
1 Joule = 1 Watt • Sec

The case temperature rise above ambient at the end of power pulse is:

(5.2) $$\Delta T_{C_{pk}} = \Delta T_{CSS}\,(1 - e^{-t_{on}/\tau})$$

where $\Delta T_{C_{pk}} = T_{C_{pk}} - T_A$
$\Delta T_{CSS} = T_{CSS} - T_A$

To account for thermal capacity, a time dependent factor r(t) is applied to the steady state case-to-ambient thermal resistance. The package thermal resistance, at a given on-time, is called transient thermal resistance and is given by:

$$R_{\theta CA}(t_{on}) = r(t_{on})\,\theta_{CA}$$

where $r(t_{on})$ = Unitless transient thermal impedance coefficient.

In terms of measurable temperatures:

(5.3) $$r(t_{on}) = \frac{\Delta T_{C_{pk}}}{\Delta T_{CSS}}$$

In model (b.) this is

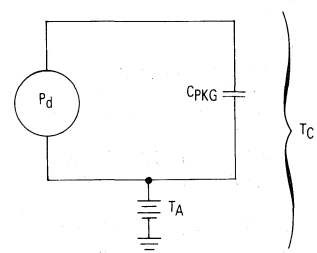
(5.4) $$r(t_{on}) = (1 - e^{-t_{on}/\tau})$$

Solving 5-4 for the package capacitance gives

(5.5) $$C_{PKG} = \frac{-t_{on}}{(\theta_{CA} \ln(1 - r(t_{on}))}$$

Use simplified model C when

$t_{on} << \tau$
$\Delta T_{C_{pk}} << \Delta T_{CSS}$



(c.) Simplified Model

(5.6) $$T_C = \frac{P_d\,t_{on}}{C_{PKG}} + T_A$$

Figure 5. Transient Thermal Response For a Single Power Pulse

Assume the case temperature changes by 40°C for a single power pulse of 66.67 W and 3 s duration. Then from equation (5.6):

$$C_{pkg} = \frac{(66.7 \text{ Watts})\,(3 \text{ seconds})}{40^\circ C} = \frac{5 \text{ Joules}}{^\circ C}$$

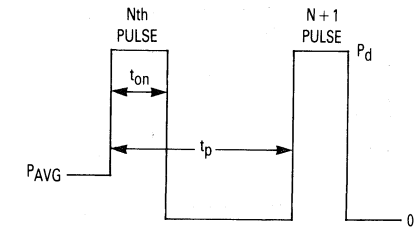
The heatsink thermal resistance can be determined by applying dc power, measuring the final case temperature, and using equation (5.0).

$$\frac{T_C - T_A}{P_D} = \frac{175\text{-}25}{5} = 30^\circ C/W$$

The application requires a 3 s on-time and 180 s period at 66.7 W. Then

$$P_{avg} = (66.7 \text{ W})\,(3/180) = 1.111 \text{ W}$$

Using equation (5.3), the theoretical steady state case temperature rise is:



$\Delta T_C (N+1) = [P_{AVG} + (P_d - P_{AVG})\, r(t_{on}+t_p) + P_d\, r(t_{on}) - P_d\, r(t_p)]\theta CA$

where $\Delta T_C (N+1)$ = maximum rise above ambient
P_d = applied average power within a pulse
P_{AVG} = average power within a period
$r(t_{on}+t_p)$ = time dependent factor for sum of t_{on} and t_p
$r(t_{on})$ = time dependent factor for t_{on}
$r(t_p)$ = time dependent factor for t_p

Figure 6. Steady State Peak Case Temperature Rise

$$T_{CSS} - T_A = (66.7\ W)\ (30°C/W) = 2000°C$$

and

$$R(t_{on}) = R\ (3\ s) = (40°C \text{ measured rise})/2000 = 0.02$$

From equation (5.4) and (5.1):

$$R\ (T_p) = (1 - e^{-180/150}) = .6988$$
$$R\ (t_{on}+T_p) = (1 - 1^{-183/150}) = .7047$$

Then from Figure 6:

$$\text{delta } T_C = (1.111 + 46.225 + 1.333 - 46.61)\ 30 = 61.8°C$$

If the ambient temperature is 25°C, T_C = 87°C.

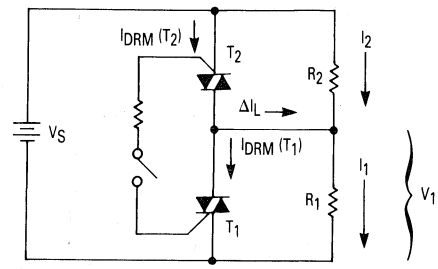
COMPENSATING FOR MAXIMUM SPECIFIED LEAKAGE

Identical value parallel resistors around each triac will prevent breakdown resulting from mismatched leakages. Figure 7 derives the method for selecting the maximum allowed resistor size. A worst case design assumes that the series pair will operate at maximum T_J and that one of the triacs leaks at the full specified value while the other has no leakage at all. A conservative design results when the tolerances in the shunt resistors place the highest possible resistor across the low leakage unit and the lowest possible resistor around the high leakage unit.

This method does not necessarily provide equal voltage balancing. It prevents triac breakover. Perfect voltage sharing requires expensive high-wattage resistors to provide large bleeder currents.

COMPENSATION FOR PROBABLE LEAKAGE

Real triacs have a leakage current greater than zero and less than the specified value. Knowledge of the leakage distribution can be used to reduce resistor power requirements. The first step is to statistically characterize the product at maximum temperature. Careful control of the temperature is critical because leakage depends strongly on it.



$$V_1 = \frac{V_S R_1}{R_1 + R_2} + \frac{\Delta I_L R_1 R_2}{R_1 + R_2}$$

Let $R_1 = R\ (1+p)$ and $R_2 = R\ (1-p)$ where
R = Nominal resistor value
p = 0.05 for 5% tolerance, etc.

$$R \leq \frac{2\ V_{DRM} - V_S\ (1+p)}{\Delta I_L\ (1-p^2)}$$

Worst case becomes:

and I_{DRM} (T_1) = 0; I_{DRM} (T_2) = Spec. max. value
ΔI_L = Spec. Max. Value

Figure 7. Maximum Allowed Resistor for Static Voltage Sharing

The process width is the leakage span at plus or minus 3 standard deviations (sigma) from the mean. To minimize the probability of out of spec parts, use a design capability index (Cp) of 2.0.

$$Cp = (\text{design } \Delta I)/(\text{process width}) \qquad (4.0)$$
$$Cp = (12 \text{ sigma})/(6 \text{ sigma})$$

Figure 2 and Figure 7 describe this. Substituting delta I_L at 6 sigma in Figure 7 gives the resistor value. The required power drops by about 4.

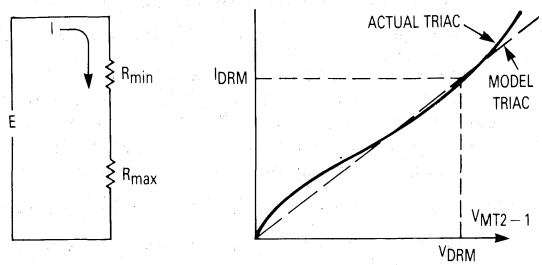
Theoretically there would be no more than 3.4 triacs per million exceeding the design tolerance even if the mean value of the leakage shifted by plus or minus 1.5 sigma.

SELECTING RESISTORS

Small resistors have low voltage ratings which can impose a lower constraint on maximum voltage than the triac. A common voltage rating for carbon resistors is:

Rated Power (W)	Maximum Voltage (V)
1/4 Watt	250 Volts
1/2	350
1	500
2	750

Series resistors are used for higher voltage.



(a) Equivalent Circuit (b) Model

$$\text{Let } V_{DRM} = E\,\frac{R_{max}}{R_{max} + R_{min}}$$

$$E = V_{DRM}\left(1 + \frac{I_{min}}{I_{max}}\right)$$

$$R_{max} = \frac{V_{DRM}}{I_{min}} \qquad R_{min} = \frac{V_{DRM}}{I_{max}} \qquad (8.0)$$

Figure 8. Maximum Voltage Sharing Without Shunt Resistor

OPERATION WITHOUT RESISTORS

Figure 8 derives the method for calculating maximum operating voltage. The voltage boost depends on the values of I_{min} and I_{max}. For example:

$$\left(1 + \frac{131\ \mu A}{683\ \mu A}\right) = 1.19$$

A 19 percent voltage boost is possible with the 6 sigma design. Testing to the measured maximum and minimum of the sample allows the boost to approach the values given in Table 1.

$$(1 + 0.835/1.228) = 1.68$$

Table 1. Normalized leakage and voltage boost factor. (Mean = 1.0)

Voltage (V)	550	650	550	550	550	550	550
T_J (°C)	25	25	100	125	125	150	150
R_{shunt}	—	—	—	—	1.5M	1.5M	510K
Sample Size	100	100	16	16	16	16	16
Maximum	1.315	1.591	1.187	1.228	1.123	1.346	1.186
Minimum	0.729	0.681	0.840	0.835	0.920	0.820	0.877
Sigma	0.116	0.172	0.106	0.113	0.055	0.132	0.084
Sample Boost	1.55	1.43	1.71	1.68	1.82	1.61	1.74
6 Sigma Boost	1.18	1.00	1.22	1.19	1.50	1.12	1.33

COMPENSATING FOR SURFACE LEAKAGE

A small low power shunt resistance will provide nearly perfect low temperature voltage sharing and will improve high temperature performance. It defines the minimum leakage current of the parallel triac-resistor combination. The design method in Figure 8 can be used by adding the resistor current to the measured maximum and minimum leakage currents of the triac sample. This is described in Table 1.

SERIES $\left(\frac{dV}{dt}\right)_s$

The series connection will provide twice the $\left(\frac{dV}{dt}\right)_s$ capability of the lowest device in the pair (Figure 9).

Dynamic matching without a snubber network depends on equality of the thyristor self capacitance. There is little variation in junction capacitance. Device gain variations introduce most of the spread in triac performance.

The blocking junction capacitance of a thyristor is a declining function of dc bias voltage. Mismatch in static blocking voltage will contribute to unequal capacitances. However, this effect is small at voltages beyond a few volts. The attachment of a heatsink at the high-impedance node formed by connection of the triac main-terminals can also contribute to imbalance by introducing stray capacitance to ground. This can be made insignificant by adding small capacitors in parallel with the triacs. Snubbers will serve the same purpose.

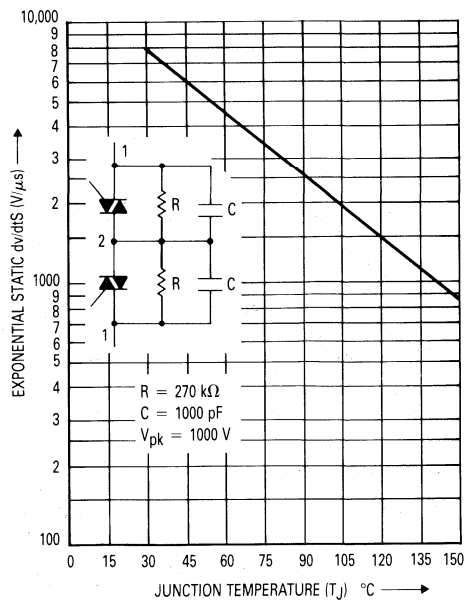


Figure 9. Exponential Static dV/dt, Series MAC15-4 Triacs

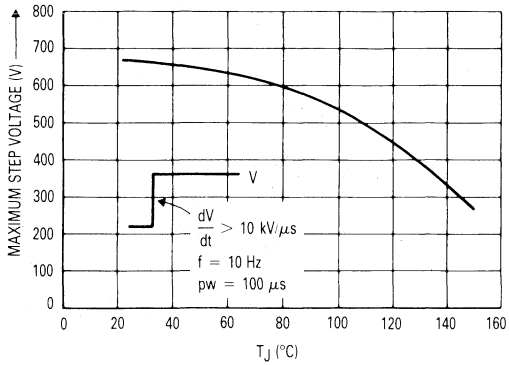


Figure 10. Step Blocking Voltage V_S T_J (Unsnubbed Series Triacs)

Triacs can tolerate very high rates of voltage rise when the peak voltage magnitude is below the threshold needed to trigger the device on. This behavior is a consequence of the voltage divider action between the device collector and gate-cathode junction capacitances. If the rise-time is made short in comparison with minority carrier lifetime, voltage and displaced charge determine whether the device triggers on or not. Series operation will extend the range of voltage and load conditions where a static $\frac{dV}{dt}$ snubber is not needed.

Figure 10 graphs the results of measurements on two series connected triacs operated without snubbers. The series connection doubled the allowed step voltage. However, this voltage remained far below the combined 1200 V breakover voltage of the pair.

Exponential $\left(\frac{dV}{dt}\right)_s$ tests performed at 1000 V and less than 2 kV/μs showed that turn-on of the series pair can occur because of breakdown or $\frac{dV}{dt}$. The former was the limiting factor at junction temperatures below 100°C. Performance improved with temperature because device gain aided voltage sharing. The triac with the highest current gain in the pair is most likely to turn-on. However, this device has the largest effective capacitance. Consequently it is exposed to less voltage and $\frac{dV}{dt}$. At higher temperatures, rate effects dominated over voltage magnitudes, and the capability of the series pair fell. $\frac{dV}{dt}$ performance of the series devices was always better than that of a single triac alone.

TURNOFF

Process tolerances cause small variations in triac turnoff time. Series operation will allow most of the reapplied blocking voltage to appear across the faster triac when a dynamic voltage sharing network is not used.

Figure 11 describes the circuit used to investigate this behavior. It is a capacitor discharge circuit with the load series resonant at 60 Hz. This method of testing is desirable because of the reduced burn and shock hazard resulting from the limited energy storage in the load capacitor.

The triacs were mounted on a temperature controlled hotplate. The single pulse non-repetitive test aids junction temperature control and allows the use of lower power rated components in the snubber and load circuit.

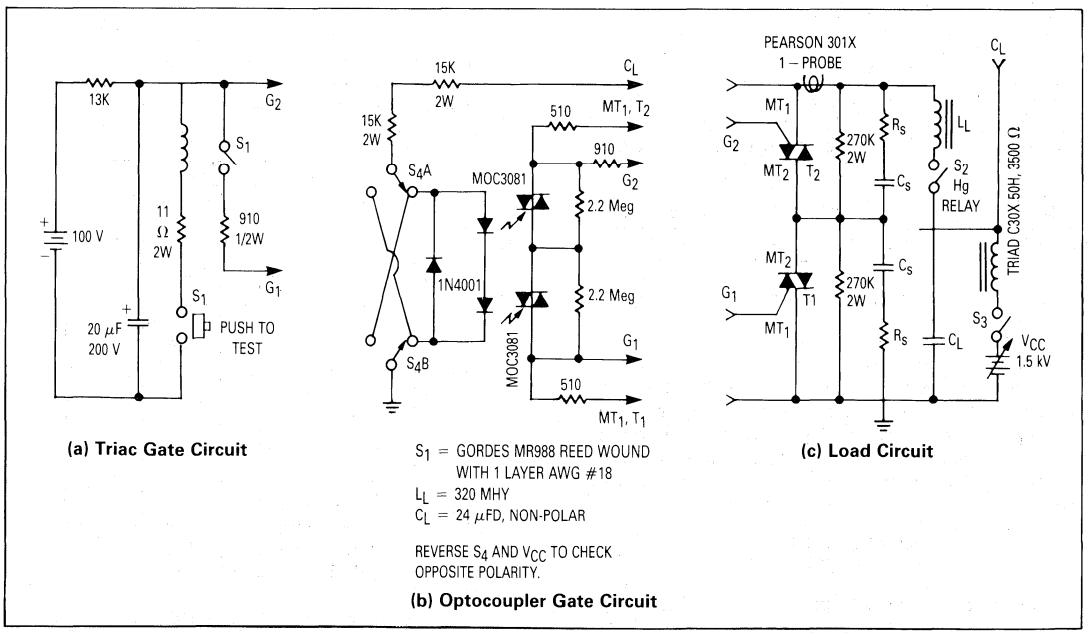


Figure 11. $\left(\frac{dV}{dt}\right)_c$ Test Circuit

Snubberless turn-off at 1200 V and 320 milli-henry resulted in 800 V peak and 100 V/μs. Although this test exceeded the ratings of the triacs, they turned off successfully.

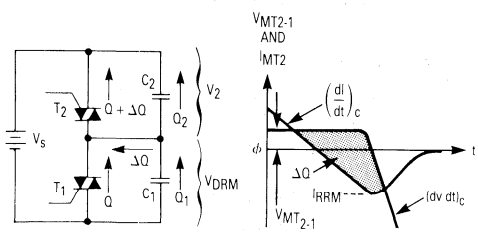
Snubberless operation is allowable when:

1. The total transient voltage across both triacs does not exceed the rating for a single device. This voltage depends on the load phase angle, self capacitance of the load and triac, damping constant, and natural resonance of the circuit.
2. The total $\left(\frac{dV}{dt}\right)_c$ across the series combination does not exceed the capability of a single device.

Maximum turn-off voltage capability and tolerance for variable loads requires the use of a snubber network to provide equal dynamic voltage sharing. Figure 12 and Figure 13 derives the minimum size snubber capacitor allowed. It is determined by the recovery charge of the triac. Measurements in fast current crossing applications suggest that the reverse recovery charge is less than 2 micro-coulombs. Recovery currents cannot be much greater than I_H or I_{GT}, or the triac would never turn-off. Recovery can be forward, reverse, or near zero current depending on conditions.

Snubber design for the series switch has the following objectives:

- Controlling the voltage peak. Resonant charging will magnify the turn-off voltage.
- Controlling the voltage rate. Peak voltage trades with voltage rate.



Worst case:

$C_2 = C(1+p)$; $C_1 = C(1-p)$; $Q_1 = 0$; $Q_2 = \Delta Q$

where C = Nominal value of capacitor
and p = 0.1 for 10% tolerance, etc.
ΔQ = Reverse recovery charge

Note that T_1 has no charge while T_2 carries full recovery charge.

For the model shown above,

$$V_S = \frac{Q_1}{C_1} + \frac{Q_2}{C_2} = \frac{Q_1}{C(1-p)} + \frac{Q_1 - \Delta Q}{C(1+p)}$$

$$C \geq \frac{\Delta Q}{2\,V_{DRM} - V_S(1+p)}$$

Figure 12. Minimum Capacitor Size for Dynamic Voltage Sharing

- Equalizing the voltage across the series devices by providing for imbalance in turn-off charge.

Designs that satisfy the first two objectives will usually provide capacitor values above the minimum size. Select the snubber for a satisfactory compromise between voltage and $\frac{dV}{dt}$. Then check the capacitor to insure that it is sufficiently large.

Snubber designs for static, commutating, and combined $\frac{dV}{dt}$ stress are shown in Table 2. Circuits switching the line or a charged capacitor across a blocking triac require the addition of a series snubber inductor. The snubber must be designed for maximum $\frac{dV}{dt}$ with the minimum circuit inductance. This contraint increases the required triac blocking voltage.

Table 2. Snubber Designs

Type	$\left(\frac{dV}{dt}\right)_c$	$\left(\frac{dV}{dt}\right)_s$	Both
L (mh)	320	0.4	320
R_L Ohm	8	0	8
R_S Ohm	1820	48	48
C_S (μf)	0.5	0.5	0.5
Damping Ratio	1.14	0.85	.035
V_{step} (V)	1200	1200	750
V_{pk} (V)	1332	1400	1423
t_{pk} (μs)	768	29.8	1230
$\frac{dV}{dt}$ (V/μs)	4.6	103	1.3

Note: Divide R_S and $\frac{dV}{dt}$ by 2, multiply C_S by 2 for each triac.

$\frac{dI}{dt}$ CAPABILITY

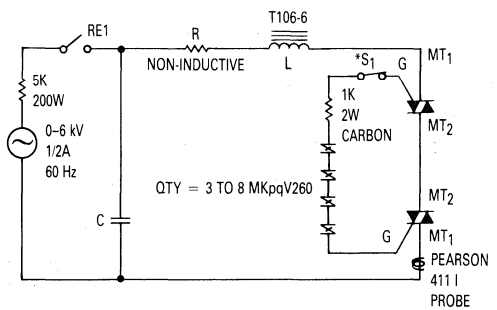
The hazard of thyristor damage by $\frac{dI}{dt}$ overstress is greater when circuit operating voltages are high because $\frac{dI}{dt}$ is proportional to voltage. Damage by short duration transients is possible even though the pulse is undetectable when observed with non-storage oscilloscopes. This type of damage can be consequence of snubber design, transients, or parasitic capacitances.

A thyristor can be triggered on by gate current, exceeding its breakdown voltage, or by exceeding its $\left(\frac{dV}{dt}\right)_s$ capability. In the latter case, a trigger current is generated by charging of the internal depletion layer capacitance in the device. This effect aids turn-on current spreading, although damage can still occur if the rate of follow on $\frac{dI}{dt}$ is high. Repetitive operation off the ac line at voltages above breakdown is a worst case condition. Quadrant 3 has a slightly slower gated turn-on time, increasing the

chance of damage in this direction. Higher operating voltages raise power density and local heating, increasing the possibility of die damage due to hot-spots and thermal run-away.

Ideally, turn-on speed mismatch should not be allowed to force the slower thyristor into breakdown. An RC snubber across each thyristor prevents this. In the worst case, one device turns on instantly while the other switches at the slowest possible turn-on time. The rate of voltage rise at the slower device is roughly $\frac{dV}{dt} = \frac{V_I R_S}{2L}$, where V_I is the maximum voltage across L. This rate should not allow the voltage to exceed V_{DRM} in less than T_{gt} to prevent breakover. But what if the thyristors are operated without a snubber, or if avalanche occurs because of a transient overvoltage condition?

The circuit in Figure 13 was constructed to investigate this behavior. The capacitor, resistor, and inductor create a pulse forming network to shape the current wave. The initial voltage on the capacitor was set by a series string of sidac bidirectional breakover devices.



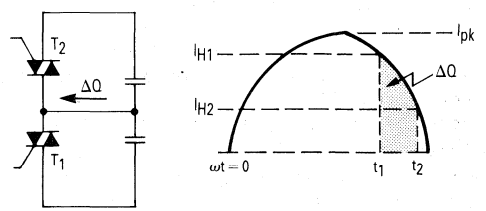
Vci V	C μFD	L μHY	R Ω	dI/dt A/μs	Rejects Tested
1000	4.06	3.4	5.7	100	0/100
1900*	1.05	7.9	5.7	179	0/195
1500	0.002	0.3	10	3000	3/10

*Open S_1 to test breakover dI/dt

Figure 13. dI/dt Test Circuit

Test results showed that operation of the triac switch was safe as long as the rate of current rise was below 200 A/μs. This was true even when the devices turned on because of breakover. However, a 0.002 μf capacitor with no series limiting impedance was sufficient to cause damage in the Q3 firing polarity.

Circuit malfunctions because of breakover will be temporary if the triac is not damaged. Test results suggest that there will be no damage when the series inductance is sufficient to hold dI/dt to acceptable values. Highly energetic transients such as those resulting from lightning strikes can cause damage to the thyristor by I^2t surge overstress. Device survival requires the use of voltage limiting devices in the circuit and $\frac{dV}{dt}$ limiting snubbers to prevent unwanted turn-on. Alternatively, a large triac capable of surviving the surge can be used.



ΔQ for turn-off at I_H

$$\Delta Q = \int_{t_1}^{t_2} I_{pk} \, \mathrm{SIN}\omega t \, dt = \frac{I_{pk}}{\omega}(\cos \omega t_1 - \cos \omega t_2)$$

$$I_{H1} = I_{pk} \, \mathrm{Sin}\omega t_1$$

$$\text{thus } t_1 = \frac{1}{\omega} \mathrm{Sin}^{-1} \frac{I_{H1}}{I_{pk}}$$

Worst case: $I_{H2} = 0$; $\phi_2 = \omega t_2 = \pi$

$$\Delta Q = \frac{I_{pk}}{\omega}\left(1 + \cos\left[\mathrm{SIN}^{-1} \frac{I_{H1}}{I_{pk}}\right]\right)$$

$$\Delta Q = \frac{I_{pk}}{\omega}\left(1 + \sqrt{1 - \left(\frac{I_{H1}}{I_{pk}}\right)^2}\right)$$

Figure 14. Forward Recovery Charge for Turn-Off at I_H

RC Snubber Networks For Thyristor Power Control and Transient Suppression

By George Templeton
Thyristor Applications Engineer

INTRODUCTION

RC networks are used to control voltage transients that could falsely turn-on a thyristor. These networks are called snubbers.

The simple snubber consists of a series resistor and capacitor placed around the thyristor. These components along with the load inductance form a series CRL circuit. Snubber theory follows from the solution of the circuit's differential equation.

Many RC combinations are capable of providing acceptable performance. However, improperly used snubbers can cause unreliable circuit operation and damage to the semiconductor device.

Both turn-on and turn-off protection may be necessary for reliability. Sometimes the thyristor must function with a range of load values. The type of thyristors used, circuit configuration, and load characteristics are influential.

Snubber design involves compromises. They include cost, voltage rate, peak voltage, and turn-on stress. Practical solutions depend on device and circuit physics.

STATIC $\frac{dV}{dt}$

WHAT IS STATIC $\frac{dV}{dt}$?

Static $\frac{dV}{dt}$ is a measure of the ability of a thyristor to retain a blocking state under the influence of a voltage transient.

$\left(\frac{dV}{dt}\right)_s$ DEVICE PHYSICS

Static $\frac{dV}{dt}$ turn-on is a consequence of the Miller effect and regeneration (Figure 1). A change in voltage across the junction capacitance induces a current through it. This current is proportional to the rate of voltage change $\left(\frac{dV}{dt}\right)$. It triggers the device on when it becomes large enough to raise the sum of the NPN and PNP transistor alphas to unity.

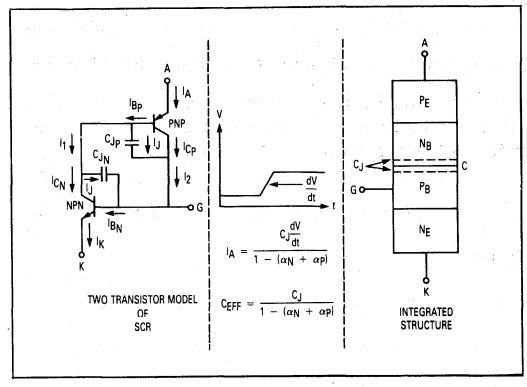


Figure 1. $\left(\frac{dV}{dt}\right)_s$ Model

CONDITIONS INFLUENCING $\left(\frac{dV}{dt}\right)_s$

Transients occurring at line crossing or when there is no initial voltage across the thyristor are worst case. The collector junction capacitance is greatest then because the depletion layer widens at higher voltage.

Small transients are incapable of charging the self-capacitance of the gate layer to its forward biased threshold voltage (Figure 2). Capacitance voltage divider action

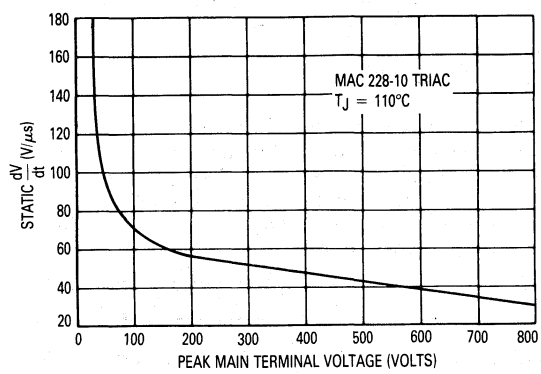


Figure 2. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Peak Voltage

between the collector and gate-cathode junctions and built-in resistors that shunt current away from the cathode emitter are responsible for this effect.

Static $\frac{dV}{dt}$ does not depend strongly on voltage for operation below the maximum voltage and temperature rating. Avalanche multiplication will increase leakage current and reduce $\frac{dV}{dt}$ capability if a transient is within roughly 50 volts of the actual device breakover voltage.

A higher rated voltage device guarantees increased $\frac{dV}{dt}$ at lower voltage. This is a consequence of the exponential rating method where a 400 V device rated at 50 V/μs has a higher $\frac{dV}{dt}$ to 200 V than a 200 V device with an identical rating. However, the same diffusion recipe usually applies for all voltages. So actual capabilities of the product are not much different.

Heat increases current gain and leakage, lowering $\left(\frac{dV}{dt}\right)_s$, the gate trigger voltage and noise immunity (Figure 3).

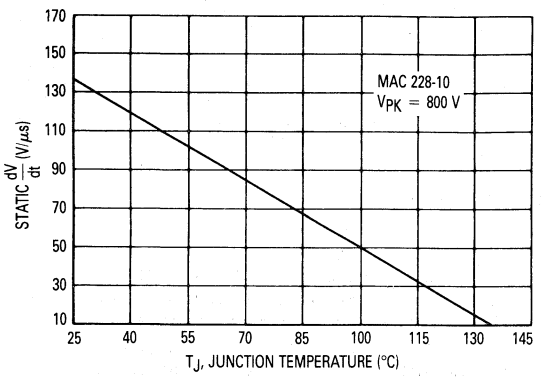


Figure 3. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Temperature

$\left(\frac{dV}{dt}\right)_s$ FAILURE MODE

Occasional unwanted turn-on by a transient may be acceptable in a heater circuit but isn't in a fire prevention sprinkler system or for the control of a large motor. Turn-on is destructive when the follow-on current amplitude or rate is excessive. If the thyristor shorts the power line or a charged capacitor, it will be damaged.

Static $\frac{dV}{dt}$ turn-on is non-destructive when series impedance limits the surge. The thyristor turns off after a half-cycle of conduction. High $\frac{dV}{dt}$ aids current spreading in the thyristor, improving its ability to withstand $\frac{dI}{dt}$. Breakdown turn-on does not have this benefit and should be prevented.

IMPROVING $\left(\frac{dV}{dt}\right)_s$

Static $\frac{dV}{dt}$ can be improved by adding an external resistor from the gate to MT1 (Figure 4). The resistor provides a path for leakage and $\frac{dV}{dt}$ induced currents that originate in the drive circuit or the thyristor itself.

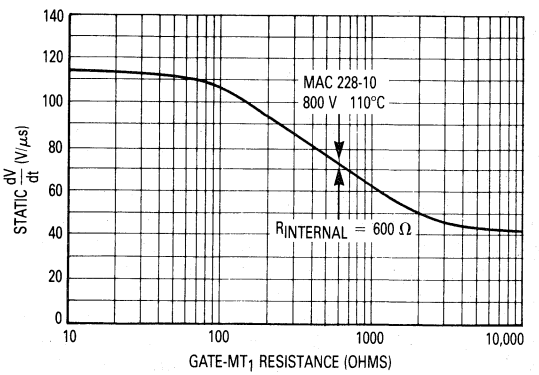


Figure 4. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Gate to MT_1 Resistance

Non-sensitive devices (Figure 5) have internal shorting resistors dispersed throughout the chip's cathode area. This design feature improves noise immunity and high temperature blocking stability at the expense of increased trigger and holding current. External resistors are optional for non-sensitive SCRs and TRIACs. They should be comparable in size to the internal shorting resistance of the device (20 to 100 ohms) to provide maximum improvement. The internal resistance of the thyristor should be measured with an ohmmeter that does not forward bias a diode junction.

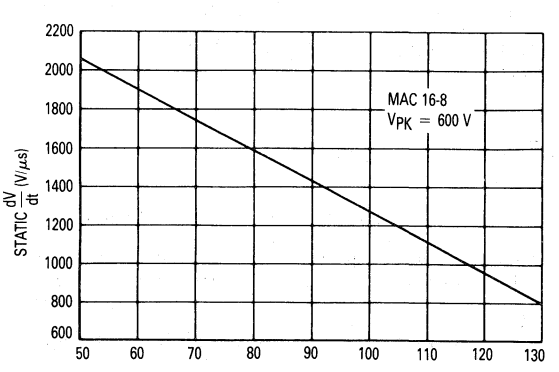


Figure 5. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Junction Temperature

Sensitive gate TRIACs run 100 to 1000 ohms. With an external resistor, their $\frac{dV}{dt}$ capability remains inferior to non-sensitive devices because lateral resistance within the gate layer reduces its benefit.

Sensitive gate SCRs ($I_{GT} < 200\ \mu A$) have no built-in resistor. They should be used with an external resistor. The recommended value of the resistor is 1000 ohms. Higher values reduce maximum operating temperature and $\left(\frac{dV}{dt}\right)_s$ (Figure 6). The capability of these parts varies by more than 100 to 1 depending on gate-cathode termination.

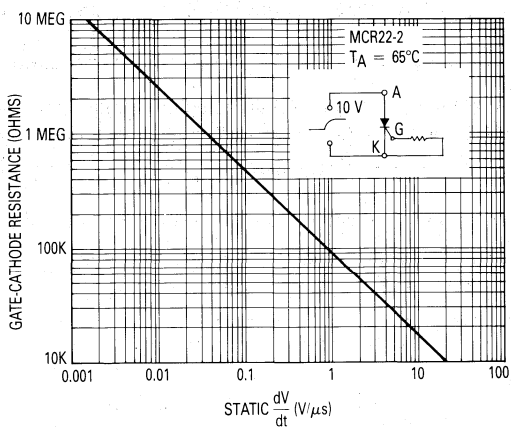


Figure 6. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Gate-Cathode Resistance

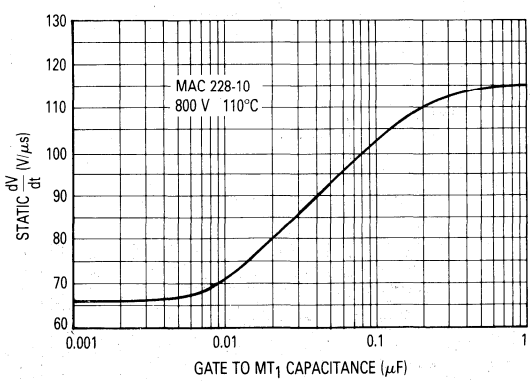


Figure 7. Exponential $\left(\frac{dV}{dt}\right)_s$ versus Gate to MT_1 Capacitance

A gate-cathode capacitor (Figure 7) provides a shunt path for transient currents in the same manner as the resistor. It also filters noise currents from the drive circuit and enhances the built-in gate-cathode capacitance voltage divider effect. The gate drive circuit needs to be able to charge the capacitor without excessive delay, but it does not need to supply continuous current as it would for a resistor that increases $\frac{dV}{dt}$ the same amount. However, the capacitor does not enhance static thermal stability.

The maximum $\left(\frac{dV}{dt}\right)_s$ improvement occurs with a short. Actual improvement stops before this because of spreading resistance in the thyristor. An external capacitor of about 0.1 μF allows the maximum enhancement at a higher value of R_{GK}.

One should keep the thyristor cool for the highest $\left(\frac{dV}{dt}\right)_s$. Also devices should be tested in the application circuit at the highest possible temperature using thyristors with the lowest measured trigger current.

TRIAC COMMUTATING $\frac{dV}{dt}$

WHAT IS COMMUTATING $\frac{dV}{dt}$?

The commutating $\frac{dV}{dt}$ rating applies when a TRIAC has been conducting and attempts to turn-off with an inductive load. The current and voltage are out of phase (Figure 8). The TRIAC attempts to turn-off as the current drops below the holding value. Now the line voltage is high and in the opposite polarity to the direction of conduction. Successful turn-off requires the voltage across the TRIAC to rise to the instantaneous line voltage at a rate slow enough to prevent retriggering of the device.

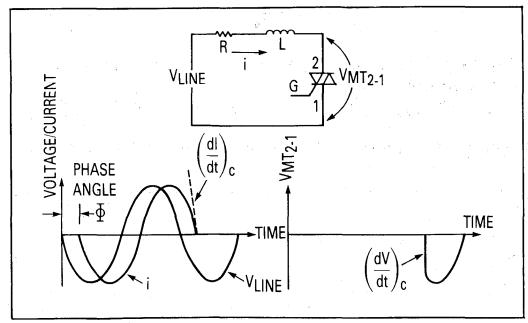


Figure 8. TRIAC Inductive Load Turn-Off $\left(\frac{dV}{dt}\right)_c$

$\left(\frac{dV}{dt}\right)_c$ DEVICE PHYSICS

A TRIAC functions like two SCRs connected in inverse-parallel. So, a transient of either polarity turns it on.

There is charge within the crystal's volume because of prior conduction (Figure 9). The charge at the boundaries

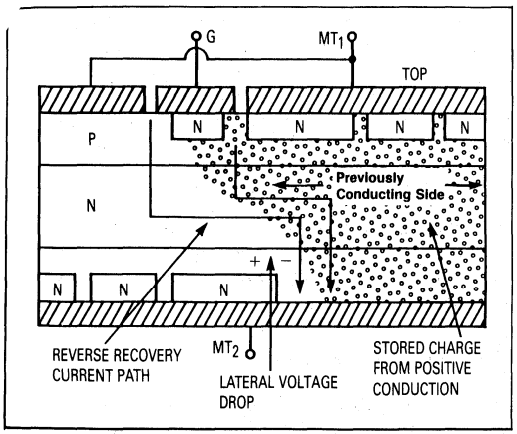


Figure 9. TRIAC Structure and Current Flow at Commutation

of the collector junction depletion layer responsible for $\left(\frac{dV}{dt}\right)_s$ is also present. TRIACs have lower $\left(\frac{dV}{dt}\right)_c$ than $\left(\frac{dV}{dt}\right)_s$ because of this additional charge.

The volume charge storage within the TRIAC depends on the peak current before turn-off and its rate of zero crossing $\left(\frac{dI}{dt}\right)_c$. In the classic circuit, the load impedance and line frequency determine $\left(\frac{dI}{dt}\right)_c$. The rate of crossing for sinusoidal currents is given by the slope of the secant line between the 50% and 0% levels as:

$$\left(\frac{dI}{dt}\right)_c = \frac{6\ f I_{TM}}{1000}\ \text{A/ms}$$

where f = line frequency and I_{TM} = maximum on-state current in the TRIAC.

Turn-off depends on both the Miller effect displacement current generated by $\frac{dV}{dt}$ across the collector capacitance and the currents resulting from internal charge storage within the volume of the device (Figure 10). If the reverse recovery current resulting from both these components is high, the lateral IR drop within the TRIAC base layer will forward bias the emitter and turn the TRIAC on. Commutating $\frac{dV}{dt}$ capability is lower when turning off from the positive direction of current conduction because of device geometry. The gate is on the top of the die and obstructs current flow.

Recombination takes place throughout the conduction period and along the back side of the current wave as it declines to zero. Turn-off capability depends on its shape. If the current amplitude is small and its zero crossing $\left(\frac{dI}{dt}\right)_c$ is low, there is little volume charge storage and turn-off becomes limited by $\left(\frac{dV}{dt}\right)_s$. At moderate current amplitudes, the volume charge begins to influence turn-off, requiring a larger snubber. When the current is large or has rapid zero crossing, $\left(\frac{dV}{dt}\right)_c$ has little influence. Commutating $\frac{dI}{dt}$ and delay time to voltage reapplication determine whether turn-off will be successful or not (Figures 11, 12).

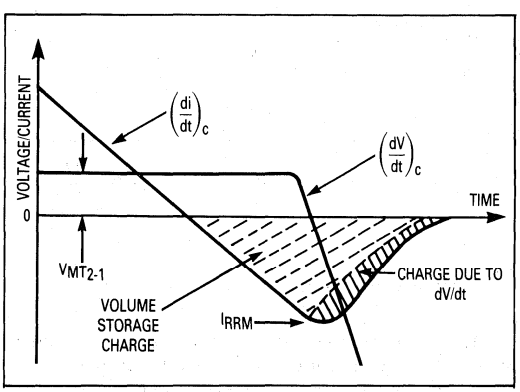


Figure 10. TRIAC Current and Voltage at Commutation

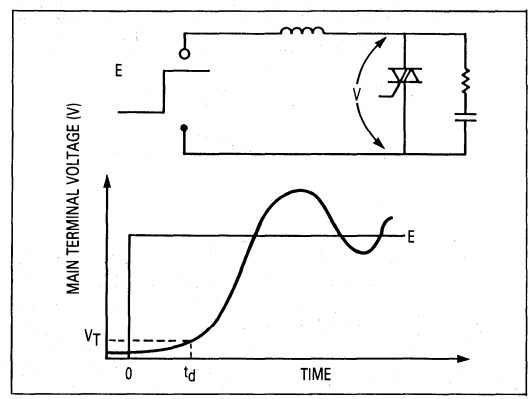


Figure 11. Snubber Delay Time

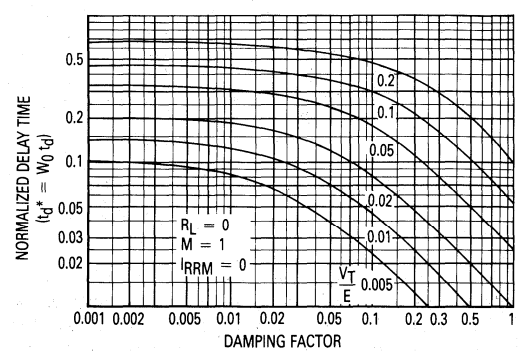


Figure 12. Delay Time To Normalized Voltage

CONDITIONS INFLUENCING $\left(\frac{dV}{dt}\right)_c$

Commutating $\frac{dV}{dt}$ depends on charge storage and recovery dynamics in addition to the variables influencing static $\frac{dV}{dt}$. High temperatures increase minority carrier life-time and the size of recovery currents, making turn-off more difficult. Loads that slow the rate of current zero-crossing aid turn-off. Those with harmonic content hinder turn-off.

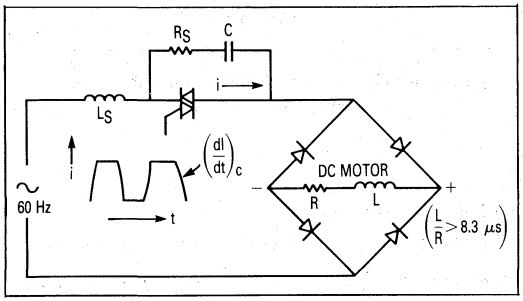


Figure 13. Phase Controlling a Motor in a Bridge

Circuit Examples

Figure 13 shows a TRIAC controlling an inductive load in a bridge. The inductive load has a time constant longer than the line period. This causes the load current to remain constant and the TRIAC current to switch rapidly as the line voltage reverses. This application is notorious for causing TRIAC turn-off difficulty because of high $\left(\frac{dI}{dt}\right)_c$.

High currents lead to high junction temperatures and rates of current crossing. Motors can have 5 to 6 times the normal current amplitude at start-up. This increases both junction temperature and the rate of current crossing, leading to turn-off problems.

The line frequency causes high rates of current crossing in 400 Hz applications. Resonant transformer circuits are doubly periodic and have current harmonics at both the primary and secondary resonance. Non-sinusoidal currents can lead to turn-off difficulty even if the current amplitude is low before zero-crossing.

$\left(\frac{dV}{dt}\right)_c$ FAILURE MODE

$\left(\frac{dV}{dt}\right)_c$ failure causes a loss of phase control. Temporary turn-on or total turn-off failure is possible. This can be destructive if the TRIAC conducts asymmetrically causing a dc current component and magnetic saturation. The winding resistance limits the current. Failure results because of excessive surge current and junction temperature.

IMPROVING $\left(\frac{dV}{dt}\right)_c$

The same steps that improve $\left(\frac{dV}{dt}\right)_s$ aid $\left(\frac{dV}{dt}\right)_c$ except when stored charge dominates turn-off. Steps that reduce the stored charge or soften the commutation are necessary then.

Larger TRIACs have better turn-off capability than smaller ones with a given load. The current density is lower in the larger device allowing recombination to claim a greater proportion of the internal charge. Also junction temperatures are lower.

TRIACs with high gate trigger currents have greater turn-off ability because of lower spreading resistance in the gate layer, reduced Miller effect, or shorter lifetime.

The rate of current crossing can be adjusted by adding a commutation softening inductor in series with the load. Small high permeability "square loop" inductors saturate causing no significant disturbance to the load current. The inductor resets as the current crosses zero introducing a large inductance into the snubber circuit at that time. This slows the current crossing and delays the reapplication of blocking voltage aiding turn-off.

The commutation inductor is a circuit element that introduces time delay, as opposed to inductance, into the circuit. It will have little influence on observed $\frac{dV}{dt}$ at the device. The following example illustrates the improvement resulting from the addition of an inductor constructed by winding 33 turns of number 18 wire on a tape wound core (52000-1A). This core is very small having an outside diameter of 3/4 inch and a thickness of 1/8 inch. The delay time can be calculated from:

$$t_s = \frac{(N\ A\ B\ 10^{-8})}{E} \text{ where:}$$

t_s = time delay to saturation in seconds.
B = saturating flux density in Gauss
A = effective core cross sectional area in cm^2
N = number of turns.

For the described inductor:

$$t_s = (33 \text{ turns})\ (0.076\ cm^2)\ (28000 \text{ Gauss})\ (1 \times 10^{-8}) / (175\ V) = 4.0\ \mu s.$$

The saturation current of the inductor does not need to be much larger than the TRIAC trigger current. Turn-off failure will result before recovery currents become greater than this value. This criterion allows sizing the inductor with the following equation:

$$I_s = \frac{H_s\ M_L}{0.4\ \pi\ N} \text{ where:}$$

H_s = MMF to saturate = 0.5 Oersted
M_L = mean magnetic path length = 4.99 cm.

$$I_s = \frac{(.5)\ (4.99)}{.4\ \pi\ 33} = 60 \text{ mA.}$$

SNUBBER PHYSICS

UNDAMPED NATURAL RESONANCE

$$\omega_0 = \frac{1}{\sqrt{LC}} \text{ Radians/second}$$

Resonance determines $\frac{dV}{dt}$ and boosts the peak capacitor voltage when the snubber resistor is small. C and L are related to one another by ω_0^2. $\frac{dV}{dt}$ scales linearly with ω_0 when the damping factor is held constant. A ten to one reduction in $\frac{dV}{dt}$ requires a 100 to 1 increase in either component.

DAMPING FACTOR

$$\rho = \frac{R}{2}\sqrt{\frac{C}{L}}$$

The damping factor is proportional to the ratio of the circuit loss and its surge impedance. It determines the trade off between $\frac{dV}{dt}$ and peak voltage. Damping factors between 0.01 and 1.0 are recommended.

The Snubber Resistor

Damping and $\frac{dV}{dt}$

When $\rho < 0.5$, the snubber resistor is small, and $\frac{dV}{dt}$ depends mostly on resonance. There is little improvement in $\frac{dV}{dt}$ for damping factors less than 0.3, but peak voltage and snubber discharge current increase. The voltage wave has a 1-COS (θ) shape with overshoot and ringing. Maximum $\frac{dV}{dt}$ occurs at a time later than t = 0. There is a time delay before the voltage rise, and the peak voltage almost doubles.

When $\rho > 0.5$, the voltage wave is nearly exponential in shape. The maximum instantaneous $\frac{dV}{dt}$ occurs at t = 0. There is little time delay and moderate voltage overshoot.

When $\rho > 1.0$, the snubber resistor is large and $\frac{dV}{dt}$ depends mostly on its value. There is some overshoot even through the circuit is overdamped.

High load inductance requires large snubber resistors and small snubber capacitors. Low inductances imply small resistors and large capacitors.

Damping and Transient Voltages

Figure 14 shows a series inductor and filter capacitor connected across the ac main line. The peak to peak voltage of a transient disturbance increases by nearly four times. Also the duration of the disturbance spreads because of ringing, increasing the chance of malfunction or damage to the voltage sensitive circuit. Closing a switch causes this behavior. The problem can be reduced by adding a damping resistor in series with the capacitor.

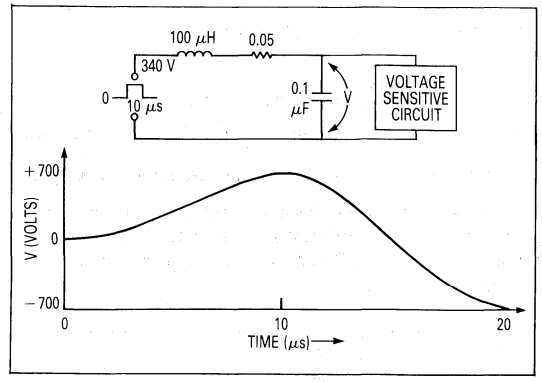


Figure 14. Undamped LC Filter Magnifies and Lengthens a Transient

$\frac{dI}{dt}$

Non-Inductive Resistor

The snubber resistor limits the capacitor discharge current and reduces $\frac{dI}{dt}$ stress. High $\frac{dI}{dt}$ destroys the thyristor even though the pulse duration is very short.

The rate of current rise is directly proportional to circuit voltage and inversely proportional to series inductance. The snubber is often the major offender because of its low inductance and close proximity to the thyristor.

With no transient suppressor, breakdown of the thyristor sets the maximum voltage on the capacitor. It is possible to exceed the highest rated voltage in the device series because high voltage devices are often used to supply low voltage specifications.

The minimum value of the snubber resistor depends on the type of thyristor, triggering quadrants, gate current amplitude, voltage, repetitive or non-repetitive operation, and required life expectancy. There is no simple way to predict the rate of current rise because it depends on turn-on speed of the thyristor, circuit layout, type and size of snubber capacitor, and inductance in the snubber resistor. The equations in Appendix D describe the circuit. However, the values required for the model are not easily obtained except by testing. Therefore, reliability should be verified in the actual application circuit.

Table 1 shows suggested minimum resistor values estimated (Appendix A) by testing a 20 piece sample from the four different TRIAC die sizes.

Table 1. Minimum Non-inductive Snubber Resistor for Four Quadrant Triggering.

TRIAC Type	Peak V_C Volts	R_S Ohms	$\frac{dI}{dt}$ A/μs
Non-Sensitive Gate	200	3.3	170
(I_{GT} > 10 mA)	300	6.8	250
8 to 40 $A_{(RMS)}$	400	11	308
	600	39	400
	800	51	400

Reducing $\frac{dI}{dt}$

TRIAC $\frac{dI}{dt}$ can be improved by avoiding quadrant 4 triggering. Most optocoupler circuits operate the TRIAC in quadrants 1 and 3. Integrated circuit drivers use quadrants 2 and 3. Zero crossing trigger devices are helpful because they prohibit triggering when the voltage is high.

Driving the gate with a high amplitude fast rise pulse increases $\frac{dI}{dt}$ capability. The gate ratings section defines the maximum allowed current.

Inductance in series with the snubber capacitor reduces $\frac{dI}{dt}$. It should not be more than five percent of the load inductance to prevent degradation of the snubber's $\frac{dV}{dt}$ suppression capability. Wirewound snubber resistors sometimes serve this purpose. Alternatively, a separate inductor can be added in series with the snubber capacitor. It can be small because it does not need to carry the load current. For example, 18 turns of AWG No. 20 wire on a T50-3 (1/2 inch) powdered iron core creates a non-saturating 6.0 μH inductor.

A 10 ohm, 0.33 μF snubber charged to 650 volts resulted in a 1000 A/μs $\frac{dI}{dt}$. Replacement of the non-inductive snubber resistor with a 20 watt wirewound unit lowered the rate of rise to a non-destructive 170 A/μs at 800 V. The inductor gave an 80 A/μs rise at 800 V with the non-inductive resistor.

The Snubber Capacitor

A damping factor of 0.3 minimizes the size of the snubber capacitor for a given value of $\frac{dV}{dt}$. This reduces the cost and physical dimensions of the capacitor. However, it raises voltage causing a counter balancing cost increase.

Snubber operation relies on the charging of the snubber capacitor. Turn-off snubbers need a minimum conduction angle long enough to discharge the capacitor. It should be at least several time constants ($R_S C_S$).

STORED ENERGY

Inductive Switching Transients

$E = \frac{1}{2} L I_0^2$ Watt-seconds or Joules

I_0 = current in Amperes flowing in the inductor at $t = 0$.

Resonant charging cannot boost the supply voltage at turn-off by more than 2. If there is an initial current flowing in the load inductance at turn-off, much higher voltages are possible. Energy storage is negligible when a TRIAC turns off because of its low holding or recovery current.

The presence of an additional switch such as a relay, thermostat or breaker allows the interruption of load current and the generation of high spike voltages at switch opening. The energy in the inductance transfers into the circuit capacitance and determines the peak voltage (Figure 15).

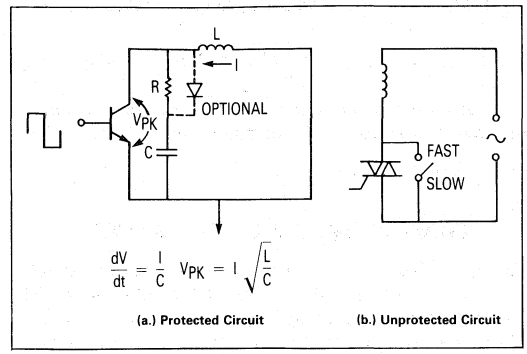


Figure 15. Interrupting Inductive Load Current

Capacitor Discharge

The energy stored in the snubber capacitor $\left(E_C = \frac{1}{2} CV^2\right)$ transfers to the snubber resistor and thyristor every time it turns on. The power loss is proportional to frequency ($P_{AV} = 120\ E_C$ @ 60 Hz).

CURRENT DIVERSION

The current flowing in the load inductor cannot change instantly. This current diverts through the snubber resistor causing a spike of theoretically infinite $\frac{dV}{dt}$ with magnitude equal to (I_{RRM} R) or (I_H R).

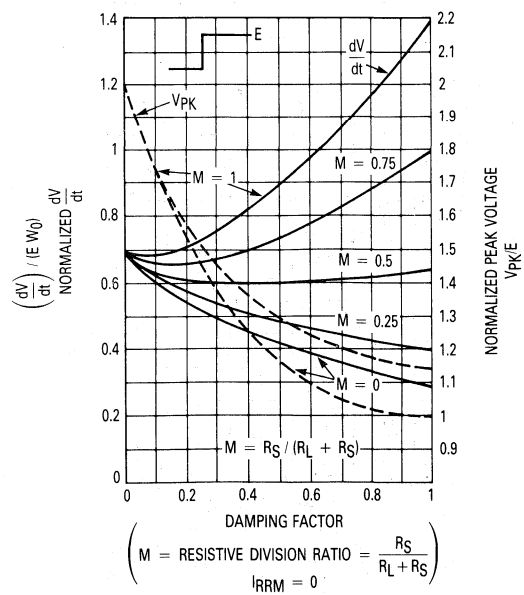


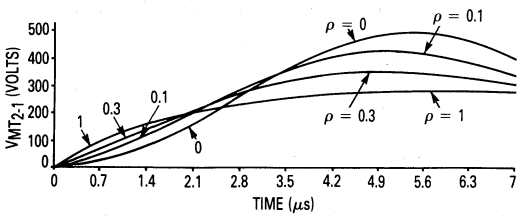
Figure 16. 0 To 63% $\frac{dV}{dt}$

LOAD PHASE ANGLE

Highly inductive loads cause increased voltage and $\left(\frac{dV}{dt}\right)_c$ at turn-off. However, they help to protect the thyristor from transients and $\left(\frac{dV}{dt}\right)_s$. The load serves as the snubber inductor and limits the rate of inrush current if the device does turn on. Resistance in the load lowers $\frac{dV}{dt}$ and V_{PK} (Figure 16).

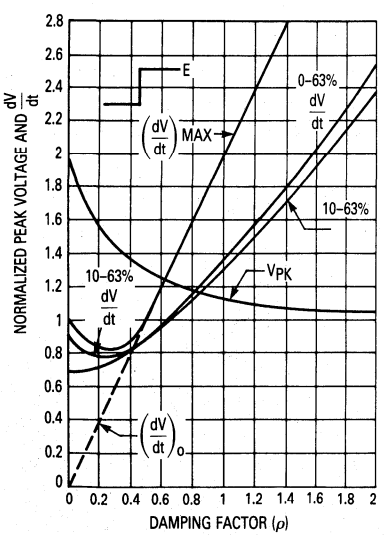
CHARACTERISTIC VOLTAGE WAVES

Damping factor and reverse recovery current determine the shape of the voltage wave. It is not exponential when the snubber damping factor is less than 0.5 (Figure 17) or when significant recovery currents are present.



$$\left[0\text{–}63\%\left(\frac{dV}{dt}\right)_s = 100\ V/\mu s,\ E = 250\ V,\ R_L = 0,\ I_{RRM} = 0\right]$$

Figure 17. Voltage Waves For Different Damping Factors



$(R_L = 0, M = 1, I_{RRM} = 0)$

NORMALIZED $\frac{dV}{dt} = \frac{dV/dt}{E\ \omega_0}$ NORMALIZED $V_{PK} = \frac{V_{PK}}{E}$

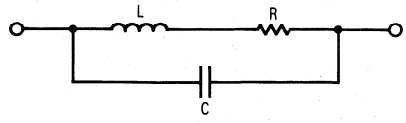
Figure 18. Trade-Off Between V_{PK} and $\frac{dV}{dt}$

A variety of wave parameters (Figure 18) describe $\frac{dV}{dt}$. Some are easy to solve for and assist understanding. These include the initial $\frac{dV}{dt}$, the maximum instantaneous $\frac{dV}{dt}$, and the average $\frac{dV}{dt}$ to the peak reapplied voltage. The 0 to 63% $\left(\frac{dV}{dt}\right)_s$ and 10 to 63% $\left(\frac{dV}{dt}\right)_c$ definitions on device data sheets are easy to measure but difficult to compute.

NON-IDEAL BEHAVIORS

CORE LOSSES

The magnetic core materials in typical 60 Hz loads introduce losses at the snubber natural frequency. They appear as a resistance in series with the load inductance and winding dc resistance (Figure 19). This causes actual $\frac{dV}{dt}$ to be less than the theoretical value.



L DEPENDS ON CURRENT AMPLITUDE, CORE SATURATION

R INCLUDES CORE LOSS, WINDING R. INCREASES WITH FREQUENCY

C WINDING CAPACITANCE. DEPENDS ON INSULATION, WIRE SIZE, GEOMETRY

Figure 19. Inductor Model

COMPLEX LOADS

Many real-world inductances are non-linear. Their core materials are not gapped causing inductance to vary with current amplitude. Small signal measurements poorly characterize them. For modeling purposes, it is best to measure them in the actual application.

Complex load circuits should be checked for transient voltages and currents at turn-on and off. With a capacitive load, turn-on at peak input voltage causes the maximum surge current. Motor starting current runs 4 to 6 times the steady state value. Generator action can boost voltages above the line value. Incandescent lamps have cold start currents 10 to 20 times the steady state value. Transformers generate voltage spikes when they are energized. Power factor correction circuits and switching devices create complex loads. In most cases, the simple CRL model allows an approximate snubber design. However, there is no substitute for testing and measuring the worst case load conditions.

SURGE CURRENTS IN INDUCTIVE CIRCUITS

Inductive loads with long L/R time constants cause asymmetric multi-cycle surges at start up (Figure 20). Triggering at zero voltage crossing is the worst case condition. The surge can be eliminated by triggering at the zero current crossing angle.

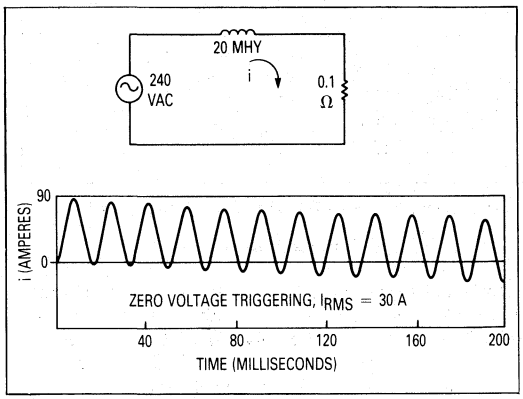


Figure 20. Start-Up Surge For Inductive Circuit

Core remanence and saturation cause surge currents. They depend on trigger angle, line impedance, core characteristics, and direction of the residual magnetization. For example, a 2.8 kVA 120 V 1:1 transformer with a 1.0 ampere load produced 160 ampere currents at start-up. Soft starting the circuit at a small conduction angle reduces this current.

Transformer cores are usually not gapped and saturate easily. A small asymmetry in the conduction angle causes magnetic saturation and multi-cycle current surges.

Steps to achieve reliable operation include:

1. Supply sufficient trigger current amplitude. TRIACs have different trigger currents depending on their quadrant of operation. Marginal gate current or optocoupler LED current causes halfwave operation.
2. Supply sufficient gate current duration to achieve latching. Inductive loads slow down the main terminal current rise. The gate current must remain above the specified I_{GT} until the main terminal current exceeds the latching value. Both a resistive bleeder around the load and the snubber discharge current help latching.
3. Use a snubber to prevent TRIAC $\left(\frac{dV}{dt}\right)_c$ failure.
4. Minimize designed-in trigger asymmetry. Triggering must be correct every half-cycle including the first. Use a storage scope to investigate circuit behavior during the first few cycles of turn-on. Alternatively, get the gate circuit up and running before energizing the load.
5. Derive the trigger synchronization from the line instead of the TRIAC main terminal voltage. This avoids regenerative interaction between the core hysteresis and the triggering angle preventing trigger runaway, halfwave operation, and core saturation.
6. Avoid high surge currents at start-up. Use a current probe to determine surge amplitude. Use a soft start circuit to reduce inrush current.

DISTRIBUTED WINDING CAPACITANCE

There are small capacitances between the turns and layers of a coil. Lumped together, they model as a single shunt capacitance. The load inductor behaves like a capacitor at frequencies above its self-resonance. It becomes ineffective in controlling $\frac{dV}{dt}$ and V_{PK} when a fast transient such as that resulting from the closing of a switch occurs. This problem can be solved by adding a small snubber across the line.

SELF-CAPACITANCE

A thyristor has self-capacitance which limits $\frac{dV}{dt}$ when the load inductance is large. Large load inductances, high power factors, and low voltages may allow snubberless operation.

SNUBBER EXAMPLES

WITHOUT INDUCTANCE

Power TRIAC Example

Figure 21 shows a transient voltage applied to a TRIAC controlling a resistive load. Theoretically there will be an instantaneous step of voltage across the TRIAC. The only elements slowing this rate are the inductance of the wiring and the self-capacitance of the thyristor. There is an exponential capacitor charging component added along with a decaying component because of the IR drop in the snubber resistor. The non-inductive snubber circuit is useful when the load resistance is much larger than the snubber resistor.

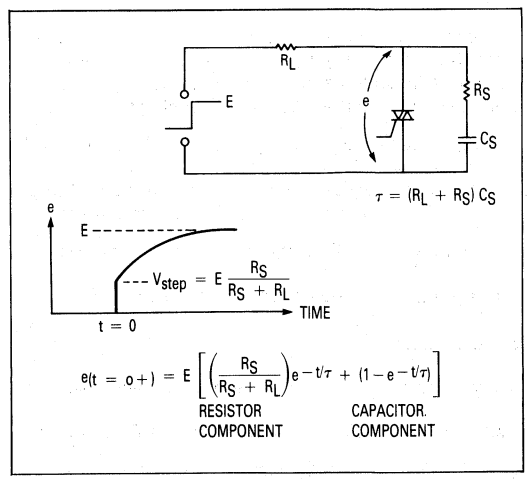


Figure 21. Non-Inductive Snubber Circuit

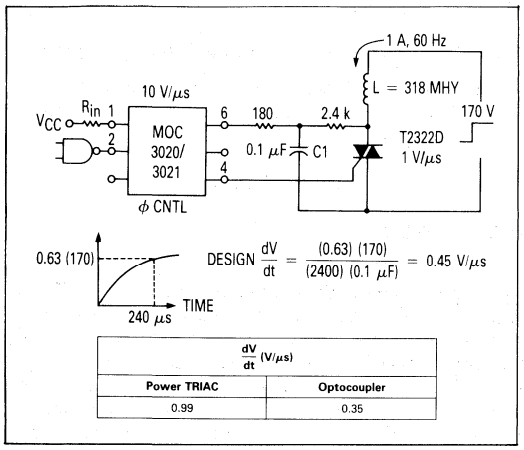
Opto-TRIAC Examples

Single Snubber, Time Constant Design

Figure 22 illustrates the use of the RC time constant design method. The optocoupler sees only the voltage across the snubber capacitor. The resistor R1 supplies the trigger current of the power TRIAC. A worst case design procedure assumes that the voltage across the power TRIAC changes instantly. The capacitor voltage rises to 63% of the maximum in one time constant. Then:

$$R_1 C_S = \tau = \frac{0.63\ E}{\left(\frac{dV}{dt}\right)_s} \text{ where } \left(\frac{dV}{dt}\right)_s \text{ is the rated static } \frac{dV}{dt}$$

for the optocoupler.



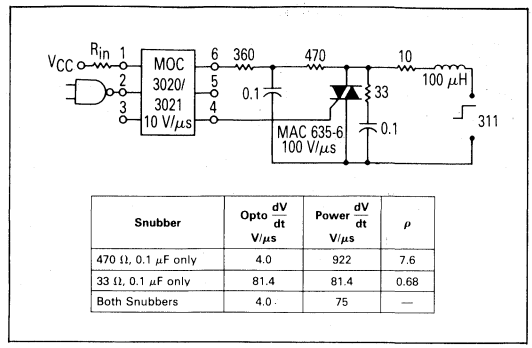
dV/dt (V/μs)	
Power TRIAC	Optocoupler
0.99	0.35

Figure 22. Single Snubber For Sensitive Gate TRIAC and Phase Controllable Optocoupler (ρ = 0.67)

The optocoupler conducts current only long enough to trigger the power device. When it turns on, the voltage between MT2 and the gate drops below the forward threshold voltage of the opto-TRIAC causing turn-off. The optocoupler sees $\left(\frac{dV}{dt}\right)_s$ when the power TRIAC turns off later in the conduction cycle at zero current crossing. Therefore, it is not necessary to design for the lower optocoupler $\left(\frac{dV}{dt}\right)_c$ rating. In this example, a single snubber designed for the optocoupler protects both devices.

Double Snubber

To minimize $\left(\frac{dV}{dt}\right)_s$ across the optocoupler, use of a sensitive gate TRIAC and a large value of R1 is recommended. This increases the damping factor and $\frac{dV}{dt}$ across the power TRIAC and may require the use of a second lower resistance snubber for the power device (Figure 23).



Snubber	Opto dV/dt V/μs	Power dV/dt V/μs	ρ
470 Ω, 0.1 μF only	4.0	922	7.6
33 Ω, 0.1 μF only	81.4	81.4	0.68
Both Snubbers	4.0	75	—

23. The Effect of Individual and Combined Snubbers on $\frac{dV}{dt}$

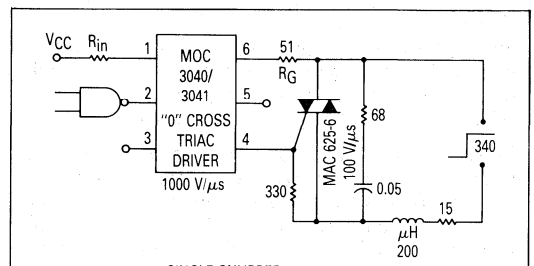
Unwanted triggering phase delay sets the maximum size for R1. This happens because the line voltage must rise sufficiently to force the power thyristor's trigger current. This limitation occurs at long conduction angles. The triggering delay angle can be calculated from:

$$\theta_d = \text{Sin}^{-1}\left(\frac{V_T}{V_{PK}}\right) \text{ where:}$$

V_T = the voltage across the thyristor needed to force the trigger current and V_{PK} = the peak line voltage.

Zero-Crossing Optocoupler Design

In Figure 24 a single snubber designed for the power TRIAC protects both devices. The time constant design method does not apply. One should refer to the discussion of the snubber with inductance for design information.



SINGLE SNUBBER ρ = 0.66

dV/dt (V/μs)	
Optocoupler	Power TRIAC
92	92

Figure 24. Snubber with Zero-Crossing Optocoupler

Optocouplers with SCRs

Anti-parallel SCR circuits result in the same $\frac{dV}{dt}$ across the optocoupler and SCR (Figure 25). Phase controllable optocouplers require the SCRs to be snubbed to their

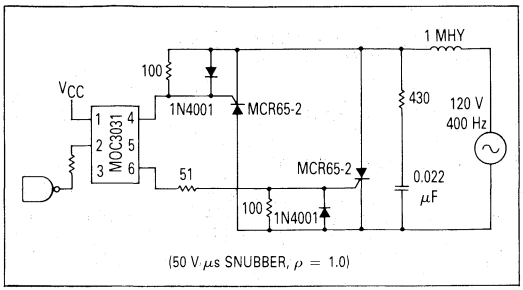


Figure 25. Anti-Parallel SCR Driver

lower $\frac{dV}{dt}$ rating. Anti-parallel SCR circuits are free from the charge storage behaviors that reduce the turn-off capability of TRIACs. Each SCR conducts for a half-cycle and has the next half cycle of the ac line in which to recover. The turn-off $\frac{dV}{dt}$ of the conducting SCR becomes a static forward blocking $\frac{dV}{dt}$ for the other device. Use the SCR data sheet $\left(\frac{dV}{dt}\right)_S$ rating in the snubber design.

A SCR used inside a rectifier bridge to control an ac load will not have a half cycle in which to recover. The available time decreases with increasing line voltage. This makes the circuit less attractive. Inductive transients can be suppressed by a snubber at the input to the bridge or across the SCR. However, the time limitation still applies.

OPTO $\left(\frac{dV}{dt}\right)_C$

Zero-crossing optocouplers can be used to switch inductive loads at currents less than 100 mA (Figure 26).

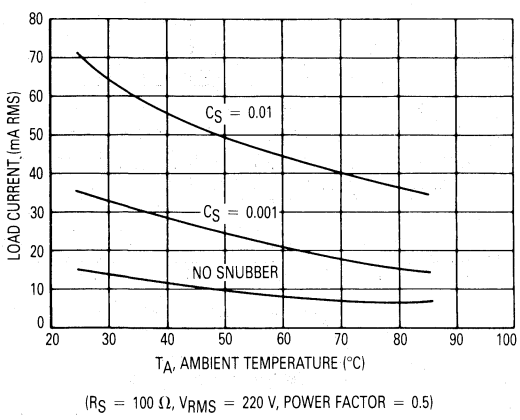


Figure 26. MOC3062 Inductive Load Current versus T_A

However a power TRIAC along with the optocoupler should be used for higher load currents.

A phase controllable optocoupler is recommended with a power device. When the load current is small, a MAC97 TRIAC is suitable.

Unusual circuit conditions sometimes lead to unwanted operation of an optocoupler in $\left(\frac{dV}{dt}\right)_C$ mode. Very large currents in the power device cause increased voltages between MT2 and the gate that hold the optocoupler on. Use of a larger TRIAC or other measures that limit inrush current solve this problem.

Very short conduction times leave residual charge in the optocoupler. A minimum conduction angle allows recovery before voltage reapplication.

1

THE SNUBBER WITH INDUCTANCE

Consider an overdamped snubber using a large capacitor whose voltage changes insignificantly during the time under consideration. The circuit reduces to an equivalent L/R series charging circuit.

The current through the snubber resistor is:

$$i = \frac{V}{R_\tau}\left(1 - e^{-\frac{t}{\tau}}\right),$$

and the voltage across the TRIAC is:

$$e = i\,R_S.$$

The voltage wave across the TRIAC has an exponential rise with maximum rate at t = 0. Taking its derivative gives its value as:

$$\left(\frac{dV}{dt}\right)_0 = \frac{V\,R_S}{L}.$$

Highly overdamped snubber circuits are not practical designs. The example illustrates several properties:

1. The initial voltage appears completely across the circuit inductance. Thus, it determines the rate of change of current through the snubber resistor and the initial $\frac{dV}{dt}$. This result does not change when there is resistance in the load and holds true for all damping factors.

2. The snubber works because the inductor controls the rate of current change through the resistor and the rate of capacitor charging. Snubber design cannot ignore the inductance. This approach suggests that the snubber capacitance is not important but that is only true for this hypothetical condition. The snubber resistor shunts the thyristor causing unacceptable leakage when the capacitor is not present. If the power loss is tolerable, $\frac{dV}{dt}$ can be controlled without the capacitor. An example is the soft-start circuit used to limit inrush current in switching power supplies (Figure 27).

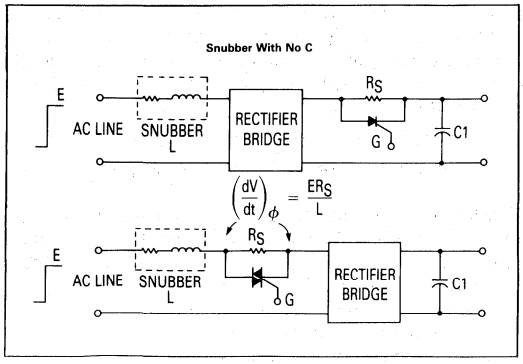


Figure 27. Surge Current Limiting For a Switching Power Supply

TRIAC DESIGN PROCEDURE $\left(\frac{dV}{dt}\right)_c$

1. Refer to Figure 18 and select a particular damping factor (ρ) giving a suitable trade-off between V_{PK} and $\frac{dV}{dt}$. Determine the normalized $\frac{dV}{dt}$ corresponding to the chosen damping factor.

The voltage E depends on the load phase angle:

$$E = \sqrt{2}\, V_{RMS} \text{ Sin } (\phi) \text{ where } \phi = \tan^{-1}\left(\frac{X_L}{R_L}\right) \text{ where}$$

ϕ = measured phase angle between line V and load I
R_L = measured dc resistance of the load.

Then

$$Z = \frac{V_{RMS}}{I_{RMS}} \sqrt{R_L{}^2 + X_L{}^2} \quad X_L = \sqrt{Z^2 - R_L{}^2} \text{ and}$$

$$L = \frac{X_L}{2\,\pi\, f_{Line}}.$$

If only the load current is known, assume a pure inductance. This gives a conservative design. Then:

$$L = \frac{V_{RMS}}{2\,\pi\, f_{Line}\, I_{RMS}} \text{ where } E = \sqrt{2}\, V_{RMS}.$$

For example:

$$E = \sqrt{2}\, 120 = 170 \text{ V}; \; L = \frac{120}{(8 \text{ A})\,(377 \text{ rps})} = 39.8 \text{ mH}.$$

Read from the graph at ρ = 0.6, V_{PK} = (1.25) 170 = 213 V. Use 400 V TRIAC. Read $\frac{dV}{dt}_{(\rho\, =\, 0.6)}$ = 1.0.

2. Apply the resonance criterion:

$$\omega_0 = \left(\text{spec } \frac{dV}{dt}\right) / \left(\frac{dV}{dt_{(P)}} E\right).$$

$$\omega_0 = \frac{5 \times 10^6 \text{ V/S}}{(1)\,(170 \text{ V})} = 29.4 \times 10^3 \text{ rps.}$$

$$C = \frac{1}{\omega_0{}^2\, L} = 0.029\ \mu\text{F}$$

3. Apply the damping criterion:

$$R_S = 2\,\rho \sqrt{\frac{L}{C}} = 2\,(0.6) \sqrt{\frac{39.8 \times 10^{-3}}{0.029 \times 10^{-6}}} = 1400 \text{ ohms.}$$

$\left(\frac{dV}{dt}\right)_c$ SAFE AREA CURVE

Figure 28 shows a MAC16 TRIAC turn-off safe operating area curve. Turn-off occurs without problem under the curve. The region is bounded by static $\frac{dV}{dt}$ at low values of $\left(\frac{dI}{dt}\right)_c$ and delay time at high currents. Reduction of the peak current permits operation at higher line frequency. This TRIAC operated at f = 400 Hz, T_J = 125°C, and I_{TM} = 6.0 amperes using a 30 ohm and 0.068 μF snubber. Low damping factors extend operation to higher $\left(\frac{dI}{dt}\right)_c$, but capacitor sizes increase. The addition of a small, saturable commutation inductor extends the allowed current rate by introducing recovery delay time.

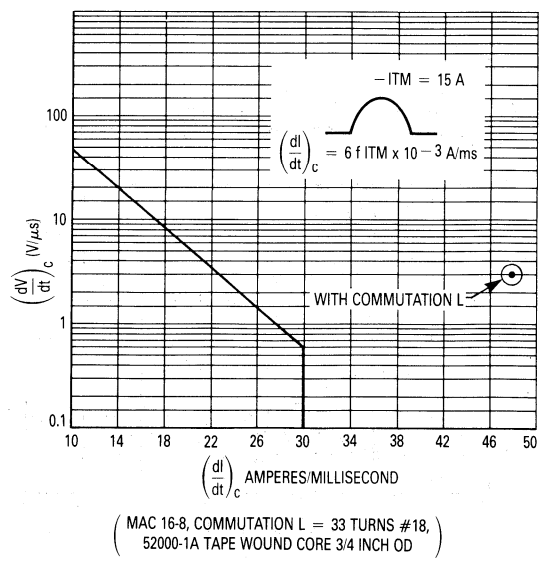


Figure 28. $\left(\frac{dV}{dt}\right)_c$ versus $\left(\frac{dI}{dt}\right)_c$ T_J = 125°C

STATIC $\frac{dV}{dt}$ DESIGN

There is usually some inductance in the ac main and power wiring. The inductance may be more than 100 μH if there is a transformer in the circuit or nearly zero when a shunt power factor correction capacitor is present. Usually the line inductance is roughly several μH. The minimum inductance must be known or defined by adding a series inductor to insure reliable operation (Figure 29).

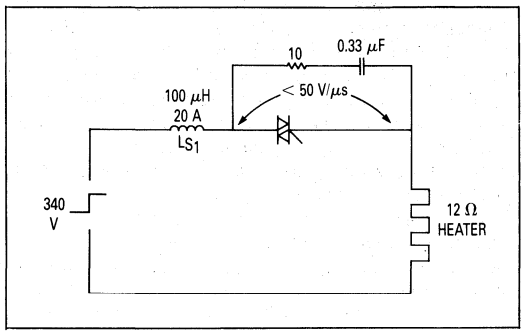


Figure 29. Snubbing For a Resistive Load

One hundred μH is a suggested value for starting the design. Plug the assumed inductance into the equation for C. Larger values of inductance result in higher snubber resistance and reduced $\frac{dI}{dt}$. For example:

Given $E = 240\sqrt{2} = 340$ V.

Pick $\rho = 0.3$.

Then from Figure 18, $V_{PK} = 1.42\ (340) = 483$ V.

Thus, it will be necessary to use a 600 V device. Using the previously stated formulas for ω_0, C and R we find:

$$\omega_0 = \frac{50 \times 10^6\ \text{V/S}}{(0.73)\ (340\ \text{V})} = 201450\ \text{rps}$$

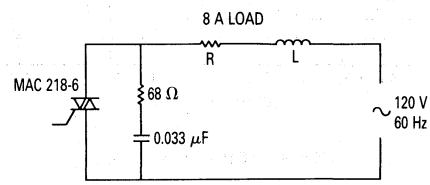
$$C = \frac{1}{(201450)^2\ (100 \times 10^{-6})} = 0.2464\ \mu\text{F}$$

$$R = 2\ (0.3)\sqrt{\frac{100 \times 10^{-6}}{0.2464 \times 10^{-6}}} = 12\ \text{ohms}$$

VARIABLE LOADS

The snubber should be designed for the smallest load inductance because $\frac{dV}{dt}$ will then be highest because of its dependence on ω_0. This requires a higher voltage device for operation with the largest inductance because of the corresponding low damping factor.

Figure 30 describes $\frac{dV}{dt}$ for an 8.0 ampere load at various power factors. The minimum inductance is a component added to prevent static $\frac{dV}{dt}$ firing with a resistive load.



$\left(\frac{dV}{dt}\right)_s = 100$ V/μs $\quad \left(\frac{dV}{dt}\right)_c = 5$ V/μs

ρ	R	L	V_{step}	V_{PK}	$\frac{dV}{dt}$
	Ω	MHY	V	V	V/μs
0.75	15	0.1	170	191	86
0.03	0	39.8	170	325	4.0
0.04	10.6	28.1	120	225	3.3
0.06	13.5	17.3	74	136	2.6

Figure 30. Snubber For a Variable Load

EXAMPLES OF SNUBBER DESIGNS

Table 2 describes snubber RC values for $\left(\frac{dV}{dt}\right)_s$. Figures 31 and 32 show possible R and C values for a 5.0 V/μs $\left(\frac{dV}{dt}\right)_c$ assuming a pure inductive load.

Table 2. Static $\frac{dV}{dt}$ Designs
($E = 340$ V, $V_{peak} = 500$ V, $\rho = 0.3$)

	5.0 V/μs		50 V/μs		100 V/μs	
L μH	C μF	R Ohm	C μF	R Ohm	C μF	R Ohm
47					0.15	10
100			0.33	10	0.1	20
220			0.15	22	0.033	47
500			0.068	51	0.015	110
1000	3.0	11	0.033	100		

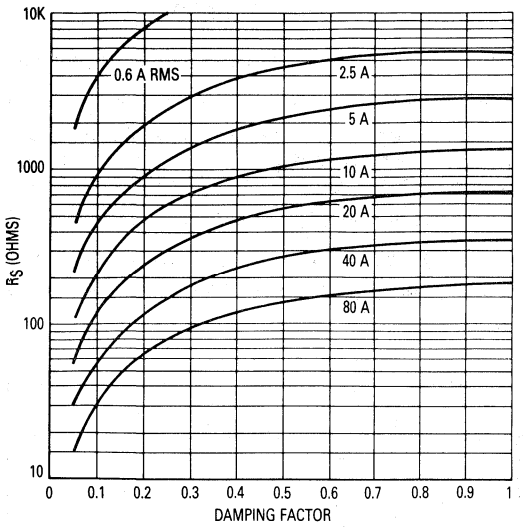


Figure 31. Snubber Resistor For $\left(\frac{dV}{dt}\right)_c$ = 5.0 V/μs

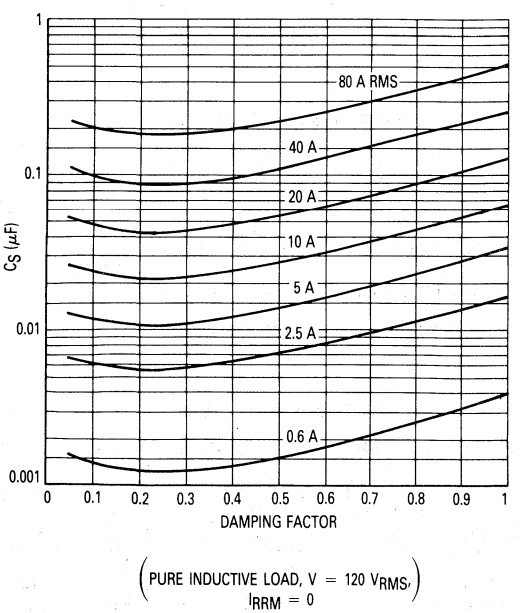


Figure 32. Snubber Capacitor For $\left(\frac{dV}{dt}\right)_c$ = 5.0 V/μs

TRANSIENT AND NOISE SUPPRESSION

Transients arise internally from normal circuit operation or externally from the environment. The latter is particularly frustrating because the transient characteristics are undefined. A statistical description applies. Greater or smaller stresses are possible. Long duration high voltage transients are much less probable than those of lower amplitude and higher frequency. Environments with infrequent lightning and load switching see transient voltages below 3.0 kV.

The natural frequencies and impedances of indoor ac wiring result in damped oscillatory surges with typical frequencies ranging from 30 kHz to 1.5 MHz. Surge amplitude depends on both the wiring and the source of surge energy. Disturbances tend to die out at locations far away from the source. Spark-over (6.0 kV in indoor ac wiring) sets the maximum voltage when transient suppressors are not present. Transients closer to the service entrance or in heavy wiring have higher amplitudes, longer durations, and more damping because of the lower inductance at those locations.

The simple CRL snubber is a low pass filter attenuating frequencies above its natural resonance. A steady state sinusoidal input voltage results in a sine wave output at the same frequency. With no snubber resistor, the rate of roll off approaches 12 dB per octave. The corner frequency is at the snubber's natural resonance. If the damping factor is low, the response peaks at this frequency. The snubber resistor degrades filter characteristics introducing an up-turn at $\omega = 1 / (RC)$. The roll-off approaches 6.0 dB/octave at frequencies above this. Inductance in the snubber resistor further reduces the roll-off rate.

Figure 33 describes the frequency response of the circuit in Figure 29. Figure 34 gives the theoretical response to a 3.0 kV 100 kHz ring-wave. The snubber reduces the peak voltage across the thyristor. However, the fast rise input causes a high $\frac{dV}{dt}$ step when series inductance is added to the snubber resistor. Limiting the input voltage with a transient suppressor reduces the step.

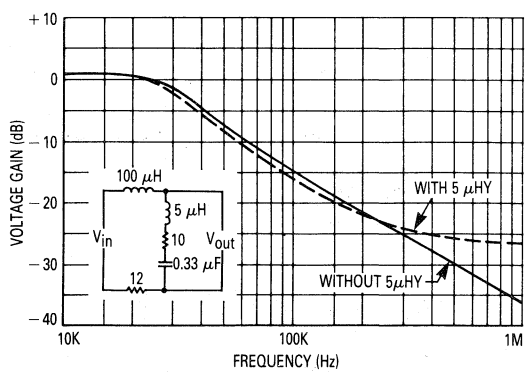


Figure 33. Snubber Frequency Response $\left(\frac{V_{out}}{V_{in}}\right)$

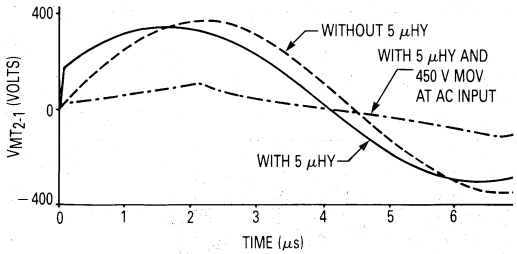


Figure 34. Theoretical Response of Figure 33 Circuit to 3.0 kV IEEE 587 Ring Wave (R_{SC} = 27.5 Ω)

The noise induced into a circuit is proportional to $\frac{dV}{dt}$ when coupling is by stray capacitance, and $\frac{dI}{dt}$ when the coupling is by mutual inductance. Best suppression requires the use of a voltage limiting device along with a rate limiting CRL snubber.

The thyristor is best protected by preventing turn-on from $\frac{dV}{dt}$ or breakover. The circuit should be designed for what can happen instead of what normally occurs.

In Figure 35, a MOV connected across the line protects many parallel circuit branches and their loads. The MOV defines the maximum input voltage and $\frac{dI}{dt}$ through the load. With the snubber, it sets the maximum $\frac{dV}{dt}$ and peak voltage across the thyristor. The MOV must be large because there is little surge limiting impedance to prevent its burn-out.

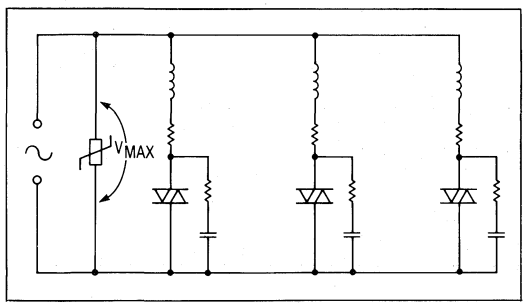


Figure 35. Limiting Line Voltage

In Figure 36, there is a separate suppressor across each thyristor. The load impedance limits the surge energy delivered from the line. This allows the use of a smaller device but omits load protection. This arrangement protects each thyristor when its load is a possible transient source.

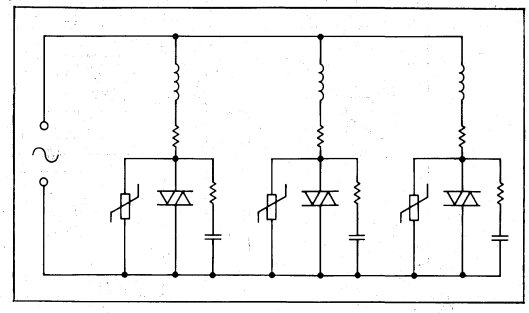


Figure 36. Limiting Thyristor Voltage

It is desirable to place the suppression device directly across the source of transient energy to prevent the induction of energy into other circuits. However, there is no protection for energy injected between the load and its controlling thyristor. Placing the suppressor directly across each thyristor positively limits maximum voltage and snubber discharge $\frac{dI}{dt}$.

EXAMPLES OF SNUBBER APPLICATIONS

In Figure 37, TRIACs switch a 3 phase motor on and off and reverse its rotation. Each TRIAC pair functions as a SPDT switch. The turn-on of one TRIAC applies the differential voltage between line phases across the blocking device without the benefit of the motor impedance to constrain the rate of voltage rise. The inductors are added to prevent static $\frac{dV}{dt}$ firing and a line-to-line short.

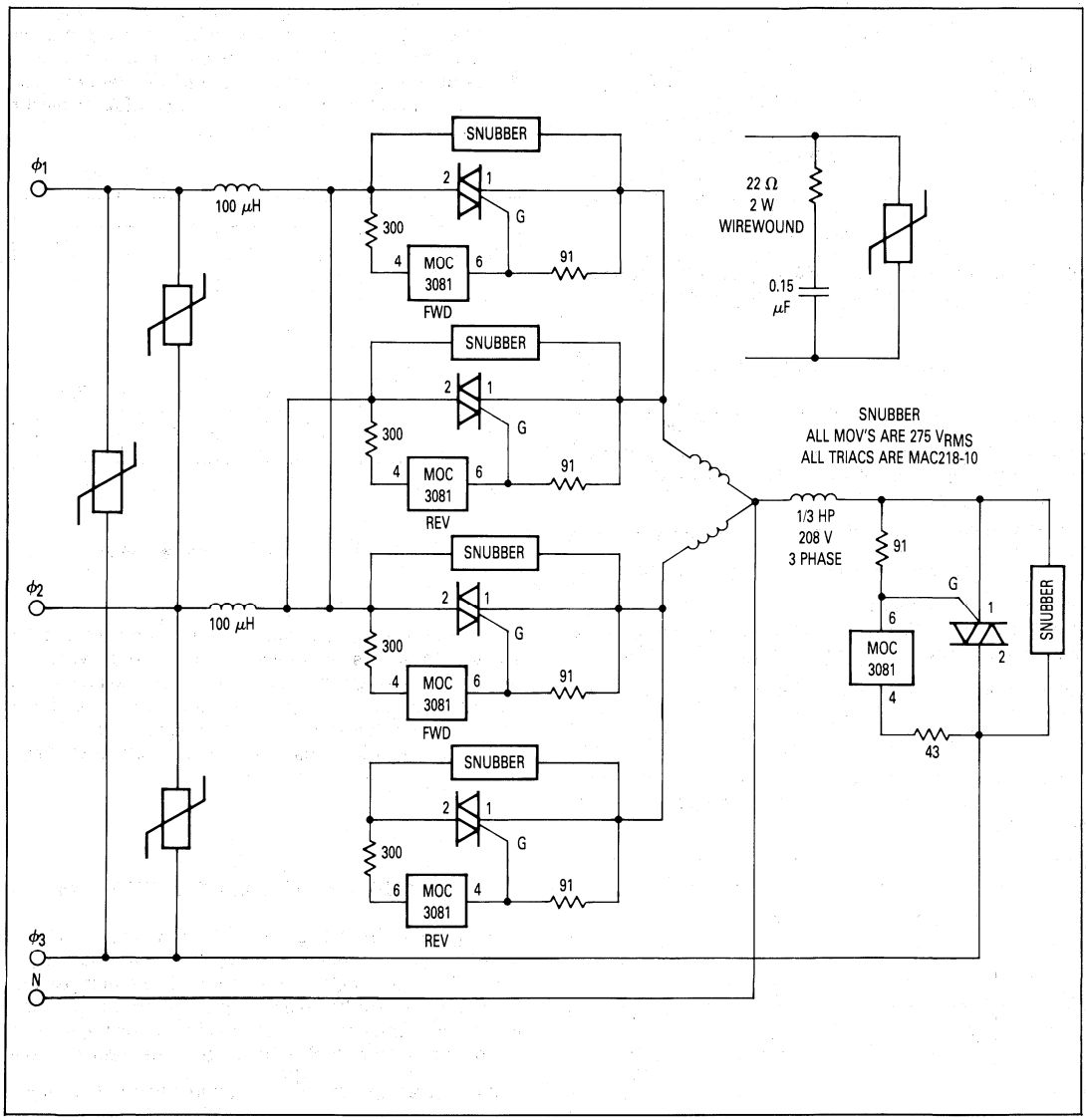


Figure 37. 3 Phase Reversing Motor

Figure 38 shows a split phase capacitor-run motor with reversing accomplished by switching the capacitor in series with one or the other winding. The forward and reverse TRIACs function as a SPDT switch. Reversing the motor applies the voltage on the capacitor abruptly across the blocking thyristor. Again, the inductor L is added to prevent $\left(\frac{dV}{dt}\right)_s$ firing of the blocking TRIAC. If turn-on occurs, the forward and reverse TRIACs short the capacitors (C_S) resulting in damage to them. It is wise to add the resistor R_S to limit the discharge current.

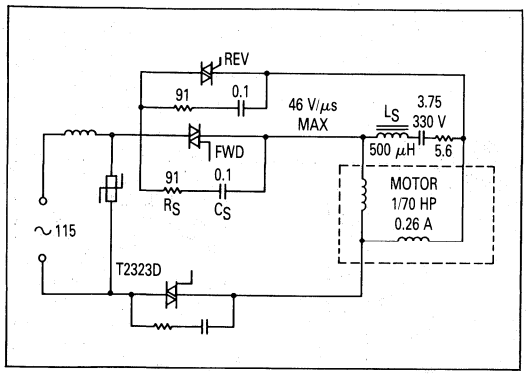


Figure 38. Split Phase Reversing Motor

Figure 39 shows a "tap changer." This circuit allows the operation of switching power supplies from a 120 or 240 vac line. When the TRIAC is on, the circuit functions as a conventional voltage doubler with diodes D1 and D2 conducting on alternate half-cycles. In this mode of operation, inrush current and $\frac{dI}{dt}$ are hazards to TRIAC reliability. Series impedance is necessary to prevent damage to the TRIAC.

The TRIAC is off when the circuit is not doubling. In this state, the TRIAC sees the difference between the line voltage and the voltage at the intersection of C1 and C2. Transients on the line cause $\left(\frac{dV}{dt}\right)_s$ firing of the TRIAC. High inrush current, $\frac{dI}{dt}$, and overvoltage damage to the filter capacitor are possibilities. Prevention requires the addition of a RC snubber across the TRIAC and an inductor in series with the line.

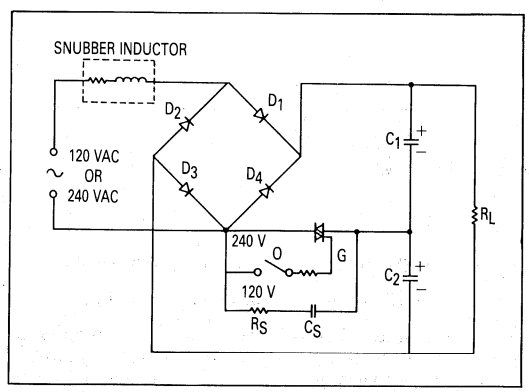


Figure 39. Tap Changer For Dual Voltage Switching Power Supply

THYRISTOR TYPES

Table 3 describes $\frac{dV}{dt}$ ratings for some thyristors.

Sensitive gate thyristors are easy to turn-on because of their low trigger current requirements. However, they have less $\frac{dV}{dt}$ capability than similar non-sensitive devices. A non-sensitive thyristor should be used for high $\frac{dV}{dt}$.

TRIAC commutating $\frac{dV}{dt}$ ratings are 5 to 20 times less than static $\frac{dV}{dt}$ ratings.

Phase controllable optocouplers have lower $\frac{dV}{dt}$ ratings than zero crossing optocouplers and power TRIACs. These should be used when a dc voltage component is present, or to prevent turn-on delay.

Zero crossing optocouplers have more $\frac{dV}{dt}$ capability than power thyristors; and they should be used in place of phase controllable devices in static switching applications.

Table 3. Thyristor $\frac{dV}{dt}$ Ratings

Device	Description	$\left(\frac{dV}{dt}\right)_s$ V/μs	$\left(\frac{dV}{dt}\right)_c$ V/μs
MAC228	8.0 A sensitive gate TRIAC	25 typical	5 typical
MAC213	12 A TRIAC	200 minimum	
MAC16	15 A TRIAC	200 minimum	
MAC321	20 A TRIAC	200 minimum	
MAC635	35 A TRIAC	100 minimum	6 minimum
MOC3041	0 Crossing TRIAC optocoupler	1000 minimum	
MOC3020	Phase controllable optocoupler	10 typical	
MCR72	8.0 A sensitive gate SCR	10 typical	
MCR218	8.0 A SCR	100 typical	

APPENDIX A
TESTING SNUBBER DISCHARGE $\frac{dI}{dt}$

The equations in Appendix D do not consider the thyristor's turn-on time or on-state resistance, thus, they predict high values of $\frac{dI}{dt}$.

Figure 40 shows the circuit used to test snubber discharge $\frac{dI}{dt}$. A MBS4991 supplies the trigger pulse while

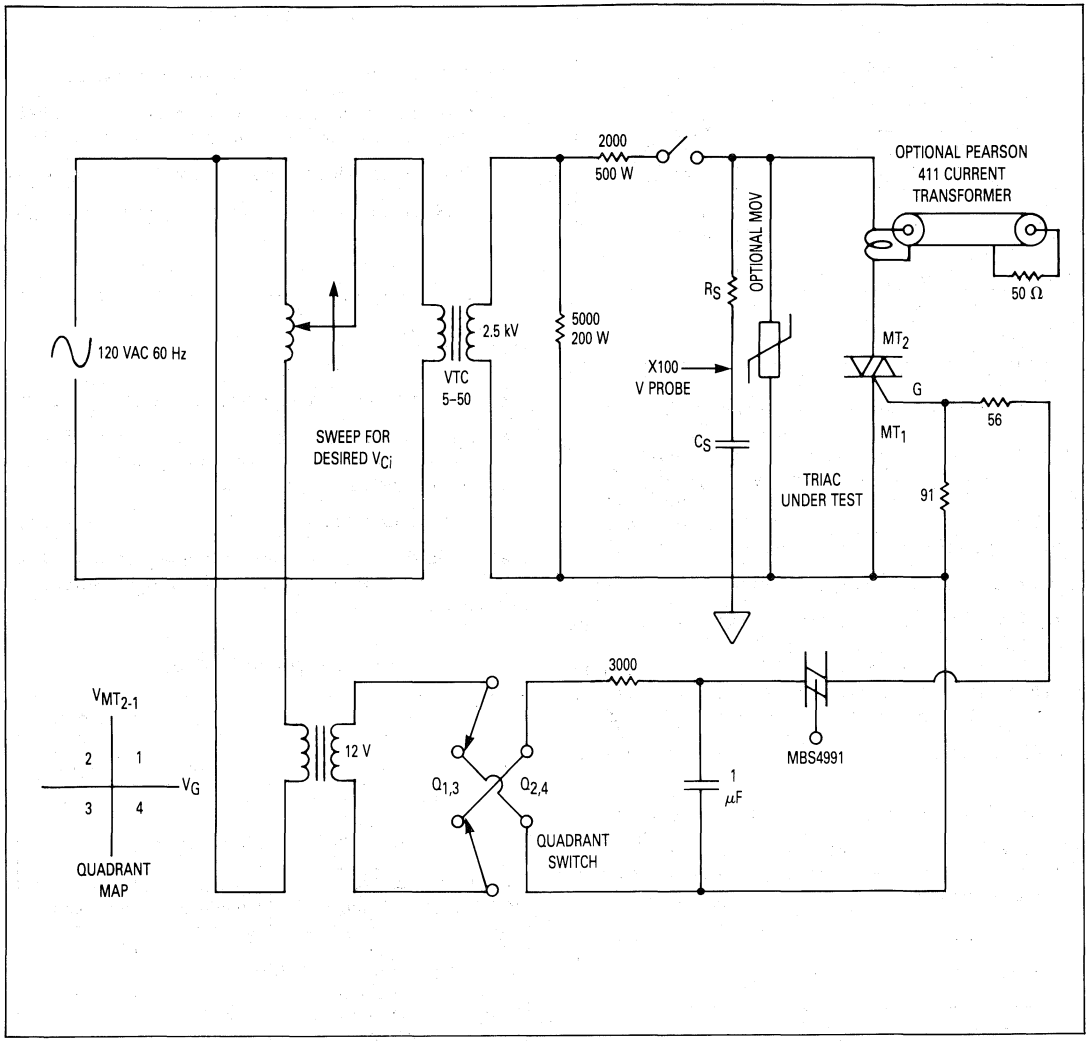


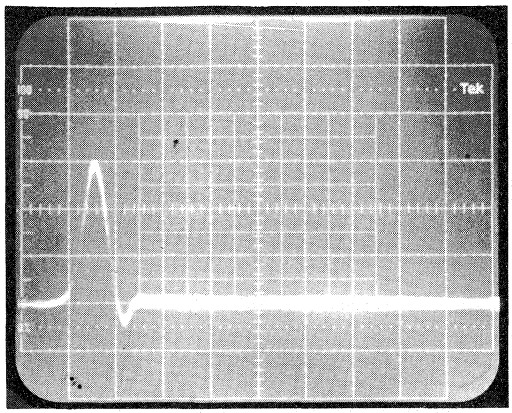
Figure 40. Snubber Discharge dI/dt Test

the quadrants of operation are switch selectable. The snubber was mounted as close to the TRIAC under test as possible to reduce inductance, and the current transformer remained in the circuit to allow results to be compared with the measured $\frac{dI}{dt}$ value.

What should the peak capacitor voltage be? A conservative approach is to test at maximum rated V_{DRM}, or the clamp voltage of the MOV.

What is the largest capacitor that can be used without limiting resistance? Figure 41 is a photo showing the current pulse resulting from a 0.001 μF capacitor charged to 800 V. The 1200 A/μs $\frac{dI}{dt}$ destroyed the TRIAC.

Is it possible for MOV self-capacitance to damage the TRIAC? A large 40 Joule, 2200 A peak current rated MOV was tested. The MOV measured 440 pF and had an 878 volt breakover voltage. Its peak discharge current (12 A) was half that of a 470 pF capacitor. This condition was safe.



HORIZONTAL SCALE — 50 ms/DIV.
VERTICAL SCALE — 10 A/DIV.
C_S = 0.001 μF, V_{Ci} = 800 V, R_S = 0, L = 250 mH, R_{TRIAC} = 10 OHMS

Figure 41. Discharge Current From 0.001 μF Capacitor

APPENDIX B

MEASURING $\left(\frac{dV}{dt}\right)_S$

Figure 42 shows a test circuit for measuring the static $\frac{dV}{dt}$ of power thyristors. A 1000 volt FET switch insures that the voltage across the device under test (D.U.T.) rises rapidly from zero. A differential preamp allows the use of a N-channel device while keeping the storage scope chassis at ground for safety purposes. The rate of voltage rise is adjusted by a variable RC time constant. The charging resistance is low to avoid waveform distortion because of the thyristor's self-capacitance but is large enough to prevent damage to the D.U.T. from turn-on $\frac{dI}{dt}$. Mounting the miniature range switches, capacitors, and G-K network close to the device under test reduces stray inductance and allows testing at more than 10 kV/μs.

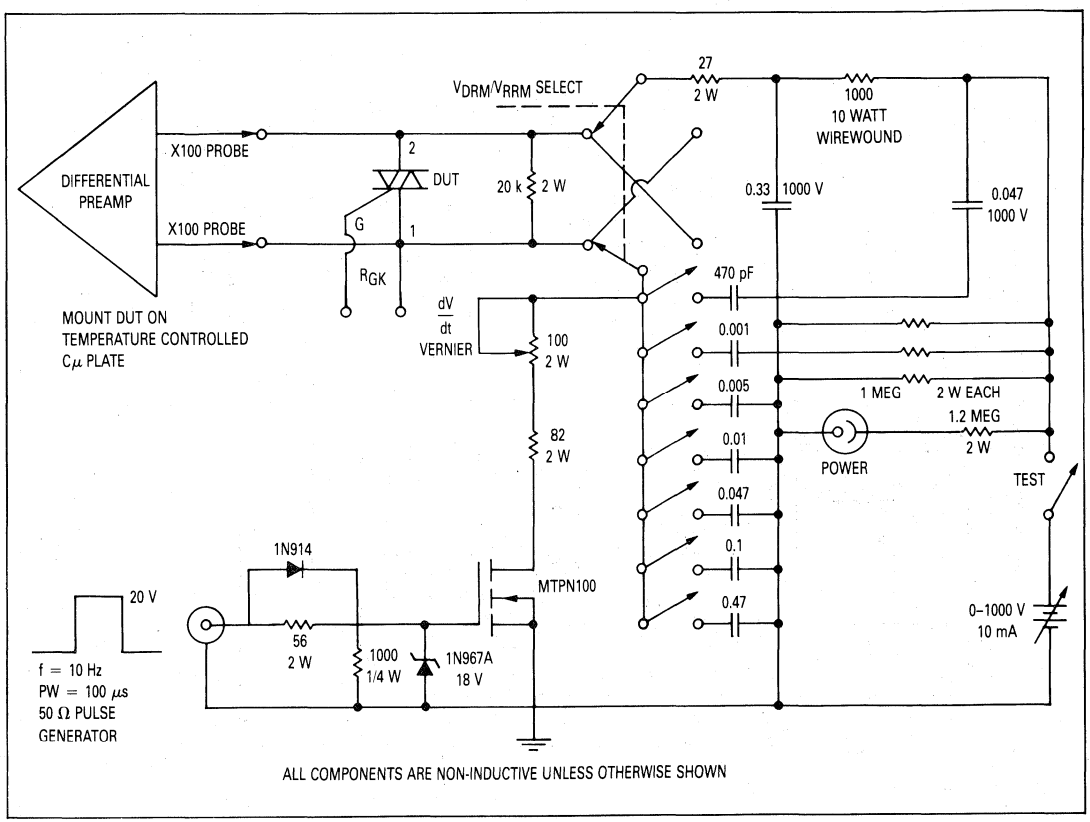


Figure 42. Circuit For Static $\frac{dV}{dt}$ Measurement of Power Thyristors

APPENDIX C

MEASURING $\left(\frac{dV}{dt}\right)_c$

A test fixture to measure commutating $\frac{dV}{dt}$ is shown in Figure 43. It is a capacitor discharge circuit with the load series resonant. The single pulse test aids temperature control and allows the use of lower power components. The limited energy in the load capacitor reduces burn and shock hazards. The conventional load and snubber circuit provides recovery and damping behaviors like those in the application.

The voltage across the load capacitor triggers the D.U.T. It terminates the gate current when the load capacitor voltage crosses zero and the TRIAC current is at its peak.

Each V_{DRM}, I_{TM} combination requires different components. Calculate their values using the equations given in Figure 43.

Commercial chokes simplify the construction of the necessary inductors. Their inductance should be adjusted by increasing the air gap in the core. Removal of the magnetic pole piece reduces inductance by 4 to 6 but extends the current without saturation.

The load capacitor consists of a parallel bank of 1500 Vdc non-polar units, with individual bleeders mounted at each capacitor for safety purposes.

An optional adjustable voltage clamp prevents TRIAC breakdown.

To measure $\left(\frac{dV}{dt}\right)_c$, synchronize the storage scope on the current waveform and verify the proper current amplitude and period. Increase the initial voltage on the capacitor to compensate for losses within the coil if necessary. Adjust the snubber until the device fails to turn off after the first half-cycle. Inspect the rate of voltage rise at the fastest passing condition.

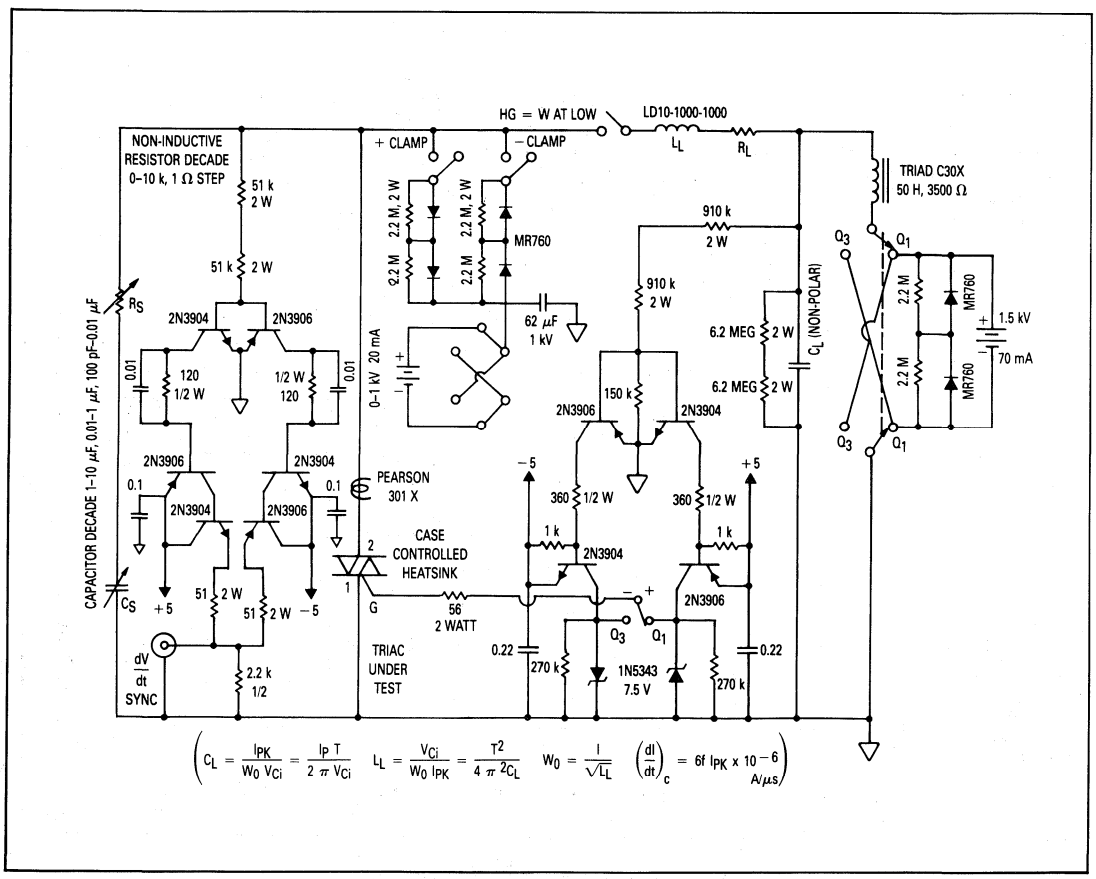


Figure 43. $\left(\frac{dV}{dt}\right)_c$ Test Circuit For Power TRIACs

APPENDIX D
$\frac{dV}{dt}$ DERIVATIONS

DEFINITIONS

1.0 $R_T = R_L + R_S$ = Total Resistance

1.1 $M = \frac{R_S}{R_T}$ = Snubber Divider Ratio

1.2 $\omega_0 = \frac{1}{\sqrt{L C_S}}$ = Undamped Natural Frequency

ω = Damped Natural Frequency

1.3 $\alpha = \frac{R_T}{2L}$ = Wave Decrement Factor

1.4 $\chi^2 = \frac{1/2\ LI^2}{1/2\ CV^2} = \frac{\text{Initial Energy In Inductor}}{\text{Final Energy In Capacitor}}$

1.5 $\chi = \frac{I}{E}\sqrt{\frac{L}{C}}$ = Initial Current Factor

1.6 $\rho = \frac{R_T}{2}\sqrt{\frac{C}{L}} = \frac{\alpha}{\omega_0}$ = Damping Factor

1.7 $V_{0L} = E - R_S I$ = Initial Voltage drop at t = 0 across the load

1.8 $\xi = \frac{I}{C_S} - \frac{E R_L}{L}$

$\left(\frac{dV}{dt}\right)_0$ = Initial instantaneous $\frac{dV}{dt}$ at t = 0, ignoring any initial instantaneous voltage step at t = 0 because of I_{RRM}

1.9 $\left(\frac{dV}{dt}\right)_0 = V_{OL}\frac{R_T}{L} + \xi$. For all damping conditions

2.0 When I = 0, $\left(\frac{dV}{dt}\right)_0 = \frac{E R_S}{L}$

$\left(\frac{dV}{dt}\right)_{max}$ = Maximum instantaneous $\frac{dV}{dt}$

t_{max} = Time of maximum instantaneous $\frac{dV}{dt}$

t_{peak} = Time of maximum instantaneous peak voltage across thyristor

Average $\frac{dV}{dt} = V_{PK}/t_{PK}$ = Slope of the secant line from t = 0 through V_{PK}

V_{PK} = Maximum instantaneous voltage across the thyristor.

CONSTANTS (depending on the damping factor):

2.1 No Damping ($\rho = 0$)

$\omega = \omega_0$

$R_T = \alpha = \rho = 0$

2.2 Underdamped ($0 < \rho < 1$)

$\omega = \sqrt{\omega_0^2 - \alpha^2} = \omega_0\sqrt{1-\rho^2}$

2.3 Critical Damped ($\rho = 1$)

$\alpha = \omega_0,\quad \omega = 0,\quad R = 2\sqrt{\frac{L}{C}},\quad C = \frac{2}{\alpha R_T}$

2.4 Overdamped ($\rho > 1$)

$\omega = \sqrt{\alpha^2 - \omega_0^2} = \omega_0\sqrt{\rho^2 - 1}$

Laplace transforms for the current and voltage in Figure 44 are:

3.0 $i(S) = \frac{E/L + SI}{S^2 + S\frac{R_T}{L} + \frac{1}{LC}};\quad e = \frac{E}{S} - \frac{S V_{0L} - \xi}{S^2 + \frac{R_T}{L}S + \frac{1}{LC}}$

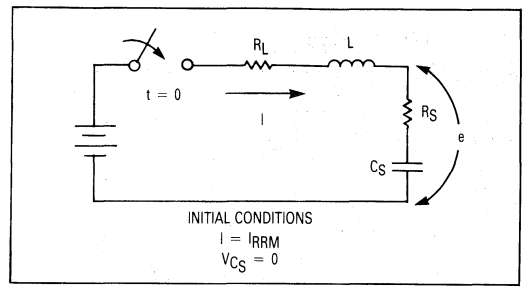


Figure 44. Equivalent Circuit for Load and Snubber

The inverse laplace transform for each of the conditions gives:

UNDERDAMPED (Typical Snubber Design)

4.0 $e = E - V_{0L}\left[\cos(\omega t) - \frac{\alpha}{\omega}\sin(\omega t)\right]e^{-\alpha t} + \frac{\xi}{\omega}\sin(\omega t)\,e^{-\alpha t}$

4.1 $\frac{de}{dt} = V_{0L}\left[2\alpha\cos(\omega t) + \frac{(\omega^2 - \alpha^2)}{\omega}\sin(\omega t)\right]e^{-\alpha t} + \xi\left[\cos(\omega t) - \frac{\alpha}{\omega}\sin(\omega t)\right]e^{-\alpha t}$

4.2 $t_{PK} = \frac{1}{\omega}\tan^{-1}\left[-\frac{2\alpha V_{0L} + \xi}{V_{0L}\left(\frac{\omega^2 - \alpha^2}{\omega}\right) - \frac{\xi\alpha}{\omega}}\right]$

When M = 0, $R_S = 0$, I = 0: $\omega t_{PK} = \pi$

4.3 $V_{PK} = E + \frac{\alpha}{\omega_0}{}^{-\alpha t_{PK}}\sqrt{\omega_0^2 V_{0L}^2 + 2\alpha\xi V_{0L} + \xi^2}$

When I = 0, $R_L = 0$, M = 1:

4.4 $\frac{V_{PK}}{E} = (1 + e^{-\alpha t_{PK}})$

Average $\frac{dV}{dt} = \frac{V_{PK}}{t_{PK}}$

4.5 $t_{max} = \frac{1}{\omega} ATN \left[\frac{\omega (2\alpha\xi - V_{0L} (\omega^2 - 3\alpha^2))}{V_{0L} (\alpha^3 - 3\alpha\omega^2) + \xi(\alpha^2 - \omega^2)} \right]$

4.6 $\left(\frac{dV}{dt}\right)_{max} = \sqrt{V_{0L}^2 \omega_0^2 + 2\alpha\xi V_{0L} + \xi^2} \quad e^{-\alpha t_{max}}$

NO DAMPING

5.0 $e = E (1 - Cos (\omega_0 t)) + \frac{I}{C \omega_0} \sin (\omega_0 t)$

5.1 $\frac{de}{dt} = E \omega_0 \sin (\omega_0 t) + \frac{I}{C} Cos (\omega_0 t)$

5.2 $\left(\frac{dV}{dt}\right)_0 = \frac{I}{C} = 0$ when $I = 0$

5.3 $t_{PK} = \frac{\pi - \tan^{-1}\left(\frac{I}{CE \omega_0}\right)}{\omega_0}$

5.4 $V_{PK} = E + \sqrt{E^2 + \frac{I^2}{\omega_0^2 C^2}}$

5.5 $\left(\frac{dV}{dt}\right)_{AVG} = \frac{V_{PK}}{t_{PK}}$

5.6 $t_{max} = \frac{1}{\omega_0} \left[\tan^{-1}\left(\frac{\omega_0 EC}{I}\right) \right] = \frac{1}{\omega_0} \frac{\pi}{2}$ when $I = 0$

5.7 $\left(\frac{dV}{dt}\right)_{max} = \frac{I}{C} \sqrt{E^2 \omega_0^2 C^2 + I^2} = \omega_0 E$ when $I = 0$

CRITICAL DAMPING

6.0 $e = E - V_{0L} (1 - \alpha t) e^{-\alpha t} + \xi t e^{-\alpha t}$

6.1 $\frac{de}{dt} = \left[\alpha V_{0L} (2 - \alpha t) + \xi (1 - \alpha t) \right] e^{-\alpha t}$

6.2 $t_{PK} = \frac{2 + \frac{\xi}{2 V_{0L}}}{\alpha + \frac{\xi}{V_{0L}}}$

6.3 $V_{PK} = E - \left[V_{0L} (1 - \alpha t_{PK}) - \xi t_{PK} \right] e^{-\alpha t_{PK}}$

6.4 Average $\frac{dV}{dt} = \frac{V_{PK}}{t_{PK}}$

When $I = 0$, $R_S = 0$, $M = 0$

e(t) rises asymptotically to E. t_{PK} and average $\frac{dV}{dt}$ do not exist.

6.5 $t_{max} = \frac{3\alpha V_{0L} + 2\xi}{\alpha^2 V_{0L} + \alpha\xi}$

When $I = 0$, $t_{max} = 0$

For $\frac{R_S}{R_T} \geqq 3/4$,

then $\frac{dV}{dt}_{max} = \left(\frac{dV}{dt}\right)_0$

6.6 $\left(\frac{dV}{dt}\right)_{max}$

$= \left[\alpha V_{0L} (2 - \alpha t_{max}) + \xi (1 - \alpha t_{max}) \right] e^{-\alpha t_{max}}$

APPENDIX E
SNUBBER DISCHARGE $\frac{dI}{dt}$ DERIVATIONS

OVERDAMPED

1.0 $i = \frac{V_{CS}}{\omega L_S} \alpha^{-\alpha t} \sinh (\omega t)$

1.1 $i_{PK} = V_{CS} \sqrt{\frac{C_S}{L_S}} \quad e^{-\alpha t_{PK}}$

1.2 $t_{PK} = \frac{1}{\omega} \tanh^{-1} \left[\frac{\omega}{\alpha} \right]$

CRITICAL DAMPED

2.0 $i = \frac{V_{CS}}{L_S} t e^{-\alpha t}$

2.1 $i_{PK} = 0.736 \frac{V_{CS}}{R_S}$

2.2 $t_{PK} = \frac{1}{\alpha}$

UNDERDAMPED

3.0 $i = \frac{V_{CS}}{\omega L_S} e^{-\alpha t} \sin (\omega t)$

3.1 $i_{PK} = V_{CS} \sqrt{\frac{C_S}{L_S}} e^{-\alpha t_{PK}}$

3.2 $t_{PK} = \frac{1}{\omega} \tan^{-1} \left(\frac{\omega}{\alpha} \right)$

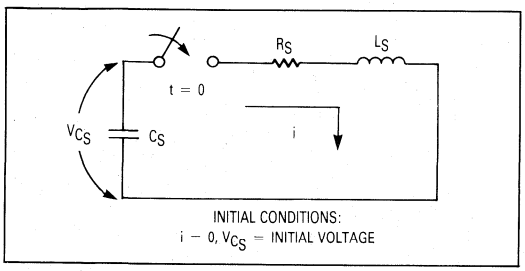


Figure 45. Equivalent Circuit for Snubber Discharge

NO DAMPING

4.0 $i = \frac{V_{CS}}{\omega L_S} \sin(\omega t)$

4.1 $i_{PK} = V_{CS} \sqrt{\frac{C_S}{L_S}}$

4.2 $t_{PK} = \frac{\pi}{2\omega}$

1

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CHAPTER 7
MOUNTING TECHNIQUES FOR THYRISTORS

INTRODUCTION

Current and power ratings of semiconductors are inseparably linked to their thermal environment. Except for lead-mounted parts used at low currents, a heat exchanger is required to prevent the junction temperature from exceeding its rated limit, thereby running the risk of a high failure rate. Furthermore, the semiconductor industry's field history indicated that the failure rate of most silicon semiconductors decreases approximately by one-half for a decrease in junction temperature from 160°C to 135°C.(1) Guidelines for designers of military power supplies impose a 110°C limit upon junction temperature.(2) Proper mounting minimizes the temperature gradient between the semiconductor case and the heat exchanger.

Most early life field failures of power semiconductors can be traced to faulty mounting procedures. With metal packaged devices, faulty mounting generally causes unnecessarily high junction temperature, resulting in reduced component lifetime, although mechanical damage has occurred on occasion from improperly mounting to a warped surface. With the widespread use of various plastic-packaged semiconductors, the prospect of mechanical damage is very significant. Mechanical damage can impair the case moisture resistance or crack the semiconductor die.

Figure 1 shows an example of doing nearly everything wrong. A tab mount TO-220 package is shown being used as a replacement for a TO-213AA (TO-66) part which was socket mounted. To use the socket, the leads are bent — an operation which, if not properly done, can crack the package, break the internal bonding wires, or crack the die. The package is fastened with a sheet-metal screw through a 1/4″ hole containing a fiber-insulating sleeve. The force used to tighten the screw tends to pull the package into the hole, possibly causing enough distortion to crack the die. In addition the contact area is small because of the area consumed by the large hole and the bowing of the package; the result is a much higher junction temperature than expected. If a rough heatsink surface and/or burrs around the hole were displayed in the illustration, most but not all poor mounting practices would be covered.

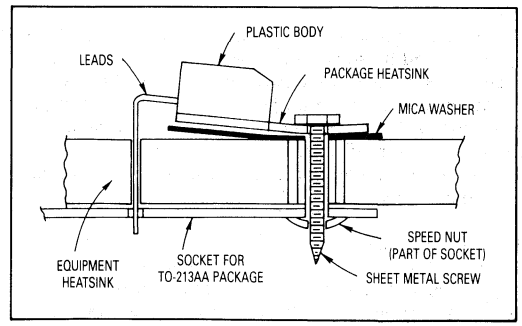

Figure 1. Extreme Case of Improperly Mounting A Semiconductor (Distortion Exaggerated)

(1) MIL-HANDBOOK — 2178, SECTION 2.2.
(2) "Navy Power Supply Reliability — Design and Manufacturing Guidelines" NAVMAT P4855-1, Dec. 1982 NAVPUBFORCEN, 5801 Tabor Ave., Philadelphia, PA 19120.

Cho-Therm is a registered trademark of Chromerics, Inc.
Grafoil is a registered trademark of Union Carbide
Kapton is a registered trademark of E.I. Dupont
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In many situations the case of the semiconductor must be electrically isolated from its mounting surface. The isolation material is, to some extent, a thermal isolator as well, which raises junction operating temperatures. In addition, the possibility of arc-over problems is introduced if high voltages are present. Various regulating agencies also impose creepage distance specifications which further complicates design. Electrical isolation thus places additional demands upon the mounting procedure.

Proper mounting procedures usually necessitate orderly attention to the following:

1. Preparing the mounting surface
2. Applying a thermal grease (if required)
3. Installing the insulator (if electrical isolation is desired)
4. Fastening the assembly
5. Connecting the terminals to the circuit

In this note, mounting procedures are discussed in general terms for several generic classes of packages. As newer packages are developed, it is probable that they will fit into the generic classes discussed in this note. Unique requirements are given on data sheets pertaining to the particular package. The following classes are defined:

Stud Mount
Flange Mount
Pressfit
Plastic Body Mount
Tab Mount
Surface Mount

Appendix A contains a brief review of thermal resistance concepts. Appendix B discusses measurement difficulties with interface thermal resistance tests. Appendix C indicates the type of accessories supplied by a number of manufacturers.

MOUNTING SURFACE PREPARATION

In general, the heatsink mounting surface should have a flatness and finish comparable to that of the semiconductor package. In lower power applications, the heatsink surface is satisfactory if it appears flat against a straight edge and is free from deep scratches. In high-power applications, a more detailed examination of the surface is required. Mounting holes and surface treatment must also be considered.

Surface Flatness

Surface flatness is determined by comparing the variance in height (Δh) of the test specimen to that of a reference standard as indicated in Figure 2. Flatness is normally specified as a fraction of the Total Indicator Reading (TIR). The mounting surface flatness, i.e, Δh/TIR, if less than 4 mils per inch, normal for extruded aluminum, is satisfactory in most cases.

Surface Finish

Surface finish is the average of the deviations both above and below the mean value of surface height. For minimum interface resistance, a finish in the range of 50 to 60 microinches is satisfactory; a finer finish is costly to achieve and does not significantly lower contact resistance. Tests conducted by Thermalloy using a copper TO-204 (TO-3) package with a typical 32-microinch finish, showed that heatsink finishes between 16 and 64 μ-in caused less than $\pm 2.5\%$ difference in interface thermal resistance when the voids and scratches were filled with a thermal joint compound.(3) Most commercially available cast or extruded heatsinks will require spotfacing when used in high-power applications. In general, milled or machined surfaces are satisfactory if prepared with tools in good working condition.

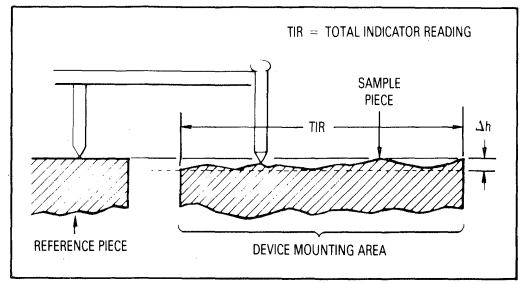


Figure 2. Surface Flatness Measurement

Mounting Holes

Mounting holes generally should only be large enough to allow clearance of the fastener. The larger thick flange type packages having mounting holes removed from the semiconductor die location, such as the TO-3, may successfully be used with larger holes to accommodate an insulating bushing, but many plastic encapsulated packages are intolerant of this condition. For these packages, a smaller screw size must be used such that the hole for the bushing does not exceed the hole in the package.

Punched mounting holes have been a source of trouble because if not properly done, the area around a punched hole is depressed in the process. This "crater" in the heatsink around the mounting hole can cause two problems. The device can be damaged by distortion of the package as the mounting pressure attempts to conform it to the shape of the heatsink indentation, or the device may only bridge the crater and leave a significant percentage of its heat-dissipating surface out of contact with the heatsink. The first effect may often be detected immediately by visual cracks in the package (if plastic), but usually an unnatural stress is imposed, which results in an early-life failure. The second effect results in hotter operation and is not manifested until much later.

Although punched holes are seldom acceptable in the relatively thick material used for extruded aluminum heatsinks, several manufacturers are capable of properly utilizing the capabilities inherent in both fine-edge blanking or sheared-through holes when applied to sheet metal as commonly used for stamped heatsinks. The holes are pierced using Class A progressive dies mounted on four-post die sets equipped with proper pressure pads and holding fixtures.

(3) Catalog #87-HS-9 (1987), page 8, Thermalloy, Inc., P.O. Box 810839, Dallas, Texas 75381-0839.

When mounting holes are drilled, a general practice with extruded aluminum, surface cleanup is important. Chamfers must be avoided because they reduce heat transfer surface and increase mounting stress. However, the edges must be broken to remove burrs which cause poor contact between device and heatsink and may puncture isolation material.

Surface Treatment

Many aluminum heatsinks are black-anodized to improve radiation ability and prevent corrosion. Anodizing results in significant electrical but negligible thermal insulation. It need only be removed from the mounting area when electrical contact is required. Heatsinks are also available which have a nickel plated copper insert under the semiconductor mounting area. No treatment of this surface is necessary.

Another treated aluminum finish is iridite, or chromate-acid dip, which offers low resistance because of its thin surface, yet has good electrical properties because it resists oxidation. It need only be cleaned of the oils and films that collect in the manufacture and storage of the sinks, a practice which should be applied to all heatsinks.

For economy, paint is sometimes used for sinks; removal of the paint where the semiconductor is attached is usually required because of paint's high thermal resistance. However, when it is necessary to insulate the semiconductor package from the heatsink, hard anodized or painted surfaces allow an easy installation for low voltage applications. Some manufacturers will provide anodized or painted surfaces meeting specific insulation voltage requirements, usually up to 400 volts.

It is also necessary that the surface be free from all foreign material, film, and oxide (freshly bared aluminum forms an oxide layer in a few seconds). Immediately prior to assembly, it is a good practice to polish the mounting area with No. 000 steel wool, followed by an acetone or alcohol rinse.

INTERFACE DECISIONS

When any significant amount of power is being dissipated, something must be done to fill the air voids between mating surfaces in the thermal path. Otherwise the interface thermal resistance will be unnecessarily high and quite dependent upon the surface finishes.

For several years, thermal joint compounds, often called grease, have been used in the interface. They have a resistivity of approximately 60°C/W/in whereas air has 1200°C/W/in. Since surfaces are highly pock-marked with minute voids, use of a compound makes a significant reduction in the interface thermal resistance of the joint. However, the grease causes a number of problems, as discussed in the following section.

To avoid using grease, manufacturers have developed dry conductive and insulating pads to replace the more traditional materials. These pads are conformal and therefore partially fill voids when under pressure.

Thermal Compounds (Grease)

Joint compounds are a formulation of fine zinc or other conductive particles in a silicone oil or other synthetic base fluid which maintains a grease-like consistency with time and temperature. Since some of these compounds do not spread well, they should be evenly applied in a very thin layer using a spatula or lintless brush, and wiped lightly to remove excess material. Some cyclic rotation of the package will help the compound spread evenly over the entire contact area. Some experimentation is necessary to determine the correct quantity; too little will not fill all the voids, while too much may permit some compound to remain between well mated metal surfaces where it will substantially increase the thermal resistance of the joint.

To determine the correct amount, several semiconductor samples and heatsinks should be assembled with different amounts of grease applied evenly to one side of each mating surface. When the amount is correct a very small amount of grease should appear around the perimeter of each mating surface as the assembly is slowly torqued to the recommended value. Examination of a dismantled assembly should reveal even wetting across each mating surface. In production, assemblers should be trained to slowly apply the specified torque even though an excessive amount of grease appears at the edges of mating surfaces. Insufficient torque causes a significant increase in the thermal resistance of the interface.

To prevent accumulation of airborne particulate matter, excess compound should be wiped away using a cloth moistened with acetone or alcohol. These solvents should not contact plastic-encapsulated devices, as they may enter the package and cause a leakage path or carry in substances which might attack the semiconductor chip.

The silicone oil used in most greases has been found to evaporate from hot surfaces with time and become deposited on other cooler surfaces. Consequently, manufacturers must determine whether a microscopically thin coating of silicone oil on the entire assembly will pose any problems. It may be necessary to enclose components using grease. The newer synthetic base greases show far less tendency to migrate or creep than those made with a silicone oil base. However, their currently observed working temperature range are less, they are slightly poorer on thermal conductivity and dielectric strength and their cost is higher.

Data showing the effect of compounds on several package types under different mounting conditions is shown in Table 1. The rougher the surface, the more valuable the grease becomes in lowering contact resistance; therefore, when mica insulating washers are used, use of grease is generally mandatory. The joint compound also improves the breakdown rating of the insulator.

Conductive Pads

Because of the difficulty of assembly using grease and the evaporation problem, some equipment manufacturers will not, or cannot, use grease. To minimize the need for grease, several vendors offer dry conductive pads which approximate performance obtained with grease. Data for a greased bare joint and a joint using Grafoil, a dry graphite compound, is shown in the data of Figure 3. Grafoil is claimed to be a replacement for grease when no electrical isolation is required; the data indicates it does indeed perform as well as grease. Another conductive pad available from Aavid is called KON-DUX. It is made with a unique, grain oriented, flake-like structure (patent pending). Highly compressible, it becomes

Table 1
Approximate Values for Interface Thermal Resistance Data from Measurements Performed in Motorola Applications Engineering Laboratory

Dry interface values are subject to wide variation because of extreme dependence upon surface conditions. Unless otherwise noted the case temperature is monitored by a thermocouple located directly under the die reached through a hole in the heatsink. (See Appendix B for a discussion of Interface Thermal Resistance Measurements.)

Package Type and Data		Interface Thermal Resistance (°C/W)						
JEDEC Outlines	Description	Test Torque In-Lb	Metal-to-Metal		With Insulator			See Note
			Dry	Lubed	Dry	Lubed	Type	
DO-203AA, TO-210AA TO-208AB	10-32 Stud 7/16″ Hex	15	0.3	0.2	1.6	0.8	3 mil Mica	
DO-203AB, TO-210AC TO-208	1/4-28 Stud 11/16″ Hex	25	0.2	0.1	0.8	0.6	5 mil Mica	
DO-208AA	Pressfit, 1/2″	—	0.15	0.1	—	—	—	
TO-204AA (TO-3)	Diamond Flange	6	0.5	0.1	1.3	0.36	3 mil Mica	1
TO-213AA (TO-66)	Diamond Flange	6	1.5	0.5	2.3	0.9	2 mil Mica	
TO-126	Thermopad 1/4″ x 3/8″	6	2.0	1.3	4.3	3.3	2 mil Mica	
TO-220AB	Thermowatt	8	1.2	1.0	3.4	1.6	2 mil Mica	1, 2

NOTES: 1. See Figures 3 and 4 for additional data on TO-3 and TO-220 packages.
2. Screw not insulated. See Figure 12.

formed to the surface roughness of both the heatsink and semiconductor. Manufacturer's data shows it to provide an interface thermal resistance better than a metal interface with filled silicone grease. Similar dry conductive pads are available from other manufacturers. They are a fairly recent development; long term problems, if they exist, have not yet become evident.

INSULATION CONSIDERATIONS

Since most power semiconductors use are vertical device construction it is common to manufacture power semiconductors with the output electrode (anode, collector or drain) electrically common to the case; the problem of isolating this terminal from ground is a common one. For lowest overall thermal resistance, which is quite important when high power must be dissipated, it is best to isolate the entire heatsink/semiconductor structure from ground, rather than to use an insulator between the semiconductor and the heatsink. Heatsink isolation is not always possible, however, because of EMI requirements, safety reasons, instances where a chassis serves as a heatsink or where a heatsink is common to several non-isolated packages. In these situations insulators are used to isolate the individual components from the heatsink. Newer packages, such as the Motorola Full Pak and EMS modules, contain the electrical isolation material within, thereby saving the equipment manufacturer the burden of addressing the isolation problem.

Insulator Thermal Resistance

When an insulator is used, thermal grease is of greater importance than with a metal-to-metal contact, because two interfaces exist instead of one and some materials, such as mica, have a hard, markedly uneven surface. With many isolation materials reduction of interface thermal resistance of between 2 to 1 and 3 to 1 are typical when grease is used.

Data obtained by Thermalloy, showing interface resistance for different insulators and torques applied to TO-204 (TO-3) and TO-220 packages, are shown in Figure 3, for bare and greased surfaces. Similar materials to those shown are available from several manufacturers. It is obvious that with some arrangements, the interface thermal resistance exceeds that of the semiconductor (junction to case).

Referring to Figure 3, one may conclude that when high power is handled, beryllium oxide is unquestionably the best. However, it is an expensive choice. (It should not be cut or abraided, as the dust is highly toxic.) Thermafilm is a filled polyimide material which is used for isolation (variation of Kapton). It is a popular material for low power applications because of its low cost ability to withstand high temperatures, and ease of handling in contrast to mica which chips and flakes easily.

A number of other insulating materials are also shown. They cover a wide range of insulation resistance, thermal resistance and ease of handling. Mica has been widely used in the past because it offers high breakdown voltage and fairly low thermal resistance at a low cost but it certainly should be used with grease.

Silicone rubber insulators have gained favor because they are somewhat conformal under pressure. Their ability to fill in most of the metal voids at the interface reduces the need for thermal grease. When first introduced, they suffered from cut-through after a few years in service. The ones presently available have solved this problem by having imbedded pads of Kapton or fiberglass. By

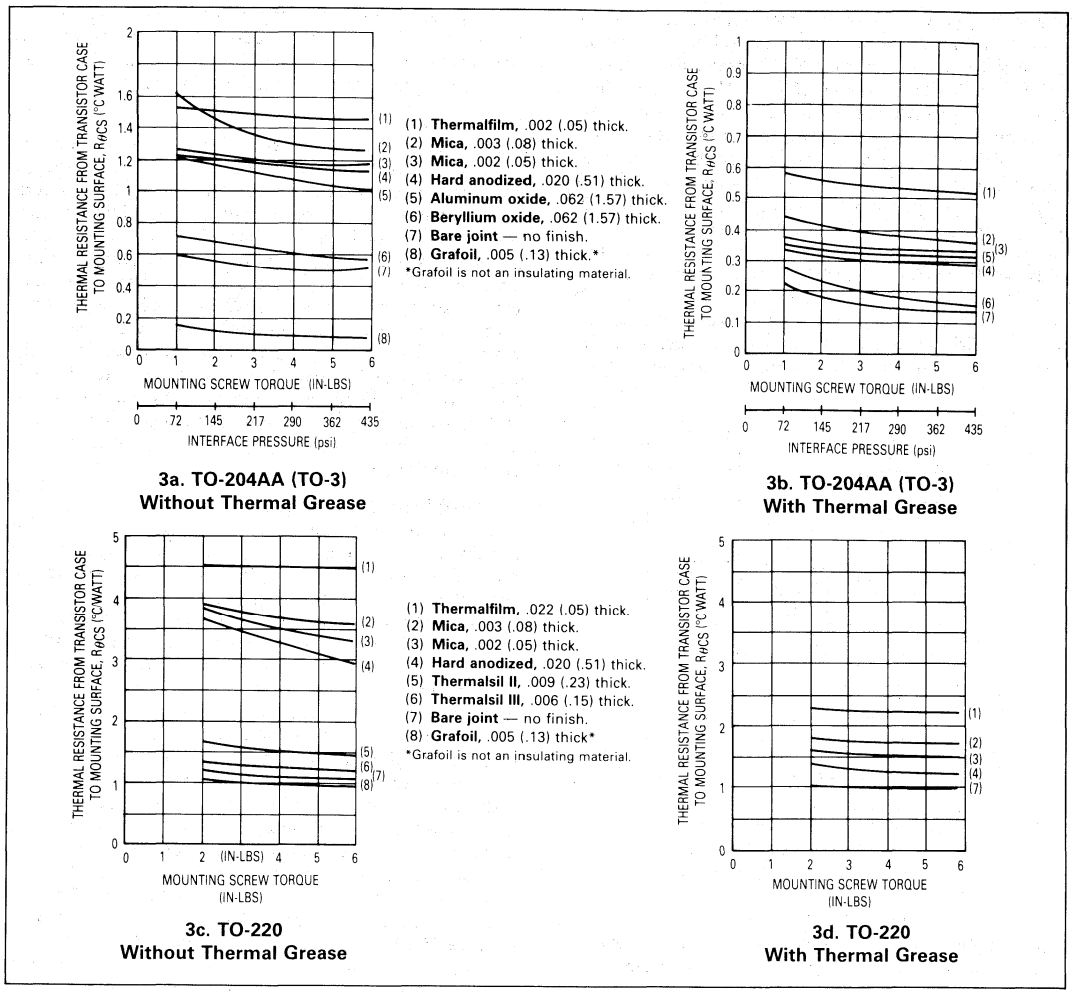


Figure 3. Interface Thermal Resistance for TO-204, TO-3 and TO-220 Packages using Different Insulating Materials as a Function of Mounting Screw Torque (Data Courtesy Thermalloy)

comparing Figures 3c and 3d, it can be noted that Thermasil, a filled silicone rubber, without grease, has about the same interface thermal resistance as greased mica for the TO-220 package.

A number of manufacturers offer silicone rubber insulators. Table 2 shows measured performance of a number of these insulators under carefully controlled, nearly identical conditions. The interface thermal resistance extremes are over 2:1 for the various materials. It is also clear that some of the insulators are much more tolerant than others of out-of-flat surfaces. Since the tests were performed, newer products have been introduced. The Bergquist K-10 pad, for example, is described as having about 2/3 the interface resistance of the Sil Pad 1000 which would place its performance close to the Chomerics 1671 pad. AAVID also offers an isolated pad called

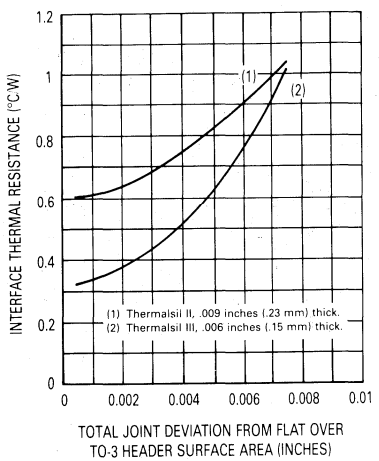
Table 2. Thermal Resistance of Silicone Rubber Pads

Manufacturer	Product	$R_{\theta CS}$ @ 3 Mils*	$R_{\theta CS}$ @ 7.5 Mils*
Wakefield	Delta Pad 173-7	.790	1.175
Bergquist	Sil Pad K-4	.752	1.470
Stockwell Rubber	1867	.742	1.015
Bergquist	Sil Pad 400-9	.735	1.205
Thermalloy	Thermalsil II	.680	1.045
Shin-Etsu	TC-30AG	.664	1.260
Bergquist	Sil Pad 400-7	.633	1.060
Chomerics	1674	.592	1.190
Wakefield	Delta Pad 174-9	.574	.755
Bergquist	Sil Pad 1000	.529	.935
Ablestik	Thermal Wafers	.500	.990
Thermalloy	Thermalsil III	.440	1.035
Chomerics	1671	.367	.655

*Test Fixture Deviation from flat from Thermalloy EIR86-1010.

Rubber-Duc, however it is only available vulcanized to a heatsink and therefore was not included in the comparison. Published data from AAVID shows $R_{\theta CS}$ below 0.3°C/W for pressures above 500 psi. However, surface flatness and other details are not specified so a comparison cannot be made with other data in this note.

The thermal resistance of some silicone rubber insulators is sensitive to surface flatness when used under a fairly rigid base package. Data for a TO-204AA (TO-3) package insulated with Thermasil is shown on Figure 4. Observe that the "worst case" encountered (7.5 mils) yields results having about twice the thermal resistance of the "typical case" (3 mils), for the more conductive insulator. In order for Thermasil III to exceed the performance of greased mica, total surface flatness must be under 2 mils, a situation that requires spot finishing.



Data courtesy of Thermalloy

Figure 4. Effect of Total Surface Flatness on Interface Resistance Using Silicon Rubber Insulators

Silicon rubber insulators have a number of unusual characteristics. Besides being affected by surface flatness and initial contact pressure, time is a factor. For example, in a study of the Cho-Therm 1688 pad thermal interface impedance dropped from 0.90°C/W to 0.70°C/W at the end of 1000 hours. Most of the change occurred during the first 200 hours where $R_{\theta CS}$ measured 0.74°C/W. The torque on the conventional mounting hardware had decreased to 3 in-lb from an initial 6 in-lb. With non-conformal materials, a reduction in torque would have increased the interface thermal resistance.

Because of the difficulties in controlling all variables affecting tests of interface thermal resistance, data from different manufacturers is not in good agreement. Table 3 shows data obtained from two sources. The relative performance is the same, except for mica which varies widely in thickness. Appendix B discusses the variables which need to be controlled. At the time of this writing ASTM Committee D9 is developing a standard for interface measurements.

The conclusions to be drawn from all this data is that some types of silicon rubber pads, mounted dry, will out perform the commonly used mica with grease. Cost may be a determining factor in making a selection.

Insulation Resistance

When using insulators, care must be taken to keep the mating surfaces clean. Small particles of foreign matter can puncture the insulation, rendering it useless or seriously lowering its dielectric strength. In addition, particularly when voltages higher than 300 V are encountered, problems with creepage may occur. Dust and other foreign material can shorten creepage distances significantly; so having a clean assembly area is important. Surface roughness and humidity also lower insulation resistance. Use of thermal grease usually raises the withstand voltage of the insulation system but excess must be removed to avoid collecting dust. Because of these factors, which are not amenable to analysis, hi-pot testing should be done on prototypes and a large margin of safety employed.

Insulated Electrode Packages

Because of the nuisance of handling and installing the accessories needed for an insulated semiconductor mounting, equipment manufacturers have longed for cost-effective insulated packages since the 1950's. The first to appear were stud mount types which usually have a layer of beryllium oxide between the stud hex and the can. Although effective, the assembly is costly and requires manual mounting and lead wire soldering to terminals on top of the case. In the late eighties, a number of electrically isolated parts became available from various semiconductor manufacturers. These offerings presently consist of multiple chips and integrated circuits as well as the more conventional single chip devices.

The newer insulated packages can be grouped into two categories. The first has insulation between the semiconductor chips and the mounting base; an exposed area of the mounting base is used to secure the part. The EMS (Energy Management Series) Modules, shown on Figure 8, Case 806 (ICePAK) and Case 388A (TO-258AA) (see Figure 11) are examples of parts in this category. The second category contains parts which have a plastic overmold covering the metal mounting base. The Full Pak,

Table 3. Performance of Silicon Rubber Insulators Tested per MIL-I-49456

Material	Measured Thermal Resistance (°C/W)	
	Thermalloy Data(1)	Berquist Data(2)
Bare Joint, greased	0.033	0.008
BeO, greased	0.082	—
Cho-Therm, 1617	0.233	—
Q Pad (non-insulated)	—	0.009
Sil-Pad, K-10	0.263	0.200
Thermasil III	0.267	—
Mica, greased	0.329	0.400
Sil-Pad 1000	0.400	0.300
Cho-therm 1674	0.433	—
Thermasil II	0.500	—
Sil-Pad 400	0.533	0.440
Sil-Pad K-4	0.583	0.440

(1) From Thermalloy EIR 87-1030
(2) From Berquist Data Sheet

Case 221C, illustrated in Figure 13, is an example of parts in the second category.

Parts in the first category — those with an exposed metal flange or tab — are mounted the same as their non-insulated counterparts. However, as with any mounting system where pressure is bearing on plastic, the overmolded type should be used with a conical compression washer, described later in this note.

FASTENER AND HARDWARE CHARACTERISTICS

Characteristics of fasteners, associated hardware, and the tools to secure them determine their suitability for use in mounting the various packages. Since many problems have arisen because of improper choices, the basic characteristics of several types of hardware are discussed next.

Compression Hardware

Normal split ring lock washers are not the best choice for mounting power semiconductors. A typical #6 washer flattens at about 50 pounds, whereas 150 to 300 pounds is needed for good heat transfer at the interface. A very useful piece of hardware is the conical, sometimes called a Belleville washer, compression washer. As shown in Figure 5, it has the ability to maintain a fairly constant pressure over a wide range of its physical deflection — generally 20% to 80%. When installing, the assembler applies torque until the washer depresses to half its original height. (Tests should be run prior to setting up the assembly line to determine the proper torque for the fastener used to achieve 50% deflection.) The washer will absorb any cyclic expansion of the package, insulating washer or other materials caused by temperature changes. Conical washers are the key to successful mounting of devices requiring strict control of the mounting force or when plastic hardware is used in the mounting scheme. They are used with the large face contacting the packages. A new variation of the conical washer includes it as part of a nut assembly. Called a Sync Nut, the patented device can be soldered to a PC board and the semiconductor mounted with a 6-32 machine screw.(4)

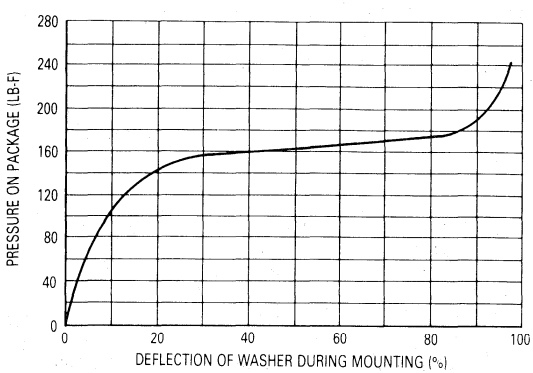


Figure 5. Characteristics of the Conical Compression Washers Designed for Use with Plastic Body Mounted Semiconductors

(4) ITW Shakeproof, St. Charles Road, Elgin, IL 60120.

Clips

Fast assembly is accomplished with clips. When only a few watts are being dissipated, the small board-mounted or free-standing heat dissipators with an integral clip, offered by several manufacturers, result in a low cost assembly. When higher power is being handled, a separate clip may be used with larger heatsinks. In order to provide proper pressure, the clip must be specially designed for a particular heatsink thickness and semiconductor package.

Clips are especially popular with plastic packages such as the TO-220 and TO-126. In addition to fast assembly, the clip provides lower interface thermal resistance than other assembly methods when it is designed for proper pressure to bear on the top of the plastic over the die. The TO-220 package usually is lifted up under the die location when mounted with a single fastener through the hole in the tab because of the high pressure at one end.

Machine Screws

Machine screws, conical washers, and nuts (or sync-nuts) can form a trouble-free fastener system for all types of packages which have mounting holes. However, proper torque is necessary. Torque ratings apply when dry; therefore, care must be exercised when using thermal grease to prevent it from getting on the threads as inconsistent torque readings result. Machine screw heads should not directly contact the surface of plastic packages types as the screw heads are not sufficiently flat to provide properly distributed force. Without a washer, cracking of the plastic case may occur.

Self-Tapping Screws

Under carefully controlled conditions, sheet-metal screws are acceptable. However, during the tapping-process with a standard screw, a volcano-like protrusion will develop in the metal being threaded; an unacceptable surface that could increase the thermal resistance may result. When standard sheet metal screws are used, they must be used in a clearance hole to engage a speed-nut. If a self tapping process is desired, the screw type must be used which roll-forms machine screw threads.

Rivets

Rivets are not a recommended fastener for any of the plastic packages. When a rugged metal flange-mount package or EMS module is being mounted directly to a heatsink, rivets can be used provided press-riveting is used. Crimping force must be applied slowly and evenly. Pop-riveting should never be used because the high crimping force could cause deformation of most semiconductor packages. Aluminum rivets are much preferred over steel because less pressure is required to set the rivet and thermal conductivity is improved.

The hollow rivet, or eyelet, is preferred over solid rivets. An adjustable, regulated pressure press is used such that a gradually increasing pressure is used to pan the eyelet. Use of sharp blows could damage the semiconductor die.

Solder

Until the advent of the surface mount assembly technique, solder was not considered a suitable fastener for power semiconductors. However, user demand has led to the development of new packages for this application. Acceptable soldering methods include conventional belt-

furnace, irons, vapor-phase reflow, and infrared reflow. It is important that the semiconductor temperature not exceed the specified maximum (usually 260°C) or the die bond to the case could be damaged. A degraded die bond has excessive thermal resistance which often leads to a failure under power cycling.

Adhesives

Adhesives are available which have coefficients of expansion compatible with copper and aluminum.[5] Highly conductive types are available; a 10 mil layer has approximately 0.3°C/W interface thermal resistance. Different types are offered: high strength types for non-field-servicable systems or low strength types for field-serviceable systems. Adhesive bonding is attractive when case mounted parts are used in wave soldering assembly because thermal greases are not compatible with the conformal coatings used and the greases foul the solder process.

Plastic Hardware

Most plastic materials will flow, but differ widely in this characteristic. When plastic materials form parts of the fastening system, compression washers are highly valuable to assure that the assembly will not loosen with time and temperature cycling. As previously discussed, loss of contact pressure will increase interface thermal resistance.

FASTENING TECHNIQUES

Each of the various classes of packages in use requires different fastening techniques. Details pertaining to each type are discussed in following sections. Some general considerations follow.

To prevent galvanic action from occurring when devices are used on aluminum heatsinks in a corrosive atmosphere, many devices are nickel- or gold-plated. Consequently, precautions must be taken not to mar the finish.

Another factor to be considered is that when a copper based part is rigidly mounted to an aluminum heatsink, a bimetallic system results which will bend with temperature changes. Not only is the thermal coefficient of expansion different for copper and aluminum, but the temperature gradient through each metal also causes each component to bend. If bending is excessive and the package is mounted by two or more screws the semiconductor chip could be damaged. Bending can be minimized by:

1. Mounting the component parallel to the heatsink fins to provide increased stiffness.
2. Allowing the heatsink holes to be a bit oversized so that some slip between surfaces can occur as temperature changes.
3. Using a highly conductive thermal grease or mounting pad between the heatsink and semiconductor to minimize the temperature gradient and allow for movement.

Stud Mount

Parts which fall into the stud-mount classification are shown in Figure 6. Mounting errors with non-insulated stud-mounted parts are generally confined to application

(5) Robert Batson, Elliot Fraunglass and James P. Moran, "Heat Dissipation Through Thermalloy Conductive Adhesives," EMTAS '83. Conference, February 1–3, Phoenix, AZ; Society of Manufacturing Engineers, One SME Drive, P.O. Box 930, Dearborn, MI 48128.

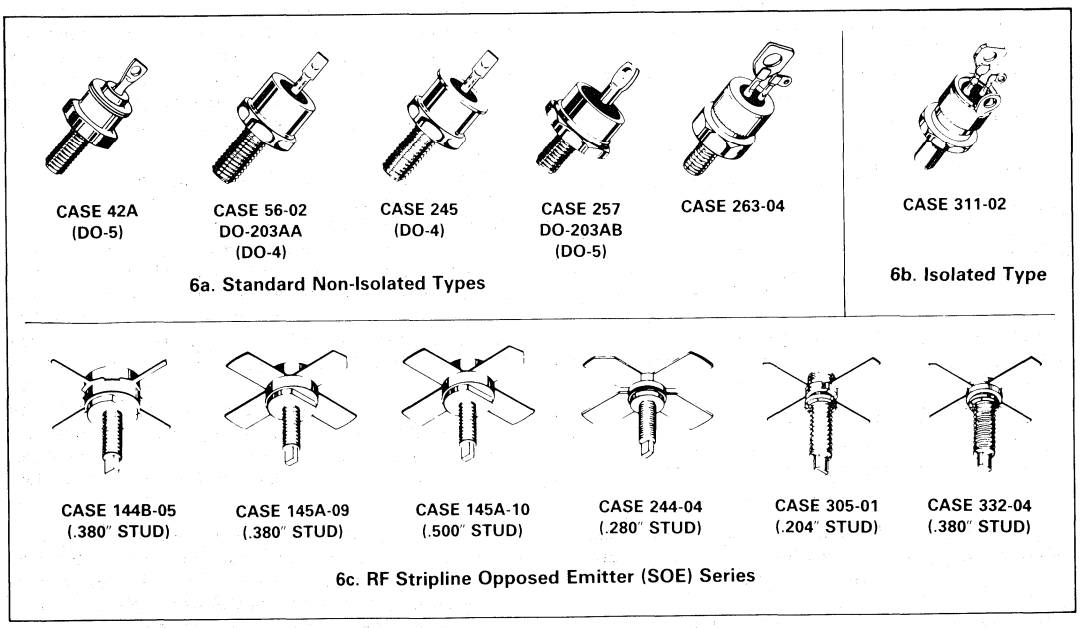


Figure 6. A Variety of Stud-Mount Parts

of excessive torque or tapping the stud into a threaded heatsink hole. Both these practices may cause a warpage of the hex base which may crack the semiconductor die. The only recommended fastening method is to use a nut and washer; the details are shown in Figure 7.

Insulated electrode packages on a stud mount base require less hardware. They are mounted the same as their non-insulated counterparts, but care must be exercised to avoid applying a shear or tension stress to the insulation layer, usually a berrylium oxide (BeO) ceramic. This requirement dictates that the leads must be attached to the circuit with flexible wire. In addition, the stud hex should be used to hold the part while the nut is torqued.

R.F. transistors in the stud-mount stripline opposed emitter (SOE) package impose some additional constraints because of the unique construction of the package. Special techniques to make connections to the stripline leads and to mount the part so no tension or shear forces are applied to any ceramic — metal interface are discussed in the section entitled "Connecting and Handling Terminals."

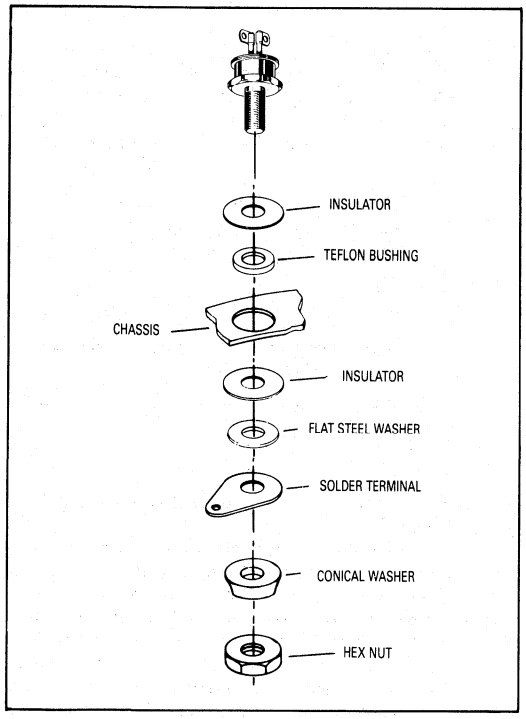
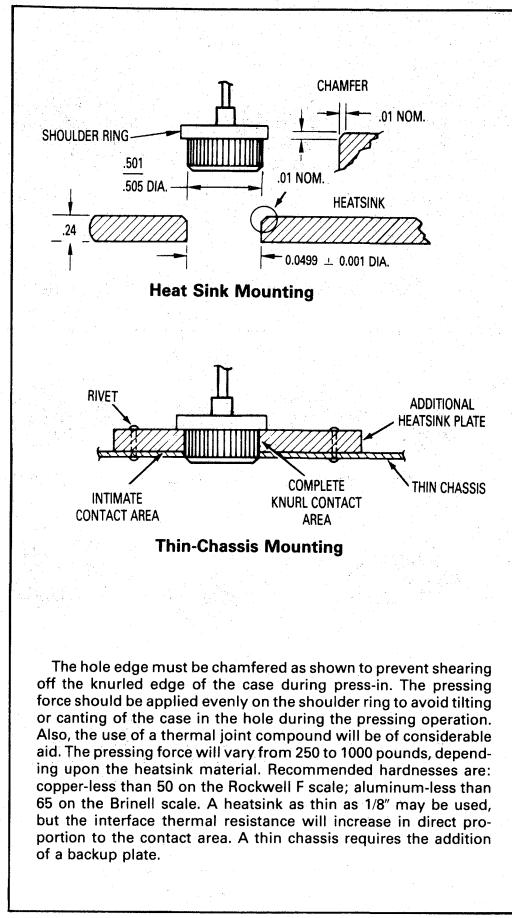


Figure 7. Isolating Hardware Used for a Non-Isolated Stud-Mount Package

Press Fit

For most applications, the press-fit case should be mounted according to the instructions shown in Figure 8. A special fixture meeting the necessary requirements must be used.



The hole edge must be chamfered as shown to prevent shearing off the knurled edge of the case during press-in. The pressing force should be applied evenly on the shoulder ring to avoid tilting or canting of the case in the hole during the pressing operation. Also, the use of a thermal joint compound will be of considerable aid. The pressing force will vary from 250 to 1000 pounds, depending upon the heatsink material. Recommended hardnesses are: copper-less than 50 on the Rockwell F scale; aluminum-less than 65 on the Brinell scale. A heatsink as thin as 1/8" may be used, but the interface thermal resistance will increase in direct proportion to the contact area. A thin chassis requires the addition of a backup plate.

Figure 8. Press-Fit Package

Flange Mount

A large variety of parts fit into the flange mount category as shown in Figure 9. Few known mounting difficulties exist with the smaller flange mount packages, such as the TO-204 (TO-3). The rugged base and distance between die and mounting holes combine to make it extremely difficult to cause any warpage unless mounted on a surface which is badly bowed or unless one side is tightened excessively before the other screw is started. It is therefore good practice to alternate tightening of the screws so that pressure is evenly applied. After the screws are finger-tight the hardware should be torqued to its final specification in at least two sequential steps. A typical mounting installation for a popular flange type part is shown in Figure 10. Machine screws (preferred) self-tapping screws, eyelets, or rivets may be used to secure the package using guidelines in the previous section. "Fastener and Hardware Characteristics."

The copper flange of the Energy Management Series (EMS) Modules is very thick. Consequently, the parts are rugged and indestructible for all practical purposes. No

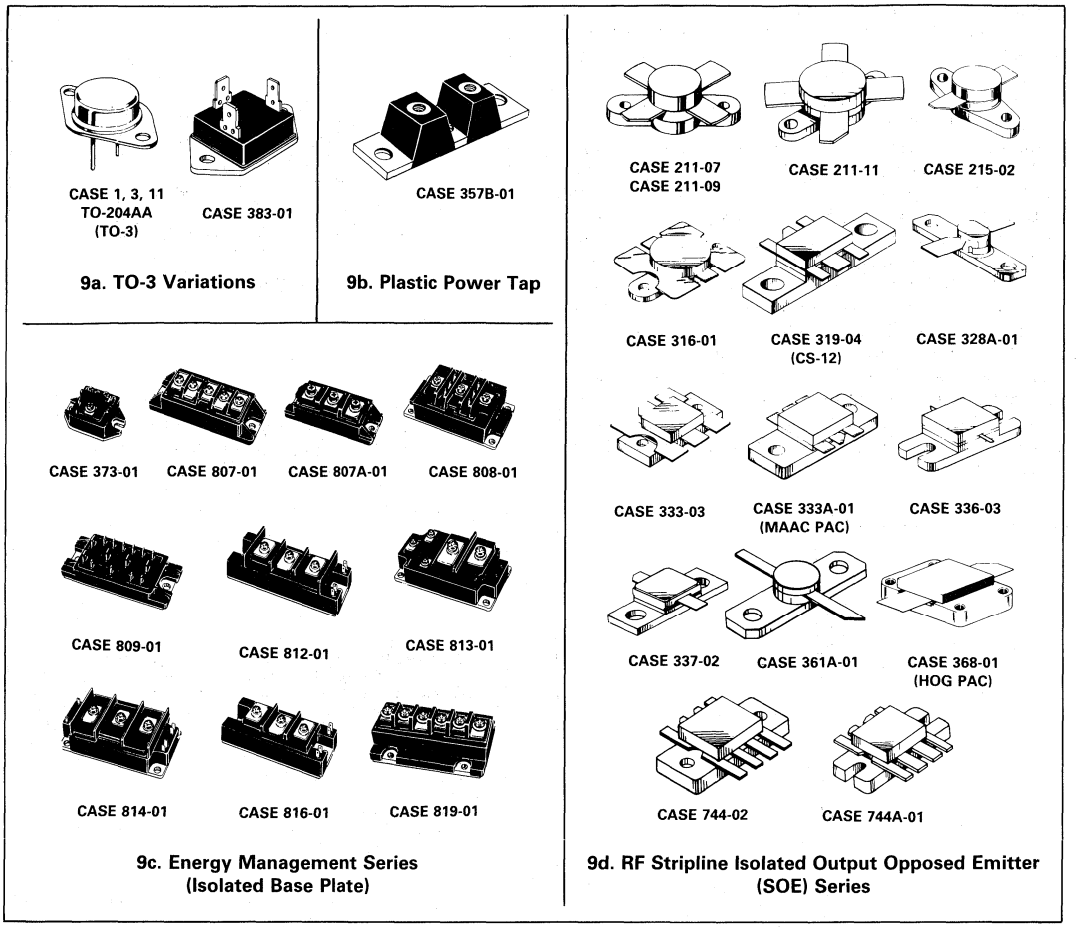


Figure 9. A Large Array of Parts Fit into the Flange-Mount Classification

special precautions are necessary when fastening these parts to a heatsink.

Some packages specify a tightening procedure. For example, with the Power Tap package, Figure 9b, final torque should be applied first to the center position.

The RF power modules (MHW series) are more sensitive to the flatness of the heatsink than other packages because a ceramic (BeO) substrate is attached to a relatively thin, fairly long, flange. The maximum allowable flange bending to avoid mechanical damage has been determined and presented in detail in EB107 "Mounting Considerations for Motorola RF Power Modules." Many of the parts can handle a combined heatsink and flange deviation from flat of 7 to 8 mils which is commonly available. Others must be held to 1.5 mils, which requires that the heatsink have nearly perfect flatness.

Specific mounting recommendations are critical to RF devices in isolated packages because of the internal ceramic substrate. The large area Case 368-1 (HOG PAC) will be used to illustrate problem areas. It is more sensitive to proper mounting techniques than most other RF power devices.

Although the data sheets contain information on recommended mounting procedures, experience indicates that they are often ignored. For example, the recommended maximum torque on the 4–40 mounting screws is 5 in/lbs. Spring and flat washers are recommended. Over torquing is a common problem. In some parts returned for failure analysis, indentions up to 10 mils deep in the mounting screw areas have been observed.

Calculations indicate that the length of the flange increases in excess of two mils with a temperature change of 75°C. In such cases, if the mounting screw torque is excessive, the flange is prevented from expanding in length, instead it bends upwards in the mid-section, cracking the BeO and the die. A similar result can also occur during the initial mounting of the device if an excessive amount of thermal compound is applied. With sufficient torque, the thermal compound will squeeze out of the mounting hole areas, but will remain under the center

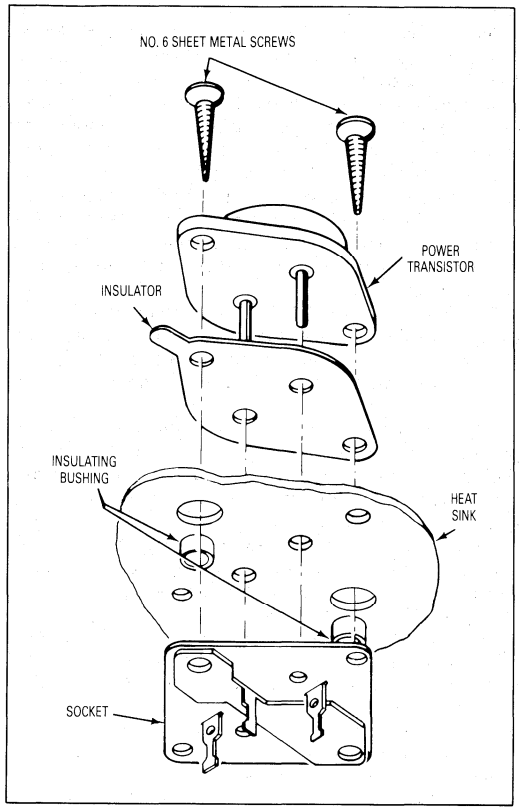


Figure 10. Hardware Used for a TO-204AA (TO-3) Flange Mount Part

of the flange, deforming it. Deformations of 2–3 mils have been measured between the center and the ends under such conditions (enough to crack internal ceramic).

Another problem arises because the thickness of the flange changes with temperature. For the 75°C temperature excursion mentioned, the increased amount is around 0.25 mils which results in further tightening of the mounting screws, thus increasing the effective torque from the initial value. With a decrease in temperature, the opposite effect occurs. Therefore thermal cycling not only causes risk of structural damage but often causes the assembly to loosen which raises the interface resistance. Use of compression hardware can eliminate this problem.

Tab Mount

The tab mount class is composed of a wide array of packages as illustrated in Figure 11. Mounting considerations for all varieties are similar to that for the popular TO-220 package, whose suggested mounting arrangements and hardware are shown in Figure 12. The rectangular washer shown in Figure 12a is used to minimize distortion of the mounting flange; excessive distortion could cause damage to the semiconductor chip. Use of the washer is only important when the size of the mounting hole exceeds 0.140 inch (6–32 clearance). Larger holes are needed to accommodate the lower insulating bushing when the screw is electrically connected to the case; however, the holes should not be larger than necessary to provide hardware clearance and should never exceed a diameter of 0.250 inch. Flange distortion is also possible if excessive torque is used during mounting. A maximum torque of 8 inch-pounds is suggested when using a 6–32 screw.

Care should be exercised to assure that the tool used to drive the mounting screw never comes in contact with the plastic body during the driving operation. Such contact can result in damage to the plastic body and internal device connections. To minimize this problem, Motorola TO-220 packages have a chamfer on one end. TO-220 packages of other manufacturers may need a spacer or combination spacer and isolation bushing to raise the screw head above the top surface of the plastic.

The popular TO-220 Package and others of similar construction lift off the mounting surface as pressure is applied to one end. (See Appendix B, Figure B1.) To counter this tendency, at least one hardware manufacturer offers a hard plastic cantilever beam which applies more even pressure on the tab.[6] In addition, it separates

(6) Catalog, Edition 18, Richco Plastic Company, 5825 N. Tripp Ave., Chicago, IL 60546.

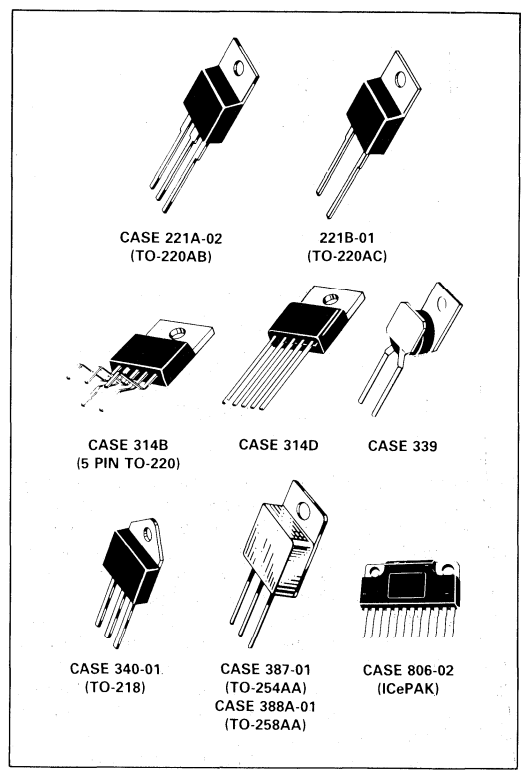


Figure 11. Several Types of Tab-Mount Parts

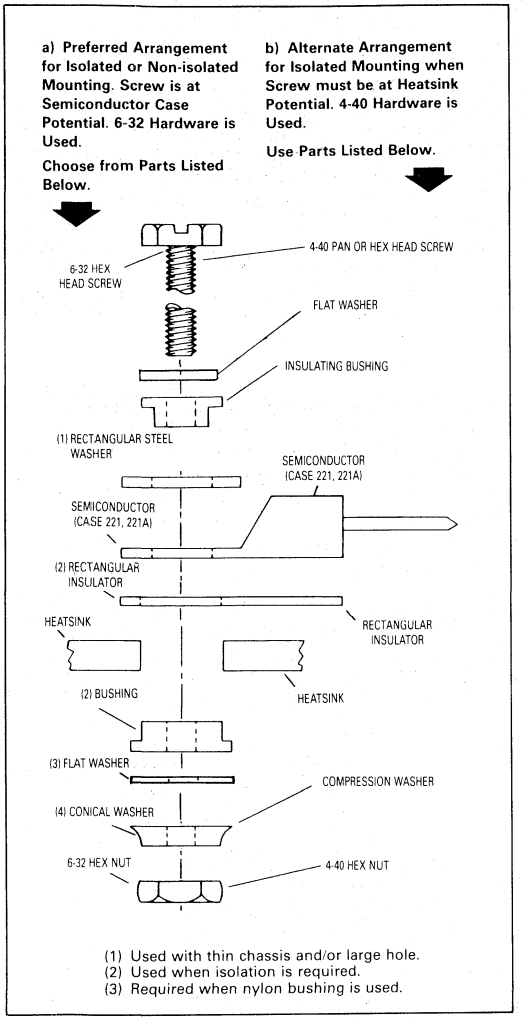


Figure 12. Mounting Arrangements for Tab Mount TO-220

the mounting screw from the metal tab. Tab mount parts may also be effectively mounted with clips as shown in Figure 15c. To obtain high pressure without cracking the case, a pressure spreader bar should be used under the clip. Interface thermal resistance with the cantilever beam or clips can be lower than with screw mounting.

The ICePAK (Case 806-02) is basically an elongated TO-220 package with isolated chips. The mounting precautions for the TO-220 consequently apply. In addition, since two mounting screws are required, the alternate tightening procedure described for the flange mount package should be used.

In situations where a tab mount package is making direct contact with the heatsink, an eyelet may be used, provided sharp blows or impact shock is avoided.

Plastic Body Mount

The Thermopad and Full Pak plastic power packages shown in Figure 13 are typical of packages in this group. They have been designed to feature minimum size with no compromise in thermal resistance. For the Thermopad (Case 77) parts this is accomplished by die-bonding the silicon chip on one side of a thin copper sheet; the opposite side is exposed as a mounting surface. The copper sheet has a hole for mounting; plastic is molded enveloping the chip but leaving the mounting hole open. The low thermal resistance of this construction is obtained at the expense of a requirement that strict attention be paid to the mounting procedure.

The Full Pak (Case 221C-01) is similar to a TO-220 except that the tab is encased in plastic. Because the mounting force is applied to plastic, the mounting procedure differs from a standard TO-220 and is similar to that of the Thermopad.

Several types of fasteners may be used to secure these packages; machine screws, eyelets, or clips are preferred. With screws or eyelets, a conical washer should be used which applies the proper force to the package over a fairly wide range of deflection and distributes the force over a fairly large surface area. Screws should not be tightened with any type of air-driven torque gun or equipment which may cause high impact. Characteristics of a suitable conical washer is shown in Figure 5.

Figure 14 shows details of mounting Case 77 devices. Clip mounting is fast and requires minimum hardware, however, the clip must be properly chosen to insure that the proper mounting force is applied. When electrical isolation is required with screw mounting, a bushing inside the mounting hole will insure that the screw threads do not contact the metal base.

The Full Pak, (Case 221C, 221D and 340B) permits the mounting procedure to be greatly simplified over that of a standard TO-220. As shown in Figure 15c, one properly chosen clip, inserted into two slotted holes in the heatsink, is all the hardware needed. Even though clip pressure is much lower than obtained with a screw, the thermal resistance is about the same for either method. This occurs because the clip bears directly on top of the die and holds the package flat while the screw causes the package to lift up somewhat under the die. (See Figure B1 of Appendix B.) The interface should consist of a layer of thermal grease or a highly conductive thermal pad. Of course, screw mounting shown in Figure 15b may also be used but a conical compression washer should be included. Both methods afford a major reduction in hardware as compared to the conventional mounting method with a TO-220 package which is shown in Figure 15a.

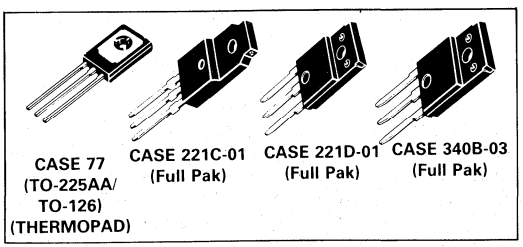


Figure 13. Plastic Body-Mount Packages

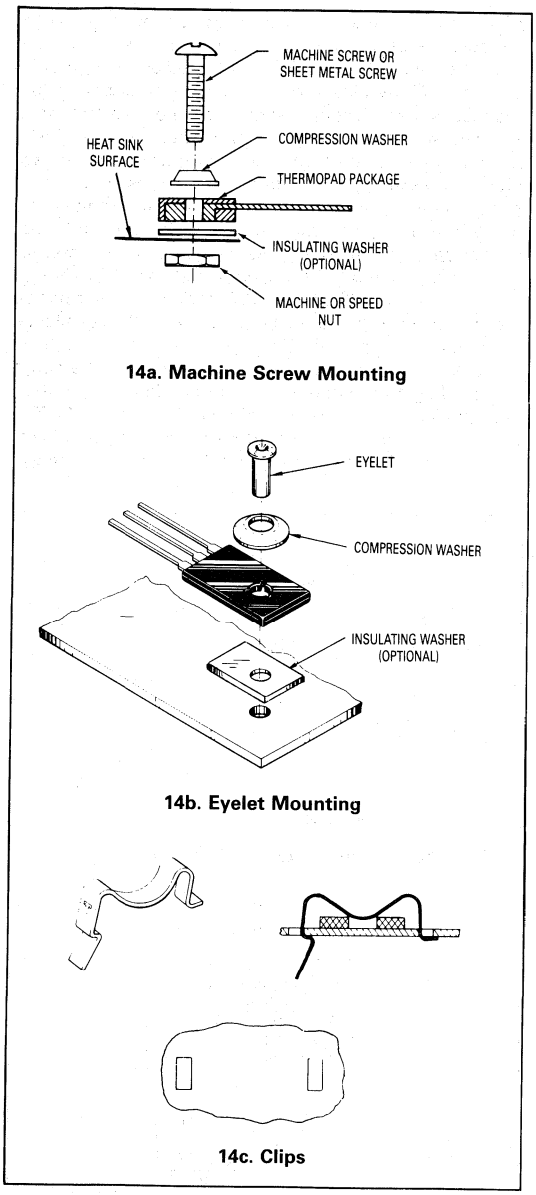


Figure 14. Recommended Mounting Arrangements for TO-225AA (TO-126) Thermopad Packages

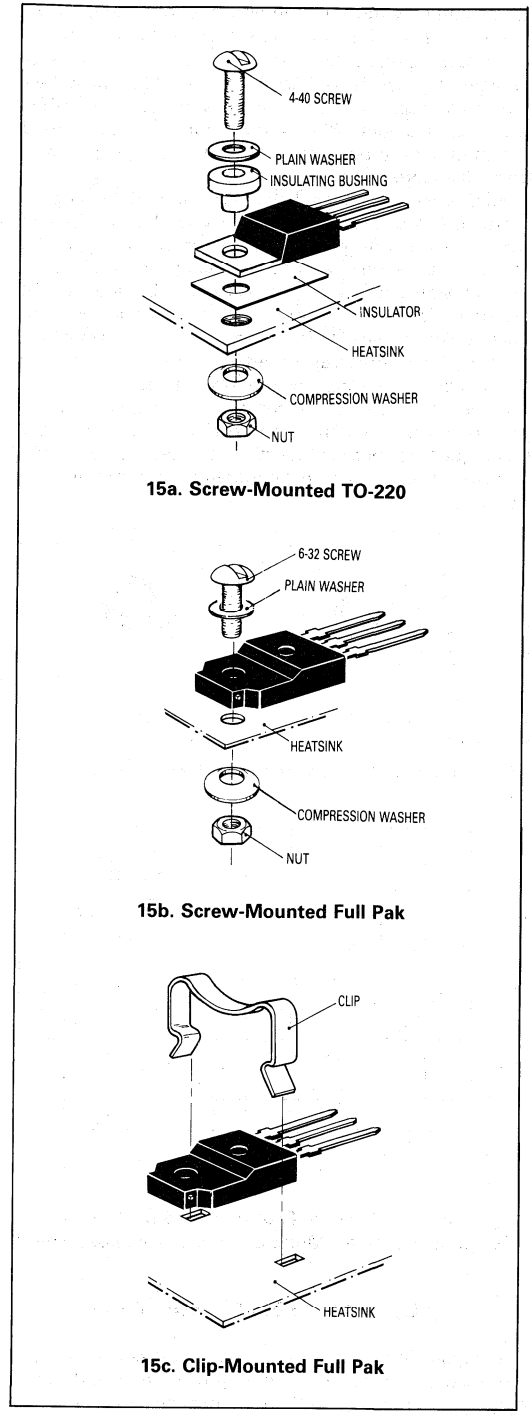


Figure 15. Mounting Arrangements for the Full Pak as Compared to a Conventional TO-220

Surface Mount

Although many of the tab mount parts have been surface mounted, special small footprint packages for mounting power semiconductors using surface mount assembly techniques have been developed. The DPAK, shown in Figure 16, for example, will accommodate a die up to 112 mils x 112 mils, and has a typical thermal resistance around 2°C/W junction to case. The thermal resis-

tance values of the solder interface is well under 1°C/W. The printed circuit board also serves as the heatsink.

Standard Glass-Epoxy 2-ounce boards do not make very good heatsinks because the thin foil has a high thermal resistance. As Figure 17 shows, thermal resistance assymtotes to about 20°C/W at 10 square inches of board area, although a point of diminishing returns occurs at about 3 square inches.

Boards are offered that have thick aluminum or copper substrates. A dielectric coating designed for low thermal resistance is overlayed with one or two ounce copper foil for the preparation of printed conductor traces. Tests run on such a product indicate that case to substrate thermal resistance is in the vicinity of 1°C/W, exact values depending upon board type.(7) The substrate may be an effective heatsink itself, or it can be attached to a conventional finned heatsink for improved performance.

Since DPAK and other surface mount packages are designed to be compatible with surface mount assembly techniques, no special precautions are needed other than to insure that maximum temperature/time profiles are not exceeded.

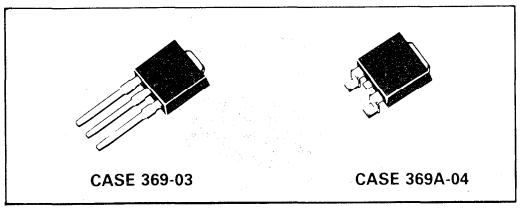


Figure 16. Surface Mount D-PAK Parts

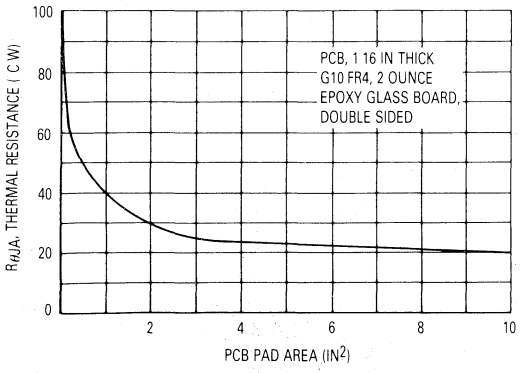
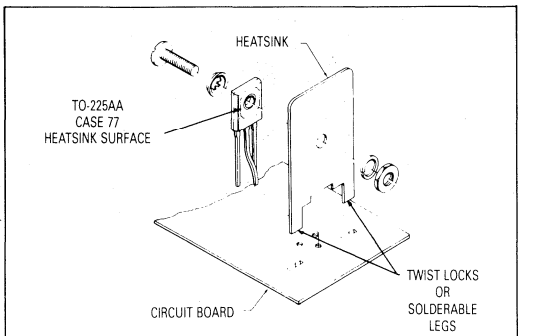


Figure 17. Effect of Footprint Area on Thermal Resistance of DPAK Mounted on a Glass-Epoxy Board

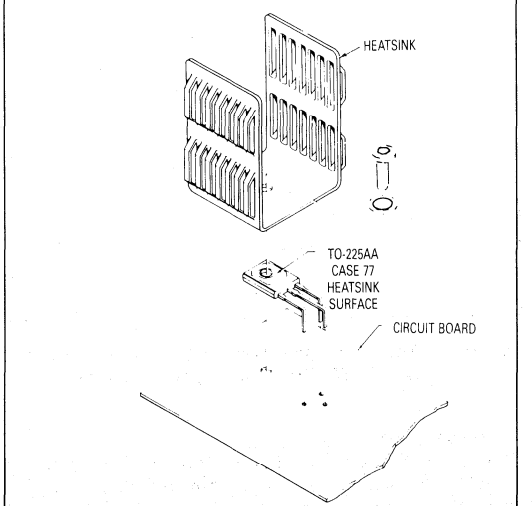
FREE AIR AND SOCKET MOUNTING

In applications where average power dissipation is on the order of a watt or so, most power semiconductors may be mounted with little or no heatsinking. The leads of the various metal power packages are not designed to support the packages; their cases must be firmly supported to avoid the possibility of cracked seals around the leads. Many plastic packages may be supported by their leads in applications where high shock and vibration stresses are not encountered and where no heatsink is used. The leads should be as short as possible to increase vibration resistance and reduce thermal resistance. As a general practice however, it is better to support the package. A plastic support for the TO-220 Package and other similar types is offered by heatsink accessory vendors.

In many situations, because its leads are fairly heavy, the CASE 77 (TO-225AA) (TO-127) package has supported a small heatsink; however, no definitive data is available. When using a small heatsink, it is good practice to have the sink rigidly mounted such that the sink or the board is providing total support for the semiconductor. Two possible arrangements are shown in Figure 18. The arrangement of part (a) could be used with any plastic package, but the scheme of part (18b) is more practical



18a. Simple Plate, Vertically Mounted



18b. Commercial Sink, Horizontally Mounted

Figure 18. Methods of Using Small Heatsinks With Plastic Semiconductor Packages

(7) Herb Fick, "Thermal Management of Surface Mount Power Devices," Powerconversion and Intelligent Motion, August 1987.

with Case 77 Thermopad devices. With the other package types, mounting the transistor on top of the heatsink is more practical.

In certain situations, in particular where semiconductor testing is required or prototypes are being developed, sockets are desirable. Manufacturers have provided sockets for many of the packages available from Motorola. The user is urged to consult manufacturers' catalogs for specific details. Sockets with Kelvin connections are necessary to obtain accurate voltage readings across semiconductor terminals.

CONNECTING AND HANDLING TERMINALS

Pins, leads, and tabs must be handled and connected properly to avoid undue mechanical stress which could cause semiconductor failure. Change in mechanical dimensions as a result of thermal cycling over operating temperature extremes must be considered. Standard metal, plastic, and RF stripline packages each have some special considerations.

Metal Packages

The pins and lugs of metal packaged devices using glass to metal seals are not designed to handle any significant bending or stress. If abused, the seals could crack. Wires may be attached using sockets, crimp connectors or solder, provided the data sheet ratings are observed. When wires are attached directly to the pins, flexible or braided leads are recommended in order to provide strain relief.

EMS Modules

The screw terminals of the EMS modules look deceptively rugged. Since the flange base is mounted to a rigid heatsink, the connection to the terminals must allow some flexibility. A rigid buss bar should not be bolted to terminals. Lugs with braid are preferred.

Plastic Packages

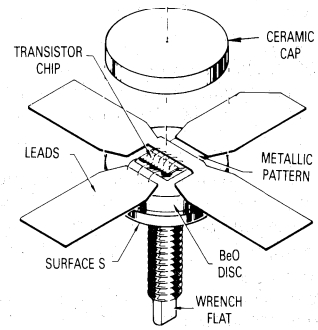
The leads of the plastic packages are somewhat flexible and can be reshaped although this is not a recommended procedure. In many cases, a heatsink can be chosen which makes lead-bending unnecessary. Numerous lead- and tab-forming options are available from Motorola on large quantity orders. Preformed leads remove the users risk of device damage caused by bending.

If, however, lead-bending is done by the user, several basic considerations should be observed. When bending the lead, support must be placed between the point of bending and the package. For forming small quantities of units, a pair of pliers may be used to clamp the leads at the case, while bending with the fingers or another pair of pliers. For production quantities, a suitable fixture should be made.

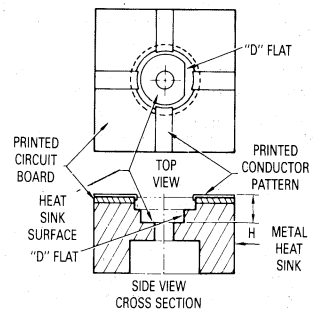
The following rules should be observed to avoid damage to the package.

1. A leadbend radius greater than 1/16 inch is advisable for TO-225AA (CASE 77) and 1/32 inch for TO-220.
2. No twisting of leads should be done at the case.
3. No axial motion of the lead should be allowed with respect to the case.

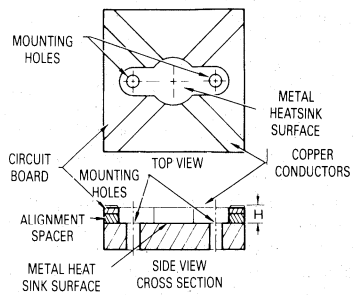
The leads of plastic packages are not designed to withstand excessive axial pull. Force in this direction greater than 4 pounds may result in permanent damage to the device. If the mounting arrangement imposes axial stress on the leads, a condition which may be caused by thermal cycling, some method of strain relief should be devised. When wires are used for connections, care should be



19a. Component Parts of a Stud Mount Stripline Package. Flange Mounted Packages are Similarly Constructed



19b. Typical Stud Type SOE Transistor Mounting Method



19c. Flange Type SOE Transistor Mounting Method

Figure 19. Mounting Details for SOE Transistors

exercised to assure that movement of the wire does not cause movement of the lead at the lead-to-plastic junctions. Highly flexible or braided wires are good for providing strain relief.

Wire-wrapping of the leads is permissible, provided that the lead is restrained between the plastic case and the point of the wrapping. The leads may be soldered; the maximum soldering temperature, however, must not exceed 260°C and must be applied for not more than 5 seconds at a distance greater than 1/8 inch from the plastic case.

Stripline Packages

The leads of stripline packages normally are soldered into a board while the case is recessed to contact a heatsink as shown in Figure 19. The following rules should be observed:

1. The device should never be mounted in such a manner as to place ceramic-to-metal joints in tension.
2. The device should never be mounted in such a manner as to apply force on the strip leads in a vertical direction towards the cap.
3. When the device is mounted in a printed circuit board with the copper stud and BeO portion of the header passing through a hole in the circuit boards, adequate clearance must be provided for the BeO to prevent shear forces from being applied to the leads.
4. Some clearance must be allowed between the leads and the circuit board when the device is secured to the heatsink.
5. The device should be properly secured into the heatsinks before its leads are attached into the circuit.
6. The leads on stud type devices must not be used to prevent device rotation during stud torque application. A wrench flat is provided for this purpose.

Figure 19b shows a cross-section of a printed circuit board and heatsink assembly for mounting a stud type stripline device. H is the distance from the top surface of the printed circuit board to the D-flat heatsink surface. If H is less than the minimum distance from the bottom of the lead material to the mounting surface of the package, there is no possibility of tensile forces in the copper stud — BeO ceramic joint. If, however, H is greater than the package dimension, considerable force is applied to the cap to BeO joint and the BeO to stud joint. Two occurrences are possible at this point. The first is a cap joint failure when the structure is heated, as might occur during the lead-soldering operation; while the second is BeO to stud failure if the force generated is high enough. Lack of contact between the device and the heatsink surface will occur as the differences between H and the package dimension become larger, this may result in device failure as power is applied.

Figure 19c shows a typical mounting technique for flange-type stripline transistors. Again, H is defined as the distance from the top of the printed circuit board to the heatsink surface. If distance H is less than the minimum distance from the bottom of transistor lead to the bottom surface of the flange, tensile forces at the various joints in the package are avoided. However, if distance H exceeds the package dimension, problems similar to those discussed for the stud type devices can occur.

CLEANING CIRCUIT BOARDS

It is important that any solvents or cleaning chemicals used in the process of degreasing or flux removal do not affect the reliability of the devices. Alcohol and unchlorinated Freon solvents are generally satisfactory for use with plastic devices, since they do not damage the package. Hydrocarbons such as gasoline and chlorinated Freon may cause the encapsulant to swell, possibly damaging the transistor die.

When using an ultrasonic cleaner for cleaning circuit boards, care should be taken with regard to ultrasonic energy and time of application. This is particularly true if any packages are free-standing without support.

THERMAL SYSTEM EVALUATION

Assuming that a suitable method of mounting the semiconductor without incurring damage has been achieved, it is important to ascertain whether the junction temperature is within bounds.

In applications where the power dissipated in the semiconductor consists of pulses at a low duty cycle, the instantaneous or peak junction temperature, not average temperature, may be the limiting condition. In this case, use must be made of transient thermal resistance data. For a full explanation of its use, see Motorola Application Note, AN569.

Other applications, notably RF power amplifiers or switches driving highly reactive loads, may create severe current crowding conditions which render the traditional concepts of thermal resistance or transient thermal impedance invalid. In this case, transistor safe operating area, thyristor di/dt limits, or equivalent ratings as applicable, must be observed.

Fortunately, in many applications, a calculation of the average junction temperature is sufficient. It is based on the concept of thermal resistance between the junction and a temperature reference point on the case. (See Appendix A.) A fine wire thermocouple should be used, such as #36 AWG, to determine case temperature. Average operating junction temperature can be computed from the following equation:

$$T_J = T_C + R_{\theta JC} \times P_D$$

where T_J = junction temperature (°C)
T_C = case temperature (°C)
$R_{\theta JC}$ = thermal resistance junction-to-case as specified on the data sheet (°C/W)
P_D = power dissipated in the device (W)

The difficulty in applying the equation often lies in determining the power dissipation. Two commonly used empirical methods are graphical integration and substitution.

Graphical Integration

Graphical integration may be performed by taking oscilloscope pictures of a complete cycle of the voltage and current waveforms, using a limit device. The pictures should be taken with the temperature stabilized. Corresponding points are then read from each photo at a suitable number of time increments. Each pair of voltage and current values are multiplied together to give instanta-

neous values of power. The results are plotted on linear graph paper, the number of squares within the curve counted, and the total divided by the number of squares along the time axis. The quotient is the average power dissipation. Oscilloscopes are available to perform these measurements and make the necessary calculations.

Substitution

This method is based upon substituting an easily measurable, smooth dc source for a complex waveform. A switching arrangement is provided which allows operating the load with the device under test, until it stabilizes in temperature. Case temperature is monitored. By throwing the switch to the "test" position, the device under test is connected to a dc power supply, while another pole of the switch supplies the normal power to the load to keep it operating at full power level. The dc supply is adjusted so that the semiconductor case temperature remains approximately constant when the switch is thrown to each position for about 10 seconds. The dc voltage and current values are multiplied together to obtain average power. It is generally necessary that a Kelvin connection be used for the device voltage measurement.

APPENDIX A
THERMAL RESISTANCE CONCEPTS

The basic equation for heat transfer under steady-state conditions is generally written as:

$$q = hA\Delta T \qquad (1)$$

where
- q = rate of heat transfer or power dissipation (P_D)
- h = heat transfer coefficient,
- A = area involved in heat transfer,
- ΔT = temperature difference between regions of heat transfer.

However, electrical engineers generally find it easier to work in terms of thermal resistance, defined as the ratio of temperature to power. From Equation 1, thermal resistance, R_θ, is

$$R_\theta = \Delta T/q = 1/hA \qquad (2)$$

The coefficient (h) depends upon the heat transfer mechanism used and various factors involved in that particular mechanism.

An analogy between Equation (2) and Ohm's Law is often made to form models of heat flow. Note that T could be thought of as a voltage thermal resistance corresponds to electrical resistance (R); and, power (q) is analogous to current (I). This gives rise to a basic thermal resistance model for a semiconductor as indicated by Figure A1.

The equivalent electrical circuit may be analyzed by using Kirchoff's Law and the following equation results:

$$T_J = P_D(R_{\theta JC} + R_{\theta CS} + R_{\theta SA}) + T_A \qquad (3)$$

where
- T_J = junction temperature,
- P_D = power dissipation
- $R_{\theta JC}$ = semiconductor thermal resistance (junction to case),
- $R_{\theta CS}$ = interface thermal resistance (case to heatsink),
- $R_{\theta SA}$ = heatsink thermal resistance (heatsink to ambient),
- T_A = ambient temperature.

The thermal resistance junction to ambient is the sum of the individual components. Each component must be minimized if the lowest junction temperature is to result.

The value for the interface thermal resistance, $R_{\theta CS}$, may be significant compared to the other thermal-resistance terms. A proper mounting procedure can minimize $R_{\theta CS}$.

The thermal resistance of the heatsink is not absolutely constant; its thermal efficiency increases as ambient temperature increases and it is also affected by orientation of the sink. The thermal resistance of the semiconductor is also variable; it is a function of biasing and temperature. Semiconductor thermal resistance specifications are normally at conditions where current density is fairly uniform. In some applications such as in RF power amplifiers and short-pulse applications, current density is not uniform and localized heating in the semiconductor chip will be the controlling factor in determining power handling ability.

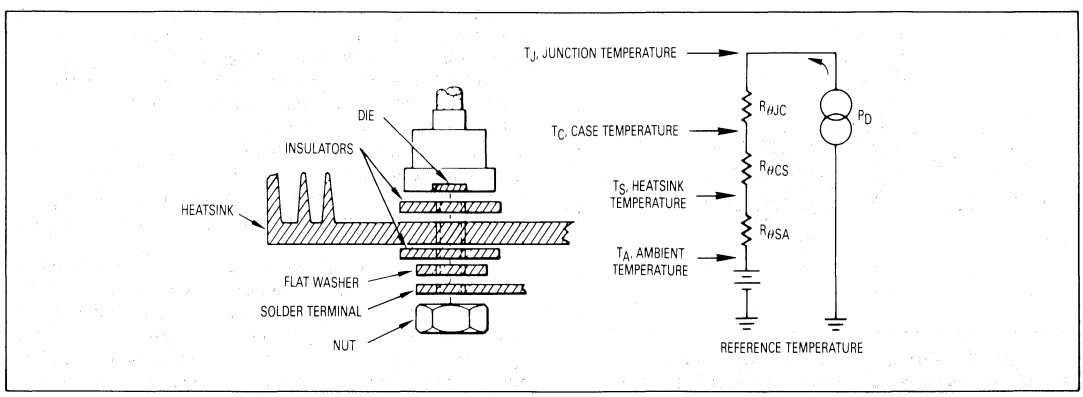


Figure A1. Basic Thermal Resistance Model Showing Thermal to Electrical Analogy for a Semiconductor

APPENDIX B
MEASUREMENT OF INTERFACE THERMAL RESISTANCE

Measuring the interface thermal resistance $R_{\theta CS}$ appears deceptively simple. All that's apparently needed is a thermocouple on the semiconductor case, a thermocouple on the heatsink, and a means of applying and measuring DC power. However, $R_{\theta CS}$ is proportional to the amount of contact area between the surfaces and consequently is affected by surface flatness and finish and the amount of pressure on the surfaces. The fastening method may also be a factor. In addition, placement of the thermocouples can have a significant influence upon the results. Consequently, values for interface thermal resistance presented by different manufacturers are not in good agreement. Fastening methods and thermocouple locations are considered in this Appendix.

When fastening the test package in place with screws, thermal conduction may take place through the screws, for example, from the flange ear on a TO-3 package directly to the heatsink. This shunt path yields values which are artificially low for the insulation material and dependent upon screw head contact area and screw material. MIL-I-49456 allows screws to be used in tests for interface thermal resistance probably because it can be argued that this is "application oriented."

Thermalloy takes pains to insulate all possible shunt conduction paths in order to more accurately evaluate insulation materials. The Motorola fixture uses an insulated clamp arrangement to secure the package which also does not provide a conduction path.

As described previously, some packages, such as a TO-220, may be mounted with either a screw through the tab or a clip bearing on the plastic body. These two methods often yield different values for interface thermal resistance. Another discrepancy can occur if the top of the package is exposed to the ambient air where radiation and convection can take place. To avoid this, the package should be covered with insulating foam. It has been estimated that a 15 to 20% error in $R_{\theta CS}$ can be incurred from this source.

Another significant cause for measurement discrepancies is the placement of the thermocouple to measure the semiconductor case temperature. Consider the TO-220 package shown in Figure B1. The mounting pressure at one end causes the other end — where the die is located — to lift off the mounting surface slightly. To improve contact, Motorola TO-220 Packages are slightly concave. Use of a spreader bar under the screw lessens the lifting, but some is inevitable with a package of this structure. Three thermocouple locations are shown:

a. The Motorola location is directly under the die reached through a hole in the heatsink. The thermocouple is held in place by a spring which forces the thermocouple into intimate contact with the bottom of the semi's case.

b. The JEDEC location is close to the die on the top surface of the package base reached through a blind hole drilled through the molded body. The thermocouple is swaged in place.

c. The Thermalloy location is on the top portion of the tab between the molded body and the mounting screw. The thermocouple is soldered into position.

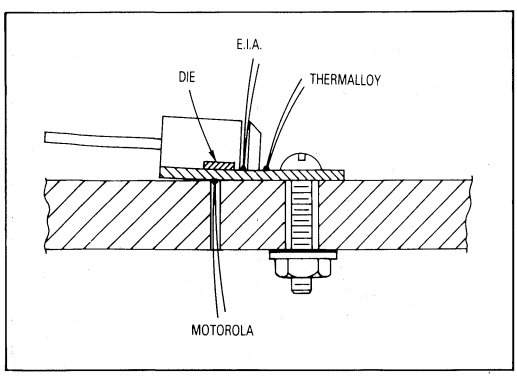


Figure B1. JEDEC TO-220 Package Mounted to Heatsink Showing Various Thermocouple Locations and Lifting Caused by Pressure at One End

Temperatures at the three locations are generally not the same. Consider the situation depicted in the figure. Because the only area of direct contact is around the mounting screw, nearly all the heat travels horizontally along the tab from the die to the contact area. Consequently, the temperature at the JEDEC location is hotter than at the Thermalloy location and the Motorola location is even hotter. Since junction-to-sink thermal resistance must be constant for a given test setup, the calculated junction-to-case thermal resistance values decrease and case-to-sink values increase as the "case" temperature thermocouple readings become warmer. Thus the choice of reference point for the "case" temperature is quite important.

There are examples where the relationship between the thermocouple temperatures are different from the previous situation. If a mica washer with grease is installed between the semiconductor package and the heatsink, tightening the screw will not bow the package; instead, the mica will be deformed. The primary heat conduction path is from the die through the mica to the heatsink. In this case, a small temperature drop will exist across the vertical dimension of the package mounting base so that the thermocouple at the EIA location will be the hottest. The thermocouple temperature at the Thermalloy location will be lower but close to the temperature at the EIA location as the lateral heat flow is generally small. The Motorola location will be coolest.

The EIA location is chosen to obtain the highest temperature on the case. It is of significance because power ratings are supposed to be based on this reference point. Unfortunately, the placement of the thermocouple is tedious and leaves the semiconductor in a condition unfit for sale.

The Motorola location is chosen to obtain the highest temperature of the case at a point where, hopefully, the case is making contact to the heatsink. Once the special heatsink to accommodate the thermocouple has been fabricated, this method lends itself to production testing and does not mark the device. However, this location is not easily accessible to the user.

The Thermalloy location is convenient and is often chosen by equipment manufacturers. However, it also blemishes the case and may yield results differing up to 1°C/W for a TO-220 package mounted to a heatsink without thermal grease and no insulator. This error is small when compared to the thermal resistance of heat dissipaters often used with this package, since power dissipation is usually a few watts. When compared to the specified junction-to-case values of some of the higher power semiconductors becoming available, however, the difference becomes significant and it is important that the semiconductor manufacturer and equipment manufacturer use the same reference point.

Another EIA method of establishing reference temperatures utilizes a soft copper washer (thermal grease is used) between the semiconductor package and the heatsink. The washer is flat to within 1 mil/inch, has a finish better than 63 μ-inch, and has an imbedded thermocouple near its center. This reference includes the interface resistance under nearly ideal conditions and is therefore application-oriented. It is also easy to use but has not become widely accepted.

A good way to improve confidence in the choice of case reference point is to also test for junction-to-case thermal resistance while testing for interface thermal resistance. If the junction-to-case values remain relatively constant as insulators are changed, torque varied, etc., then the case reference point is satisfactory.

APPENDIX C
Sources of Accessories

Manufacturer	Joint Compound	Adhesives	Insulators						Heatsinks
			BeO	AlO_2	Anodize	Mica	Plastic Film	Silicone Rubber	
Aavid Eng.	X	X	—	—	—	—	—	X	X
AHAM-TOR	—	—	—	—	—	—	—	—	X
Astrodynamis	X	—	—	—	—	—	—	—	X
Delbert Blinn	—	—	X	—	X	X	X	X	X
IERC	X	—	—	—	—	—	—	—	X
Staver	—	—	—	—	—	—	—	—	X
Thermalloy	X	X	X	X	X	X	X	X	X
Tran-tec	—	—	X	X	X	X	—	X	X
Wakefield Eng.	X	X	X	—	X	—	—	X	X

Other sources for silicone rubber pads: Chomerics, Berquist

Suppliers Addresses

Aavid Engineering, Inc., 30 Cook Court, Laconia, New Hampshire 03246 (603) 524-4443

AHAM-TOR Heatsinks, 27901 Front Street, Rancho, California 92390 (714) 676-4151

Astro Dynamics, Inc., 2 Gill St., Woburn, Massachusetts 01801 (617) 935-4944

Berquist, 5300 Edina Industrial Blvd., Minneapolis, Minnesota 55435 (612) 835-2322

Chomerics, Inc., 16 Flagstone Drive, Hudson, New Hampshire 03051 1-800-633-8800

Delbert Blinn Company, P.O. Box 2007, Pomona, California 91769 (714) 629-3900

International Electronic Research Corporation, 135 West Magnolia Boulevard, Burbank, California 91502 (213) 849-2481

The Staver Company, Inc., 41-51 Saxon Avenue, Bay Shore, Long Island, New York 11706 (516) 666-8000

Thermalloy, Inc., P.O. Box 34829, 2021 West Valley View Lane, Dallas, Texas 75234 (214) 243-4321

Tran-tec Corporation, P.O. Box 1044, Columbus, Nebraska 68601 (402) 564-2748

Wakefield Engineering, Inc., Wakefield, Massachusetts 01880 (617) 245-5900

PREFACE

When the JEDEC registration system for package outlines started in 1957, numbers were assigned sequentially whenever manufacturers wished to establish a package as an industry standard. As minor variations developed from these industry standards, either a new, non-related number was issued by JEDEC or manufacturers would attempt to relate the part to an industry standard via some appended description.

In an attempt to ease confusion, JEDEC established the present system in late 1968 in which new packages are assigned into a category, based on their general physical appearance. Differences between specific packages in a category are denoted by suffix letters. The older package designations were re-registered to the new system as time permitted.

For example the venerable TO-3 has many variations. Can heights differ and it is available with 30, 40, 50, and 60 mil pins, with and without lugs. It is now classified in the TO-204 family. The TO-204AA conforms to the original outline for the TO-3 having 40 mil pins while the TO-204AE has 60 mil pins, for example.

The new numbers for the old parts really haven't caught on very well. It seems that the DO-4, DO-5 and TO-3 still convey sufficient meaning for general verbal communication.

Motorola Case Number	JEDEC Outline		Notes	Mounting Class	See Page
	Original System	Revised System			
001	TO-3	TO-204AA		Flange	9
003	TO-3		2	Flange	9
009	TO-61	TO-210AC		Stud	8
011	TO-3	TO-204AA	—	Flange	9
011A	TO-3	—	2	Flange	9
012	TO-3	—	2	Flange	9
036	TO-60	TO-210AB	—	Stud	8
042A	DO-5	DO-203AB	—	Stud	8
044	DO-4	DO-203AA	—	Stud	8
054	TO-3	—	2	Flange	9
056	DO-4	—	—	Stud	8
058	DO-5	—	2	Stud	8
61-03				Flange	9
63-02	TO-64	TO-208AB		Stud	8
63-03	TO-64	TO-2088AB		Stud	8
077	TO-126	TO-225AA	—	Plastic	12
080	TO-66	TO-213AA	—	Flange	9
086	—	TO-208	1	Stud	8
086L	—	TO-298	1	Stud	8
144B-05				Stud	8
145A-09				Stud	8
145A-10				Stud	8
145C	TO-232		1	Stud	8
157	—	DO-203	1	Stud	8
160-03	TO-59	TO-210AA	—	Stud	8
167	—	DO-203	1	Stud	8
174-04				Pressfit	9
175-03				Stud	8
197	—	TO-204AE	—	Flange	9
211-07				Flange	9
211-09				Flange	9

Notes: 1. Would fit within this family outline if registered with JEDEC.
2. Not within all JEDEC dimensions.

Motorola Case Number	JEDEC Outline		Notes	Mounting Class	See Page
	Original System	Revised System			
211-11				Flange	9
215-02				Flange	9
221	—	TO-220AB	—	Tab	11
221C-02				Plastic	12
221D-01	—	—	Isolated TO-220	Plastic	12
235	—	TO-208	1	Stud	8
235-03				Stud	8
238	—	TO-208	1	Stud	8
239	—	TO-208	—	Stud	8
244-04				Stud	8
245	DO-4	—	—	Stud	8
257-01	DO-5	—	—	Stud	8
263	—	TO-208	—	Stud	8
263-04				Stud	8
283	DO-4	—	—	Stud	8
289	—	TO-209	1	Stud	8
305-01				Stud	8
310-02				Pressfit	9
311-01			Isolated	Stud	8
311-02				Pressfit	9
311-02				Stud	8
314B-01				Tab	11
314D-01				Tab	11
316-01				Flange	9
319-04				Flange	9
328A-01				Flange	9
332-04				Stud	8
333-03				Flange	9
333A-01				Flange	9
336-03				Flange	9

Motorola Case Number	JEDEC Outline		Notes	Mounting Class	See Page
	Original System	Revised System			
337-02				Flange	9
340		TO-218AC		Tab	11
340A-02				Plastic	12
340B-03			Isolated TO-218	Plastic	12
342-01				Flange	9
357B-01				Flange	9
361-01				Flange	9
368-01				Flange	9
369-03		TO-251		Insertion	14
369A-04		TO-252		Surface	13
373-01			Isolated	Flange	9
383-01			Isolated	Flange	10
387-01		TO-254AA	Isolated 2	Tab	11
388A-01		TO-258AA	Isolated 2	Tab	11
744-02				Flange	9
744A-01				Flange	9
806-02			Isolated	Flange	9
807-01			Isolated	Flange	9
807-02			Isolated	Flange	9
807A-01			Isolated	Flange	9
808-01			Isolated	Flange	9
809-01			Isolated	Flange	9
812-01			Isolated	Flange	9
813-01			Isolated	Flange	9
814-01			Isolated	Flange	9
814A-01			Isolated	Flange	9
084B-01			Isolated	Flange	9
816-01			Isolated	Flange	9
819-01			Isolated	Flange	9
043-02	DO-21	DO-208AA		Pressfit	9

CHAPTER 8
RELIABILITY AND QUALITY

As the price of semiconductor devices decreases, reliability and quality have become increasingly important in selecting a vendor. In many cases these considerations even outweigh price, delivery and service.

The reason is that the cost of device fallout and warranty repairs can easily equal or exceed the original cost of the devices. Consider the example shown in Figure 8.1. Although the case is simplistic, the prices and costs are realistic by today's standards. In this case, the cost of failures raised the device cost from 15 cents to 21 cents, an increase of 40%. Clearly, then, investing in quality and reliability can pay big dividends.

With nearly three decades of experience as a major semiconductor supplier, Motorola is the largest manufacturer of discrete semiconductors in the world today. Since semiconductor prices are strongly influenced by manufacturing volume, this leadership has permitted Motorola to be strongly competitive in the marketplace while making massive investments in equipment, processes and procedures to guarantee that the company's after-purchase costs will be among the lowest in the industry. As a result of the procedures, Motorola is projecting an outgoing quality level of less than 3 PPM by 1992.

Given:
Purchase = 100,000 components @ 15¢ each
Assumptions: Line Fallout = 0.1%
Warranty Failures = 0.01%

Components Cost =	100,000 x 15¢ =	$15,000
Line Fallout Cost = @ $40 per repair	100 x $40 =	4,000
Warranty Cost = @ $200 per repair	10 x $200 =	2,000
		$21,000

Adjusted Cost
Per Component = $21,000 ÷ 100,000 = 21¢

Definitions:
Line Fallout = Module or subassembly failure requiring troubleshooting, parts replacement and retesting
Warranty Failure = System field failure requiring in warranty repair

Figure 8.1. Component Costs to the User (including line fallout and warranty costs)

Quality and reliability are two essential elements in order for a semiconductor company to be successful in the marketplace today. Quality and reliability are interrelated because reliability is quality extended over the expected life of the product.

Quality is the assurance that a product will fulfill customers' expectations.

Reliability is the probability that a product will perform its intended function satisfactorily for a prescribed life under certain stated conditions.

The quality and reliability of Motorola thyristors are achieved with a four step program:

1) Thoroughly tested designs and materials
2) Stringent in-process controls and inspections
3) Process average testing along with 100% quality assurance redundant testing
4) Reliability verifications through audits and reliability studies

ESSENTIALS OF RELIABILITY

Paramount in the mind of every semiconductor user is the question of device performance versus time. After the applicability of a particular device has been established, its effectiveness depends on the length of trouble free service it can offer. The reliability of a device is exactly that — an expression of how well it will serve the customer. Reliability can be redefined as the probability of failure free performance, under a given manufacturer's specifications, for a given period of time. The failure rate of semiconductors in general, when plotted versus a long period of time, exhibit what has been called the "bath tub curve" (Figure 8.2).

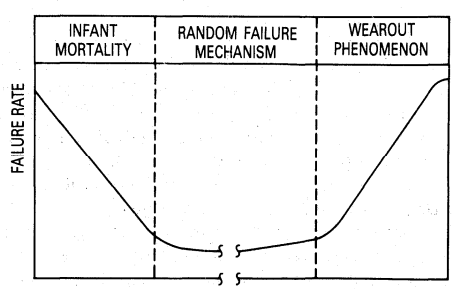


Figure 8.2. Failure Rate of Semiconductor

RELIABILITY MECHANICS

Since reliability evaluations usually involve only samples of an entire population of devices, the concept of the central limit theorem applies and a failure rate is calculated using the λ^2 distribution through the equation:

$$\lambda \leq \frac{\lambda^2 (\alpha, 2r + 2)}{2\ nt}$$

λ^2 = chi squared distribution

where $\alpha = \frac{100 - cl}{100}$

λ = Failure rate

cl = Confidence limit in percent

r = Number of rejects

n = Number of devices

t = Duration of tests

The confidence limit is the degree of conservatism desired in the calculation. The central limit theorem states that the values of any sample of units out of a large population will produce a normal distribution. A 50% confidence limit is termed the best estimate, and is the mean of this distribution. A 90% confidence limit is a very conservative value and results in a higher λ which represents the point at which 90% of the area of the distribution is to the left of that value (Figure 8.3).

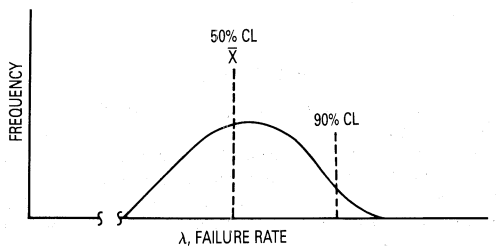


Figure 8.3. Confidence Limits and the Distribution of Sample Failure Rates

The term (2r + 2) is called the degrees of freedom and is an expression of the number of rejects in a form suitable to λ^2 tables. The number of rejects is a critical factor since the definition of rejects often differs between manufacturers. Due to the increasing chance of a test not being representative of the entire population as sample size and test time are decreased, the λ^2 calculation produces surprisingly high values of λ for short test durations even though the true long term failure rate may be quite low. For this reason relatively large amounts of data must be gathered to demonstrate the real long term failure rate. Since this would require years of testing on thousands of devices, methods of accelerated testing have been developed.

Years of semiconductor device testing have shown that temperature will accelerate failures and that this behavior fits the form of the Arrhenius equation:

$$R(t) = R_o(t)e^{-o/KT}$$

Where R(t) = reaction rate as a function of time and temperature

R_o = A constant

t = Time

T = Absolute temperature, °Kelvin (°C + 273°)

o = Activation energy in electron volts (ev)

K = Boltzman's constant = 8.62×10^{-5} ev/°K

This equation can also be put in the form:

AF = Acceleration factor

T2 = User temperature

T1 = Actual test temperature

The Arrhenius equation states that reaction rate increases exponentially with the temperature. This produces a straight line when plotted on log-linear paper with a slope expressed by o. o may be physically interpreted as the energy threshold of a particular reaction or failure mechanism. The overall activation energy exhibited by Motorola thyristors is 1 ev.

RELIABILITY QUALIFICATIONS/EVALUATIONS OUTLINE:

Some of the functions of Motorola Reliability and Quality Assurance Engineering are to evaluate new products for introduction, process changes (whether minor or major), and product line updates to verify the integrity and reliability of conformance, thereby ensuring satisfactory performance in the field. The reliability evaluations may be subjected to a series of extensive reliability testing, such as in the tests performed section, or special tests, depending on the nature of the qualification requirement.

AVERAGE OUTGOING QUALITY (AOQ)

With the industry trend to average outgoing qualities (AOQ) of less than 100 PPM, the role of device final test, and final outgoing quality assurance have become a key ingredient to success. At Motorola, all parts are 100% tested to process average limits then the yields are monitored closely by product engineers, and abnormal areas of fallout are held for engineering investigation. Motorola also 100% redundant tests all dc parameters again after marking the device to further reduce any mixing problems associated with the first test. Prior to shipping, the parts are again sampled, tested to a tight sampling plan by our Quality Assurance department, and finally our outgoing final inspection checks for correct paperwork, mixed product, visual and mechanical inspections prior to packaging to the customers.

AVERAGE OUTGOING QUALITY (AOQ)

$$AOQ = \text{Process Average} \times \text{Probability of Acceptance} \times 10^6 \text{ (PPM)}$$

$$\text{Process Average} = \frac{\text{No. of Reject Devices}}{\text{No. of Devices Tested}}$$

$$\text{Probability of Acceptance} = \left(1 - \frac{\text{No. of Lots Rejected}}{\text{No. of Lots Tested}}\right)$$

$$10^6 = \text{To Convert to Parts Per Million}$$

$$AOQ = \frac{\text{No. of Reject Devices}}{\text{No. of Devices Tested}} \times \left(1 - \frac{\text{No. of Lots Rejected}}{\text{No. of Lots Tested}}\right) \times 10^6 \text{(PPM)}$$

Current AOQ levels (1988) are less than 50 PPM. The projected goal, by 1992, is less than 3 PPM, a defect rate so low that it becomes virtually invisible to the user. Figure 8.4 shows AOQ of Motorola Thyristors.

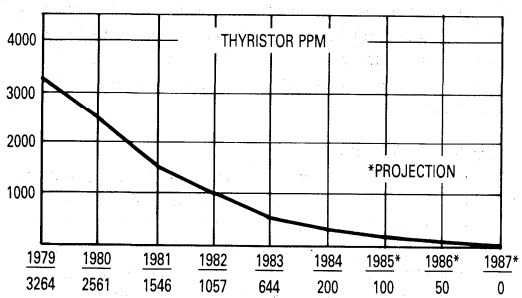
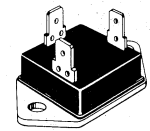


Figure 8.4. Average Outgoing Quality (AOQ) of Motorola Thyristors

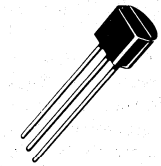
THYRISTOR RELIABILITY

The reliability data described herein applies to Motorola's extensive offering of thyristor products for low and medium current applications. The line includes not only the pervasive Silicon Controlled Rectifiers (SCRs) and TRIACs, but also a variety of Unijunction Transistors (UJTs), Programmable Unijunction Transistors (PUTs), Silicon Bidirectional Switches (SBSs), SIDACs and other associated devices used for SCR and TRIAC triggering purposes. Moreover, these devices are available in different package styles with overlapping current ranges to provide an integral chip-and-package structure that yields lowest cost, consistent with the overriding consideration of high reliability.

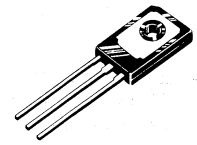
The various packages and the range of electrical specifications associated with the resultant products are shown in Figure 8.5.



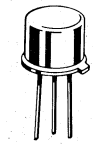
Case 383-01
Devices Available:
SCRs, UJTs, PUTs
Current Range: to 0.5 A
Voltage Range: 25 to 200 V



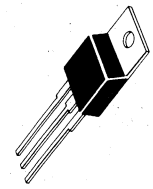
Case 29-04/TO-226AA (TO-92)
Devices Available:
SCRs, TRIACs, UJTs, PUTs, SBSs
Current Range: to 0.8 A
Voltage Range: 25 to 600 V



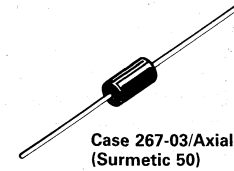
Case 77-05/TO-225AA (TO-126)
Devices Available:
SCRs, TRIACs
Current Range: to 4 A
Voltage Range: 25 to 600 V



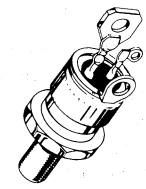
Case 79-04/TO-205AD (TO-39)
Devices Available:
SCRs
Current Range: to 1.6 A
Voltage Range: 25 to 600 V



Case 221A-04/TO-220AB
Devices Available:
SCRs, TRIACs
Current Range: to 40 A
Voltage Range: 50 to 1000 V



Case 267-03/Axial Lead (Surmetic 50)
Devices Available:
SIDAC
Voltage Range: 104 to 240 V



Case 311-02/ Press Fit/Stud
Devices Available:
SCRs, TRIACs
Current Range: to 55 A
Voltage Range: 25 to 800 V

Figure 8.5. Motorola Thyristor Packages

To evaluate the reliability of these structures, production line samples from each type of package are being subjected to a battery of accelerated reliability tests deliberately designed to induce long-term failure. Though the tests are being conducted on a continuing basis, the results so far are both meaningful and impressive. They are detailed on the following pages in the hope that they will provide for the readers a greater awareness of the potential for thyristors in their individual application.

THYRISTOR CONSTRUCTION THROUGH A TIME TESTED DESIGN AND ADVANCED PROCESSING METHODS

A pioneer in discrete semiconductor components and the world's largest supplier thereof, Motorola has pyramided continual process and material improvements into thyristor products whose inherent reliability meets the most critical requirements of the market.

These improvements are directed towards long-term reliability in the most strenuous applications and the most adverse environments.

1

DIE GLASSIVATION

All Motorola thyristor die are glass-sealed with a Motorola patented passivation process making the sensitive junctions impervious to moisture and impurity penetration. This imparts to low-cost plastic devices the same freedom from external contamination formerly associated only with hermetically sealed metal packages. Thus, metal encapsulation is required primarily for higher current devices that would normally exceed the power-dissipation capabilities of plastic packages — or for applications that specify the hermetic package.

HIGHLY MECHANIZED ASSEMBLY

Motorola uses many techniques to ensure quality manufacturing of its thyristor products. A few of the newest techniques are (1) all diffusion steps are performed by microprocessor-controlled diffusion tubes, (2) four point probes are used to verify resistivity at each step during wafer processing, (3) all chips are individually probed to their electrical specifications, (4) wafer preparation for assembly includes laser scribing to separate the die while in wafer form, (5) assembly furnaces are automated to perform automatic insertion and removal of assembly fixtures.

VOID-FREE PLASTIC ENCAPSULATION

A fifth generation plastic package material, combined with improved copper piece-part designs, maximize package integrity during thermal stresses. The void-free encapsulation process imparts to the plastic package a mechanical reliability (ability to withstand shock and vibration) even beyond that of metal packaged devices.

IN-PROCESS CONTROLS AND INSPECTIONS

INCOMING INSPECTIONS

Apparently routine procedures, inspection of incoming parts and materials, are actually among the most critical segments of the quality and reliability assurance program. That's because small deviations from materials specifications can traverse the entire production cycle before being detected by outgoing Quality Control, and, if undetected, could affect long-term reliability. At Motorola, piece-part control involves the services of three separate laboratories . . . Radiology, Electron Optics and Product Analysis. All three are utilized to insure product integrity:

Raw Wafer Quality, in terms of defects, orientation, flatness and resistivity;

Physical Dimensions, to tightly specified tolerances;
Metal Hardness, to highly controlled limits;
Gaseous Purity and Doping Level;
Mold Compounds, for void-free plastic encapsulation.

IN-PROCESS INSPECTIONS

As illustrated in Figure 8.6, every major manufacturing step is followed by an appropriate in-process QA inspection. Quality control in wafer processing, assembly and final test impart to Motorola standard thyristors a reliability level that easily exceeds most industrial, consumer and military requirements . . . built-in quality assurance aimed at insuring failure-free shipments of Motorola products.

RELIABILITY AUDITS

Reliability audits are performed following assembly. Reliability audits are used to detect process shifts which can have an adverse effect on long-term reliability. Extreme stress testing on a real-time basis, for each product run, uncovers process abnormalities that may have escaped the stringent in-process controls. Typical tests include HTRB/FB (high-temperature reverse bias and forward bias) storage life and temperature cycling. When abnormalities are detected, steps are taken to correct the process.

OUTGOING QC

The most stringent in-process controls do not guarantee strict adherence to tight electrical specifications. Motorola's 100% electrical parametric test does — by eliminating all devices that do not conform to the specified characteristics. Additional parametric tests, on a sampling basis, provide data for continued improvement of product quality. And to help insure safe arrival after shipment, antistatic handling and packaging methods are employed to assure that the product quality that has been built in stays that way.

From rigid incoming inspection of piece parts and materials to stringent outgoing quality verification, assembly and process controls encompass an elaborate system of test and inspection stations that ensure step-by-step adherence to a prescribed procedure designed to yield a high standard of quality.

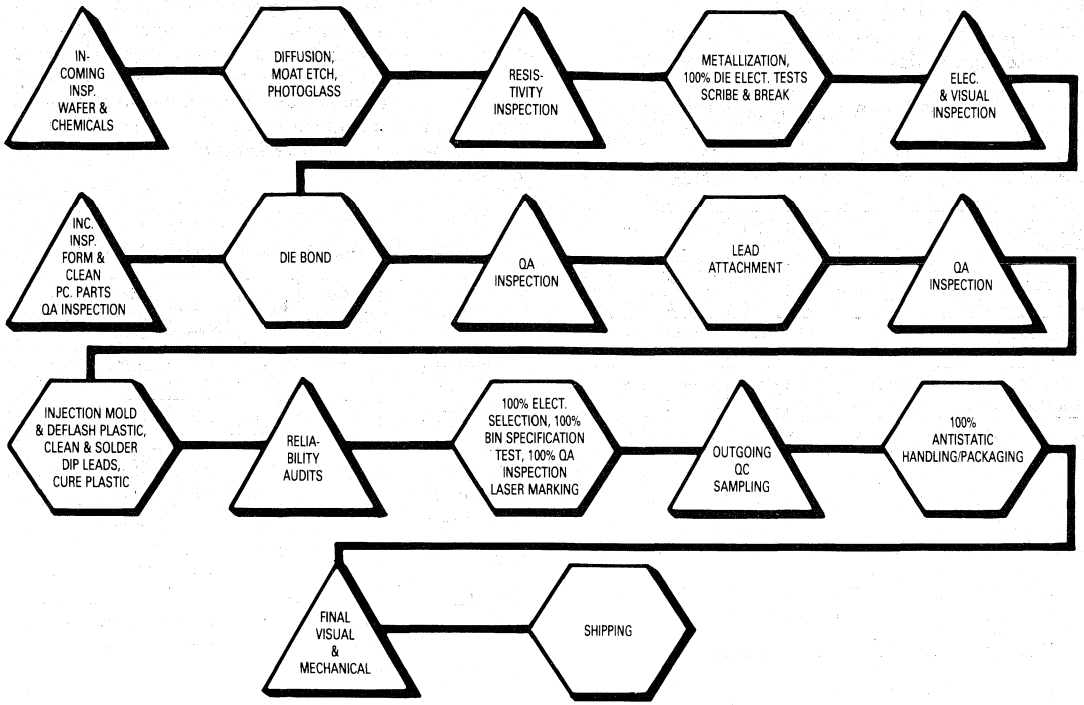


Figure 8.6. In-Process Quality Assurance Inspection Points for Thyristors

RELIABILITY TESTS

Only actual use of millions of devices, under a thousand different operating conditions, can conclusively establish the reliability of devices under the extremes of time, temperature, humidity, shock, vibration and the myriads of other adverse variables likely to be encountered in practice. But thorough testing, in conjunction with rigorous statistical analysis, is the next-best thing. The series of torture tests described in this document instills a high confidence level regarding thyristor reliability. The tests are conducted at maximum device ratings and are designed to deliberately stress the devices in their most susceptible failure modes. The severity of the tests compresses into a relatively short test cycle the equivalent of the stresses encountered during years of operation under more normal conditions. The results not only indicate the degree of reliability in terms of anticipated failures; they trigger subsequent investigations into failure modes and failure mechanisms that serve as the basis of continual improvements. And they represent a clear-cut endorsement that, for Motorola thyristors, low-cost and high quality are compatible attributes.

BLOCKING LIFE TEST

This test is used as an indicator of long-term operating reliability and overall junction stability (quality). All semiconductor junctions exhibit some leakage current under reverse-bias conditions. Thyristors, in addition, exhibit leakage current under forward-bias conditions in the off state. As a normal property of semiconductors, this junction leakage current increases proportionally with temperature in a very predictable fashion.

Leakage current can also change as a function of time — particularly under high-temperature operation. Moreover, this undesirable "drift" can produce catastrophic failures when devices are operated at, or in excess of, rated temperature limits for prolonged periods.

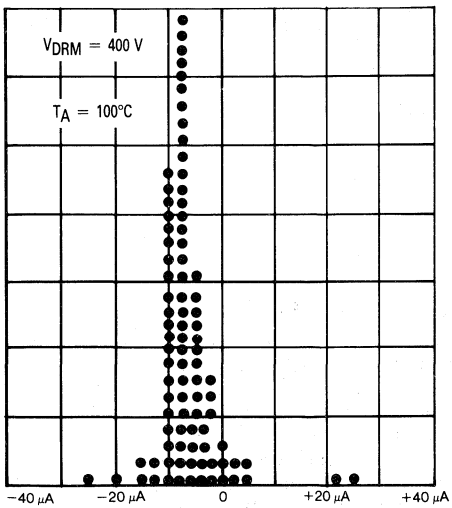
The blocking life test operates representative numbers of devices at rated (high) temperature and reverse-bias voltage limits to define device quality (as measured by leakage drifts) and reliability (as indicated by the number of catastrophic failures*). The results of these tests are shown in Table 8.I. Table 8.II shows leakage-current drift after 1000 hours HTRB.

Table 8.I. Blocking Life Test High Temperature Reverse Bias (HTRB) and High Temperature Forward Bias (HTFB)

Case	Test Conditions T_A @ Rated Voltage	Sample Size	Duration (Hours)	Total Device Hours	Catastrophic Failures*
29-04/TO-226AA (TO-92)	100°C	1000	1000	1,000,000	1
77-07/TO-225AA (TO-126)	110°C	1000	1000	1,000,000	0
221A-04/TO-220AB	100°C	1000	1000	1,000,000	0
311-02/Press Fit/Stud	100°C	200	1000	200,000	0
79-04/TO-205AD (TO-39)	110°C	150	1000	150,000	0
267-03/Axial Lead (Surmetic 50)	125°C	150	1000	150,000	0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

Table 8.II. Leakage-Current Drift after 1000 Hours HTRB



The favorable blocking-life-test drift results shown here are attributed to Motorola's unique "glassivated junction" process which imparts a high degree of stability to the devices.

HIGH TEMPERATURE STORAGE LIFE TEST

This test consists of placing devices in a high-temperature chamber. Devices are tested electrically prior to exposure to the high temperature, at various time intervals during the test, and at the completion of testing. Electrical readout results indicate the stability of the devices, their potential to withstand high temperatures, and the internal manufacturing integrity of the package. Readouts at the various intervals offer information as to the time period in which failures occur. Although devices are not exposed to such extreme high temperatures in the field, the purpose of this test is to accelerate any failure mechanisms that could occur during long periods at actual storage temperatures. Results of this test are shown in Table 8.III.

Table 8.III. High Temperature Storage Life

Case	Test Conditions	Sample Size	Duration (Hours)	Total Device Hours	Catastrophic Failures*
29-04/TO-226AA (TO-92)	T_A 150°C	1000-2000	400	1,500,000	0
77-07/TO-225AA (TO-126)	**	1000-2000	350	550,000	0
221A-04/TO-220		1000	300	300,000	0
311-02/Press Fit/Stud		1000	125	125,000	0
79-04/TO-205AD (TO-39)		1000-2000	100	150,000	0
267-03/Axial Lead (Surmetic 50)		1000	100	100,000	0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

**Same for all.

STRESS TESTING — POWER CYCLING AND THERMAL SHOCK

POWER CYCLING TEST

How do the devices hold up when they are repeatedly cycled from the off state to the on state and back to the off state under conditions that force them to maximum rated junction temperature during each cycle? The Power Cycling Test was devised to provide the answers.

In this test, devices are subjected to full-rated, on-state power until rated junction temperature (T_J) has been reached. The devices are then turned off and T_J decreases to near ambient, at which time the cycle is repeated.

This test is important to determine the integrity of the chip and lead frame assembly since it repeatedly stresses the devices to their maximum ratings. It is unlikely that these worst-case conditions would be continuously encountered in actual use. Any reduction in T_J results in an exponential increase in operating longevity. Table 8.IV shows the results of power cycling.

Table 8.IV. Power Cycling

Case	Test Conditions	Sample Size	Duration (Cycles)	Total Device Cycles	Catastrophic Failures*
29-04/TO-226AA (TO-92)	$I_{T(rms)}$ = Maximum Rating ΔT_J = 70°C (30°C to 100°C) or ΔT_J = 95°C (30°C to 125°C) Depending on Maximum T_J Rating Force Air Cooling **t_{off} = Adjust per device type	100 200 100	5,000 10,000 20,000	500,000 3,000,000 2,000,000	0 1 0
77-07/TO-225AA (TO-126)		622	10,000	6,220,000	0
221A-04/TO-220		200 200 100	5,000 10,000 20,000	1,000,000 2,000,000 2,000,000	0 0 0
311-02/Press Fit/Stud 79-04/TO-205AD (TO-39)		50 75	10,000 10,000	500,000 750,000	0 0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.
**Typical 4 seconds OFF.

THERMAL SHOCK CONDITIONS BEYOND THE NORM

Excesses in temperature not only cause variations in electrical characteristics, they can raise havoc with the mechanical system. Under temperature extremes, contraction and expansion of the chip and package can cause physical dislocations of mechanical interfaces and induce catastrophic failure.

To evaluate the integrity of Motorola thyristors under the most adverse temperature conditions, they are subjected to two different thermal shock tests: liquid-to-liquid and air-to-air.

LIQUID-TO-LIQUID (GLASS STRAIN)

This thermal shock test is performed for the same reasons as the temperature cycling test, however, the extreme changes in temperature are more sudden. This is an especially useful tool for evaluating the metal to nonmetal interface. Results of this test are shown in Table 8.V.

AIR-TO-AIR (TEMPERATURE CYCLING)

This thermal shock test is conducted to determine the ability of the devices to withstand exposure to extreme high and low temperature environments and to the shock of alternate exposures to the temperature extremes. Results of this test are shown in Table 8.VI.

Table 8.V. Liquid-to-Liquid

Case	Test Conditions	Sample Size	Number of Cycles	Catastrophic Failures*
29-04/TO-226AA (TO-92)	Mil Std 750, Method 1056-1	950	300	0
77-07/TO-225AA	0°C to +100°C	200	300	0
221A-04/TO-220 (TO-126)	Dwell Time — 15 seconds	300	300	◀0
79-04/TO-205AD (TO-39)	Immediate Transfer	75	100	0
311-02/Press Fit/Stud		75	100	0
267-03/Axial Lead (Surmetic 50)		75	300	0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

Table 8.VI. Air-to-Air

Case	Test Conditions	Sample Size	Number of Cycles	Total Device Cycles	Catastrophic Failures*
29-04/TO-226AA (TO-92)	−40°C or −65°C	900	400	360,000	0
77-07/TO-225AA (TO-126)	to +150°C	500	400	200,000	0
221A-04/TO-220	Dwell — 15 minutes	400	400	160,000	0
311-02/Press Fit/Stud	at each extreme	150	400	60,000	0
79-04/TO-205AD (TO-39)	Immediate Transfer	75	50	3,750	0
267-03/Axial Lead (Surmetic 50)		100	400	40,000	

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

1

ENVIRONMENTAL TESTING

MOISTURE TESTS

Humidity has been a traditional enemy of semiconductors, particularly plastic packaged devices. Most moisture-related degradations result, directly or indirectly, from penetration of moisture vapor through passivating materials, and from surface corrosion. At Motorola, this erstwhile problem has been effectively controlled through the use of a unique junction "glassivation" process and selection of package materials. The resistance to moisture-related failures is indicated by the tests described here.

BIASED HUMIDITY TEST

This test was devised to determine the resistance of component parts and constituent materials to the combined deteriorative effects of prolonged operation in a high-temperature/high-humidity environment. H^3TRB test results are shown in Table 8.VII.

MOISTURE RESISTANCE TEST

This test evaluates the deteriorative effects of temperature cycling and high humidity under accelerated conditions simulating tropical environments. During this test, devices are subjected to alternate periods of condensation and drying, which "breathe" moisture into nonhermetic packages and accelerate the development of corrosion. Increased effectiveness is further obtained by use of high temperatures which intensify the effects of humidity. Test results are shown in Table 8.VIII.

Table 8.VII. Biased Humidity Test
High humidity, high Temperature, reverse bias (H^3TRB)

Case	Test Conditions	Sample Size	Duration Hours	Total Device Cycles	Catastrophic Failures*
29-04/TO-226AA (TO-82)	Relative Humidity 85% T_A = 85°C	400	500–1000	300,000	0
77-07/TO-225AA	Reverse Voltage-Rated	200	500–1000	150,000	0
221A-04/TO-220	or 200 V Maximum	100	500–1000	75,000	0
267-03/Axial Lead (Surmetic 50)		30	1000	30,000	0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

Table 8.VIII. Moisture Resistance Test

Case	Test Conditions	Sample Size	Duration Hours	Catastrophic Failures*
29-04/TO-226AA (TO-92)	Mil-Std-750 Method 1021	100	960	0
77-07/TO-225AA (TO-126)	Relative Humidity 92–98%	100	960	0
221A-04/TO-220	Temperature Cycle — Cycle Time 24 Hours	50	720	0
267-03/Axial Lead (Surmetic 50)		40	960	0

*Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

MECHANICAL TESTING

VIBRATION TEST (VARIABLE FREQUENCY)

Vibration, by causing loosening of parts or relative motion between parts in the sample, can produce objectionable operating characteristics, noise, wear and physical distortion, and often results in fatigue and failure of mechanical parts.

This test determines the effects of vibration within the predominate frequency ranges and magnitudes that may be encountered by thyristors during field service. Most vibration encountered in field service is not of a simple harmonic nature. Nevertheless, tests based on vibrations of this type have proved satisfactory for determining critical frequencies, modes of vibration and other data necessary for planning protective steps against the effects of undue vibration. Test results are shown in Table 8.IX.

Table 8.IX

Test Conditions Mil-S-750, Method 2056	Case	Duration (Minutes)	Sample Size	Catastrophic Failures*
20 G's in X_1 Y_1 Z_1 Axis	77-07	48	100	0
	221A-04	48	150	0
	29-04	48	340	0
	79-04	48	—	—
	311-02		25	0
	267-03	48	84	0

* Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

SOLVENTS TEST

Almost invariably, in actual practice, plastic thyristors are subjected to a cleaning solution to remove residue flux or solvents utilized during circuit manufacturing. Frequently, the cleaning solution is highly chlorinated and the devices are completely immersed in an elevated temperature bath. Motorola plastic thyristors subjected to hot chlorinated baths have proved highly resistant to the potentially damaging effects of this chemical.

Alcohols, water baths, freon, etc., produce no known latent failure mechanisms. Tests have indicated that Motorola plastic packages will withstand all cleaning agents commonly used. Solvents' test results are shown in Table 8.X.

Table 8.X.

Test Conditions	Case	Number of Minutes	Sample Size	Catastrophic Failures*
Chlorothene Immersion at T_A = 50 °C	221A-04	30	100	0
	77-07	30	50	0
	29-04	30	50	0
	267-03	30	50	0

* Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

CONSTANT ACCELERATION TEST

The Constant Acceleration test is used to determine the effects of a centrifugal force on semiconductor devices. This test is designed to indicate types of structural and mechanical weaknesses not necessarily detected in shock and vibration tests. The results of this test are shown in Table 8.XI.

Table 8.XI

Test Conditions Mil-S 750, Method 2006	Case	Duration (Minutes)	Sample Size	Catastrophic Failures*
10,000 G's in Y_1 Y_2 Axis	77-07	1	100	0
	29-04	1	340	0
	221A-04	1	150	0
	267-03	1	84	0

* Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

DROP SHOCK TEST

Drop shock testing simulates the stress encountered by a device during rough handling, transportation or field operation. Shocks of this type may disturb operating characteristics or may cause catastrophic failures. Motorola plastic thyristors exhibited extremely good resistance to mechanical shock. Results are shown in Table 8.XII.

Table 8.XII

Test Conditions Mil-STD-202, Method 202	Case	Sample Size	Catastrophic Failures*
X_1 Y_1 Axis 15 G's, 10 Blows	77-07	100	1
	221A-04	150	0
	29-04	340	0
	311-02	25	0
	267-03	84	0

* Failures are at maximum rated values. The severe nature of these tests is normally not seen under actual conditions.

CHAPTER 9
USING THYRISTOR DIE FOR HYBRID ASSEMBLY

1

Substantial savings in weight and volume can be achieved by hybrid packaging techniques. Most Motorola thyristor packaged devices are in die form for custom hybrid assembly. The same advanced thyristor processing techniques available in packaged form is available in die form.

DIE CHARACTERISTICS

Currently 8 die sizes are available. Four in the low current range (< 8 amps) and 4 in the medium current range (> 8 amps). Table 9.I lists all low current devices and their die sizes while Table 9.II lists all medium current devices (for die layout and bond pad dimensions consult the factory).

Table 9.I. Low Current (< 8 Amps)

Die Metallization: Al (front), Au (back)

Part Number	*Die Size (Mil^2)
MCRC100-4 thru -6	42
MCR106-2 thru -6	70
MAC97-4 thru -6	42
2C6071 thru 6074	84
2C6027	20
MBSC4991	20

*All Thyristor Die are square unless indicated.

Table 9.II. Medium Current (> 8 Amps)

Die metallization: Solderable TiNiAg (Front and Back)

Part Number	*Die Size (Mil^2)
MCRC218-2 thru -8	92
MCRC72-2 thru -6	92
MCRC67-2 thru -6	92
2C6394 thru 6398	92
MCRC68-2 thru -6	92
2C6400 thru 6404	150
MCRC70-2 thru 70-6	150
MCRC264-2 thru -8	180
CC228A thru CC228M	150
MCRC265-2 thru -8	210
TC2500B thru N	120
MACC218-4 thru -8	120
2C6342 thru 6348	120
MACC228A4 thru A8	120
2C6342A thru 6348A	150
MACC15A4 thru A8	150
MACC223A4 thru A8	180
TC6421B thru N	210
MACC224-4 thru -8	210

*All Thyristor Die are square unless indicated.

ELECTRICAL CHARACTERISTICS

All die are individually probed at room temperature. However due to limitations when probing in wafer (die) form some of the specifications of the equivalent packaged device cannot be tested and guaranteed in wafer (die) form. These parameters include $R_{\theta JC}$, V_{TM} at high current, switching times, and all elevated and reduced temperature tests. The above parameters depend upon

assembly techniques of the individual user. Each die is 100% tested by state-of-the-art computer test equipment and visually inspected per MIL-STD 19500 and MIL-STD 750 prior to shipment. Electrical specifications for most devices listed in Tables 9.I and 9.II can be found by consulting the corresponding data sheet for the packaged part.

Example:	MCRC218-2 thru 10	See MCR218 Data Sheet
	CC228A thru M	See C228 Data Sheet

1

VISUAL INSPECTION OF THYRISTOR DIE

All Motorola thyristor die meet the visual inspection criteria of MIL-Standard 750B, Method 2073, with the exception of specific criteria listed below. All thyristor dice are visually screened to a 1% AQL level.

DIE BACKING

All standard thyristor die come with the following metallization: low current (< 8 amps) — aluminum front metal and gold back metal, and medium current (> 8 amps) solderable titanium nickel silver on both front and back metal. This metallization is suitable for solder preform mounting. Commonly used header materials such as copper, nickel plated copper, gold plated molybdenum, beryllia, and alumina are acceptable. The substrate must be free of all oxides prior to assembly. Mounting is generally accomplished in a profiled belt furnace (hydrogen atmosphere is recommended).

WIRE BONDING AND CLIP ATTACHMENT

Electrical connection for low current thyristors can be accomplished by ultrasonic wire bonding, using AlMg* wire having an elongation of 10%. Caution should be exercised during wire bonding that the bond footprint remains within the bonding pad area. Wire bond settings should be optimized and a wire pull test performed (see Method 2037, MIL-STD 750B) to monitor wire bond strength uniformity. Destructive sample testing and 100% non-destructive testing is recommended. Electrical connection for medium current thyristors can be accomplished through the solder clip method as the pad metallization is a solderable TiNiAg. Assembly techniques such as a profiled belt furnace (hydrogen atmosphere) is recommended for attachment of the clips. Clips are recommended in order to withstand the high surge currents.

*Wire sizes of 15 mils and greater are pure Al.

ENCAPSULATION

Before encapsulation, the assembly must be kept in a moisture-free environment. For a non-hermetic package, a high grade electronic coating such as Dow Corning RTV3140 should be applied (coating is optional with hermetic package). Before encapsulation, a 150°C two-hour bake should be performed to remove any surface moisture, and any capping of hermetic packages must be performed in a dry nitrogen atmosphere.

HANDLING AND SHIPPING

Thyristor die are packaged several ways.

MULTI-PAK

This is a 2″ x 2″ or a 4″ x 4″ waffle type carrier with a separate carrier for each die, holding 100 to 400 die depending upon die size. The multi-pak is shown in Figure 9.1.

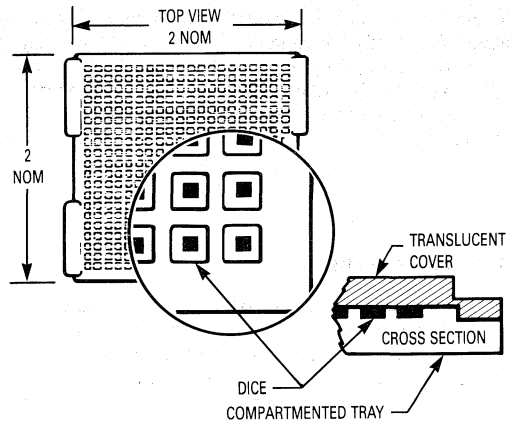


Figure 9.1. Multi-Pak

WAFER PAK

The die are 100% electrically tested (rejects are inked) but the wafer is left unsawed. No visual inspection is performed.

VACUUM PAK

The die are 100% electrically tested (rejects are inked). A wafer or a quarter wafer is scribed and broken with the wafer left on the mylar. This is then sealed in a vacuum bag. No visual inspection is performed.

Upon opening the packaging medium, dice should be stored in a nitrogen atmosphere to prevent oxidation of bond areas prior to assembly. All dice should be handled with teflon tipped probes to prevent any mechanical damage.

APPENDICES

APPENDIX I

USING THE TWO TRANSISTOR ANALYSIS

DEFINITIONS:

I_C ≡ Collector current
I_B ≡ Base current
I_{CS} ≡ Collector leakage current (saturation component)
I_A ≡ Anode current
I_K ≡ Cathode current
α ≡ Current amplification factor
I_G ≡ Gate current

The subscript "i" indicates the appropriate transistor.

FOR TRANSISTOR #1:

$I_{C1} = \alpha_1 I_A + I_{CSI}$

and

$I_{B1} = I_A - I_{C1}$

Combining these equations,

$$I_{B1} = (1 - \alpha_1) I_A - I_{CS1} \qquad (1)$$

LIKEWISE, FOR TRANSISTOR #2

$$I_{C2} = \alpha_2 I_K + I_{CS2} \qquad (2)$$

$I_{B1} = I_{C2}$

and by combining Equations (1) and (2) and substituting $I_K = I_A + I_G$, it is found that

$$I_A = \frac{\alpha_2 I_G + I_{CS1} + I_{CS2}}{1 - \alpha_1 - \alpha_2} \qquad (3)$$

Equation (3) relates I_A to I_G, and note that as $\alpha_1 + \alpha_2 = 1$, I_A goes to infinity. I_A can be put in terms of I_K and α's as follows:

$I_{B1} = I_{C2}$

Combining equations (1) and (2):

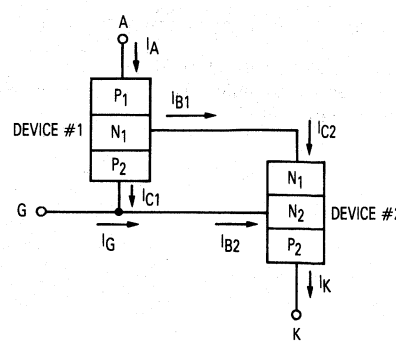
$$I_A = \frac{I_{CS1} + I_{CS2}}{1 - \alpha_1 - (\frac{I_K}{I_A})\alpha_2}$$

I_A — ∞ if denominator approaches zero, i.e., if

$$\frac{I_K}{I_A} = \frac{1 - \alpha_1}{\alpha_2}$$

Note that just prior to turn-on there is a majority carrier build-up in the P_2 "base." If the gate bias is small there will actually be hole current flowing out from P_2 into the gate circuit so that I_G is negative, $I_K = I_A + I_G$ is less than I_A so: (see Figure 3.2 for the directions of current components)

$\frac{I_K}{I_A} < 1$ which corresponds to $\alpha_1 + \alpha_2 > 1$



Schematic Diagram of the Two Transistor Model of a Thyristor

APPENDIX II

CHARGE AND PULSE WIDTH

In the region of large pulse widths using current triggering, where transit time effects are not a factor, we can consider the input gate charge for triggering, Q_{in}, as consisting of three components:

1) Triggering charge Q_{tr}, assumed to be constant.
2) Charge lost in recombination, Q_r, during current regeneration prior to turn-on.
3) Charge drained, Q_{dr}, which is by-passed through the built-in gate cathode shunt resistance (the presence of this shunting resistance is required to increase the dv/dt capability of the device).

Mathematically, we have

$$Q_{in} = Q_{tr} + Q_{dr} + Q_r = I_G\tau \qquad (1)$$

Q_r is assumed to be proportional to Q_{in}; to be exact,

$$Q_r = Q_{in}\,(1 - \exp^{-\tau/\tau_1}) \qquad (2)$$

where I_G = gate current,

τ = pulse width of gate current,

τ_1 = effective life time of minority carriers in the bases

The voltage across the gate to cathode P-N junction during forward bias is given by V_{GK} (usually 0.6 V for silicon).* The gate shunt resistance is R_s (for the MCR729, typically 100 ohms), so the drained charge can be expressed by

$$Q_{dr} = \frac{V_{GC}}{R_s}\,\tau \qquad (3)$$

Combining equations (1), (2), and (3), we get

$$Q_{in} = I_G\tau = (Q_{tr} + \frac{V_{GC}}{R_s}\tau)\ \exp.\ \tau/\tau_1 \qquad (4)$$

Note that at region A and C of Figure 3.4(b) Q_{in} has an increasing trend with pulse width as qualitatively described by Equation (4).

Assume life time at the temperature range of operation increases as some power of temperature

$$\tau_1 = KT^m \qquad (5)$$

where K and m are positive real numbers. Combining Equations (4) and (5), we can get the slope of Q_{in} with respect to temperature to be

$$\text{slope} = \frac{dQ_{in}}{dT} = -m(Q_{tr} + \frac{V_{GC}}{R_s}\tau)\,\frac{\tau}{\tau_1}\,\frac{\exp.\ \tau/\tau_1}{T} \qquad (6)$$

In reality, Q_{tr} is not independent of temperature, in which case the Equation (6) must be modified by adding an additional term to become:

$$\text{slope} = -m(Q_{tr} + \frac{V_{GC}}{R_s}\tau)\,\frac{\tau}{\tau_1}\,\frac{\exp.\tau/\tau_1}{T} + \frac{dQ_{tr}}{dT}\exp\tau/\tau_1 \qquad (7)$$

Physically, not only does Q_{tr} decrease with temperature so that dQ_{tr}/dT is a negative number, but also $|dQ_{tr}/dT|$ decreases with temperature as does $|d\alpha/dT|$ in the temperature range of interest.

Equation (6) [or (7)] indicates two things:

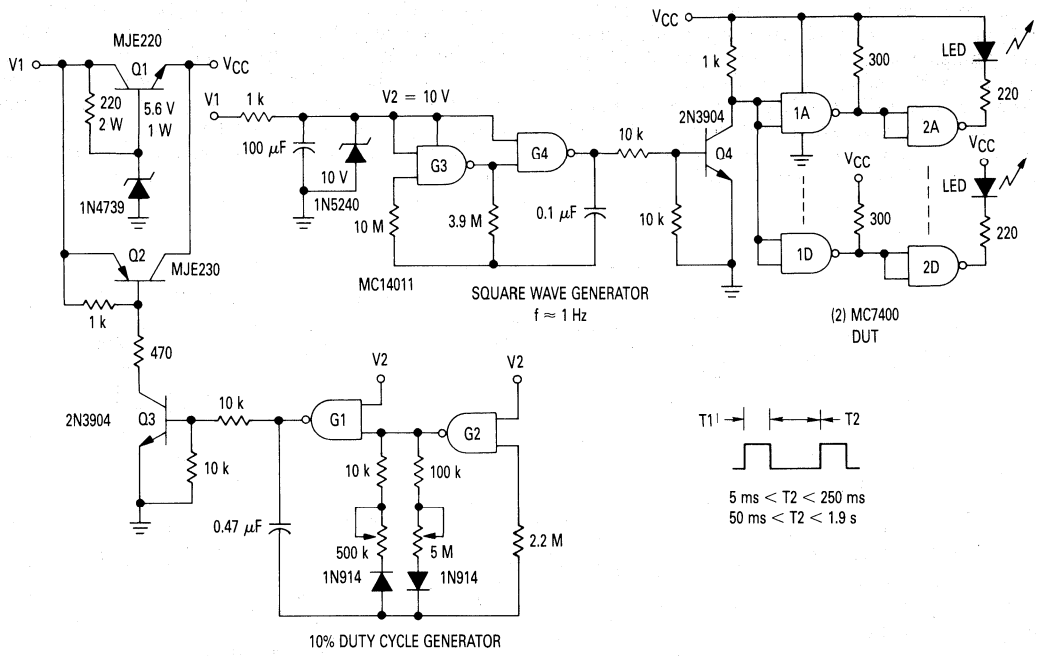
1) The rate of change of input trigger charge decreases as temperature (life time) increases.
2) The larger the pulse width of gate trigger current, the faster the rate of change of Q_{in} with respect to change in temperature. Figure 3.11 shows these trends.

*V_{GC} is not independent of I_G. For example, for the MCR729 the saturation V_{GC} is typically 1 V, but at lower I_G's the V_{GC} is also smaller, e.g. for I_G = 5 mA, V_{GC} is typically 0.3 V.

APPENDIX III

TTL SOA TEST CIRCUIT

Using the illustrated test circuit, the two TTL packages (quad, 2-input NAND gates) to be tested were powered by the simple, series regulator that is periodically shorted by the clamp transistor, Q2, at 10% duty cycle rate. By varying the input to the regulator V1 and the clamp pulse width, various power levels can be supplied to the TTL load. Thus, as an example, V_{CC} could be at 5 V for 90 ms and 10 V for 10 ms, simulating a transient on the bus or a possibly shorted power supply pass transistor for that duration. These energy levels are progressively increased until the gate (or gates) fail, as detected by the status of the output LEDs, the voltage and current waveforms and the device case temperature.



TTL SOA Test Circuit

APPENDIX IV

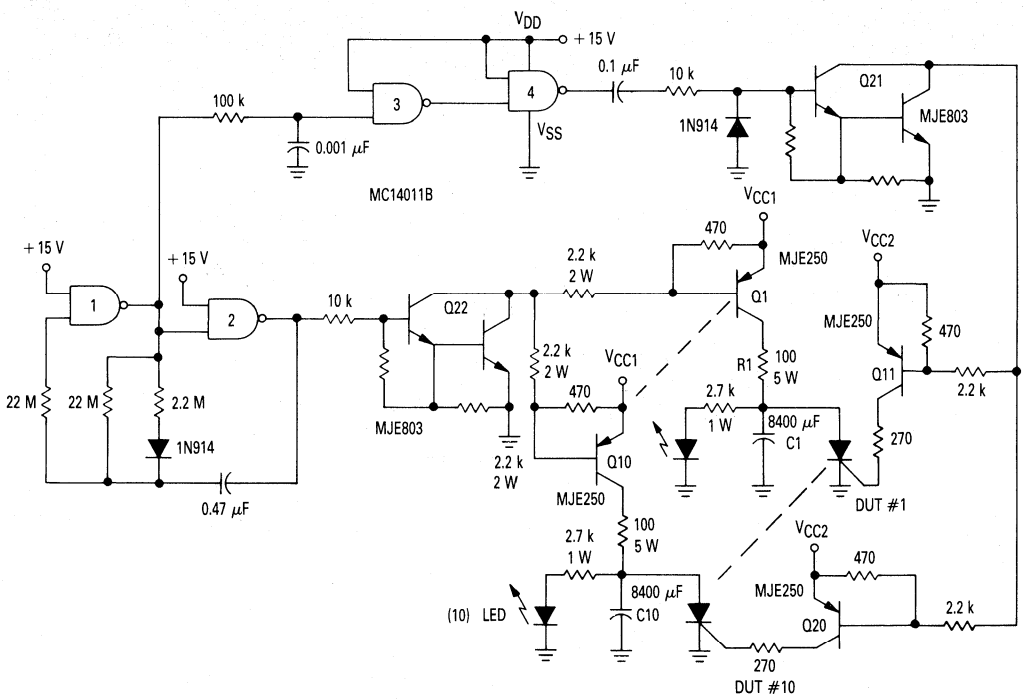
SCR CROWBAR LIFE TESTING

This crowbar life test fixture can simultaneously test ten SCRs under various crowbar energy and gate drive conditions and works as follows.

The CMOS Astable M.V. (Gates 1 and 2) generate an asymmetric Gate 2 output of about ten seconds high, one second low. This pulse is amplified by Darlington Q22 to turn on the capacitor charging transistors Q1–Q10 for the ten seconds. The capacitors for crowbarring are thus charged in about four seconds to whatever power supply voltage to which V_{CC1} is set. The charging transistors are then turned off for one second and the SCRs are fired by an approximately 100 μs delayed trigger derived from Gates 3 and 4. The R-C network on Gate 3 input integrates the complementary pulse from Gate 1, resulting in the delay, thus insuring non-coincident firing of the test circuit. The shaped pulse out of Gate 4 is differentiated and the positive-going pulse is amplified by Q21 and the following ten SCR gate drivers (Q11–Q20) to form the approximate 2 ms wide, 1 μs rise time, SCR gate triggers, I_{GT}. I_{GT} is set by the collector resistors of the respective gate drivers and the supply voltage, V_{CC2}; thus, for $I_{GT} \approx 100$ mA, $V_{CC2} \approx 30$ V, etc.

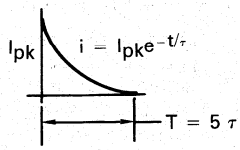
The LEDs across the storage capacitors show the state of the voltage on the capacitors and help determine whether the circuit is functioning properly. The timing sequence would be an off LED for the one-second capacitor dump period followed by an increasingly brighter LED during the capacitor charge time. Monitoring the current of V_{CC1} will also indicate proper operation.

The fixture's maximum energy limits are set by the working voltage of the capacitors and breakdown voltage of the transistors. For this illustration, the 60 V, 8400 μF capacitors (ESR $\approx$ 20 mΩ) produced a peak current of about 2500 A lasting for about 0.5 ms when V_{CC1} equals 60 V. Other energy values (lower i_{pk}, greater t_w) can be obtained by placing a current limiting resistor between the positive side of the capacitor and the crowbar SCR anode.



APPENDIX V

DERIVATION OF THE RMS CURRENT OF AN EXPONENTIALLY DECAYING CURRENT WAVEFORM



$$I_{rms} = \sqrt{\frac{1}{T}\int_0^T i^2(t)dt}$$

$$= \left[\frac{1}{T}\int_0^T (I_{pk}e^{-t/\tau})^2 dt\right]^{1/2}$$

$$= \left[\frac{I_{pk}^2}{T}\,\frac{e^{-2t/\tau}}{(-2/\tau)}\Bigg|_0^T\right]^{1/2}$$

$$= \left[\frac{I_{pk}^2}{T}\left(\frac{-\tau}{2}\right)(e^{-2T/\tau} - e^0)\right]^{1/2}$$

where $T = 5\,\tau$,

$$= \left[-\frac{I_{pk}^2}{10}(e^{-10} - 1)\right]^{1/2}$$

$$I_{rms} = \frac{I_{pk}}{\sqrt{10}} = 0.316\,I_{pk}$$

APPENDIX VI

DERIVATION OF I^2t FOR VARIOUS TIMES

Thermal Equation $\Delta t = Z(\theta)P_D$

where $Z(\theta) = r(t)R_{\theta JC}$

and $r(t) = K\sqrt{t}$

Therefore, for the same Δt,

$$\Delta t = K\sqrt{t_1}\,R_{\theta JC}P_{D_1} = K\sqrt{t_2}\,R_{\theta JC}P_{D_2},$$

$$\frac{P_{D_1}}{P_{D_2}} = \sqrt{\frac{t_2}{t_1}} = \frac{I_1^2R}{I_2^2R},$$

$$\frac{I_1^2}{I_2^2} = \sqrt{\frac{t_2}{t_1}}$$

Multiplying both sides by (t_1/t_2),

$$\frac{I_1^2t_1}{I_2^2t_2} = \left(\frac{t_2}{t_1}\right)^{1/2}\frac{t_1}{t_2} = \left(\frac{t_1}{t_2}\right)^{1/2}.$$

$$I_1^2t_1 = I_2^2t_2\sqrt{\frac{t_1}{t_2}}$$

1

APPENDIX VII

THERMAL RESISTANCE CONCEPTS

The basic equation for heat transfer under steady-state conditions is generally written as:

$$q = hA\Delta T \qquad (1)$$

where q = rate of heat transfer or power dissipation (P_D),
h = heat transfer coefficient,
A = area involved in heat transfer,
ΔT = temperature difference between regions of heat transfer.

However, electrical engineers generally find it easier to work in terms of thermal resistance, defined as the ratio of temperature to power. From Equation (1), thermal resistance, R_θ, is

$$R_\theta = \Delta T/q = 1/hA \qquad (2)$$

The coefficient (h) depends upon the heat transfer mechanism used and various factors involved in that particular mechanism.

An analogy between Equation (2) and Ohm's Law is often made to form models of heat flow. Note that ΔT could be thought of as a voltage; thermal resistance corresponds to electrical resistance (R); and, power (q) is analogous to current (I). This gives rise to a basic thermal resistance model for a semiconductor (indicated by Figure A3).

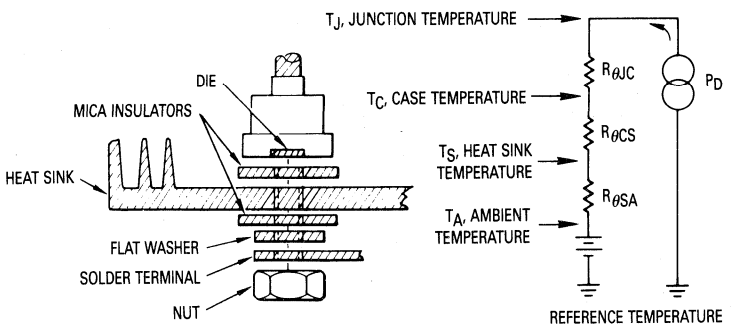
The equivalent electrical circuit may be analyzed by using Kirchoff's Law and the following equation results:

$$T_J = P_D(R_{\theta JC} + R_{\theta CS} + R_{\theta SA}) + T_A \qquad (3)$$

where T_J = junction temperature,
P_D = power dissipation,
$R_{\theta JC}$ = semiconductor thermal resistance (junction to case),
$R_{\theta CS}$ = interface thermal resistance (case to heat sink),
$R_{\theta SA}$ = heat sink thermal resistance (heat sink to ambient),
T_A = ambient temperature.

The thermal resistance junction to ambient is the sum of the individual components. Each component must be minimized if the lowest junction temperature is to result. The value for the interface thermal resistance, $R_{\theta CS}$, is affected by the mounting procedure and may be significant compared to the other thermal-resistance terms.

The thermal resistance of the heat sink is not constant; it decreases as ambient temperature increases and is affected by orientation of the sink. The thermal resistance of the semiconductor is also variable; it is a function of biasing and temperature. In some applications such as in RF power amplifiers and short-pulse applications, the concept may be invalid because of localized heating in the semiconductor chip.



Basic Thermal Resistance Model Showing Thermal to Electrical Analogy for a Semiconductor

APPENDIX VIII

DERIVATION OF RFI DESIGN EQUATIONS

The relationship of flux to voltage and time is $E = N\frac{d\phi}{dt}$ or $E = NA_c\frac{dB}{dt}$ since $\phi = BA_c$ and A_c is a constant. Rearranging this equation and integrating we get:

$$\int E\,dt = NA_c\,(B_2 - B_1) = NA_c\,\Delta B \qquad (1)$$

which says that the volt-second integral required determines the size of the core. In an L-R circuit such as we have with a thyristor control circuit, the volt-second characteristic is the area under an exponential decay. A conservative estimate of the area under the curve may be obtained by considering a triangle whose height is the peak line voltage and the base is the allowable switching time. The area is then 1/2 bh or $\frac{E_p t_r}{2}$. Substituting in Equation (1):

$$\frac{E_p t_r}{2} = N\,A_c\,\Delta B \qquad (2)$$

where:

E_p is the peak line voltage

t_r is the allowable current rise time

N is the number of turns on the coil

A_c is the usable core area in cm^2

ΔB is the maximum usable flux density of the core material in W/m^2

Rewriting Equation (2) to change ΔB from W/m^2 to gauss, substituting $\sqrt{2}\,E_{rms}$ for E_p and solving for N, we get:

$$N = \frac{\sqrt{2}\,E_{rms}\,t_r}{2\,A_c\,\Delta B} \times 10^8 = \frac{0.707\,E_{rms}\,t_r \times 10^8}{B_{MAX}\,A_c}$$

A_c in this equation is in cm^2. To change to in^2, multiply A_c by 6.452. Then:

$$N = \frac{10.93\,E_{rms}\,t_r \times 10^6}{B_{MAX} \times A_c} \qquad (3)$$

where:

N is total turns

E_{rms} is line voltage

t_r is allowable current rise time in seconds

B_{MAX} is maximum usable flux density of core material

A_c is usable core area in square inches

Window area necessary is:

$$A_w = N\,A_{wire} \times 3 \qquad (4)$$

The factor of 3 is an approximation which allows for insulation and winding space not occupied by wire. Substituting equation (3) in (4):

$$A_w = \frac{10.93\,E_{rms}\,t_r \times 10^6}{B_{MAX}\,A_c}\,A_{wire} \times 3$$

(The factor 10.93 may be rounded to 11 since two significant digits are all that are necessary.)

The factor A_cA_w can easily be found for most cores and is an easy method for selecting a core.

$$A_cA_w = \frac{33\,E_{rms}\,t_r A_{wire} \times 10^6}{B_{MAX}}$$

In this equation, the core area is in in^2. To work with circular mils, multiply by 0.78×10^{-6} so that:

$$A_cA_w = \frac{26\,E_{rms}\,t_r A_{wire}}{B_{MAX}}$$

where A_{wire} is the wire area in circular mils.

Inductance of an iron core inductor is

$$L = \frac{3.19\,N^2\,A_c\,10^{-8}}{l_g + \frac{l_c}{\mu}}$$

Rearranging terms,

$$l_g = \frac{3.19\,N^2\,A_c\,10^{-8}}{L} - \frac{l_c}{\mu}$$

APPENDIX IX

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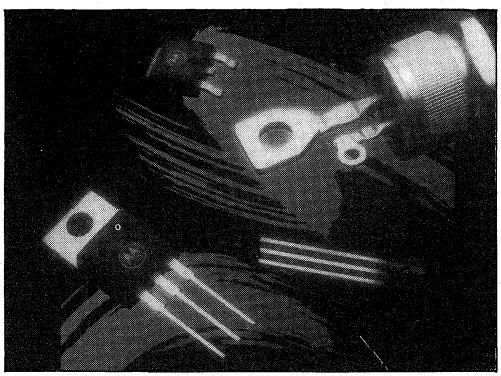

In Brief . . .

Motorola's broad line of Thyristors include. . . .

- A full line of TRIACs and SCRs covering a forward current range from 0.5 to 55 amperes and blocking voltages from 25 to 800 volts.
- Two basic package categories — plastic for lowest cost — including fully insulated plastic Case 221C-02 (TO-220 Full Pak); metal for hermetically-sealed requirements in high-reliability projects.
- An extensive line of trigger devices that includes PUTs, SBS — even optically-coupled TRIAC drivers from Motorola's optoelectronic product line.

Selector Guide

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MCR100 Series	2-3
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MCR225FP Series	2-8
MCR229 Series	2-10
MCR230 Series	2-8
MCR230( )3 Series	2-9
MCR231 (3) Series	2-9
MCR264 Series	2-10
MCR310 Series	2-6
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MCR703A Series	2-5
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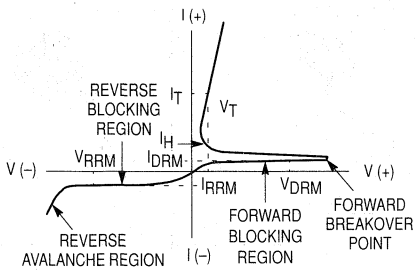
SCRs

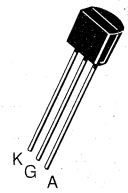
Silicon Controlled Rectifiers



Thyristors — SCR's

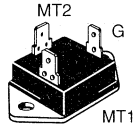
Metal/Plastic Packages

0.5 to 55 Amperes RMS
25–1000 Volts

On-State (RMS) Current		
0.8 AMP	1.5 AMPS	
T_C = 58°C	T_C = 50°C	
		
Sensitive Gate		V_{DRM} V_{RRM} (Volts)
Case 29-04 TO-226AA (TO-92) Style 10		
MCR102 2N5060 BRX44/BRY55-30*		25
MCR103 2N5061 BRX45/BRY55-609*	MCR22-2	50
MCR100-3 2N5062 BRX46/BRY55-100*	MCR22-3	100
MCR100-4 2N5064 BRX47/BRY55-200*	MCR22-4	200
MCR100-6 BRX49/BRY55-400*	MCR22-6	400
BRY55-500*		500
MCR100-8 BRY55-600*	MCR22-8	600
Maximum Electrical Characteristics		
10	15 150(1)	I_{TSM} (Amps) 60 Hz
0.2		I_{GT} (mA)
0.8		V_{GT} (V)
–65 to +110	–40 to +125	T_J Operating Range (°C)

ATTENTION:
PACKAGE INNOVATION

Isolated TRIAC Mold Type

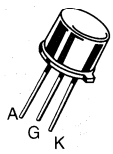
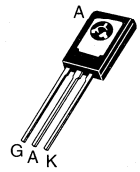
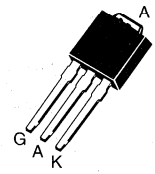


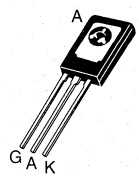
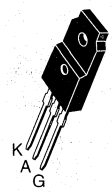
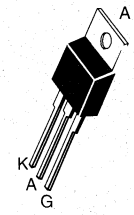
CASE 383-01

Features a TO-3 isolated mounting with a high isolation voltage of 2.5 kVrms min. This package also offers quick disconnect lead attachments, is plastic encapsulated to provide economical cost and is UL recognized. See pages 2-22 and 2-23 for the MAC625 and MAC635 series.

(1) Exponential decay 2 μs wide at 5 time constants, f = 12 Hz.
* European Part Numbers. Package is Case 29 with Leadform 18. Case style is 3.

SCR's (continued)

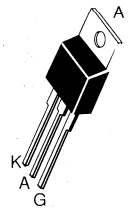
V_{DRM} V_{RRM} (Volts)	On-State (RMS) Current				
	1.6 AMPS		4 AMPS		
	T_C = 80°C	T_C = 65°C	T_C = 93°C	T_C = 30°C	
					
		Sensitive Gate			
	Case 79-04 TO-205AD (TO-39) Style 3		Case 77-07 TO-126 Style 2		Case 369 Style 5
50	2N1595		MCR106-2 2N6237	C106F	
100	2N1596		MCR106-3 2N6238	C106A	MCR703A1
200	2N1597	MCR1906-4	MCR106-4 2N6239	C106B	MCR704A1
400	2N1599	MCR1906-6	MCR106-6 2N6240	C106D	MCR706A1
600		MCR1906-8	MCR106-8 2N6241	C106M	MCR708A1
800					
Maximum Electrical Characteristics					
I_{TSM} (Amps) 60 Hz	15		25	20	25
I_{GT} (mA)	10	1	0.2		0.075
V_{GT} (V)	3	1		0.8	1
T_J Operating Range (°C)	–65 to +125	–65 to +110	–40 to +110		

On-State (RMS) Current					
4 AMPS	6 AMPS	8 AMPS			
T_C = 30°C		T_C = 70°C	T_C = 75°C		
					
Surface Mount	Sensitive Gate	Isolated			
Case 369A Style 5	Case 77-07 TO-126 Style 2	Case 221C-02 Style 2	Case 221A-04 TO-220AB Style 3		V_{DRM} V_{RRM} (Volts)
	MCR506-2	MCR218-2FP	MCR218-2	C122F1	50
MCR703A(1)	MCR506-3		MCR218-3	C122A1	100
MCR704A(1)	MCR506-4	MCR218-4FP	MCR218-4	C122B1	200
MCR706A(1)	MCR506-6	MCR218-6FP	MCR218-6	C122D1	400
MCR708A(1)	MCR506-8	MCR218-8FP	MCR218-8	C122M1	600
		MCR218-10FP	MCR218-10	C122N1	800
Maximum Electrical Characteristics					
25	40	80		90	I_{TSM} (Amps) 60 Hz
0.075	0.2	25		25	I_{GT} (mA)
1		1.5			V_{GT} (V)
–40 to +110		–40 to +125		–40 to +100	T_J Operating Range (°C)

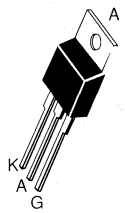
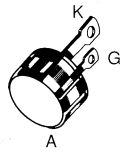
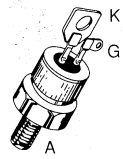
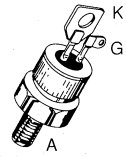
(1) For tape and reel, add suffix "RL."

2

SCR's (continued)

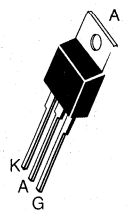
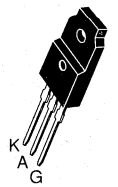
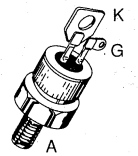
V_{DRM} V_{RRM} (Volts)	On-State (RMS) Current				
	8 AMPS	10 AMPS		12 AMPS	
	T_C = 83°C	T_C = 75°C		T_C = 85°C	T_C = 90°C
	Sensitive Gate				
	Case 221A-04 TO-220AB Style 3				
50	MCR72-2	MCR310-2	S2800F	MCR68-2	2N6394
100	MCR72-3	MCR310-3	S2800A	MCR68-3	2N6395
200	MCR72-4	MCR310-4	S2800B		2N6396
400	MCR72-6	MCR310-6	S2800D	MCR68-6	2N6397
600	MCR72-8	MCR310-8	S2800M		2N6398
800	MCR72-10	MCR310-10	S2800N		2N6399
Maximum Electrical Characteristics					
I_{TSM} (Amps) 60 Hz	100			300(1)	100
I_{GT} (mA)	0.2		15	30	
V_{GT} (V)	1.5				
T_J Operating Range (°C)	–40 to +110		–40 to +100	–40 to +125	

(1) Peak capacitor discharge current for t_w = 1 ms. t_w is defined as five time constants of an exponentially decaying current pulse (crowbar applications).

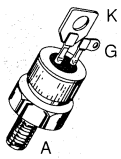
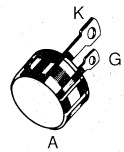
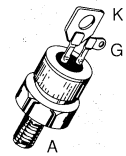
On-State (RMS) Current					
16 AMPS	20 AMPS				
T_C = 90°C	T_C = 65°C			T_C = 67°C	
			Isolated		
Case 221A-04 TO-220AB Style 3	Case 174-04 Style 1	Case 263-04 Style 1	Case 311-02 Style 1	Case 263-04 Style 1	V_{DRM} V_{RRM} (Volts)
2N6400		2N5168		MCR3918-2A	**50**
2N6401	MCR6200A	S6210A	2N6167 S6220A	MCR3918-3A	**100**
2N6402	MCR6200B	S6210B	2N6168 S6220B	MCR3918-4A	**200**
2N6403	MCR6200D	S6210D	2N6169 S6220D	MCR3918-6A	**400**
2N6404	MCR6200M	S6210M	2N6170 S6220M	MCR3918-8A	**600**
2N6405				MCR3918-10A	**800**
Maximum Electrical Characteristics					
160	MCR6200/S6210/S6220/2N 200/200/200/240			240	I_{TSM} (Amps) 60 Hz
30	MCR6200/S6210/S6220/2N 15/15/15/40			40	I_{GT} (mA)
1.5	MCR6200/S6210/S6220/2N . . . . 2/2/2/1.5 or 1.6			1.5	V_{GT} (V)
–40 to +125	–40 to +100		–40 to +100	–40 to +125	T_J Operating Range (°C)

2

SCR's (continued)

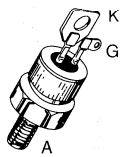
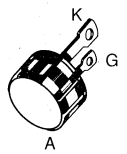
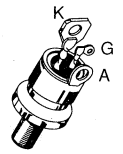
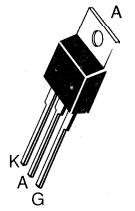
V_{DRM} V_{RRM} (Volts)	On-State (RMS) Current				
	25 AMPS				
	T_C = 85°C			T_C = 65°C	T_C = 60°C
					
			Isolated		
	Case 221A-04 TO-220AB Style 3		Case 221C-02 Style 3	Case 263-04 Style 1	
50	2N6504	MCR69-2	MCR225-2FP	2N682	MCR230F
100	2N6505	MCR69-3		2N683	MCR230A
200	2N6506		MCR225-4FP	2N685	MCR230B
400	2N6507	MCR69-6	MCR225-6FP	2N688	MCR230D
600	2N6508		MCR225-8FP	2N690	MCR230M
800	2N6509		MCR225-10FP	2N692	
Maximum Electrical Characteristics					
I_{TSM} (Amps) 60 Hz	300	750(1)	300	150	250
I_{GT} (mA)	40	30	40		25
V_{GT} (V)	1.5			2	
T_J Operating Range (°C)	–40 to +125			–65 to +125	–40 to +125

(1) Peak capacitor discharge current for t_w = 1 ms. t_w is defined as five time constants of an exponentially decaying current pulse (crowbar applications).

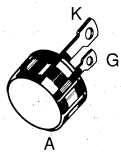
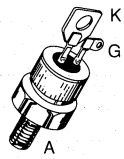
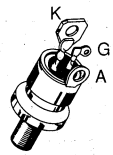
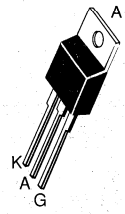
On-State (RMS) Current					
25 AMPS				35 AMPS	
T_C = 60°C	T_C = 65°C		T_C = 70°C	T_C = 65°C	
					
	Isolated				
Case 263-04 Style 1	Case 311-02 Style 1		Case 174-04 Style 1	Case 263-04 Style 1	V_{DRM} V_{RRM} (Volts)
MCR231F	MCR230F3	MCR231F3			50
MCR231A	MCR230A3	MCR231A3	2N3870		100
MCR231B	MCR230B3	MCR231B3	2N3871	MCR3935-4A	200
MCR231D	MCR230D3	MCR231D3	2N3872	MCR3935-6A	400
MCR231M	MCR230M3	MCR231M3	2N3873	MCR3935-8A	600
			MCR3835-10	MCR3935-10A	800
Maximum Electrical Characteristics					
250			350		I_{TSM} (Amps) 60 Hz
9	25	9	40		I_{GT} (mA)
2	1.5		MCR3835/MCR3935/2N 1.5/1.5/1.6		V_{GT} (V)
–40 to +125					T_J Operating Range (°C)

2

SCR's (continued)

V_{DRM} V_{RRM} (Volts)	On-State (RMS) Current				
	35 AMPS				40 AMPS
	T_C = 65°C				T_C = 80°C
					
			Isolated		
	Case 263-04	Case 174-04 Style 1	Case 311-02 Style 1		Case 221A-04 TO-220AB Style 3
50	MCR70-2A				
100	MCR70-3A	MCR229A	2N6171	C228A3	
200		MCR229B	2N6172	C228B3	MCR264-4
400	MCR70-6A	MCR229D	2N6173	C228D3	MCR264-6
600		MCR229M	2N6174	C228M3	MCR264-8
800					MCR264-10
Maximum Electrical Characteristics					
I_{TSM} (Amps) 60 Hz	350 850(1)	300	350	300	400
I_{GT} (mA)	30	40			50
V_{GT} (V)	1.5	2.5	1.6	2.5	1.5
T_J Operating Range (°C)	–40 to +125				

(1) Peak capacitor discharge current for t_w = 1 ms. t_w is defined as five time constants of an exponentially decaying current pulse (crowbar applications).

On-State (RMS) Current					
55 AMPS					
T_C = 75°C		T_C = 70°C		T_C = 85°C	
					
		Isolated			
Case 174-04 Style 1	Case 263-04 Style 1	Case 311-02 Style 1	Case 221A-04 TO-220AB Style 3	Case 263-04 Style 1	V_{DRM} V_{RRM} (Volts)
MCR63-2A	MCR64-2		MCR265-2	MCR71-2	**50**
MCR63-3A	MCR64-3	MCR65-3		MCR71-3	**100**
MCR63-4A	MCR64-4	MCR65-4	MCR265-4		**200**
MCR63-6A	MCR64-6	MCR65-6	MCR265-6	MCR71-6	**400**
MCR63-8A	MCR64-8	MCR65-8	MCR265-8		**600**
MCR63-10A	MCR64-10	MCR65-10	MCR265-10		**800**
Maximum Electrical Characteristics					
550				550 1700(1)	**I_{TSM} (Amps) 60 Hz**
40			50	30	**I_{GT} (mA)**
3			1.5		**V_{GT} (V)**
–40 to +125					**T_J Operating Range (°C)**

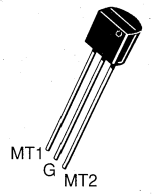
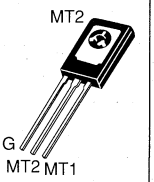
(1) Peak capacitor discharge current for t_w = 1 ms. t_w is defined as five time constants of an exponentially decaying current pulse (crowbar applications).

2

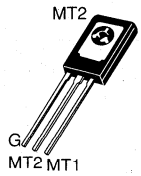
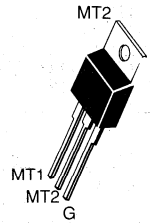
TRIACs

Metal/Plastic Packages

0.6 to 40 Amperes
25 to 800 Volts

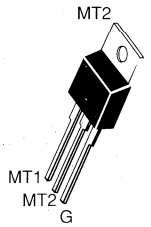
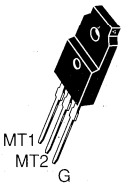
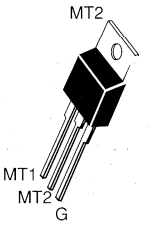
V_{DRM} (Volts)	On-State (RMS) Current				
	0.6 AMP		0.8 AMP		2.5 AMPS
	T_C = 50°C		T_C = 60°C		T_C = 70°C
					
	Sensitive Gate				
	Case 29-04 TO-226AA (TO-92) Style 12				Case 77-07 Style 5
200	MAC97-4	MAC97A4	LMAC94A2	LMAC94B2	T2322B
400	MAC97-6	MAC97A6	LMAC94A3	LMAC94B3	T2322D
600	MAC97-8	MAC97A8	LMAC94A4	LMAC94B4	T2322M
800					
Maximum Electrical Characteristics					
I_{TSM} (Amps)	8				25
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 10 10 10 10	 5 5 5 7	 10 10 10 10	 25 25 25 25	 10 10 10 10
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2 2 2 2.5				 2.2 2.2 2.2 2.2
T_J Operating Range (°C)	–40 to +110				

NOTE: Industry Standards, with a variety of Custom Specifications and Leadforms available.

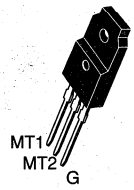
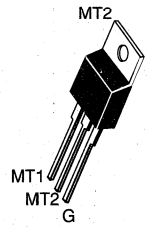
On-State (RMS) Current					
2.5 AMPS	**4 AMPS**			**6 AMPS**	
T_C = 70°C	**T_C = 85°C**			**T_C = 80°C**	
					
Sensitive Gate					
Case 77-07 Style 5				**Case 221A-04 TO-220AB Style 4**	**V_{DRM} (Volts)**
T2323B	2N6071	2N6071A	2N6071B	T2500B	**200**
T2323D	2N6073	2N6073A	2N6073B	T2500D	**400**
T2323M	2N6075	2N6075A	2N6075B	T2500M	**600**
				T2500N	**800**
Maximum Electrical Characteristics					
25	30			60	**I_{TSM} (Amps)**
25 40 25 40	30 — 30 —	5 5 5 10	3 3 3 5	25 60 25 60	**I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)**
2.2 2.2 2.2 2.2	@ –40°C 2.5 — 2.5 —	@ –40°C 2.5 2.5 2.5 2.5		2.5 2.5 2.5 2.5	**V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)**
–40 to +110		–40 to +100			**T_J Operating Range (°C)**

2

TRIACs (continued)

	On-State (RMS) Current				
	6 AMPS			8 AMPS	
	T_C = 80°C				
		 Isolated			
V_{DRM} (Volts)	Case 221A-04 TO-220AB Style 4	Case 221C-02 Style 3	Case 221A-06 TO-220AB Style 4		
200	T2801B	T2500BFP	SC141B	SC143B	MAC218-4 MAC218A4
400	T2801D	T2500DFP	SC141D	SC143D	MAC218-6 MAC218A6
600	T2801M	T2500MFP	SC141M	SC143M	MAC218-8 MAC218A8
800		T2500NFP	SC141N		MAC218-10 MAC218A10
Maximum Electrical Characteristics					
I_{TSM} (Amps)	80	100	80	100	
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 80 — 80	 25 60 25 60	 50 50 50 —		 50 50 50 75*
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2.5 2.5 2.5 2.5		 2.5 2.5 2.5 —		 2 2 2 2.5*
T_J Operating Range (°C)	–40 to +100		–40 to +110		–40 to +125

* Applied to A-version only.
Non A-version is unspecified.

On-State (RMS) Current					
8 AMPS					
T_C = 80°C					
					
Isolated	**High Noise Immunity**				
Case 221C-02 Style 3	**Case 221A-04 TO-220AB Style 4**				**V_{DRM} (Volts)**
MAC218-4FP MAC218A4FP	MAC219-4	2N6342 2N6346	T2800B	T2802B	**200**
MAC218-6FP MAC218A6FP	MAC219-6	2N6343 2N6347	T2800D	T2802D	**400**
MAC218-8FP MAC218A8FP	MAC219-8	2N6344 2N6348	T2800M	T2802M	**600**
MAC218-10FP MAC218A10FP	MAC219-10	2N6345 2N6349			**800**
Maximum Electrical Characteristics					
100					**I_{TSM} (Amps)**
50 50 50 75*	100 100 100 —	50 75# 50 75#	25 60 25 60	50 — 50 —	**I_{GT} @ 25°C (mA)** **MT2(+)G(+)** **MT2(+)G(–)** **MT2(–)G(–)** **MT2(–)G(+)**
2 2 2 2.5*	2 2 2 —	2 2.5# 2.5 2.5#	2.5 2.5 2.5 2.5	2.5 — 2.5 —	**V_{GT} @ 25°C (V)** **MT2(+)G(+)** **MT2(+)G(–)** **MT2(–)G(–)** **MT2(–)G(+)**
–40 to +125			–40 to +100		**T_J Operating Range (°C)**

* Applied to A-version only.
Non A-version is unspecified.
# Denotes 2N6346-49 Series only.

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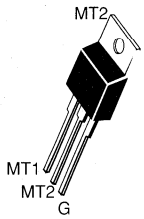
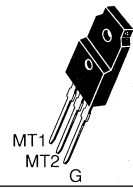
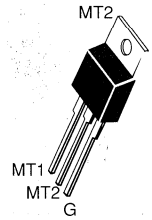
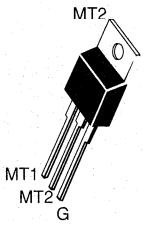
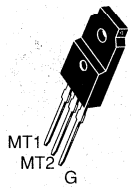
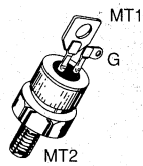
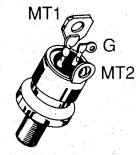
TRIACs (continued)

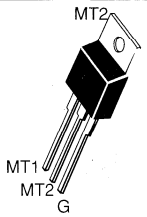
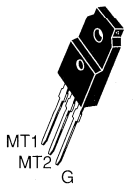
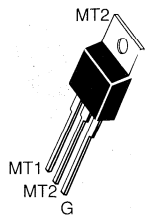
V_{DRM} (Volts)	On-State (RMS) Current				
	8 AMPS				10 AMPS
	T_C = 80°C				T_C = 70°C
					
			Isolated		
	Sensitive Gate				
	Case 221A-04 TO-220AB Style 4		Case 221C-02 Style 3		Case 221A-04 TO-220AB Style 4
200	MAC228-4 MAC228A4	MAC229-4 MAC229A4	MAC228-4FP MAC228A4FP	MAC229-4FP MAC229A4FP	MAC210-4 MAC210A4
400	MAC228-6 MAC228A6	MAC229-6 MAC229A6	MAC228-6FP MAC228A6FP	MAC229-6FP MAC229A6FP	MAC210-6 MAC210A6
600	MAC228-8 MAC228A8	MAC229-8 MAC228A8	MAC228-8FP MAC228A8FP	MAC229-8FP MAC2298FP	MAC210-8 MAC210A8
800	MAC228-10 MAC228A10	MAC229-10 MAC229A10	MAC228-10FP MAC228A10FP	MAC229-10FP MAC229A10FP	MAC210-10 MAC210A10
Maximum Electrical Characteristics					
I_{TSM} (Amps)	80				100
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 5 5 5 10*	 10 10 10 20*	 5 5 5 10*	 10 10 10 20*	 50 50 50 75*
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2 2 2 2.5*				
T_J Operating Range (°C)	–40 to +110				–40 to +125

* Applied to A-version only.
Non A-version is unspecified.

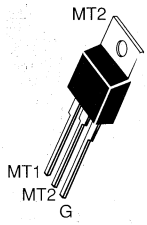
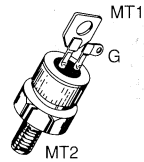
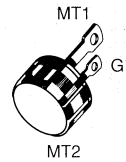
On-State (RMS) Current					
10 AMPS					
T_C = 80°C	T_C = 70°C	T_C = 85°C	T_C = 78°C	T_C = 85°C	
					
	Isolated			Isolated	
Case 221A-04 TO-220AB Style 4	Case 221C-02 Style 3	Case 263-04 Style 2		Case 311-02 Style 2	V_{DRM} (Volts)
SC146B	MAC210-4FP MAC210A4FP	MAC5569	MAC245B	MAC4121B	**200**
SC146D	MAC210-6FP MAC210A6FP	MAC5570	MAC245D	MAC4121D	**400**
SC146M	MAC210-8FP MAC210A8FP	MAC4111M	MAC245M	MAC4121M	**600**
SC146N	MAC210-10FP MAC210A10FP		MAC245N	MAC4121N	**800**
Maximum Electrical Characteristics					
120	100				I_{TSM} (Amps)
50 50 50 —	50 50 50 75*	25 40 25 40	50 50 50 —	25 40 25 40	I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
2.5 2.5 2.5 —	2 2 2 2.5*	2.5 2.5 2.5 2.5	2.5 2.5 2.5 —	2.5 2.5 2.5 2.5	V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
–40 to +125		–65 to +125	–40 to +125	–65 to +125	T_J Operating Range (°C)

* Applied to A-version only.
Non A-version is unspecified.

2

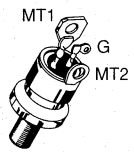
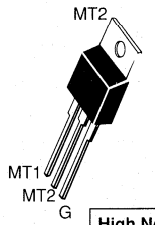
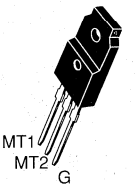
	On-State (RMS) Current				
	12 AMPS				
	T_C = 75°C	T_C = 85°C			T_C = 80°C
					
	Sensitive Gate	Isolated		High Noise Immunity	
V_{DRM} (Volts)	Case 221A-04 TO-220AB Style 4	Case 221C-02 Style 3	Case 221A-04 TO-220AB Style 4		
200	MAC310-4 MAC310A4	MAC212-4FP MAC212A4FP	MAC212-4 MAC212A4	MAC213-4	SC149B
400	MAC310-6 MAC310A6	MAC212-6FP MAC212A6FP	MAC212-6 MAC212A6	MAC213-6	SC149D
600	MAC310-8 MAC318A8	MAC212-8FP MAC212A8FP	MAC212-8 MAC212A8	MAC213-8	SC149M
800	MAC310-10 MAC310A10	MAC212-10FP MAC212A10FP	MAC212-10 MAC212A10	MAC213-10	
Maximum Electrical Characteristics					
I_{TSM} (Amps)	100				120
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 5 5 5 10*	 50 50 50 75*	 50 50 50 75*	 100 100 100 —	 50 50 50 —
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2 2 2 2.5*	 2 2 2 2.5*		 2 2 2 —	 2.5 2.5 2.5 —
T_J Operating Range (°C)	–40 to +125				

* Applied to A-version only.
Non A-version is unspecified.

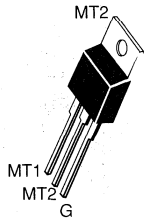
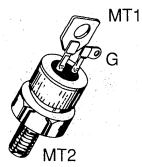
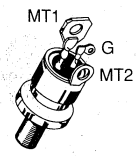
On-State (RMS) Current					
12 AMPS		15 Amps			
T_C = 80°C		T_C = 85°C	T_C = 80°C		
					
Case 221A-04 TO-220AB Style 4		Case 263-04 Style 2	Case 174-04 Style 3		V_{DRM} (Volts)
2N6342A	2N6346A	MAC250B	2N5571	SC251B	200
2N6343A	2N6347A	MAC250D	2N5572	SC251D	400
2N6344A	2N6348A	MAC250M	T4100M	SC251M	600
2N6345A	2N6349	MAC250N			800
Maximum Electrical Characteristics					
120		100			I_{TSM} (Amps)
50 — 50 —	50 75 50 75	50 50 50 —	50 80 50 80		I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
2 — 2 —	2 2.5 2 2.5	2.5 2.5 2.5 —	2.5 2.5 2.5 2.5		V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
–40 to +125		–40 to +115	–40 to +100		T_J Operating Range (°C)

2

TRIACs (continued)

	On-State (RMS) Current				
	15 AMPS				20 AMPS
	T_C = 75°C	T_C = 90°C			T_C = 75°C
					
	Isolated		High Noise Immunity	Isolated UL Listed (1)	
V_{DRM} (Volts)	Case 311-02 Style 2	Case 221A-04 TO-220AB Style 4		Case 221C-02 Style 3	
200	MAC4120B	MAC15-4 MAC15A4	MAC16-4	MAC15-4FP MAC15A4FP	MAC320-4FP MAC320A4FP
400	MAC4120D	MAC15-6 MAC15A6	MAC16-6	MAC15-6FP MAC15A6FP	MAC320-6FP MAC320A6FP
600	MAC4120M	MAC15-8 MAC15A8	MAC16-8	MAC15-8FP MAC15A8FP	MAC320-8FP MAC320A8FP
800	MAC4120N	MAC15-10 MAC15A10	MAC16-10	MAC15-10FP MAC15A10FP	MAC320-10FP MAC320A10FP
Maximum Electrical Characteristics					
I_{TSM} (Amps)	100	150			
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 50 80 50 80	 50 50 50 75*	 100 100 100 —	 50 50 50 75*	
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2.5 2.5 2.5 2.5	 2 2 2 2.5*	 2 2 2 —	 2 2 2 2.5*	
T_J Operating Range (°C)	–40 to +100	–40 to +125			

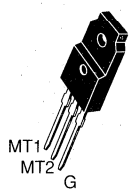
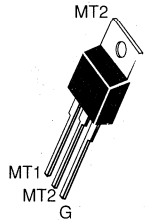
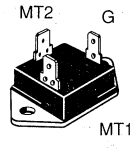
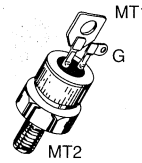
(1) UL #E69369
* Applied to A-version only.
Non A-version is unspecified.

On-State (RMS) Current					
20 AMPS	25 AMPS				
T_C = 75°C		T_C = 80°C	T_C = 75°C	T_C = 80°C	
					
	High Noise Immunity		Isolated		
Case 221A-04 TO-220AB Style 4		Case 263-04 Style 2	Case 311-02 Style 2	Case 174-04 Style 3	V_{DRM} (Volts)
MAC320-4 MAC320A4	MAC321-4	SC260B	SC260B3	MAC261B	**200**
MAC320-6 MAC320A6	MAC321-6	SC260D	SC260D3	MAC261D	**400**
MAC320-8 MAC320A8	MAC321-8	SC260M	SC260M3	MAC261M	**600**
MAC320-10 MAC320A10	MAC321-10				**800**
Maximum Electrical Characteristics					
150		250			I_{TSM} (Amps)
50 50 50 75*	100 100 100 —	50 50 50 —			I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
2 2 2 2.5*	2 2 2 —	2.5 2.5 2.5 —			V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
–40 to +125		–40 to +115			T_J Operating Range (°C)

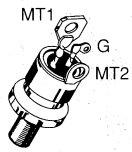
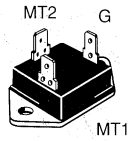
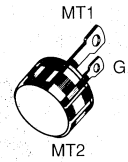
* Applied to A-version only.
Non A-version is unspecified.

2

TRIACs (continued)

	On-State (RMS) Current				
	25 AMPS			30 AMPS	
	T_C = 80°C		T_C = 85°C		T_C = 60°C
					
	Isolated		Isolated UL Listed (1)		
V_{DRM} (Volts)	Case 221C-02 Style 3	Case 221A-04 TO-220AB Style 4	Case 383-01 Style 1	Case 263-04 Style 2	
200	MAC223-4FP MAC223A4FP	MAC223-4 MAC223A4	MAC625-4	2N6160	T6411B
400	MAC223-6FP MAC223A6FP	MAC223-6 MAC223A6	MAC625-6	2N6161	T6411D
600	MAC223-8FP MAC223A8FP	MAC223-8 MAC223A8	MAC625-8	2N6162	T6411M
800	MAC223-10FP MAC223A10FP	MAC223-10 MAC223A10			T6411N
Maximum Electrical Characteristics					
I_{TSM} (Amps)	250				300
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 50 50 50 75*		 50 50 50 —	 60 70 70 100	 50 80 50 80
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2 2 2 2.5*		 3 3 3 —	 2 2.1 2.1 2.5	 2.5 2.5 2.5 2.5
T_J Operating Range (°C)	–40 to +125			–65 to +125	–65 to +100

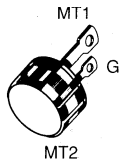
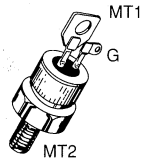
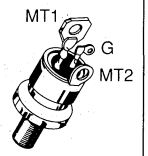
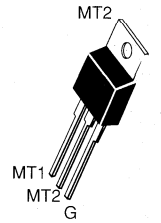
(1) UL #E69369
* Applied to A-version only.
Non A-version is unspecified.

On-State (RMS) Current					
30 AMPS		35 AMPS	40 AMPS		
T_C = 85°C	T_C = 55°C	T_C = 58°C	T_C = 65°C	T_C = 70°C	
					
Isolated		Isolated UL Listed (1)			
Case 311-02 Style 2		Case 383-01 Style 1	Case 174-04 Style 3		V_{DRM} (Volts)
2N6163	T6421B	MAC635-4	MAC6401B	MAC5441	**200**
2N6164	T6421D	MAC635-6	MAC6401D	MAC5442	**400**
2N6165	T6421M	MAC635-8	MAC6401M	MAC5443	**600**
	T6421N		MAC6401N		**800**
Maximum Electrical Characteristics					
250	300	330	300		I_{TSM} (Amps)
60 70 70 100	50 80 50 80	50 50 50 —	50 80 50 80	70 70 70 100	I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
2 2.1 2.1 2.5	2.5 2.5 2.5 2.5	3 3 3 —	2.5 2.5 2.5 2.5	2 2 2 2.5	V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)
–40 to +100	–65 to +100	–40 to +125	–65 to +125		T_J Operating Range (°C)

(1) UL #E69369

2

TRIACs (continued)

	On-State (RMS) Current			
	40 AMPS			
	T_C = 70°C	T_C = 65°C	T_C = 60°C	T_C = 75°C
				
			Isolated	
V_{DRM} (Volts)	Case 174-04 Style 3	Case 263-04 Style 2	Case 311-02 Style 2	Case 221A-04 TO-220AB Style 4
200	MAC6400B	2N5444	T6420B	MAC224-4 MAC224A4
400	MAC6400D	2N5445	T6420D	MAC224-6 MAC224A6
600	MAC6400M	2N5446	T6420M	MAC224-8 MAC224A8
800	MAC6400N		T6420N	MAC224-10 MAC224A10
Maximum Electrical Characteristics				
I_{TSM} (Amps)	300			350
I_{GT} @ 25°C (mA) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 50 80 50 80	 70 70 70 100	 50 80 50 80	 50 50 50 75*
V_{GT} @ 25°C (V) MT2(+)G(+) MT2(+)G(–) MT2(–)G(–) MT2(–)G(+)	 2.5 2.5 2.5 2.5	 2 2 2 2.5	 2.5 2.5 2.5 2.5	 2 2 2 2.5*
T_J Operating Range (°C)	–65 to +125	–65 to +110		–40 to +125

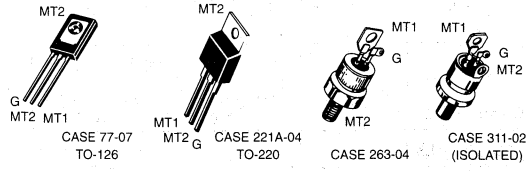
* Applied to A-version only.
Non A-version is unspecified.

TRIACs

Optically Isolated

Triac Driver/Triac Combinations

This series of Triac Drivers consists of infrared LEDs optically coupled to photodetectors with Triac output. 7500 V isolation between input and output allows safe, economical triggering of higher power triacs from logic sources with output as low as 3 volts, 10 mA. Associated voltage-compatible triacs provide matched pairs for a variety of voltage/current requirements.



Triac Drivers — all in Case 730A

Peak Blocking Voltage Volts	LED Trigger Current I_{FT} mA, Max	Device
250	30	MOC3009
	15	3010
	10	3011
400	30	3020
	15	3021
For Zero Crossover Firing		
250	30	MOC3030*
	15	3031*
400	30	3040
	15	3031
600	30	3060
	15	3061

*Underwriters' Laboratories Recognition, File No. E54915.

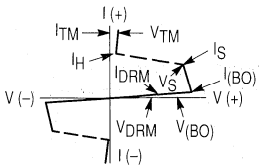
Triacs

Output Current I_{RMS} A, Max	Peak Blocking Voltage Volts			Case
	250	400	600	
4	MAC3010-4	MAC3020-4	—	77
8	-8	-8	—	221A
15	-15	-15	—	221A
25	-25	-25	—	221A
40	-40	-40	—	263
40	-40I	-40I	—	311
For Zero Crossover Firing				
4	MAC3030-4	MAC3040-4	MAC3060-4	77
8	-8	-8	-8	221A
15	-15	-15	-15	221A
25	-25	-25	-25	221A
40	-40	-40		
40	-40I	-40I		

2

Thyristor Triggers

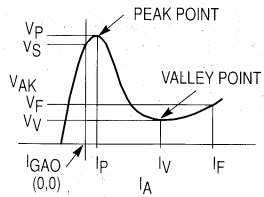
SIDACs



High voltage trigger devices similar in operation to a Triac. Upon reaching the breakover voltage in voltage either direction, the device switches to a low-voltage on-state.

Device Type	I_{TSM} Amps	V_{BO} Volts Min	V_{BO} Volts Max
Case 267-03/1			
MK1V115	20	104	115
MK1V125	20	110	125
MK1V135	20	120	135
MK1V240	20	220	250
MK1V260	20	240	270
MK1V270	20	250	280
Case 59-04/1			
MKP9V120	4	110	125
MKP9V130	4	120	135
MKP9V240	4	220	250
MKP9V260	4	240	270
MKP9V270	4	250	280

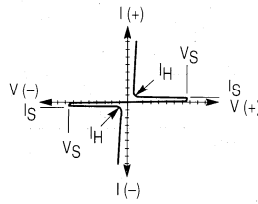
Programmable Unijunction Transistors — PUT



Similar to UJTs, except that I_V, I_P and intrinsic standoff voltage are programmable (adjustable) by means of external voltage divider. This stabilizes circuit performance for variations in device parameters. General operating frequency range is from 0.01 Hz to 10 kHz, making them suitable for long-duration timer circuits.

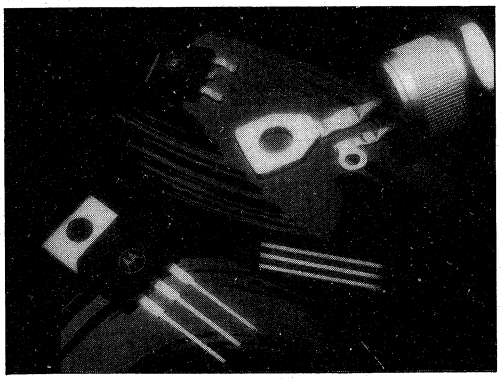
Device Type	I_P R_G = 10 kΩ (μA Max)	I_P R_G = 1 MΩ (μA Max)	I_{GAO} @ 40 V nA Max	I_V R_G = 10 kΩ μA Min	I_V R_G = 1 MΩ μA Max
Plastic TO-92 (Case 29-04/16)					
2N6027	5	2	10	70	50
2N6028	1	0.15	10	25	25

Silicon Bidirectional Switch (SBS)



This versatile trigger device exhibits highly symmetrical bidirectional switching characteristics which can be modified by means of a gate lead. Requires a gate trigger current of only 250 μAdc for triggering.

Device Type	V_S Volts Min	V_S Volts Max	I_S μA Max	I_H mA Max
Plastic TO-92/TO-226AA (Case 29-04/12)				
MBS4991	6	10	500	1.5
MBS4992	7.5	9	120	0.5
MBS4993	7.5	9	250	0.75



Data Sheets

3

Silicon Controlled Rectifiers

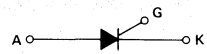
Reverse Blocking Triode Thyristors

2N682 thru 2N692

SCRs
25 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire for Greater Parameter Uniformity and Stability
- Blocking Voltage to 800 Volts



CASE 263-04
STYLE 1

3

MAXIMUM RATINGS (T_J = 125°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Repetitive Off-State Blocking Voltage, Note 1	V_{RRM} or V_{DRM}		Volts
2N682		50	
2N683		100	
2N685		200	
2N688		400	
2N690		600	
2N692		800	
*Peak Non-Repetitive Reverse Voltage	V_{RSM}		Volts
2N682		75	
2N683		150	
2N685		300	
2N688		500	
2N690		720	
2N692		960	
*RMS On-State Current (All Conduction Angles)	$I_{T(RMS)}$	25	Amps
*Average On-State Current (T_C = 65°C)	$I_{T(AV)}$	16	Amps
*Peak Non-Repetitive Surge Current (One cycle, 60 Hz, preceded and followed by rated current and voltage)	I_{TSM}	150	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	93	A^2s
*Peak Gate Power	P_{GM}	5	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt

*Indicates JEDEC Registered Data

(cont.)

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued (T_J = 125°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Forward Gate Current 2N682–2N688 2N690, 2N692	I_{GM}	2 1.2	Amps
*Peak Gate Voltage — Forward Reverse	V_{FGM} V_{RGM}	10 5	Volts
*Operating Junction Temperature Range	T_J	−65 to +125	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Average Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open, T_J = 125°C) 2N682–2N683 2N685 2N688 2N690 2N692	$I_{D(AV)}$, $I_{R(AV)}$	 — — — — —	 — — — — —	 6.5 6 4 2.5 2	mA
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	 — —	 — —	 10 20	 μA mA
*Peak On-State Voltage (I_{TM} = 50.3 A peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc) (V_{AK} = 12 Vdc, R_L = 50 Ω) *(V_{AK} = 12 Vdc, R_L = 50 Ω, T_C = −65°C)	I_{GT}	 — —	 — —	 40 80	mA
Gate Trigger Voltage (Continuous dc) (V_{AK} = 12 Vdc, R_L = 50 Ω) *(V_{AK} = 12 Vdc, R_L = 50 Ω, T_J = −65°C)	V_{GT}	 — —	 0.65 —	 2 3	Volts
*Gate Non-Trigger Voltage (Rated V_{DRM}, R_L = 50 Ω, T_J = 125°C)	V_{GD}	0.25	—	—	Volts
Holding Current (V_{AK} = 12 Vdc, Gate Open)	I_H	—	7.3	50	mA
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Exponential Waveform, T_J = 125°C, Gate Open)	dv/dt	—	30	—	V/μs

*Indicates JEDEC Registered Data.

3

TYPICAL CHARACTERISTICS

FIGURE 1 – AVERAGE CURRENT DERATING

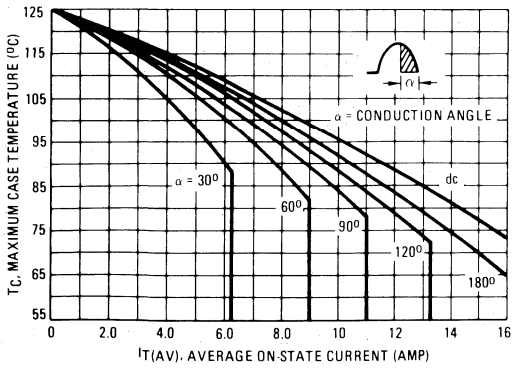


FIGURE 2 – MAXIMUM ON-STATE POWER DISSIPATION

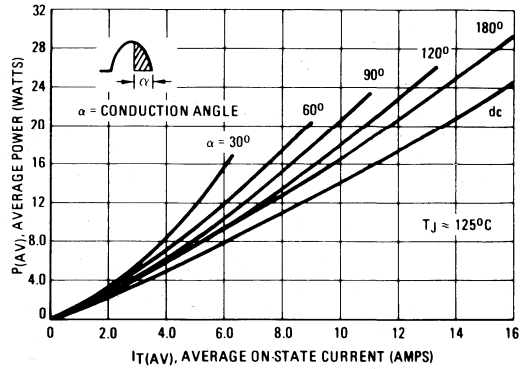


FIGURE 3 – ON-STATE CHARACTERISTICS

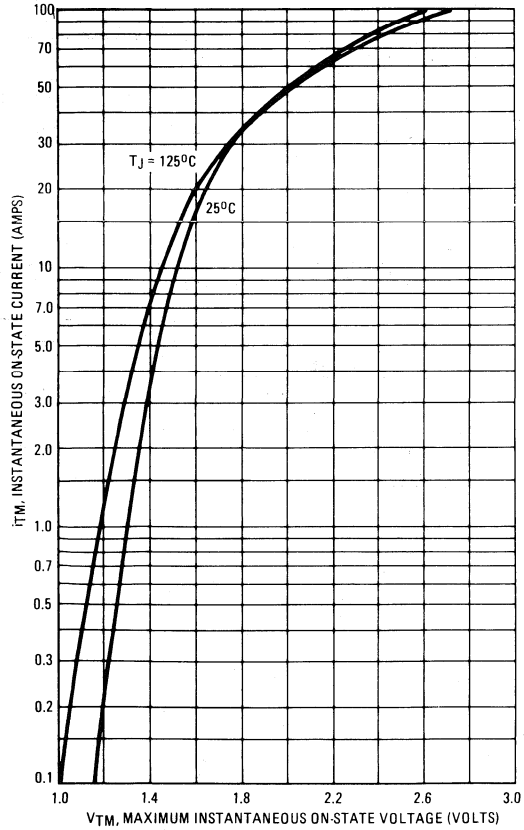


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

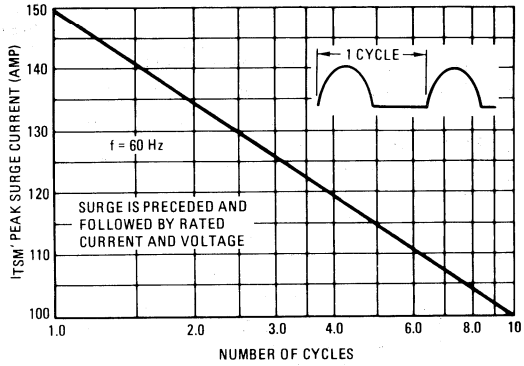


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FIGURE 5 – THERMAL RESPONSE

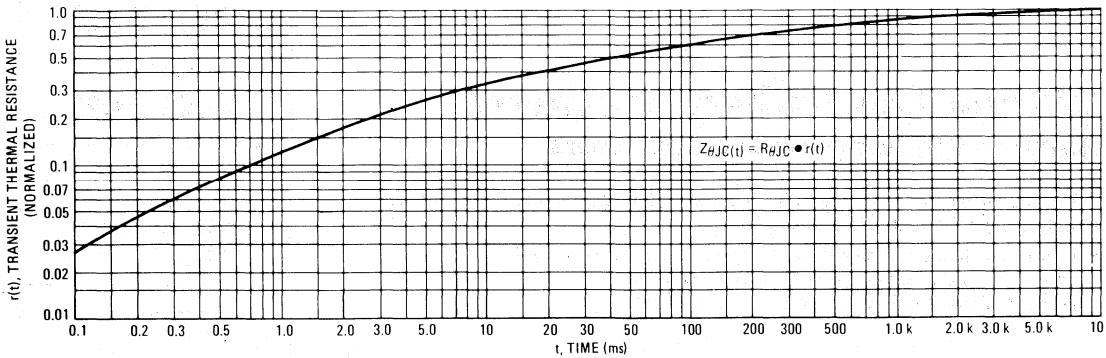


FIGURE 6 – PULSE TRIGGER CURRENT

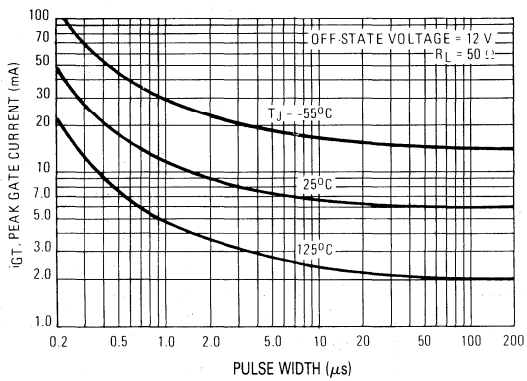


FIGURE 7 – GATE TRIGGER CURRENT

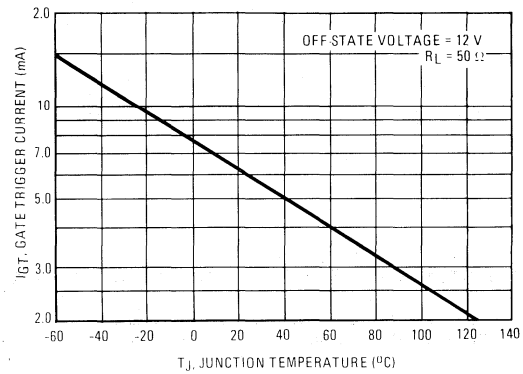


FIGURE 8 – GATE TRIGGER VOLTAGE

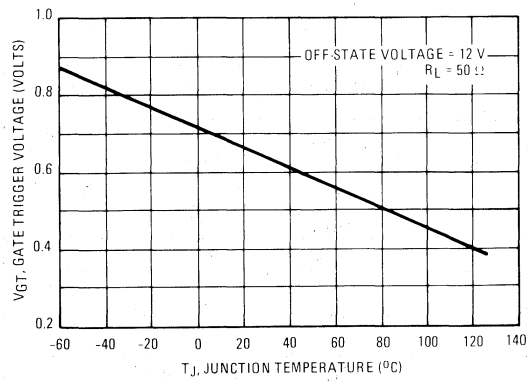
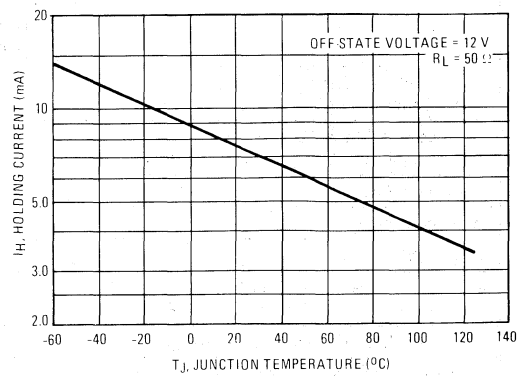


FIGURE 9 – HOLDING CURRENT



3

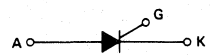
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

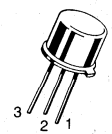
2N1595 thru 2N1599

SCRs
1.6 AMPERES RMS
50 thru 400 VOLTS

These devices are glassivated planar construction designed for gating operation in mA/μA signal or detection circuits.

- Low-Level Gate Characteristics — I_{GT} = 10 mA (Max) @ 25°C
- Low Holding Current — I_H = 5 mA (Typ) @ 25°C
- Glass-to-Metal Bond for Maximum Hermetic Seal





CASE 79-04
(TO-205AD)
STYLE 3

3

***MAXIMUM RATINGS** (T_J = 125°C unless otherwise noted, R_{GC} = 1 kΩ.)

Rating		Symbol	Value	Unit
Repetitive Peak Reverse Blocking Voltage, Note 1	2N1595 2N1596 2N1597 2N1599	V_{RRM}	50 100 200 400	Volts
Repetitive Peak Forward Blocking Voltage, Note 1	2N1595 2N1596 2N1597 2N1599	V_{DRM}	50 100 200 400	Volts
RMS On-State Current (All Conduction Angles)		$I_{T(RMS)}$	1.6	Amps
Peak Non-Repetitive Surge Current (One Cycle, 60 Hz, T_J = −65 to +125°C)		I_{TSM}	15	Amps
Peak Gate Power		P_{GM}	0.1	Watt
Average Gate Power		$P_{G(AV)}$	0.01	Watt
Peak Gate Current		I_{GM}	0.1	Amp
Peak Gate Voltage — Forward Reverse		V_{GFM} V_{GRM}	10 10	Volts
Operating Junction Temperature Range		T_J	−65 to +125	°C
Storage Temperature Range		T_{stg}	−65 to +150	°C

*Indicates JEDEC Registered Data.

Note 1. V_{DRM} or V_{RRM} for all types can be applied on a continuous DC basis without incurring damage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 6	μA mA
*Peak On-State Voltage (I_F = 1 Adc, Pulsed, 1 ms (Max), Duty Cycle ≤ 1%)	V_{TM}	—	1.1	2	Volts
*Gate Trigger Current (Continuous dc) (V_D = 7 V, R_L = 12 Ohms)	I_{GT}	—	2	10	mA
*Gate Trigger Voltage (Continuous dc) (V_D = 7 V, R_L = 12 Ohms) (V_D = 7 V, R_L = 12 Ohms, T_J = 125°C)	V_{GT}	 — 0.2	 0.7 —	 3 —	Volts
Reverse Gate Current (V_{GK} = 10 V)	I_{GR}	—	17	—	mA
Holding Current (V_D = 7 V)	I_H	—	5	—	mA
Turn-On Time (I_{GT} = 10 mA, I_F = 1 A) (I_{GT} = 20 mA, I_F = 1 A)	t_{gt}	 — —	 0.8 0.6	 — —	μs
Turn-Off Time (I_F = 1 A, I_R = 1 A, dv/dt = 20 V/μs, T_J = 125°C)	t_q	—	10	—	μs

*Indicates JEDEC Registered Data.

CURRENT DERATING

FIGURE 1 – CASE TEMPERATURE REFERENCE

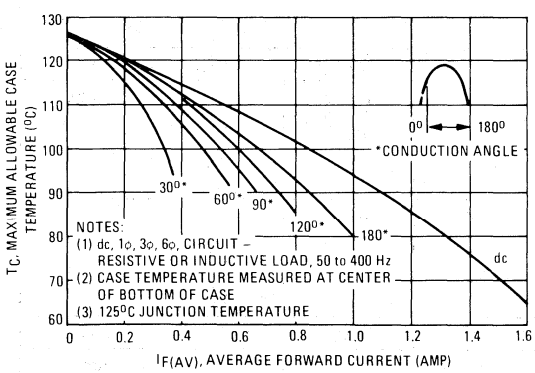
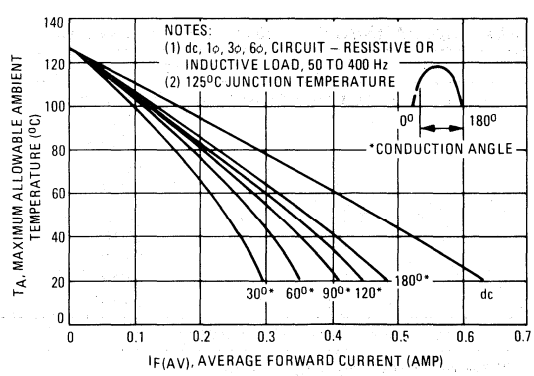


FIGURE 2 – AMBIENT TEMPERATURE REFERENCE



3

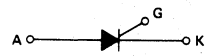
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

2N1843 thru 2N1849

SCRs
16 AMPERES RMS
50 thru 400 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions with Center Gate Geometry for Greater Parameter Uniformity and Stability
- Blocking Voltage to 400 Volts



CASE 263-04
STYLE 1

3

MAXIMUM RATINGS (T_J = 100°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Repetitive Forward or Reverse Blocking Voltage, Note 1 2N1843 2N1844 2N1846 2N1849	V_{DRM} or V_{RRM}	 50 100 200 400	Volts
*Non-Repetitive Peak Reverse Voltage 2N1843 2N1844 2N1846 2N1849	V_{RSM}	 75 150 300 500	Volts
*Average On-State Current (T_C = 35°C)	$I_{T(AV)}$	10	Amps
*Peak Non-Repetitive Surge Current (One cycle, 60 Hz, preceded and followed by rated current and voltage)	I_{TSM}	125	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	65	A^2s
*Peak Gate Power	P_{GM}	5	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt
*Peak Forward Gate Current	I_{GM}	2	Amps
*Peak Gate Voltage — Forward Reverse	V_{FGM} V_{RGM}	10 5	Volts
*Operating Junction Temperature Range	T_J	−40 to +100	°C
*Storage Temperature Range	T_{stg}	−40 to +125	°C

*Indicates JEDEC Registered Data.

(cont.)

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Average Forward or Reverse Blocking Current (V_D = Rated V_{DRM}, V_R = Rated V_{RRM}, T_C = 35°C)	$I_{D(AV)}$, $I_{R(AV)}$				mA
2N1843		—	—	19	
2N1844		—	—	12.5	
2N1846		—	—	6	
2N1849		—	—	4	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	6	mA
*Peak On-State Voltage (I_{TM} = 31.4 A peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	2.5	Volts
Gate Trigger Current (Continuous dc)	I_{GT}				mA
(V_D = 12 Vdc, R_L = 50 Ω)		—	6	80	
*(V_D = 12 Vdc, R_L = 50 Ω, T_C = −40°C)		—	—	150	
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volts
(V_D = 12 Vdc, R_L = 50 Ω)		—	0.65	—	
*(V_D = 12 Vdc, R_L = 50 Ω, T_C = −40°C)		—	—	3.5	
*(V_D Rated V_{DRM}, R_L = 50 Ω, T_C = 100°C)		0.3	—	—	
Holding Current (V_D = 12 Vdc, Gate Open)	I_H	—	7	—	mA
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 100°C, Gate Open)	dv/dt	—	30	—	V/μs

*Indicates JEDEC Registered Data.

3

FIGURE 1 – AVERAGE CURRENT DERATING

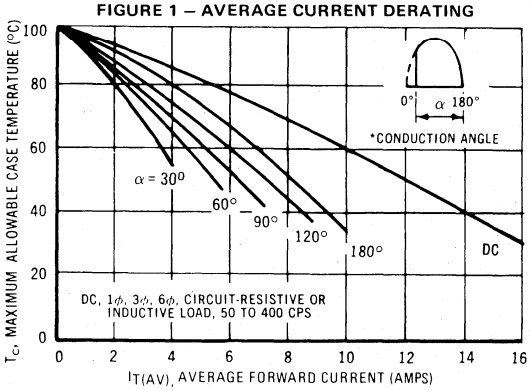


FIGURE 2 – GATE TRIGGER CURRENT

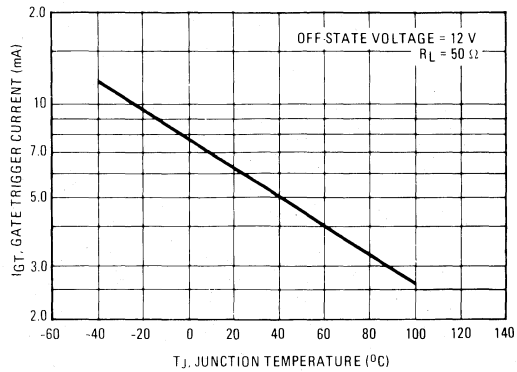


FIGURE 3 – GATE TRIGGER VOLTAGE

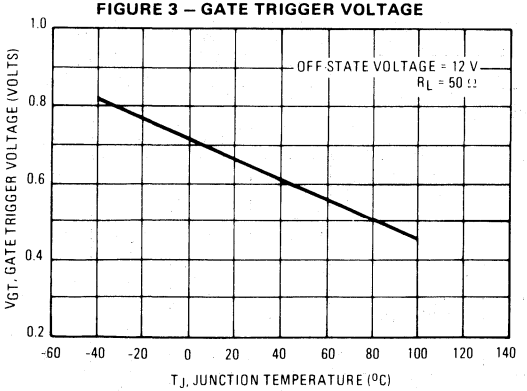
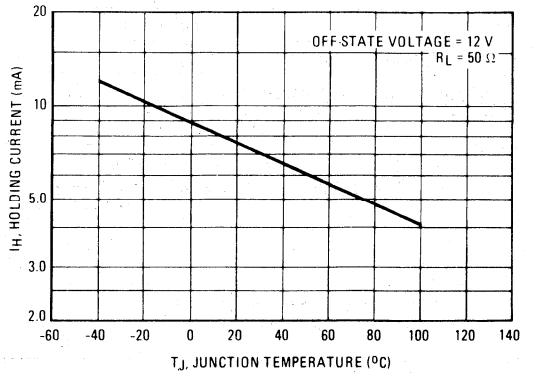


FIGURE 4 – HOLDING CURRENT



3

Silicon Controlled Rectifiers

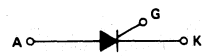
Reverse Blocking Triode Thyristors

2N1843A thru 2N1849A

SCRs
16 AMPERES RMS
50 thru 400 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions with Center Gate Geometry for Greater Parameter Uniformity and Stability
- Blocking Voltage to 400 Volts
- Junction Temperature Rated @ 125°C



CASE 263-04
STYLE 1

MAXIMUM RATINGS (T_J = 100°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Repetitive Forward or Reverse Blocking Voltage, Note 1 2N1843A 2N1844A 2N1846A 2N1849A	V_{DRM} or V_{RRM}	 50 100 200 400	Volts
*Non-Repetitive Peak Reverse Voltage 2N1843A 2N1844A 2N1846A 2N1849A	V_{RSM}	 75 150 300 500	Volts
*Average On-State Current (T_C = 80°C)	$I_{T(AV)}$	10	Amps
*Peak Non-Repetitive Surge Current (One cycle, 60 Hz, preceded and followed by rated current and voltage)	I_{TSM}	125	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	65	A^2s
*Peak Gate Power	P_{GM}	5	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt
*Peak Forward Gate Current	I_{GM}	2	Amps
*Peak Gate Voltage — Forward Reverse	V_{FGM} V_{RGM}	10 5	Volts
*Operating Junction Temperature Range	T_J	−65 to +125	°C
*Storage Temperature Range	T_{stg}	−65 to +125	°C

*Indicates JEDEC Registered Data.

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 125°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Average Forward or Reverse Blocking Current (V_D = Rated V_{DRM} or V_R = Rated V_{RRM}, gate open, T_C = 125°C)	$I_{D(AV)}$, $I_{R(AV)}$				mA
2N1843A		—	—	19	
2N1844A		—	—	12.5	
2N1846A		—	—	6	
2N1849A		—	—	4	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	6	mA
*Peak On-State Voltage (I_{TM} = 31.4 A peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	2.5	Volts
Gate Trigger Current (Continuous dc)	I_{GT}				mA
(V_D = 12 Vdc, R_L = 50 Ω)		—	6	80	
*(V_D = 12 Vdc, R_L = 50 Ω, T_C = −65°C)		—	—	150	
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volts
(V_D = 12 Vdc, R_L = 50 Ω)		—	0.65	—	
*(V_D = 12 Vdc, R_L = 50 Ω, T_C = −40°C)		—	—	3.5	
*(V_D = 12 Vdc, R_L = 50 Ω, T_C = −65°C)		—	—	3.7	
*(V_D = Rated V_{DRM}, R_L = 50 Ω, T_C = 125°C)		0.25	—	—	
Holding Current (V_D = 12 Vdc, Gate Open)	I_H	—	7	—	mA
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 125°C, Gate Open)	dv/dt	—	30	—	V/μs

*Indicates JEDEC Registered Data.

3

FIGURE 1 – AVERAGE CURRENT DERATING

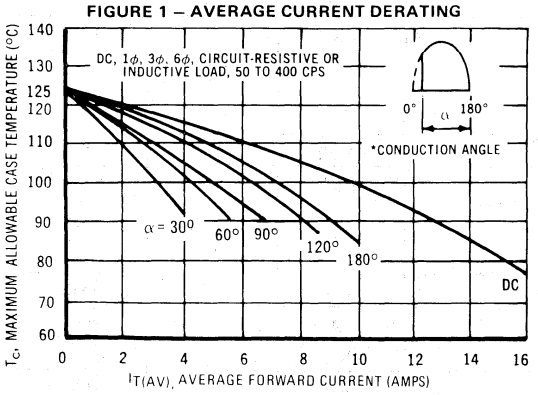


FIGURE 2 – GATE TRIGGER CURRENT

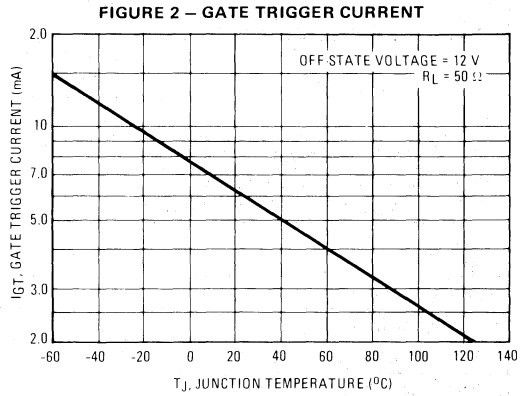


FIGURE 3 – GATE TRIGGER VOLTAGE

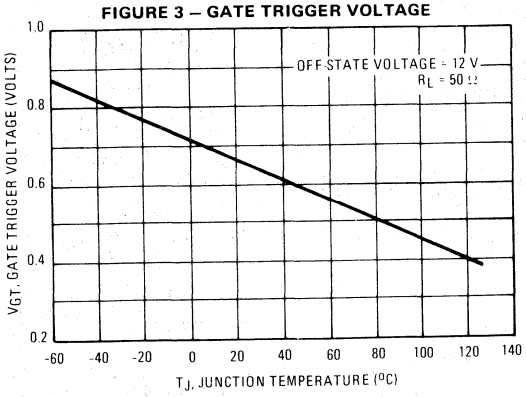
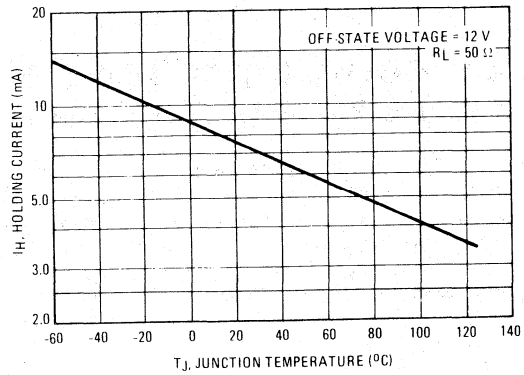


FIGURE 4 – HOLDING CURRENT



3

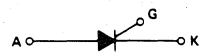
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

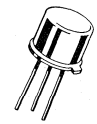
2N2323 thru 2N2329

SCRs
1.6 AMPERES RMS
50 thru 400 VOLTS

. . . all diffused PNP devices designed for gating operation in mA/μA signal or detection circuits.

- Low-Level Gate Characteristics — I_{GT} = 200 μA (Max) @ 25°C
- Low Holding Current — I_H = 2 mA (Max) @ 25°C
- Anode Common to Case
- Glass-to-Metal Bond for Maximum Hermetic Seal





CASE 79-04
(TO-205AD)
STYLE 3

3

***MAXIMUM RATINGS** (T_C = 25°C unless otherwise noted, R_{GK} = 1000 ohms.)

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage, Notes 1 and 2 2N2323 2N2324 2N2326 2N2329	V_{DRM} or V_{RRM}	 50 100 200 400	Volts
Non-Repetitive Peak Reverse Blocking Voltage (t ≤ 5 ms, Notes and 2) 2N2323 2N2324 2N2326 2N2329	V_{RSM}	 75 150 300 500	Volts
RMS On-State Current (All Conduction Angles)	$I_{T(RMS)}$	1.6	Amps
Average On-State Current T_C = 85°C T_A = 30°C	$I_{T(AV)}$	1 0.45	Amp
Peak Non-Repetitive Surge Current (One cycle, 60 Hz, T_C = 80°C) Preceded and followed by rated current and voltage	I_{TSM}	15	Amps

*Indiciates JEDEC Registered Data.

(cont.)

Notes: 1. Thyristor devices shall not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

2. Thyristor devices shall not have a positive bias applied to the gate concurrently with a negative potential applied to the anode.

***MAXIMUM RATINGS — continued** (T_C = 25°C unless otherwise noted, R_{GK} = 1000 ohms.)

Rating	Symbol	Value	Unit
Peak Gate Power	P_{GM}	0.1	Watt
Average Gate Power	$P_{G(AV)}$	0.01	Watt
Peak Gate Current	I_{GM}	0.1	Amp
Peak Gate Voltage	V_{GM}	6	Volts
Operating Junction Temperature Range	T_J	−65 to +125	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Lead Solder Temperature (>1/16″ from case, 10 s max)	—	+230	°C

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted, R_{GK} = 1000 ohms.)

Characteristic	Symbol	Min	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	10	μA
T_J = 125°C		—	100	μA
Peak On-State Voltage (I_{TM} = 1 A peak)	V_{TM}	—	1.5	Volts
(I_{TM} = 3.14 A Peak, T_C = 85°C)*		—	2	
Gate Trigger Current (Continuous dc), Note 1 (V_D = 6 Vdc, R_L = 100 ohms)	I_{GT}	—	200	μA
(V_D = 6 Vdc, R_L = 100 ohms, T_C = −65°C)		—	350*	
Gate Trigger Voltage (Continuous dc) (V_D = 6 Vdc, R_L = 100 ohms)	V_{GT}	—	0.8	Volts
(V_D = 6 Vdc, R_L = 100 ohms, T_C = −65°C)*		—	1	
(V_D = Rated V_{DRM}, R_L = 100 ohms, T_J = 125°C)*		0.1	—	
Holding Current (V_D = 6 Vdc)	I_H	—	2	mA
(V_D = 6 Vdc, T_C = −65°C)*		—	3	
(V_D = 6 Vdc, T_C = 125°C)*		0.15	—	

*Indicates JEDEC Registered Data.

Note 1. R_{GK} current is not included in measurement.

3

CURRENT DERATING

FIGURE 1 — CASE TEMPERATURE

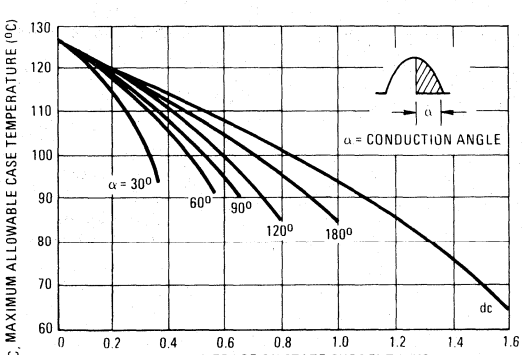
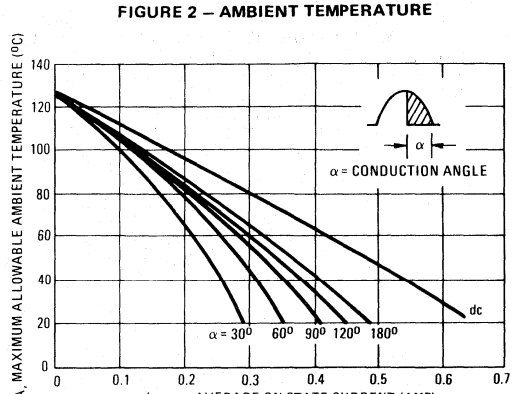


FIGURE 2 — AMBIENT TEMPERATURE



Silicon Controlled Rectifiers

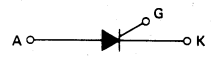
Reverse Blocking Triode Thyristors

. . . designed for industrial and consumer applications such as power supplies; battery chargers; temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 350 Amp
- Practical Level Triggering and Holding Characteristics — 4 and 5.2 mA (Typ) @ T_C = 25°C
- Rugged Construction in Either Pressfit, Stud or Isolated Stud Package

2N3870 thru 2N3873 MCR3896 thru MCR3899 2N6171 thru 2N6174

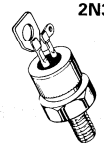
SCRs
35 AMPERES RMS
100 thru 600 VOLTS



MAXIMUM RATINGS (T_C = 100°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Repetitive Forward or Reverse Blocking Voltage, Note 1 (T_J = −40 to +100°C, 1/2 Sine Wave, 50 to 400 Hz, Gate Open)	V_{RRM} or V_{DRM}		Volts
2N3870, MCR3896, 2N6171		100	
2N3871, MCR3897, 2N6172		200	
2N3872, MCR3898, 2N6173		400	
2N3873, MCR3899, 2N6174		600	
*Peak Non-Repetitive Forward or Reverse Blocking Voltage (t ≤ 5 ms)	V_{RSM} or V_{DSM}		Volts
2N3870, MCR3896, 2N6171		150	
2N3871, MCR3897, 2N6172		330	
2N3872, MCR3898, 2N6173		660	
2N3873, MCR3899, 2N6174		700	
*Average On-State Current, Note 2	$I_{T(AV)}$		Amps
(T_C = −40 to +65°C)		22	
(T_C = +85°C)		11	
*Peak Non-Repetitive Surge Current (One cycle, 60 Hz) (T_C = +65°C)	I_{TSM}	350	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	510	A^2s

CASE 174-04
(TO-203AA)
STYLE 1
2N3870 thru 2N3873



CASE 263-04
STYLE 1
MCR3896 thru MCR3899

CASE 311-02
STYLE 1
(Stud Isolated)
2N6171 thru 2N6174

3

MAXIMUM RATINGS (T_C = 100°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Peak Gate Power	P_{GM}	20	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt
*Peak Forward Gate Current	I_{GM}	2	Amps
Peak Gate Voltage	V_{GM}	10	Volts
*Operating Junction Temperature Range	T_J	−40 to +100	°C
*Storage Temperature Range	T_{stg}	−40 to +150	°C
Stud Torque	—	30	in. lb.

*Indicates JEDEC Registered Data.

***THERMAL CHARACTERISTICS**

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$		°C/W
2N3870 thru 2N3873, MCR3896 thru MCR3899		0.9	
2N6171 thru 2N6174		1	

* Indicates JEDEC Registered Data.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current	I_{DRM}, I_{RRM}				mA
(Rated V_{DRM} or V_{RRM}, gate open, T_C = 100°C)					
2N3870, MCR3896, 2N6171		—	1	2	
2N3871, MCR3897, 2N6172		—	1	2.5	
2N3872, MCR3898, 2N6173		—	1	3	
2N3873, MCR3899, 2N6174		—	1	4	
(Rated V_{DRM} or V_{RRM}, gate open, T_C = 25°C) All Devices		—	—	10	μA
*Peak On-State Voltage (I_{TM} = 69 A Peak)	V_{TM}	—	1.5	1.85	Volts
*Gate Trigger Current (Continuous dc) *T_C = −40°C	I_{GT}	—	9	80	mA
(V_D = 12 V, R_L = 24 ohms) T_C = 25°C		—	4	40	
*Gate Trigger Voltage (Continuous dc) *T_C = −40°C	V_{GT}	—	0.9	3	Volts
(V_D = q2 V, R_L = 24 ohms) T_C = 25°C		—	0.69	1.6	
*Holding Current (Gate Open) *T_C = −40°C	I_H	—	14	90	mA
(V_D = 12 V, I_{TM} = 200 mA) T_C = 25°C		—	5.2	50	
*Gate Controlled Turn-On Time (t_d + t_r) (I_{TM} = 41 Adc, V_D = rated V_{DRM}, I_{GT} = 40 mAdc, Rise Time ≤ 0.05 μs, Pulse Width = 10 μs)	t_{gt}	—	—	1.5	μs
Circuit Commutated Turn-Off Time	t_q				μs
(I_{TM} = 10 A, I_R = 10 A)		—	25	—	
(I_{TM} = 10 A, I_R = 10 A, T_C = 100°C)		—	35	—	
Forward Voltage Application Rate (T_C = 100°C, V_D = Rated V_{DRM})	dv/dt	—	50	—	V/μs

*Indicates JEDEC Registered Data.

3

FIGURE 1 – AVERAGE CURRENT DERATING

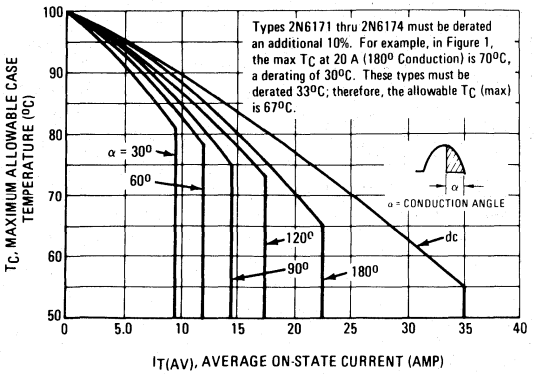


FIGURE 2 – ON-STATE POWER DISSIPATION

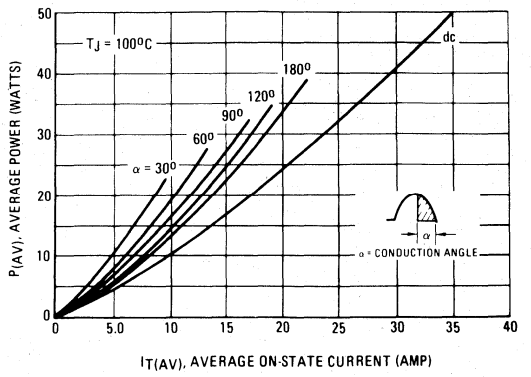


FIGURE 3 – ON-STATE CHARACTERISTICS

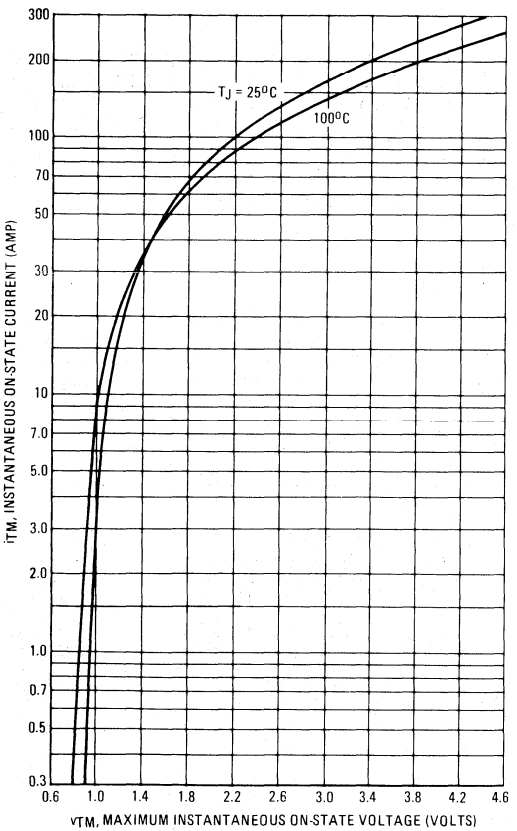


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

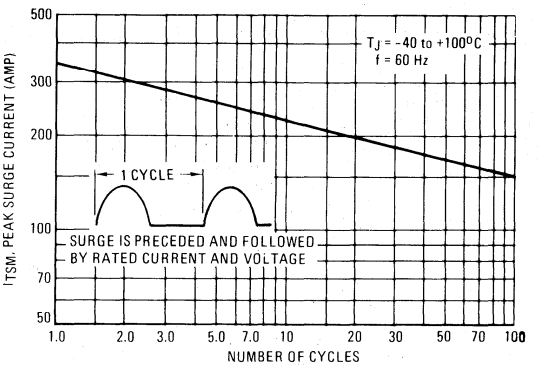


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FIGURE 5 – TYPICAL THERMAL RESPONSE

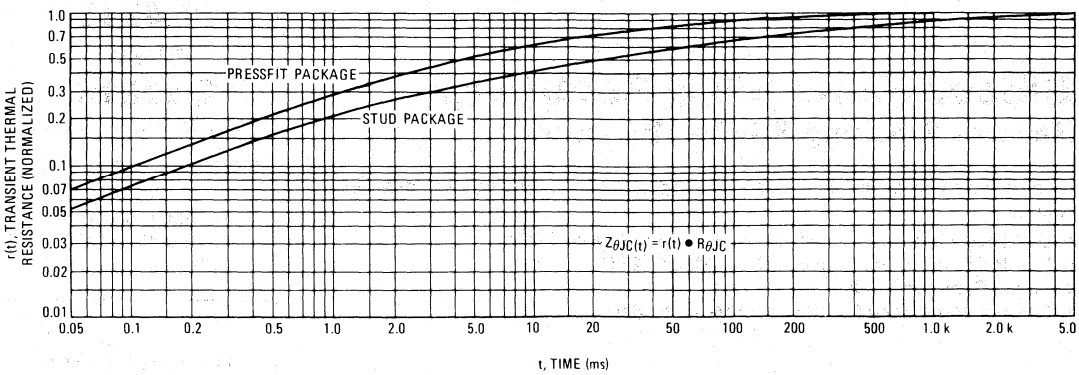


FIGURE 6 – PULSE TRIGGER CURRENT

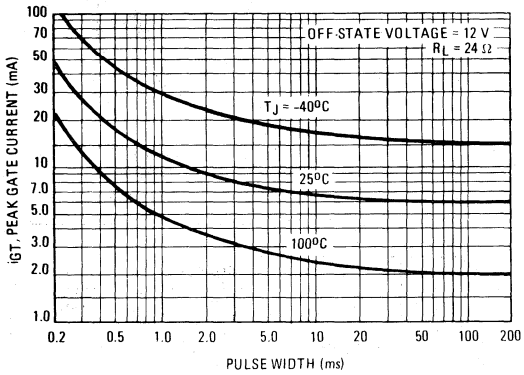


FIGURE 7 – GATE TRIGGER CURRENT

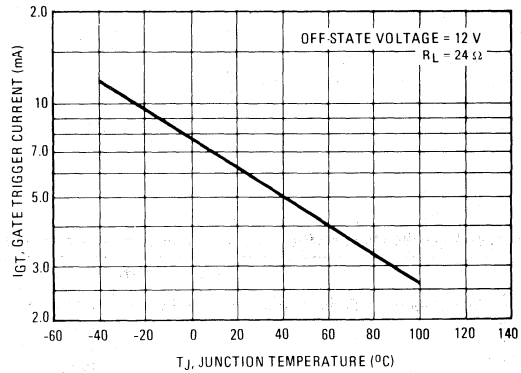


FIGURE 8 – GATE TRIGGER VOLTAGE

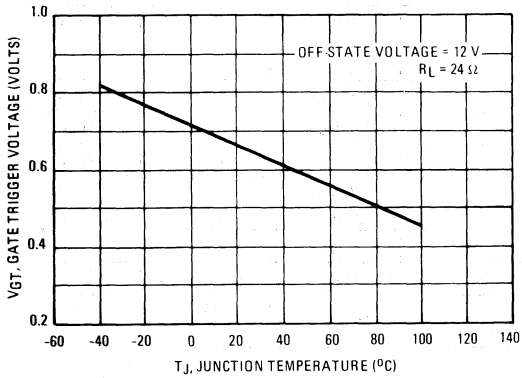
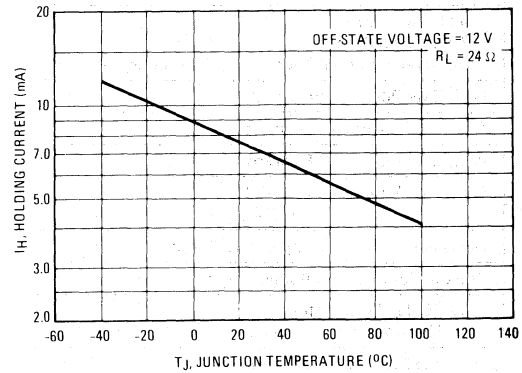


FIGURE 9 – HOLDING CURRENT



3

Silicon Controlled Rectifiers

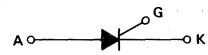
Reverse Blocking Triode Thyristors

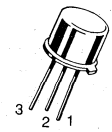
2N4213 thru 2N4219

SCRs
1.6 AMPERES RMS
50 thru 400 VOLTS

. . . all diffused PNPN devices designed for operation in mA/μA signal or detection circuits.

- Low-Level Gate Characteristics — I_{GT} = 100 μA Max @ 25°C
- Low Holding Current — I_{HX} = 3 mA Max @ 25°C
- Anode Common To Case
- Glass-to-Metal Bond for Maximum Hermetic Seal





CASE 79-04
(TO-205AD)
STYLE 3

3

***MAXIMUM RATINGS** (T_J = 125°C unless otherwise noted.)

Characteristic	Symbol	Rating	Unit
Peak Repetitive Forward and Reverse Blocking Voltage, Note 1 2N4213 2N4214 2N4216 2N4219	V_{DRM} or V_{RRM}	 50 100 200 400	Volts
Forward Current RMS (All Conduction Angles)	$I_{T(RMS)}$	1.6	Amps
Peak Surge Current (One Cycle, 60 Hz) No Repetition until Thermal Equilibrium is Restored	I_{TSM}	15	Amps
Peak Gate Power — Forward	P_{GFM}	0.1	Watt
Average Gate Power — Forward	$P_{GF(AV)}$	0.01	Watt
Peak Gate Current — Forward	I_{GFM}	0.1	Amp
Peak Gate Voltage — Forward Reverse	V_{GFM} V_{GRM}	6 6	Volts
Operating Junction Temperature Range	T_J	−65 to +125	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Lead Solder Temperature (>1/16" from case, 10 s max)	—	+230	°C

*Indicates JEDEC Registered Values.

Note 1. V_{DRM} and V_{RRM} can be applied for all types on a continuous dc basis without incurring damage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted, R_{GK} = 1000 ohms.), Note 1

Characteristic	Symbol	Min	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C T_C = 125°C	I_{DRM}, I_{RRM}	— —	10 200	μA μA
*Forward "On" Voltage (I_{TM} = 1 Adc peak)	V_{TM}	—	1.5	Volts
Gate Trigger Current (Continuous dc), Note 2 (V_D = 7 V, R_L = 100 ohms) (T_C = 25°C) (T_C = −65°C)	I_{GT}	— —	100 300	μAdc
Gate Trigger Voltage (Continuous dc) (V_D = 7 V, R_L = 100 ohms, T_C = 25°C) *(V_D = 7 V, R_L = 100 ohms, T_C = −65°C) *(V_D = Rated V_{DRM}, R_L = 100 ohms, T_C = 125°C)	V_{GT}	— — 0.1	0.8 1 —	Volt
Holding Current (V_D = 7 V) T_C = 25°C *T_C = −65°C	I_{HX}		3 7	mA
Turn-On Time	t_{on}	Circuit dependent, consult manufacturer		
Turn-Off Time	t_{off}			

*Indicates JEDEC Registered Values.

Notes: 1. Thyristor devices shall not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

Thyristor devices shall not have a positive bias applied to the gate concurrently with a negative potential applied to the anode.

2. R_{GK} current is not included in measurement.

3

FIGURE 1 — CASE TEMPERATURE vs CURRENT

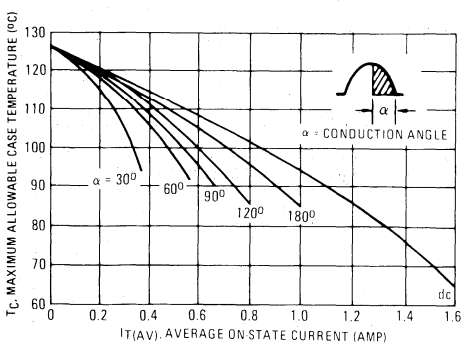
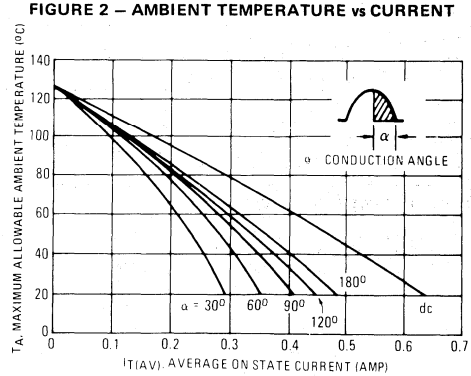


FIGURE 2 — AMBIENT TEMPERATURE vs CURRENT



NOT RECOMMENDED FOR NEW DESIGNS

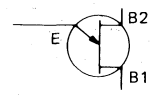
PN Unijunction Transistors
Silicon Unijunction Transistors

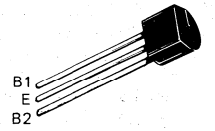
2N4870
2N4871

PN UJTs

. . . designed for pulse and timing circuits, sensing circuits, and thyristor trigger circuits. These devices feature:

- Low Peak Point Current — 1 μA Typical
- Low Emitter Reverse Current — 5 nA Typical
- Passivated Surface for Reliability and Uniformity
- One-Piece Injection-Molded Unibloc‡ Plastic Package for Economy and Reliability
- High η for greater bandwidth





CASE 29-04
(TO-226AA)
STYLE 9

3

MAXIMUM RATINGS (T_A = 25° unless otherwise noted.)

Rating	Symbol	Value	Unit
RMS Power Dissipation, Note 1	P_D	300	mW
RMS Emitter Current	I_e	50	mA
Peak-Pulse Emitter Current, Note 2	i_e	1.5	Amp
Emitter Reverse Voltage	V_{B2E}	30	Volts
Interbase Voltage, Note 3	V_{B2B1}	35	Volts
Operating Junction Temperature Range	T_J	−55 to +125	°C
Storage Temperature Range	T_{stg}	−55 to +150	°C

Notes: 1. Derate 3 mW/°C increase in ambient temperature.
2. Duty cycle ≤ 1%, PRR = 10 PPS (see Figure 5).
3. Based upon power dissipation at T_A = 25°C.

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted.)

Characteristic		Fig. No.	Symbol	Min	Typ	Max	Unit
Intrinsic Standoff Ratio, Note 1 (V_{B2B1} = 10 V)	2N4870 2N4871	4, 7	η	0.56 0.70	— —	0.75 0.85	—
Interbase Resistance (V_{B2B1} = 3 V, I_E = 0)		10, 11	R_{BB}	4	6	9.1	k ohms
Interbase Resistance Temperature Coefficient (V_{B2B1} = 3 V, I_E = 0, T_A = −65 to +125°C)		11	αR_{BB}	0.10	—	0.90	%/°C
Emitter Saturation Voltage, Note 2 (V_{B2B1} = 10 V, I_E = 50 mA)			$V_{EB1(sat)}$	—	2.5	—	Volts
Modulated Interbase Current (V_{B2B1} = 10 V, I_E = 50 mA)			$I_{B2(mod)}$	—	15	—	mA
Emitter Reverse Current (V_{B2E} = 30 V, I_{B1} = 0)		6	I_{EB2O}	—	0.005	1	μA
Peak-Point Emitter Current (V_{B2B1} = 25 V)		8, 9	I_P	—	1	5	μA
Valley-Point Current, Note 2 (V_{B2B1} = 20 V, R_{B2} = 100 ohms)	2N4870 2N4871	12, 13	I_V	2 4	5 7	— —	mA
Base-One Peak Pulse Voltage	2N4870 2N4871	3, 16	V_{OB1}	3 5	6 8	— —	Volts

Notes: 1. η, Intrinsic standoff ratio, is defined in terms of the peak-point voltage, V_P, by means of the equation: $V_P = \eta V_{B2B1} + V_F$, where V_F is about 0.49 volt at 25°C @ I_F = 10 μA and decreases with temperature at about 2.5 mV/°C. The test circuit is shown in Figure 4. Components R_1, C_1, and the UJT form a relaxation oscillator; the remaining circuitry serves as a peak-voltage detector. The forward drop of Diode D_1 compensates for V_R. To use, the "cal" button is pushed, and R_3 is adjusted to make the current meter, M_1, read full scale. When the "cal" button is released, the value of η is read directly from the meter, if full scale on the meter reads 1.

2. Use pulse techniques: PW ≈ 300 μs, duty cycle ≤ 2% to avoid internal heating, which may result in erroneous readings.

3

FIGURE 1 – UNIJUNCTION TRANSISTOR SYMBOL AND NOMENCLATURE

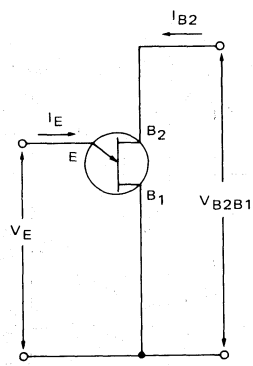


FIGURE 2 – STATIC EMITTER CHARACTERISTICS CURVES

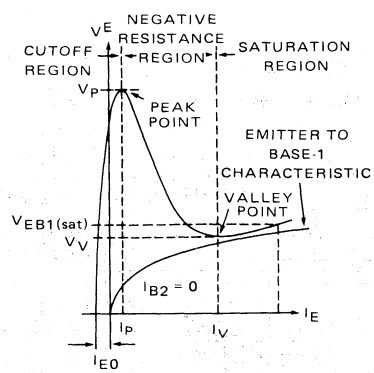


FIGURE 3 – V_{OB1} TEST CIRCUIT

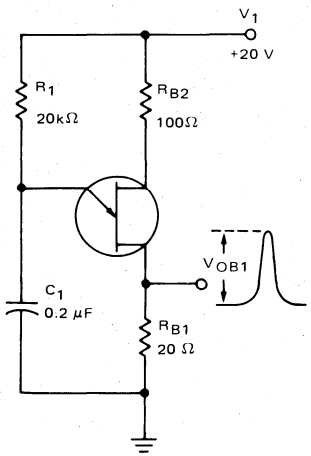


FIGURE 4 – η TEST CIRCUIT

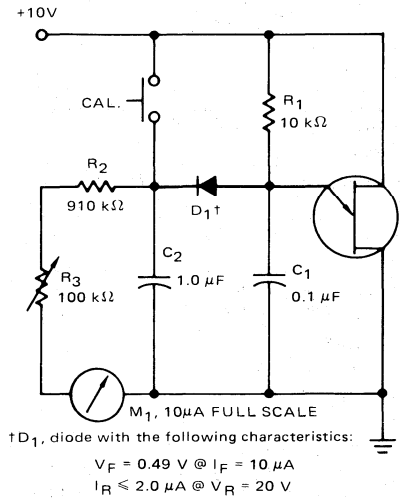
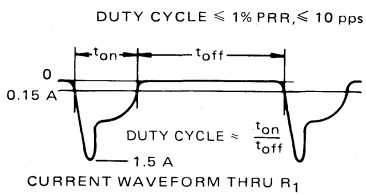
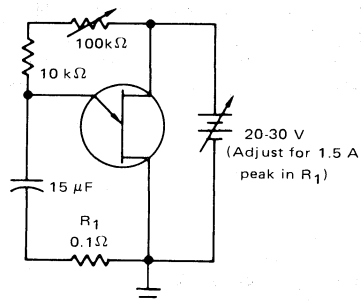


FIGURE 5 – PRR TEST CIRCUIT AND WAVEFORM





TYPICAL CHARACTERISTICS

FIGURE 6 – EMITTER REVERSE CURRENT

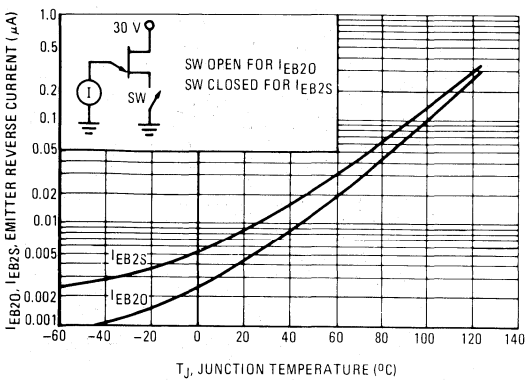
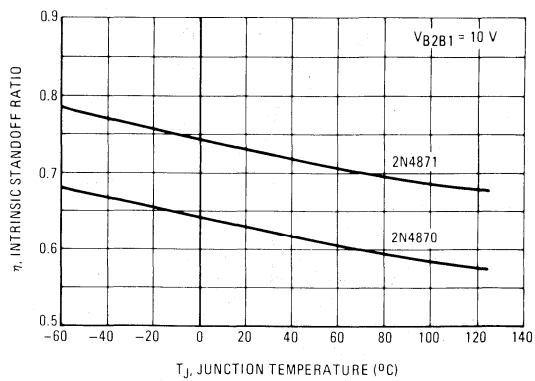


FIGURE 7 – INTRINSIC STANDOFF RATIO



PEAK POINT CURRENT

FIGURE 8 – EFFECT OF VOLTAGE

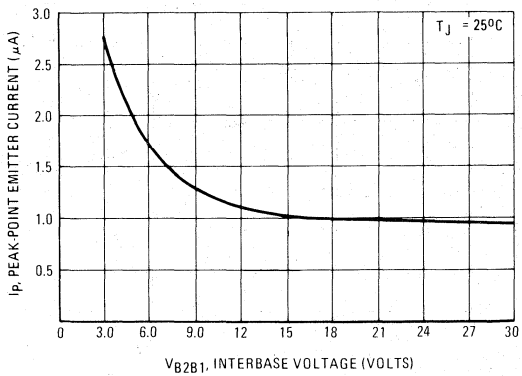
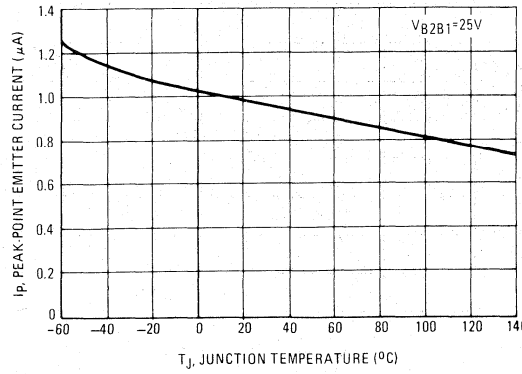


FIGURE 9 – EFFECT OF TEMPERATURE



INTERBASE RESISTANCE

FIGURE 10 – EFFECT OF VOLTAGE

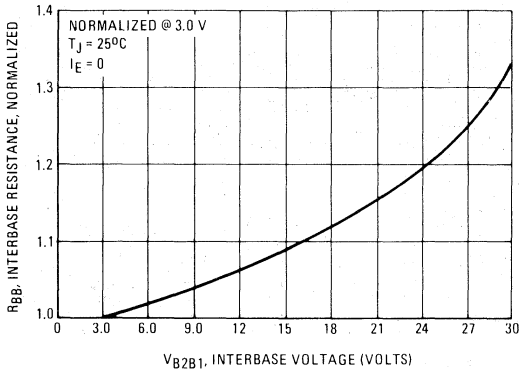
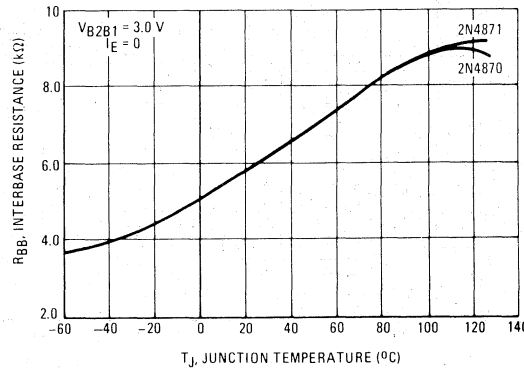


FIGURE 11 – EFFECT OF TEMPERATURE



TYPICAL CHARACTERISTICS

VALLEY CURRENT

FIGURE 12 – EFFECT OF VOLTAGE

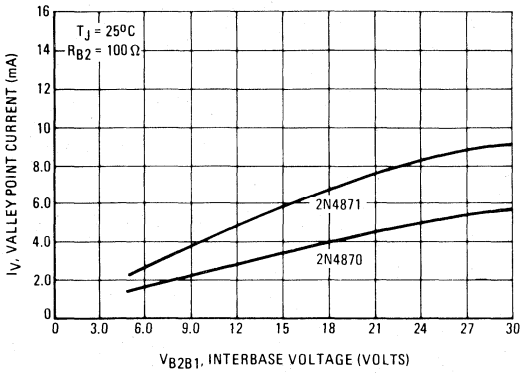
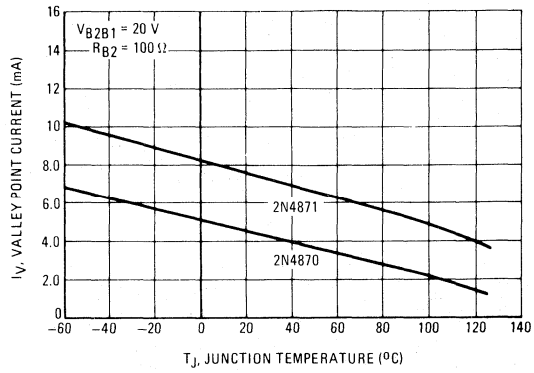


FIGURE 13 – EFFECT OF TEMPERATURE



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VALLEY VOLTAGE

FIGURE 14 – EFFECT OF VOLTAGE

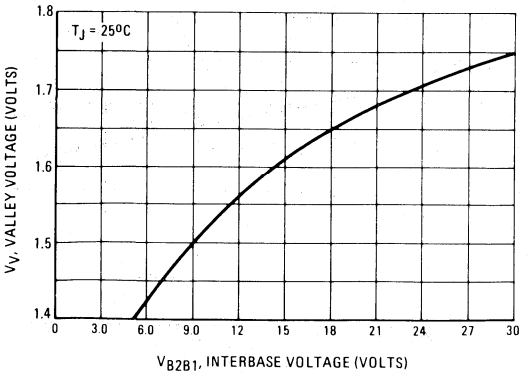


FIGURE 15 – EFFECT OF TEMPERATURE

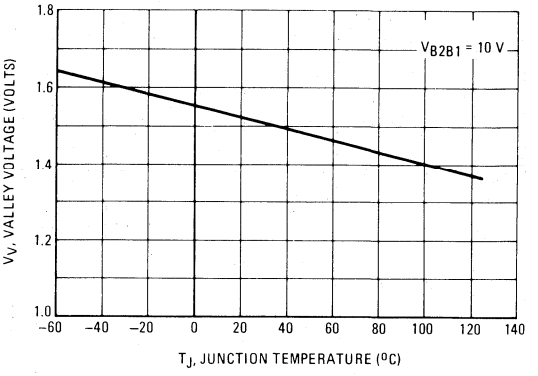
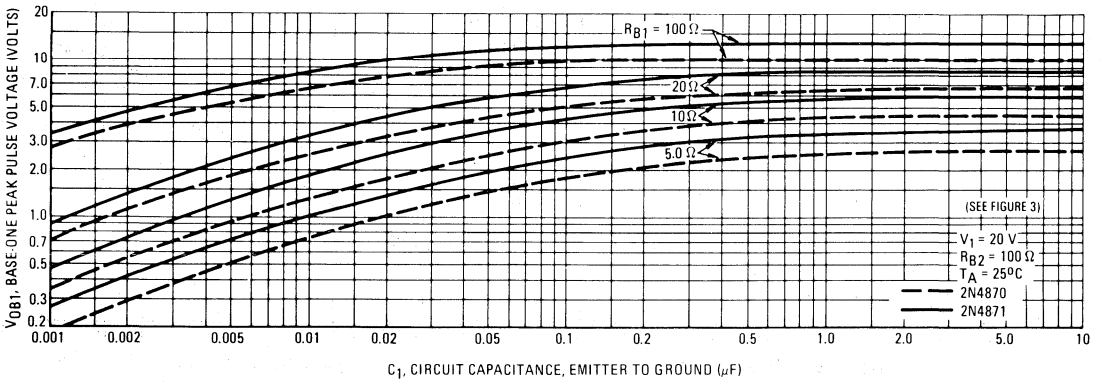


FIGURE 16 – OUTPUT VOLTAGE



3

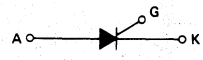
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

2N5060 thru 2N5064

SCRs
0.8 AMPERES RMS
30 thru 200 VOLTS

. . . Annular PNPN devices designed for high volume consumer applications such as relay and lamp drivers, small motor controls, gate drivers for larger thyristors, and sensing and detection circuits. Supplied in an inexpensive plastic TO-226AA (TO-92) package which is readily adaptable for use in automatic insertion equipment.

- Sensitive Gate Trigger Current — 200 μA Maximum
- Low Reverse and Forward Blocking Current — 50 μA Maximum, T_C = 125°C
- Low Holding Current — 5 mA Maximum
- Passivated Surface for Reliability and Uniformity
- Also Available with TO-5 or TO-206AA (TO-18) Leadform



CASE 29-04
(TO-226AA)
STYLE 10

3

MAXIMUM RATINGS (1)

Rating	Symbol	Value	Unit
*Peak Repetitive Reverse Blocking Voltage, Note 1 (R_{GK} = 1000 ohms, T_C = +125°C) 2N5060	V_{DRM} or V_{RRM}	30	Volts
2N5061		60	
2N5062		100	
2N5063		150	
2N5064		200	
On-State Current RMS (All Conduction Angles)	$I_{T(RMS)}$	0.8	Amp
*Average On-State Current (T_C = 67°C)	$I_{T(AV)}$	0.51	Amp
(T_C = 102°C)		0.255	
*Peak Non-Repetitive Surge Current, T_A = 25°C (1/2 cycle, Sine Wave, 60 Hz)	I_{TSM}	10	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	0.4	A^2s
*Peak Gate Power, T_A = 25°C	P_{GM}	0.1	Watt
*Average Gate Power, T_A = 25°C	$P_{G(AV)}$	0.01	Watt
*Peak Forward Gate Current, T_A = 25°C (300 μs, 120 PPS)	I_{FGM}	1	Amp
*Peak Reverse Gate Voltage	V_{RGM}	5	Volts

*Indicates JEDEC Registered Data.

Note 1. Ratings apply for zero or negative gate voltage. Device ratings exclude having a positive bias applied to the gate concurrently with a negative potential on the anode. Devices should not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
*Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−65 to +125	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
*Lead Solder Temperature (Lead Length ≥ 1/16″ from case, 10 s Max)	—	+230*	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case, Note 1	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.), Note 2

Characteristic		Symbol	Min	Typ	Max	Unit
*Peak Forward Blocking Voltage, Note 3 (T_C = 125°C, R_{GK} = 1000 Ohms)		V_{DRM}				Volts
	2N5060		30	—	—	
	2N5061		60	—	—	
	2N5062		100	—	—	
	2N5063		150	—	—	
	2N5064		200	—	—	
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, R_{GK} = 1000 Ohms)	T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
	T_C = 125°C		—	—	50	μA
*Forward "On" Voltage, Note 4 (I_{TM} = 1.2 A peak @ T_A = 25°C)		V_{TM}	—	—	1.7	Volts
Gate Trigger Current (Continuous dc), Note 5 *(Anode Voltage = 7 Vdc, R_L = 100 Ohms, R_{GK} = 1000 Ohms)	T_C = 25°C	I_{GT}	—	—	200	μA
	T_C = −65°C		—	—	350	
Gate Trigger Voltage (Continuous dc) *(Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	V_{GT}	—		0.8	Volts
	T_C = −65°C		—		1.2	
(Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms)	T_C = 125°C	V_{GD}	0.1		—	
Holding Current *(Anode Voltage = 7 Vdc, initiating current = 20 mA)	T_C = 25°C	I_H	—	—	5	mA
	T_C = −65°C		—	—	10	
Turn-On Time						μs
Delay Time		t_d	—	3	—	
Rise Time (I_{GT} = 1 mA, R_{GK} = 1 Ohm, V_D = Rated V_{DRM}, Forward Current = 1 A, di/dt = 6 A/μs		t_r	—	0.2	—	
Turn-Off Time (Forward Current = 1 A pulse, Pulse Width = 50 μs, 0.1% Duty Cycle, di/dt = 6 A/μs, dv/dt = 20 V/μs, I_{GT} = 1 mA, R_{GK} = 1 k Ohm)	2N5060, 2N5061	tq	—	10	—	μs
	2N5062, 5063, 5064		—	30	—	
Forward Voltage Application Rate (Rated V_{DRM}, R_{GK} = 1 k, Exponential)		dv/dt	—	30	—	V/μs

*Indicates JEDEC Registered Data.

Notes:
1. This measurement is made with the case mounted "flat side down" on a heat sink and held in position by means of a metal clamp over the curved surface.
2. For electrical characteristics for Gate-to-cathode resistance other than 1000 ohms see Motorola Bulletin EB-30.
3. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage but positive gate voltage shall not be applied concurrently with a negative potential on the anode. When checking forward or reverse blocking capability, thyristor devices should not be tested with a constant current source in a manner that the voltage applied exceeds the rated blocking voltage.
4. Forward current applied for 1 ms maximum duration, duty cycle ≤ 1%.
5. R_{GK} current is not included in measurement.

CURRENT DERATING

FIGURE 1 – MAXIMUM CASE TEMPERATURE

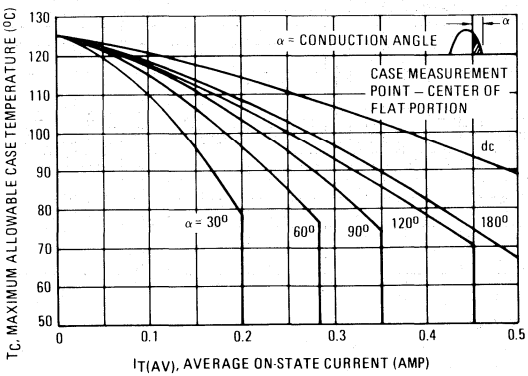


FIGURE 2 – MAXIMUM AMBIENT TEMPERATURE

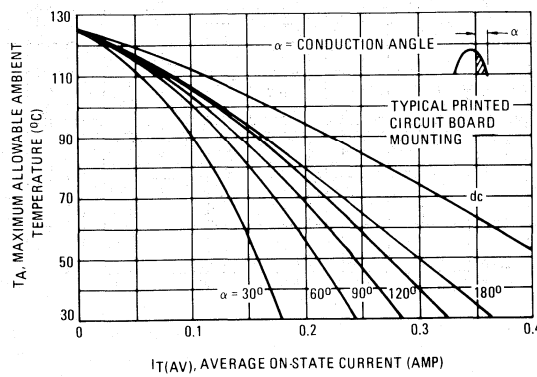


FIGURE 3 – TYPICAL FORWARD VOLTAGE

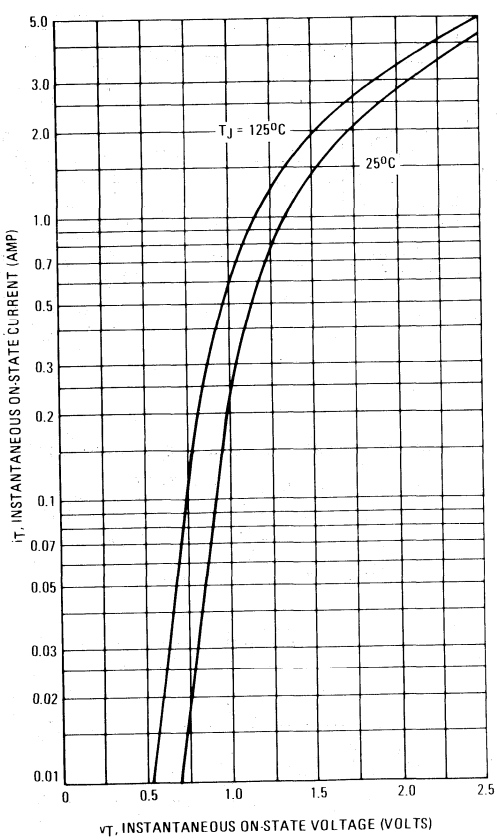


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

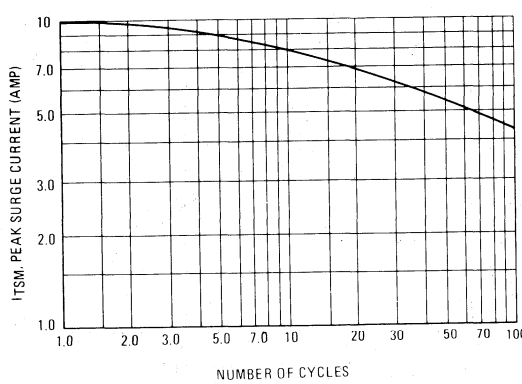


FIGURE 5 – POWER DISSIPATION

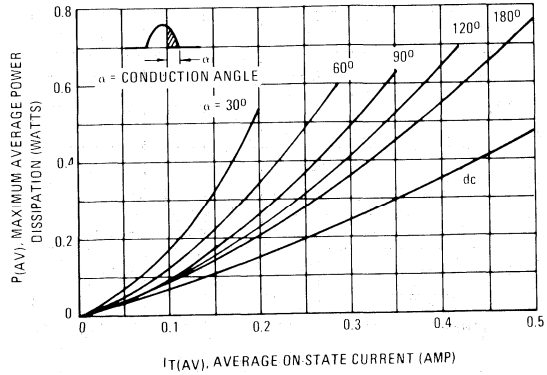


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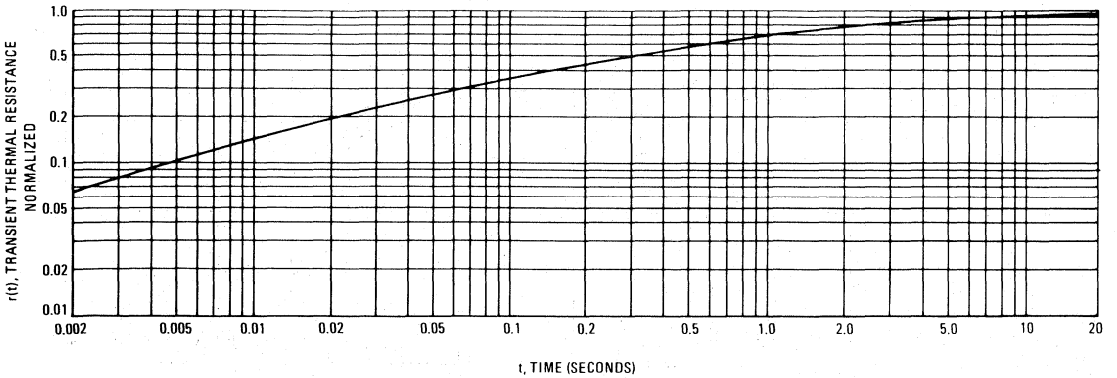
FIGURE 6 – THERMAL RESPONSE



TYPICAL CHARACTERISTICS

FIGURE 7 – GATE TRIGGER VOLTAGE

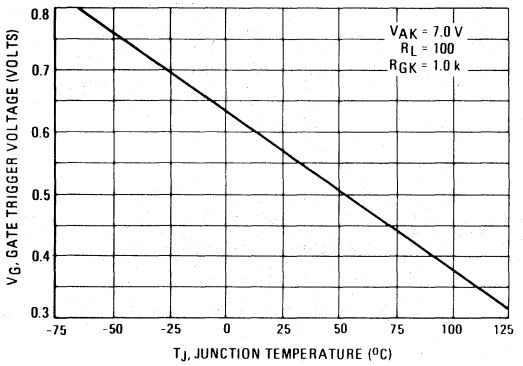


FIGURE 8 – GATE TRIGGER CURRENT

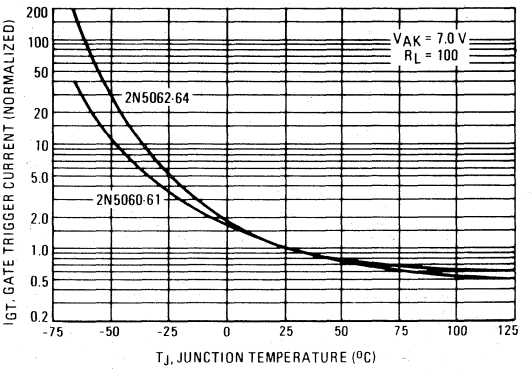


FIGURE 9 – HOLDING CURRENT

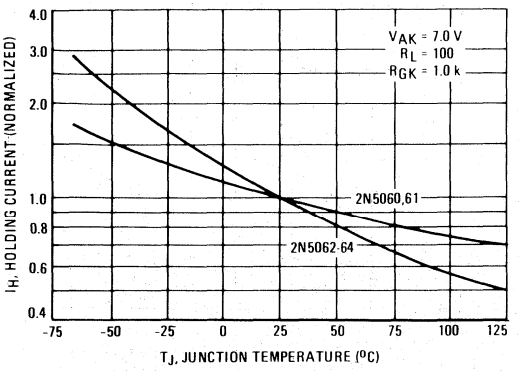
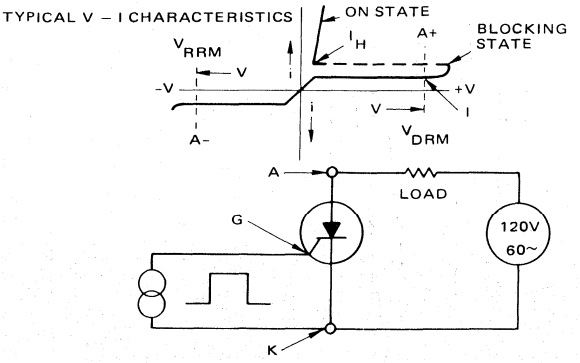


FIGURE 10 – CHARACTERISTICS AND SYMBOLS



Silicon Controlled Rectifiers

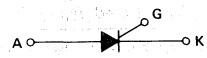
Reverse Blocking Triode Thyristors

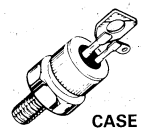
2N5168 thru 2N5171 MCR5164 thru MCR5167

SCRs
20 AMPERES RMS
50 thru 600 VOLTS

. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Supplied in Either Pressfit or Stud Package
- High Surge Current Rating — I_{TSM} = 240 Amps
- Low On-State Voltage — 1.2 V (Typ) @ I_{TM} = 20 Amps
- Practical Level Triggering and Holding Characteristics — 40 mA (Max) and 50 mA (Max) @ T_C = 25°C





CASE 263-04
STYLE 1
2N5168 thru 2N5171

CASE 174-04
(TO-203AA)
STYLE 1
MCR5164 thru MCR5167

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Forward and *Repetitive Reverse Blocking Voltage, Notes 1 and 2 MCR5164, 2N5168 MCR5165, 2N5169 MCR5166, 2N5170 MCR5167, 2N5171	V_{DRM} or V_{RRM}	 50 200 400 600	Volts
*Non-Repetitive Peak Reverse Blocking Voltage MCR5164, 2N5168 MCR5165, 2N5169 MCR5166, 2N5170 MCR5167, 2N5171	V_{RSM}	 75 300 500 700	Volts
On-State Current RMS	$I_{T(RMS)}$	20	Amps
Average On-State Current (T_C = 67°C)	$I_{T(AV)}$	13	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	240	A^2s
*Peak Non-Repetitive Surge Current (One cycle, 60 Hz, T_C = 67°C) Preceded and followed by rated current and voltage	I_{TSM}	240	Amps
*Peak Gate Power (Maximum Pulse Width = 10 μs)	P_{GM}	5	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt
*Peak Forward Gate Current (Maximum Pulse Width = 10 μs)	I_{GM}	2	Amps
Peak Gate Voltage	V_{GM}	10	Volts
*Operating Junction Temperature Range	T_J	−40 to +100	°C
*Storage Temperature Range	T_{stg}	−40 to +150	°C
Stud Torque 2N5168–2N5171		30	in. lb.

*Indicates JEDEC registered data.

Notes: 1. V_{DRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage applied exceeds the rated blocking voltage.

2. Devices should not be operated with a positive bias applied to the gate concurrent with a negative potential applied to the anode.

3

THERMAL CHARACTERISTICS

Characteristic	Symbol	Typ	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$			°C/W
MCR5164, 65, 66, 67		1	1.5	
2N5168, 69, 70, 71		1.1	1.6	

Note 2. Devices should not be operated with a positive bias applied to the gate concurrent with a negative potential applied to the anode.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	10	μA
T_C = 100°C		—	5	mA
Gate Trigger Current (Continuous dc), Note 1 (V_D = 7 Vdc, R_L = 100 Ω)	I_{GT}	—	40	mA
*(V_D = 7 Vdc, R_L = 100 Ω, T_C = −40°C)		—	75	
Gate Trigger Voltage (Continuous dc) (V_D = 7 Vdc, gate open)	V_{GT}	—	1.5	Volts
*(V_D = 7 Vdc, R_L = 100 Ω, T_C = −40°C)		—	2.5	
*(V_D = Rated V_{DRM}, R_L = 100 Ω, T_J = 100°C)		0.2	—	
Peak On-State Voltage (Pulse Width = 1 ms max, duty cycle ≤ 1%) (I_{TM} = 20 A)	V_{TM}	—	1.5	Volts
*(I_{TM} = 41 A)		8	1.7	
Holding Current (V_D = 7 Vdc, gate open)	I_H	—	50	mA
*(V_D = 7 Vdc, gate open, T_C = −40°C)		—	90	
		Typical		
Gate Controlled Turn-On Time (t_d + t_r) (I_{TM} = 20 A, I_{GT} = 40 mAdc, V_D = Rated V_{DRM})	t_{gt}	1		μs
Circuit Commutated Turn-Off Time (I_{TM} = 10 A, I_R = 10 A)	t_q	20		μs
(I_{TM} = 10 A, I_R = 10 A, T_J = 100°C) (V_D = V_{DRM} = rated voltage) (dv/dt = 30 V/μs)		30		
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Wave Form, Gate open, T_J = 100°C)	dv/dt	50		V/μs

*Indicates JEDEC registered data.

Note 1. Devices should not be operated with a positive bias applied to the gate concurrent with a negative potential applied to the anode.

3

EFFECT OF TEMPERATURE UPON TYPICAL TRIGGER CHARACTERISTICS

FIGURE 1 – GATE TRIGGER CURRENT

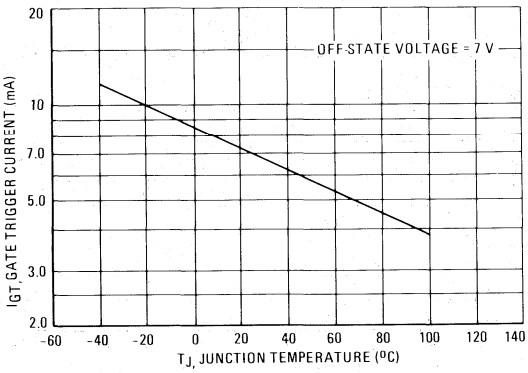
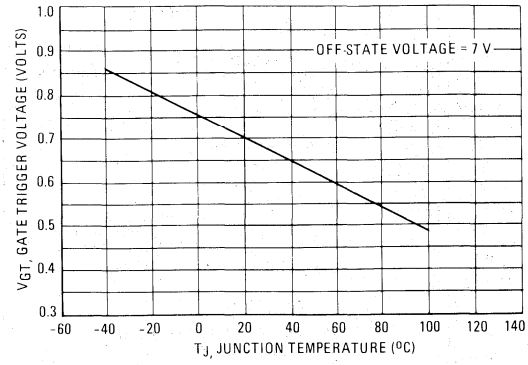


FIGURE 2 – GATE TRIGGER VOLTAGE



MAXIMUM ALLOWABLE NON-REPETITIVE SURGE CURRENT

FIGURE 3 – 60 Hz SURGES

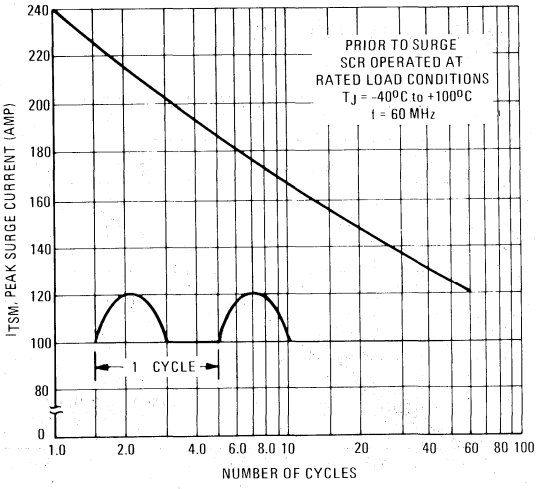


FIGURE 4 – SUB-CYCLE SURGES

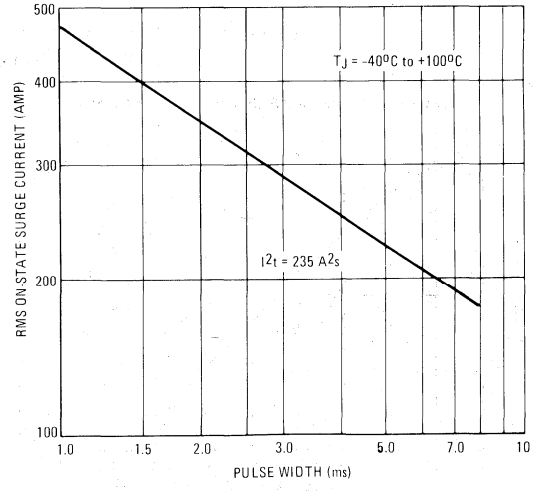


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FIGURE 5 – GATE TRIGGER CHARACTERISTICS

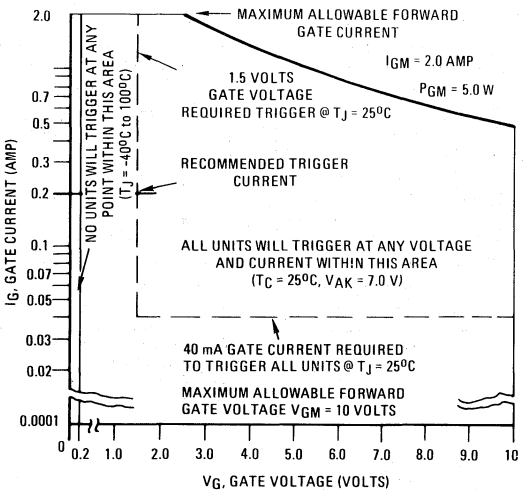
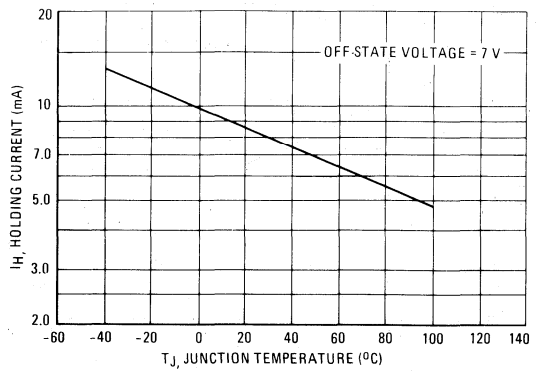


FIGURE 6 – EFFECT OF TEMPERATURE ON TYPICAL HOLDING CURRENT



DERATING AND DISSIPATION FOR RESISTIVE AND INDUCTIVE LOADS (f = 60 to 400 Hz, SINE WAVE)

3

FIGURE 7 – AVERAGE CURRENT DERATING

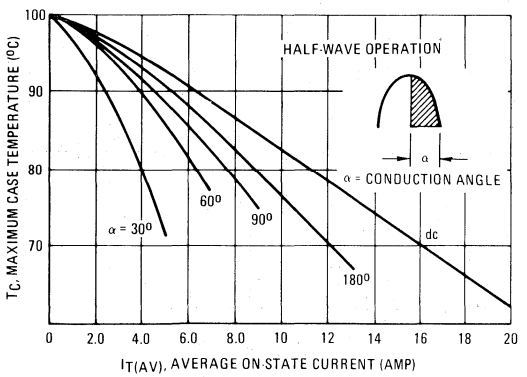


FIGURE 8 – ON-STATE POWER DISSIPATION

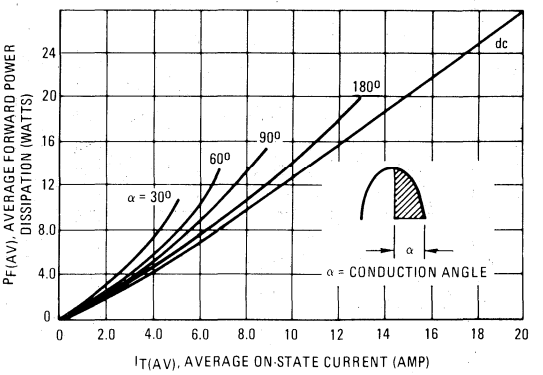


FIGURE 9 – ON-STATE CHARACTERISTICS

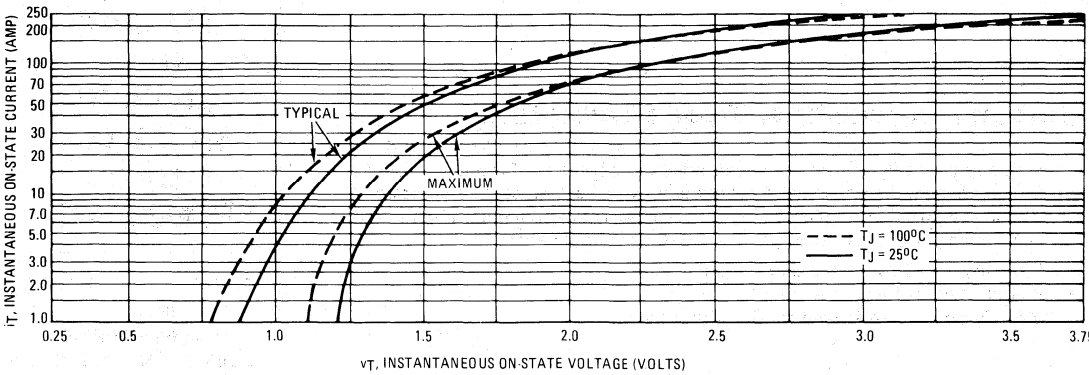


FIGURE 10 – TYPICAL THERMAL RESISTANCE OF PLATES

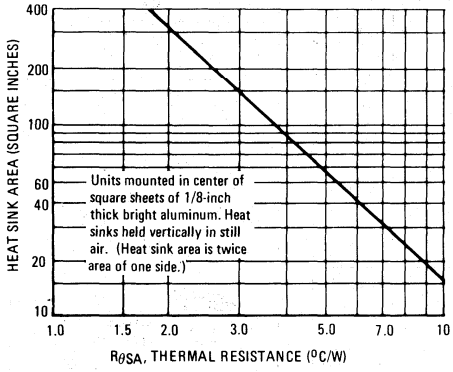
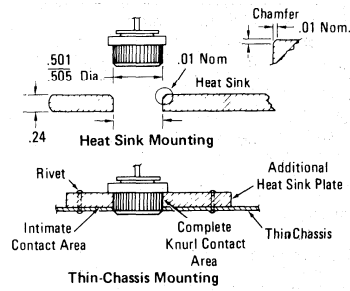


FIGURE 11 – MOUNTING DETAILS FOR PRESSFIT THYRISTORS



3

Triacs

Silicon Bidirectional Triode Thyristors

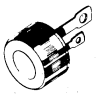
2N5444 thru 2N5446 MAC5441 thru MAC5443

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire
- Isolated Stud for Ease of Assembly
- Gate Triggering Guaranteed In All 4 Quadrants

TRIACs
40 AMPERES RMS
200 thru 600 VOLTS

CASE 263-04
STYLE 2
2N5444 thru 2N5446



CASE 174-04
(TO-203AA)
STYLE 3
MAC5441 thru MAC5443

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −65 to +110°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
*Peak Principal Voltage MAC5441, 2N5444 MAC5442, 2N5445 MAC5443, 2N5446		 200 400 600	
*RMS On-State Current (T_C per Figure 2) (T_C = +100°C) Full Sine Wave, 50 to 60 Hz	$I_{T(RMS)}$	 40 20	Amps
*Peak Non-Repetitive Surge Current (One Full Cycle of surge current at 60 Hz, preceded and followed by a 40 A RMS current, T_C = 100°C)	I_{TSM}	300	Amps
*Peak Gate Power (Pulse Width = 10 μs Max)	P_{GM}	40	Watts
*Average Gate Power	$P_{G(AV)}$	0.75	Watt
*Peak Gate Current (10 μs Max)	I_{GM}	4	Amps
*Peak Gate Voltage	V_{GM}	30	Volts
*Operating Junction Temperature Range	T_J	−65 to +110	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
*Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case MAC5441, MAC5442, MAC5443 2N5444, 2N5445, 2N5446	$R_{\theta JC}$	 0.8 0.9	°C/W

*Indicates JEDEC Registered Data.

3

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and either polarity of MT2 to MT1 voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	0.5	4	mA
*Peak On-State Voltage (I_{TM} = 56 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.65	1.85	Volts
Gate Trigger Current (Continuous dc), Note 1 (Main Terminal Voltage = 12 Vdc, R_L = 50 Ohms)	I_{GT}				mA
MT2(+), G(+)		—	—	70	
MT2(+), G(−)		—	—	70	
MT2(−), G(−)		—	—	70	
MT2(−), G(+)		—	—	100	
*MT2(+), G(+); MT2(−), G(−) T_C = −65°C		—	—	125	
*MT2(+), G(−); MT2(−), G(+) T_C = −65°C		—	—	240	
*Gate Trigger Voltage (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 50 Ohms)	V_{GT}				Volts
MT2(+), G(+)					
MT2(+), G(−)		—	—	2	
MT2(−), G(−)		—	—	2	
MT2(−), G(+)		—	—	2	
*All Quadrants, T_C = −65°C		—	—	2.5	
*Main Terminal Voltage = Rated V_{DRM} = R_L = 10 k ohms,		—	—	3.4	
T_C = 100°C		0.2	—	—	
*Holding Current (Main Terminal Voltage = 12 Vdc, Gate Open) (Initiating Current = 150 mA) T_C = 25°C	I_H	—	—	70	mA
*T_C = −65°C		—	—	100	
*Turn-On Time (Main Terminal Voltage = Rated V_{DRM}, I_{TM} = 56 A, Gate Source Voltage = 12 V, R_S = 12 Ohms, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1	2	μs
*Critical Rate-of-Rise of Commutation Voltage (Rated V_{DRM}, I_{TM} = 40 A, Commutating di/dt = 22 A/ms, gate energized)	dv/dt(c)				V/μs
T_C = 70°C MAC5441, MAC5442, MAC5443		5	30	—	
= 65°C 2N5444, 2N5445, 2N5446		5	30	—	
		5	30	—	
Critical Rate-of-Rise of Off State Voltage (Rated V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 110°C)	dv/dt				V/μs
MAC5441, 2N5444		50			
MAC5442, 2N5445		30			
MAC5443, 2N5446		20			

*Indicates JEDEC Registered Data for 2N5541 thru 2N5446.

Note 1. All voltage polarities referenced to main terminal 1.

3

FIGURE 1 – ON-STATE POWER DISSIPATION

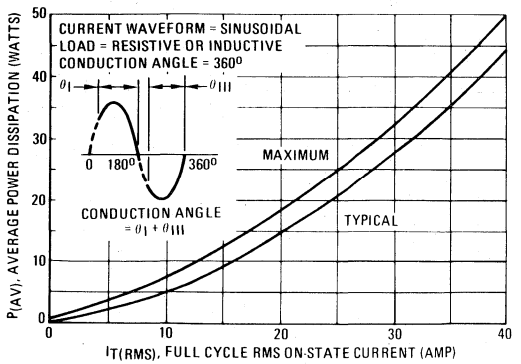


FIGURE 2 – RMS CURRENT DERATING

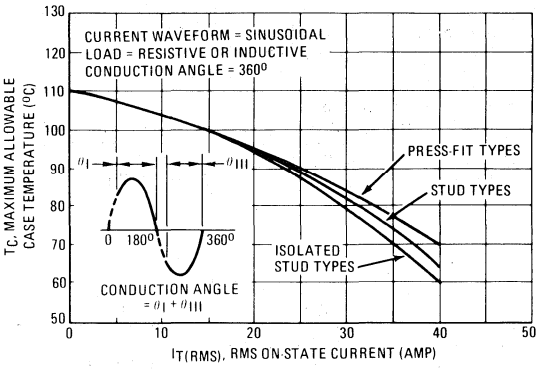


FIGURE 3 – TYPICAL GATE TRIGGER VOLTAGE

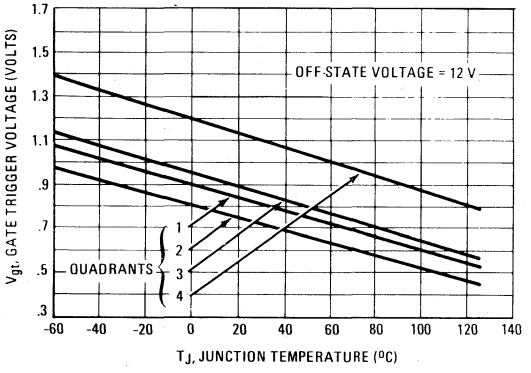


FIGURE 4 – TYPICAL GATE TRIGGER CURRENT

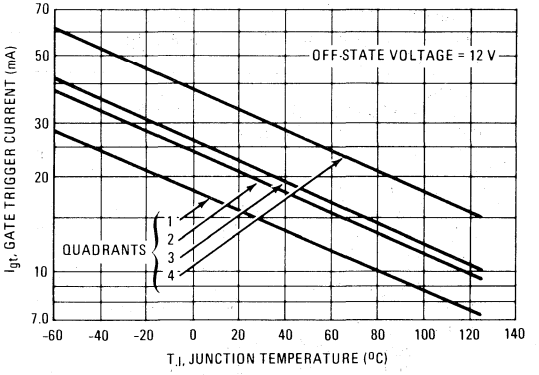


FIGURE 5 – TYPICAL THERMAL RESPONSE

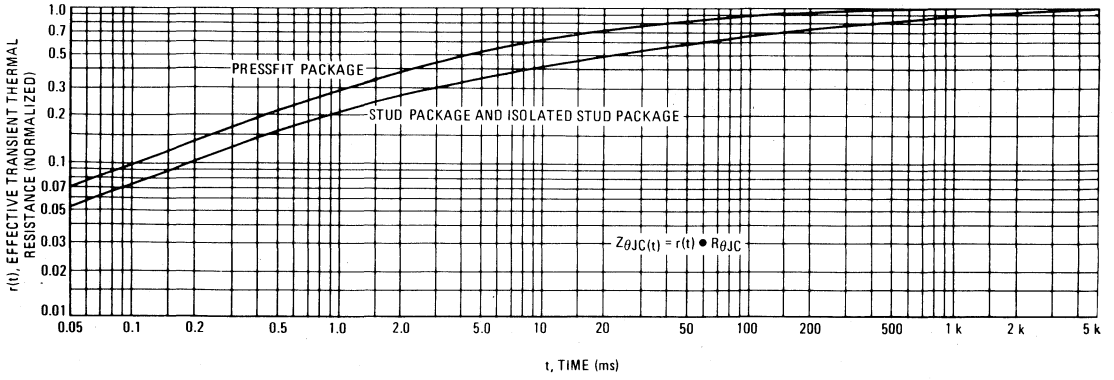


FIGURE 6 – ON-STATE CHARACTERISTICS

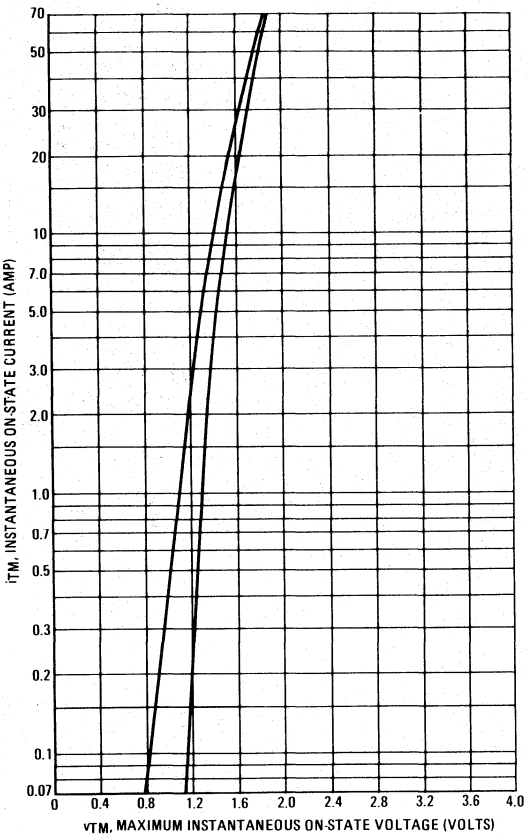


FIGURE 7 – TYPICAL HOLDING CURRENT

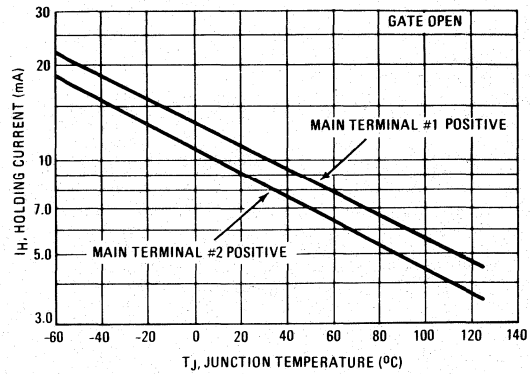
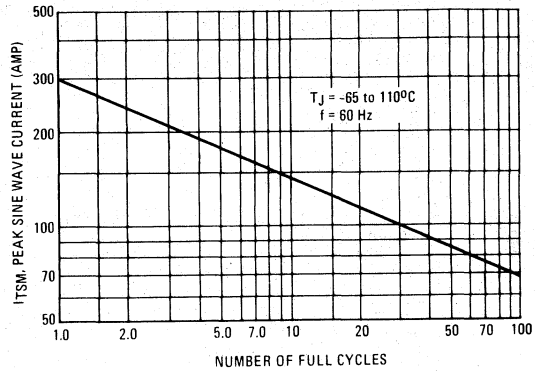


FIGURE 8 – MAXIMUM ALLOWABLE SURGE CURRENT



3

Triacs

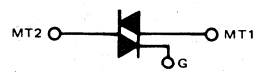
Bidirectional Triode Thyristors

2N5567
2N5568
MAC5569
MAC5570
MAC4111M
MAC4121
Series

. . . designed primarily for industrial and military applications for the fullwave control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Pressfit, Stud and Isolated Stud Packages
- Gate Triggering Guaranteed In All 4 Quadrants

TRIACs
10 AMPERES RMS
200 thru 800 VOLTS



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −65 to +100°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
2N5567, MAC5569, MAC4121B		200	
2N5568, MAC5570, MAC4121D		400	
MAC4111M, MAC4121M		600	
MAC4121N		800	
*Peak Gate Voltage	V_{GM}	20	Volts
*RMS On-State Current (T_C = −65 to +85°C)	$I_{T(RMS)}$	10	Amps
(T_C = +90°C) Full cycle, Sine Wave, 50 to 60 Hz		6.7	
*Peak Non-Repetitive Surge Current (One Full cycle of surge current at 60 Hz, preceded and followed by rated current, T_C = 85°C)	I_{TSM}	100	Amps
Circuit Fusing Considerations (t − 8.3 ms)	I^2t	40	A^2s
Peak Gate Power *(T_C = 85°C, Pulse Width = 1 μs)	P_{GM}	16	Watts
*Average Gate Power (T_C = 85°C, Pulse Width = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
*Operating Junction Temperature Range	T_J	−65 to +100	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$		°C/W
Stud and Pressfit		1	
Isolated Stud		1.1	

*Indicates JEDEC Registered Data.

CASE 174-04
(TO-203AA)
STYLE 3
2N5567
2N5568

CASE 263-04
STYLE 2
MAC5569
MAC5570
MAC4111M

CASE 311-02
STYLE 2
MAC4121 SERIES

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	2	mA
*Peak On-State Voltage (I_{TM} = 14.2 A Peak, Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.65	Volts
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 12 Ohms)	I_{GT}				mA
MT2(+), G(+); MT2(−), G(−)		—	—	25	
MT2(+), G(−); MT2(−), G(+)		—	—	40	
*MT2(+), G(+); MT2(−), G(−), T_C = −65°C		—	—	100	
*MT2(+), G(−); MT2(−), G(+), T_C = −65°C		—	—	150	
Gate Trigger Voltage (Continuous dc) (All Quadrants) (V_D = 12 Vdc, R_L = 12 Ohms T_C = 25°C	V_{GT}	—	—	2.5	Volts
*T_C = −65°C		—	—	4	
(V_D = Rated V_{DRM}, R_L = 125 Ω) T_C = 100°C		0.2	—	—	
Holding Current (V_D = 12 Vdc, Gate Open) T_C = 25°C	I_H	—	—	30	mA
*T_C = −65°C		—	—	200	
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 15 A Peak, I_{GT} = 160 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs) MT2(+), G(+); MT2(−), G(−)	t_{gt}	—	1	2.5	μs
*Critical Rate-of-Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 14.2 A Peak, Commutating di/dt = 5.4 A/ms, gate unenergized) T_C = 85°C	dv/dt(c)	2	10	—	V/μs
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 100°C)	dv/dt				V/μs
*2N5567, *MAC5569, MAC4121B		30	150	—	
*2N5568, *MAC5570, MAC4121D		20	100	—	
MAC4111M, MAC4121M		10	75	—	
MAC4121N		10	75	—	

*Indicates JEDEC Registered Data.
Note 1. All Voltage polarities referenced to main terminal 1.

3

FIGURE 1 — RMS CURRENT DERATING
(Isolated Stud)

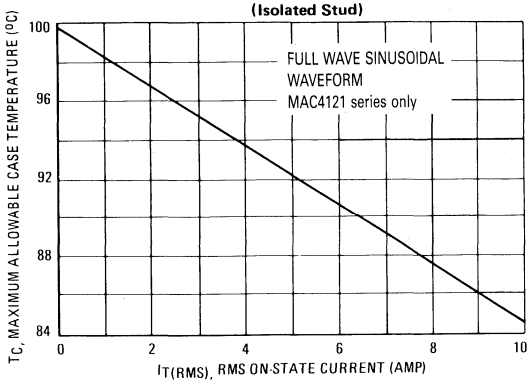


FIGURE 2 — RMS CURRENT DERATING
(Pressfit and Stud)

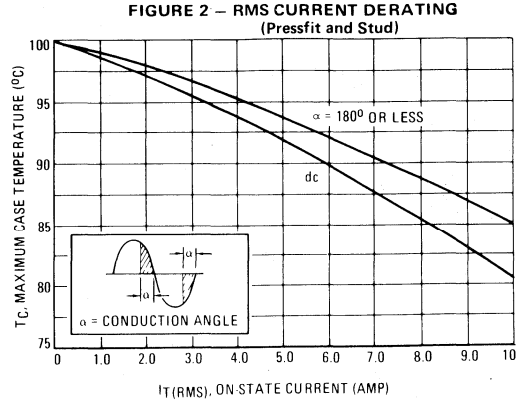


FIGURE 3 – POWER DISSIPATION
(Isolated Stud)

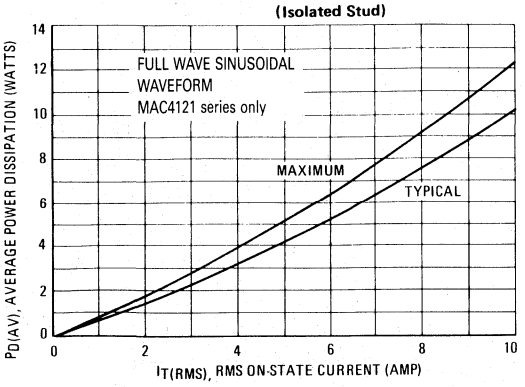


FIGURE 4 – POWER DISSIPATION
(Pressfit and Stud)

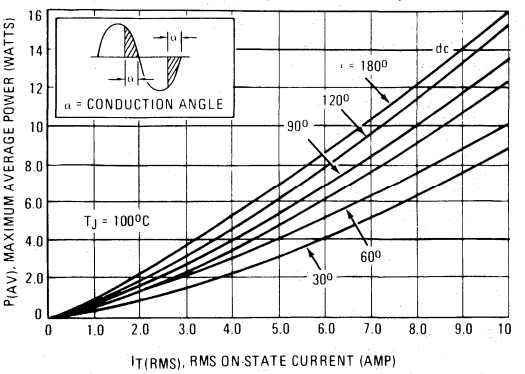


FIGURE 5 – TYPICAL GATE TRIGGER VOLTAGE

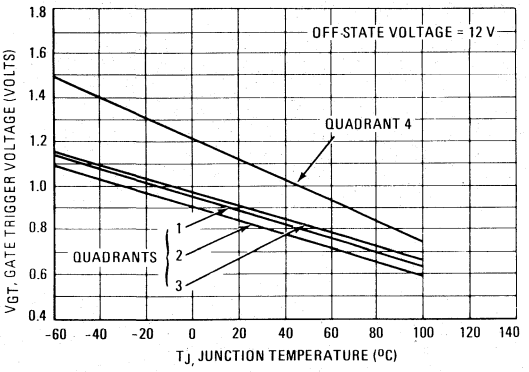


FIGURE 6 – TYPICAL GATE TRIGGER CURRENT

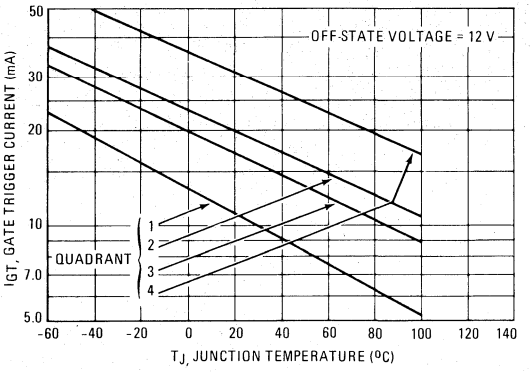


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FIGURE 7 – ON-STATE CHARACTERISTICS

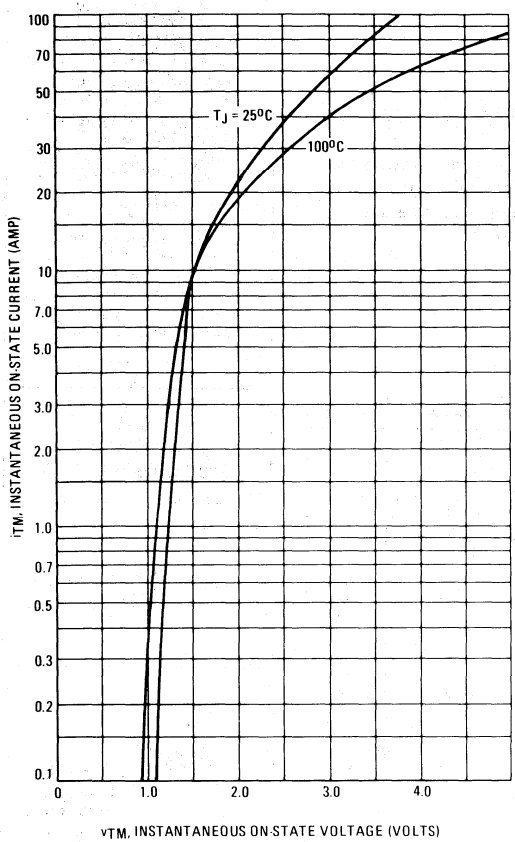


FIGURE 8 – TYPICAL HOLDING CURRENT

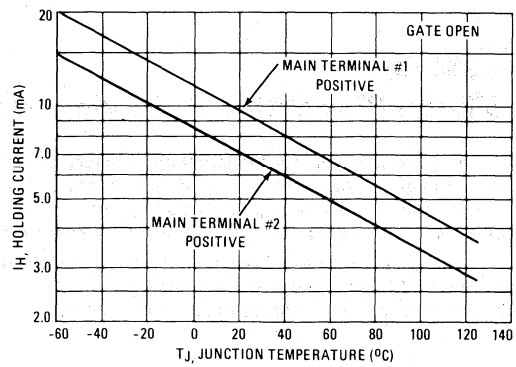


FIGURE 9 – MAXIMUM NON-REPETITIVE SURGE CURRENT

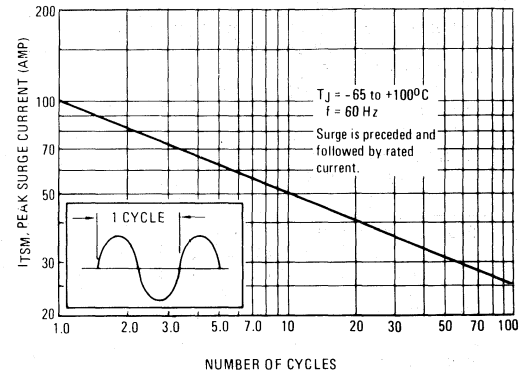
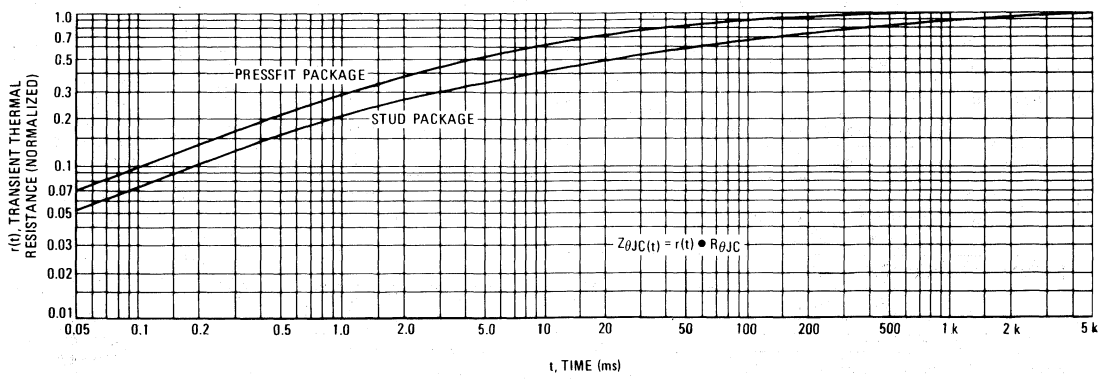


FIGURE 10 – TYPICAL THERMAL RESPONSE



3

Triacs

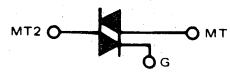
Silicon Bidirectional Triode Thyristors

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Pressfit, Stud and Isolated Stud Packages
- Gate Triggering Guaranteed In All 4 Quadrants

2N5571
2N5572
MAC5573
MAC5574
2N6145
thru
2N6147
T4100M
MAC4110M

TRIACs
15 AMPERES RMS
200 thru 600 VOLTS



MAXIMUM RATINGS

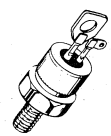
Rating	Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −65 to +100°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
2N5571, MAC5573, 2N6145		200	
2N5572, MAC5574, 2N6146		400	
T4100M, MAC4110M, 2N6147		600	
*Peak Gate Voltage	V_{GM}	20	Volts
*RMS On-State Current (T_C = −65 to +80°C)	$I_{T(RMS)}$	15	Amps
(T_C = +85°C)		10	
*Peak Non-Repetitive Surge Current (One Full cycle of surge current at 60 Hz, preceded and followed by rated current, T_C = 85°C)	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power *(T_C = 80°C, Pulse Width = 1 μs) 2N5571, 2N5572, MAC5573, MAC5574 T4100M, MAC4110M	P_{GM}	16	Watts
*(T_C = 80°C, Pulse Width = 2 μs) 2N6145 thru 2N6147		20	
*Average Gate Power (T_C = 80°C, Pulse Width = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
*Peak Gate Current	I_{GM}	2	Amps
*Operating Junction Temperature Range	T_J	−65 to +100	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque MAC5573, MAC5574, MAC4110M *2N6145, 2N6146, 2N6147	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	1	°C/W

*Indicates JEDEC Registered Data.

CASE 174-04
(TO-203AA)
STYLE 3
2N5571
2N5572
T4100M



CASE 263-04
STYLE 2
MAC5573
MAC5574
MAC4110M

CASE 311-02
STYLE 2
2N6145
2N6146
2N6147

3

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	2	mA
*Peak On-State Voltage (I_{TM} = 21 A Peak, Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.8	Volts
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 30 Ohms)	I_{GT}				mA
MT2(+), G(+); MT2(−), G(−)		—	—	50	
MT2(+), G(−); MT2(−), G(+)		—	—	80	
*MT2(+), G(+); MT2(−), G(−), T_C = −65°C		—	—	150	
*MT2(+), G(−); MT2(−), G(+), T_C = −65°C		—	—	200	
Gate Trigger Voltage (Continuous dc) (All Quadrants) (V_D = 12 Vdc, R_L = 30 Ohms) T_C = 25°C	V_{GT}	—	—	2.5	Volts
*T_C = −65°C		—	—	4	
*(V_D = Rated V_{DRM}, R_L = 10 k ohms, T_C = +100°C)		0.2	—	—	
Holding Current (V_D − 12 Vdc, Gate Open) (Initiating Current = 500 mA) T_C = 25°C	I_H	—	—	75	mA
*T_C = −65°C		—	—	300	
Gate Controlled Turn-On Time (Rated V_{DRM}, I_{TM} = 21 A Peak, I_{GT} = 160 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1	2	μs
*Critical Rate-of-Rise of Commutation Voltage (Rated V_{DRM}, I_{TM} = 21 A Peak, Commutating di/dt = 8 A/ms, gate unenergized T_C = 80°C 2N5571, 2N5572, T4100M, MAC4110M	dv/dt(c)	2	10	—	V/μs
T_C = 75°C 2N6145 thru 2N6147		2	10	—	
Critical Rate-of-Rise of Off-State Voltage (Rated V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 100°C) *2N5571, MAC5573, 2N6145	dv/dt	30	150	—	V/μs
*2N5572, MAC5574, 2N6146		20	100	—	
T4100M, MAC4110M, 2N6147		10	75	—	

*Indicates JEDEC Registered Data.
Note 1. All Voltage polarities referenced to main terminal 1.

3

FIGURE 1 — RMS CURRENT DERATING

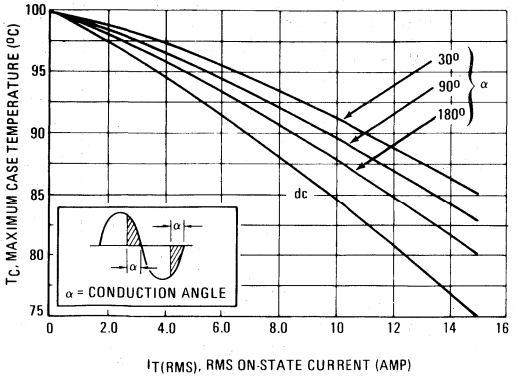


FIGURE 2 — ON-STATE POWER DISSIPATION

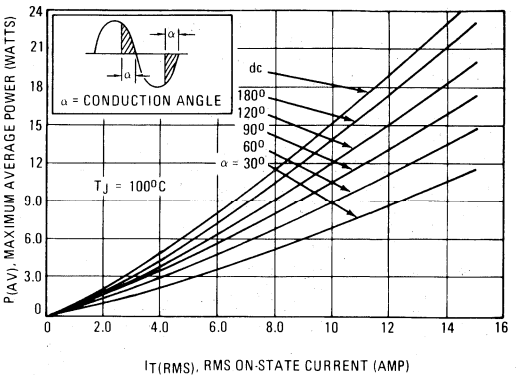


FIGURE 3 — TYPICAL GATE TRIGGER VOLTAGE

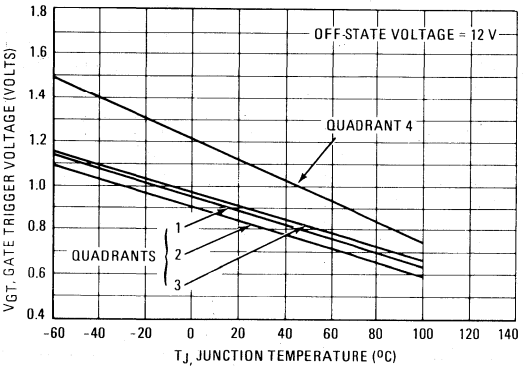


FIGURE 4 — TYPICAL GATE TRIGGER CURRENT

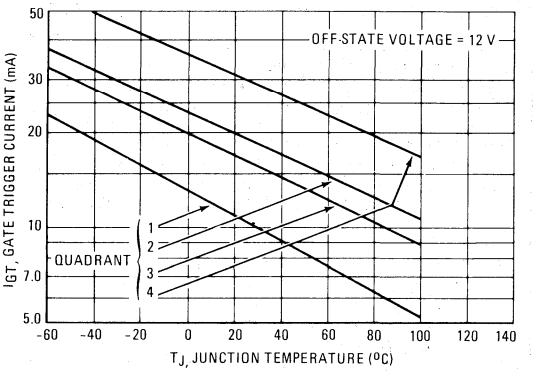


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FIGURE 5 – MAXIMUM ON-STATE CHARACTERISTICS

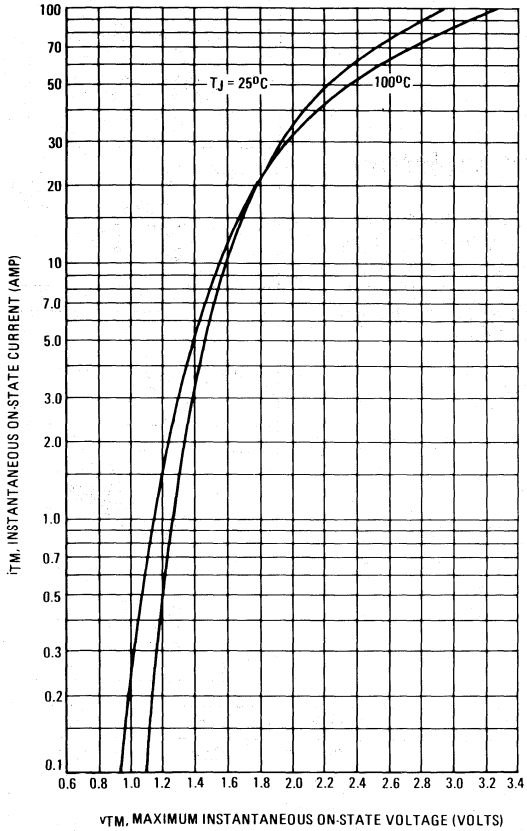


FIGURE 6 – TYPICAL HOLDING CURRENT

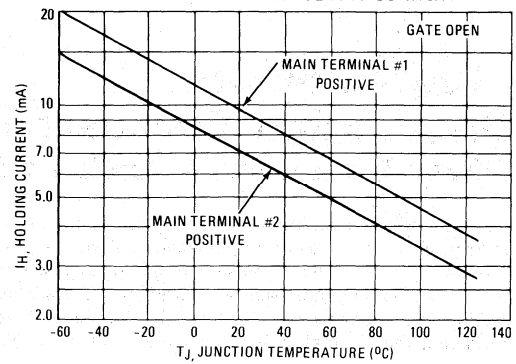


FIGURE 7 – MAXIMUM NON-REPETITIVE SURGE CURRENT

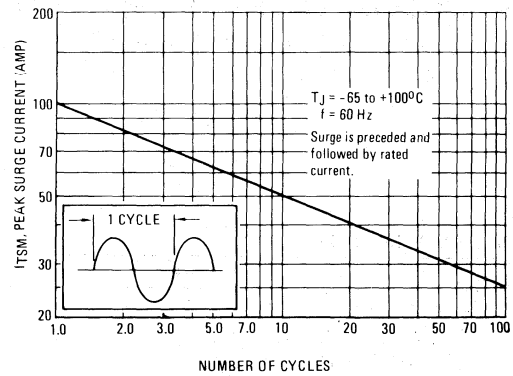
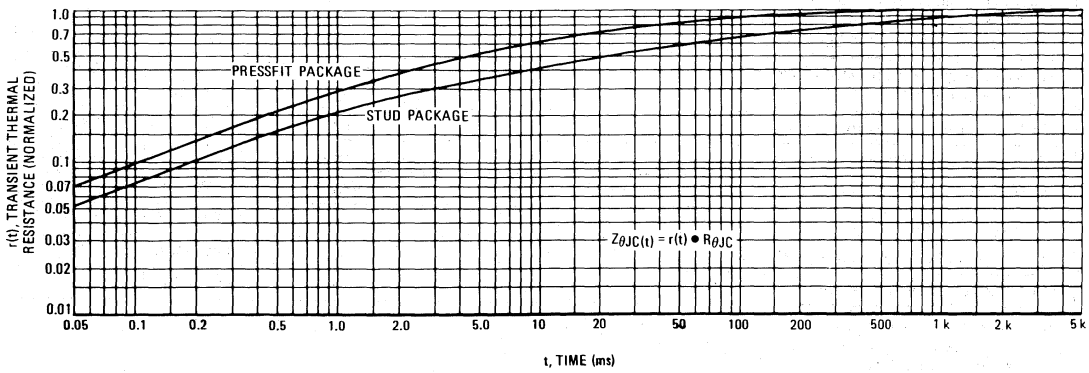


FIGURE 8 – TYPICAL THERMAL RESPONSE



3

Programmable Unijunction Transistors

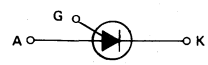
Silicon Programmable Unijunction Transistors

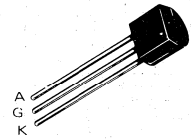
2N6027
2N6028

PUTs
40 VOLTS
300 mW

. . . designed to enable the engineer to "program" unijunction characteristics such as R_{BB}, η, I_V, and I_P by merely selecting two resistor values. Application includes thyristor-trigger, oscillator, pulse and timing circuits. These devices may also be used in special thyristor applications due to the availability of an anode gate. Supplied in an inexpensive TO-92 plastic package for high-volume requirements, this package is readily adaptable for use in automatic insertion equipment.

- Programmable — R_{BB}, η, I_V and I_P.
- Low On-State Voltage — 1.5 Volts Maximum @ I_F = 50 mA
- Low Gate to Anode Leakage Current — 10 nA Maximum
- High Peak Output Voltage — 11 Volts Typical
- Low Offset Voltage — 0.35 Volt Typical (R_G = 10 k ohms)





CASE 29-04
(TO-226AA)
STYLE 16

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Power Dissipation Derate Above 25°C	P_F $1/\theta_{JA}$	300 4	mW mW/°C
*DC Forward Anode Current Derate Above 25°C	I_T	150 2.67	mA mA/°C
*DC Gate Current	I_G	±50	mA
Repetitive Peak Forward Current 100 μs Pulse Width, 1% Duty Cycle *20 μs Pulse Width, 1% Duty Cycle	I_{TRM}	 1 2	Amps
Non-Repetitive Peak Forward Current 10 μs Pulse Width	I_{TSM}	5	Amps
*Gate to Cathode Forward Voltage	V_{GKF}	40	Volts
*Gate to Cathode Reverse Voltage	V_{GKR}	5	Volts
*Gate to Anode Reverse Voltage	V_{GAR}	40	Volts
*Anode to Cathode Voltage, Note 1	V_{AK}	±40	Volts
Operating Junction Temperature Range	T_J	−50 to +100	°C
*Storage Temperature Range	T_{stg}	−55 to +150	°C

*Indicates JEDEC Registered Data

Note 1. Anode positive, R_{GA} = 1000 ohms
Anode negative, R_{GA} = open

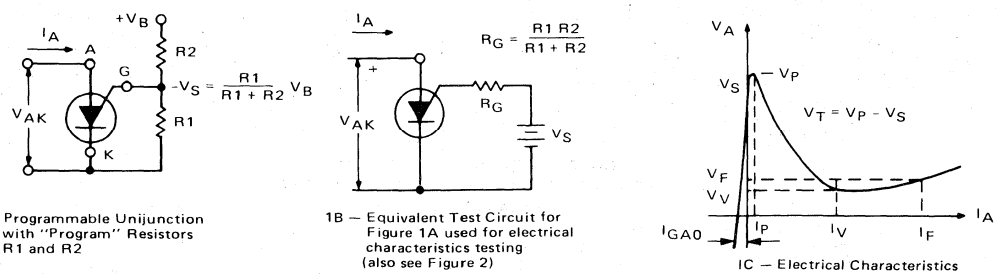
ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted.)

Characteristic		Fig. No.	Symbol	Min	Typ	Max	Unit
*Peak Current		2,9,11	I_P				μA
(V_S = 10 Vdc, R_G = 1 MΩ)	2N6027			—	1.25	2	
	2N6028			—	0.08	0.15	
(V_S = 10 Vdc, R_G = 10 k ohms)	2N6027			—	4	5	
	2N6028			—	0.70	1	
*Offset Voltage		1	V_T				Volts
(V_S = 10 Vdc, R_G = 1 MΩ)	2N6027			0.2	0.70	1.6	
	2N6028			0.2	0.50	0.6	
(V_S = 10 Vdc, R_G = 10 k ohms)	(Both Types)			0.2	0.35	0.6	
*Valley Current		1,4,5	I_V				μA
(V_S = 10 Vdc, R_G = 1 MΩ)	2N6027			—	18	50	
	2N6028			—	18	25	
(V_S = 10 Vdc, R_G = 10 k ohms)	2N6027			70	150	—	
	2N6028			25	150	—	
(V_S = 10 Vdc, R_G = 200 ohms)	2N6027			1.5	—	—	mA
	2N6028			1	—	—	
*Gate to Anode Leakage Current		—	I_{GAO}				nAdc
(V_S = 40 Vdc, T_A = 25°C, Cathode Open)				—	1	10	
(V_S = 40 Vdc, T_A = 75°C, Cathode Open)				—	3	—	
Gate to Cathode Leakage Current (V_S = 40 Vdc, Anode to Cathode Shorted)		—	I_{GKS}	—	5	50	nAdc
*Forward Voltage (I_F = 50 mA Peak)		1,6	V_F	—	0.8	1.5	Volts
*Peak Output Voltage (V_G = 20 Vdc, C_C = 0.2 μF)		3,7	V_o	6	11	—	Volt
Pulse Voltage Rise Time (V_B = 20 Vdc, C_C = 0.2 μF)		3	t_r	—	40	80	ns

*Indicates JEDEC Registered Data.

3

FIGURE 1 – ELECTRICAL CHARACTERIZATION



1A – Programmable Unijunction with "Program" Resistors R1 and R2

1B – Equivalent Test Circuit for Figure 1A used for electrical characteristics testing (also see Figure 2)

1C – Electrical Characteristics

FIGURE 2 – PEAK CURRENT (I_P) TEST CIRCUIT

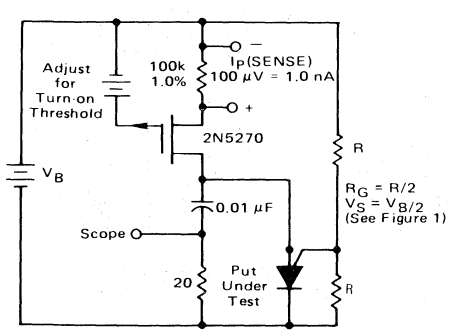
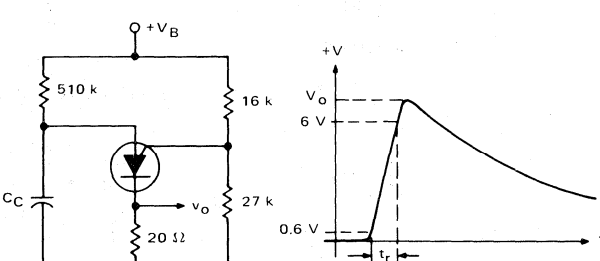


FIGURE 3 – V_o AND t_r TEST CIRCUIT



TYPICAL VALLEY CURRENT BEHAVIOR

FIGURE 4 – EFFECT OF SUPPLY VOLTAGE

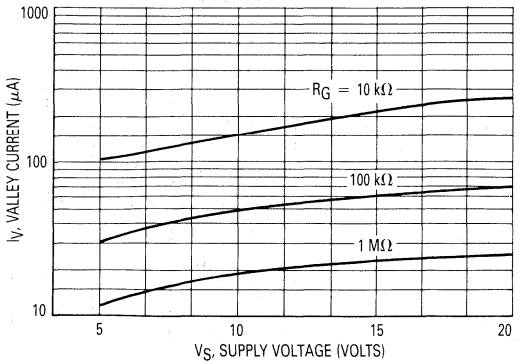


FIGURE 5 – EFFECT OF TEMPERATURE

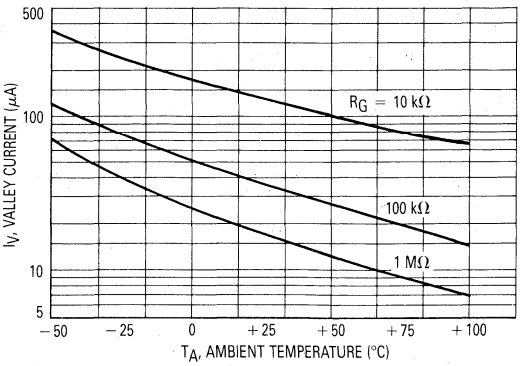


FIGURE 6 – FORWARD VOLTAGE

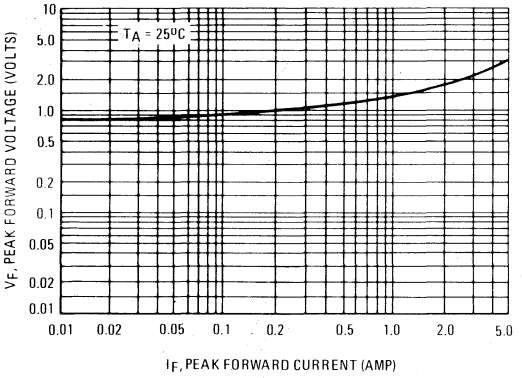


FIGURE 7 – PEAK OUTPUT VOLTAGE

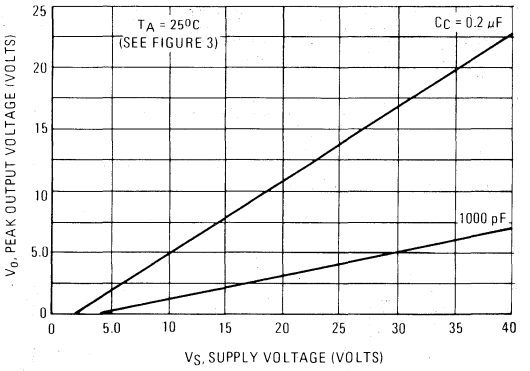
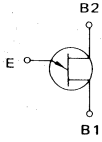
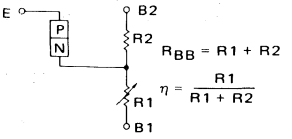


FIGURE 8 – STANDARD UNIJUNCTION COMPARED TO PROGRAMMABLE UNIJUNCTION

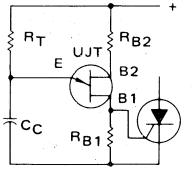
STANDARD UNIJUNCTION



Circuit Symbol

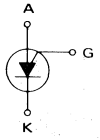


Equivalent Circuit

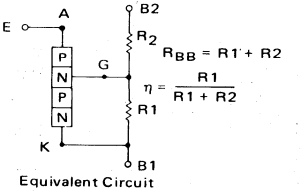


Typical Application

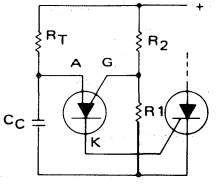
PROGRAMMABLE UNIJUNCTION



Circuit Symbol



Equivalent Circuit with External "Program" Resistors R1 and R2



Typical Application

TYPICAL PEAK CURRENT BEHAVIOR

2N6027

FIGURE 9 – EFFECT OF SUPPLY VOLTAGE AND R_G

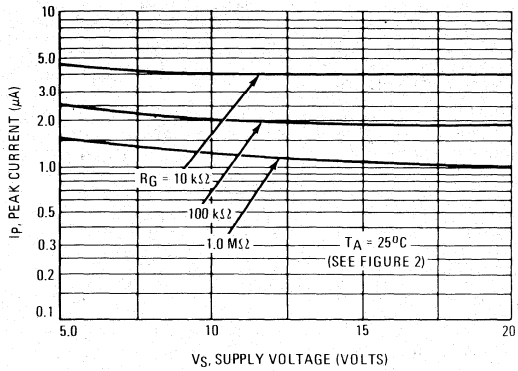
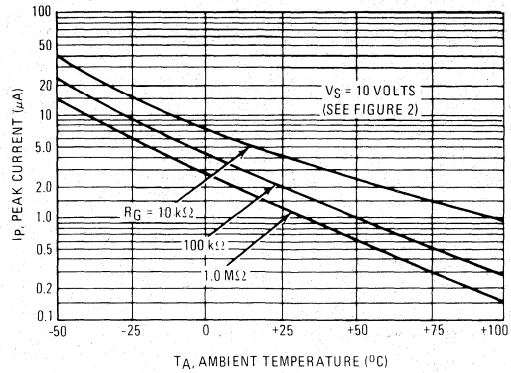


FIGURE 10 – EFFECT OF TEMPERATURE AND R_G



2N6028

FIGURE 11 – EFFECT OF SUPPLY VOLTAGE AND R_G

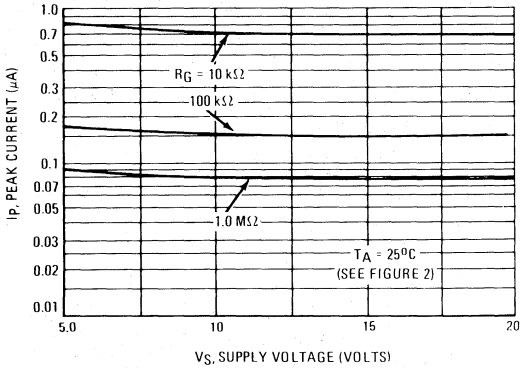
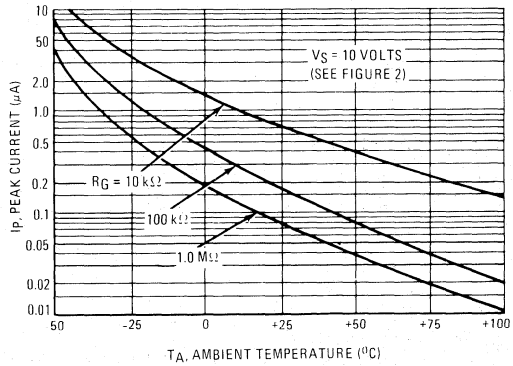


FIGURE 12 – EFFECT OF TEMPERATURE AND R_G



3

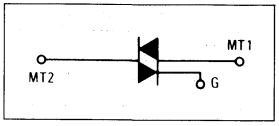
Sensitive Gate Triacs
Silicon Bidirectional Thyristors

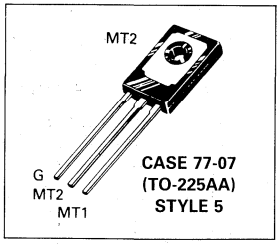
2N6071,A,B
2N6073,A,B
2N6075,A,B

TRIACs
4 AMPERES RMS
200 thru 600 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Sensitive Gate Triggering (A and B versions) Uniquely Compatible for Direct Coupling to TTL, HTL, CMOS and Operational Amplifier Integrated Circuit Logic Functions
- Gate Triggering 2 Mode — 2N6071, 2N6073, 2N6075
  4 Mode — 2N6071,A,B, 2N6073,A,B, 2N6075,A,B
- Blocking Voltages to 600 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability





3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Repetitive Peak Off-State Voltage, Note 1 (T_J = 110°C) 2N6071,A,B / 2N6073,A,B / 2N6075,A,B	V_{DRM}	200 / 400 / 600	Volts
*On-State Current RMS (T_C = 85°C)	$I_{T(RMS)}$	4	Amps
*Peak Surge Current (One Full cycle, 60 Hz, T_J = −40 to +110°C)	I_{TSM}	30	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	3.7	A^2s
*Peak Gate Power	P_{GM}	10	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt
*Peak Gate Voltage	V_{GM}	5	Volts

*Indicates JEDEC Registered Data.

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Operating Junction Temperature Range	T_J	−40 to +110	°C
*Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque (6-32 Screw), Note 1	—	8	in. lb.

*Indicates JEDEC Registered Data.

Note 1. Torque rating applies with use of compression washer (B52200F006). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Main terminal 2 and heatsink contact pad are common.

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +200°C, for 10 seconds. Consult factory for lead bending options.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	3.5	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	75	°C/W

*Indicates JEDEC Registered Data.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	2	mA
*On-State Voltage (Either Direction) (I_{TM} = 6 A Peak)	V_{TM}	—	—	2	Volts
*Peak Gate Trigger Voltage (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms, T_J = −40°C)	V_{GT}				Volts
MT2(+), G(+); MT2(−), G(−) All Types		—	1.4	2.5	
MT2(+), G(−); MT2(−), G(+) 2N6071,A,B, 2N6073,A,B 2N6075,A,B		—	1.4	2.5	
(Main Terminal Voltage = Rated V_{DRM}, R_L = 10 k ohms, T_J = 110°C)					
MT2(+), G(+); MT2(−), G(−) All Types		0.2	—	—	
MT2(+), G(−); MT2(−), G(+) 2N6071,A,B, 2N6073,A,B, 2N6075,A,B		0.2	—	—	
*Holding Current (Either Direction) (Main Terminal Voltage = 12 Vdc, Gate Open, T_J = −40°C)	I_H				mA
(Initiating Current = 1 Adc) 2N6071, 2N6073, 2N6075		—	—	70	
2N6071A,B, 2N6073A,B, 2N6075A,B		—	—	30	
(T_J = 25°C) 2N6071, 2N6073, 2N6075		—	—	30	
2N6071A,B, 2N6073A,B, 2N6075A,B		—	—	15	
Turn-On Time (Either Direction) (I_{TM} = 14 Adc, I_{GT} = 100 mAdc)	t_{on}	—	1.5	—	μs
Blocking Voltage Application Rate at Commutation @ V_{DRM}, T_J = 85°C, Gate Open	dv/dt	—	5	—	V/μs

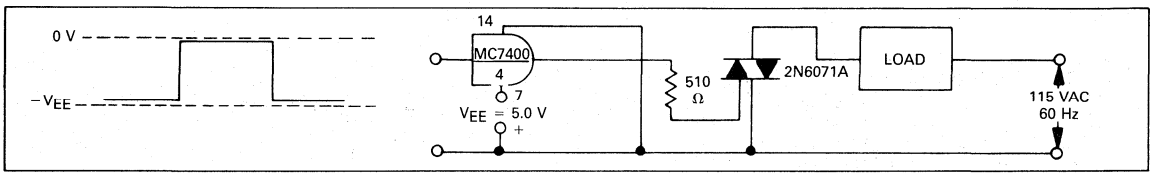
*Indicates JEDEC Registered Data.

3

			QUADRANT (See Definition Below)			
	Type	I_{GT} @ T_J	I mA	II mA	III mA	IV mA
Gate Trigger Current (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 100 ohms)	2N6071 2N6073 2N6075	+25°C	30	—	30	—
		−40°C	60	—	60	—
Maximum Value	2N6071A 2N6073A 2N6075A	+25°C	5	5	5	10
		−40°C	20	20	20	30
	2N6071B 2N6073B 2N6075B	+25°C	3	3	3	5
		−40°C	15	15	15	20

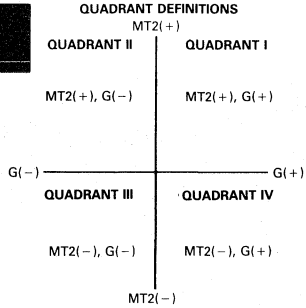
*Indicates JEDEC Registered Data.

SAMPLE APPLICATION:
TTL-SENSITIVE GATE 4 AMPERE TRIAC
TRIGGERS IN MODES II AND III



3

QUADRANT DEFINITIONS



Trigger devices are recommended for gating on Triacs. They provide:

1. Consistent predictable turn-on points.
2. Simplified circuitry.
3. Fast turn-on time for cooler, more efficient and reliable operation.

For 2N6071, 2N6073, 2N6075

ELECTRICAL CHARACTERISTICS of RECOMMENDED BIDIRECTIONAL SWITCHES

Usage	General		Lamp Dimmer
Part Number	MBS4991	MBS4993	MBS100
V_S	6.0 – 10 V	7.5 – 9.0 V	3.0 – 5.0 V
I_S	350 μA Max	250 μA Max	100 – 400 μA
$V_{S1} - V_{S2}$	0.5 V Max	0.2 V Max	0.35 V Max
Temperature Coefficient	0.02%/°C Typ		

See AN-526 for Theory and Characteristics of Silicon Bidirectional Switches.

SENSITIVE GATE LOGIC REFERENCE

IC Logic Functions	Firing Quadrant			
	I	II	III	IV
TTL		2N6071A Series	2N6071A Series	
HTL		2N6071A Series	2N6071A Series	
CMOS (NAND)	2N6071B Series			2N6071B Series
CMOS (Buffer)		2N6071B Series	2N6071B Series	
Operational Amplifier	2N6071A Series			2N6071A Series
Zero Voltage Switch		2N6071A Series	2N6071A Series	

FIGURE 1 – AVERAGE CURRENT DERATING

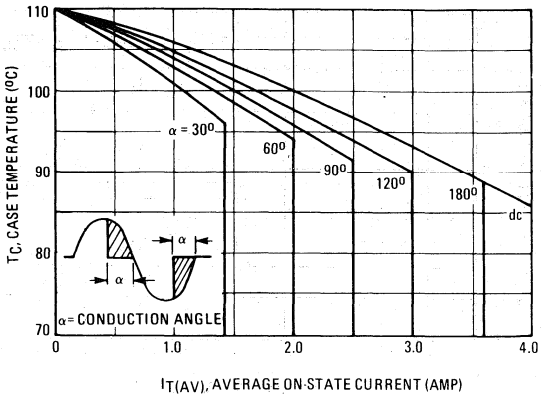


FIGURE 2 – RMS CURRENT DERATING

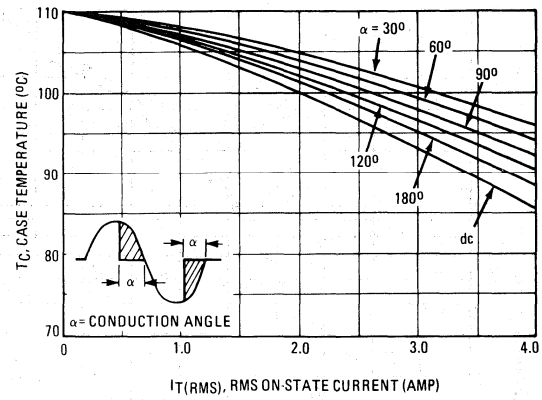


FIGURE 3 – POWER DISSIPATION

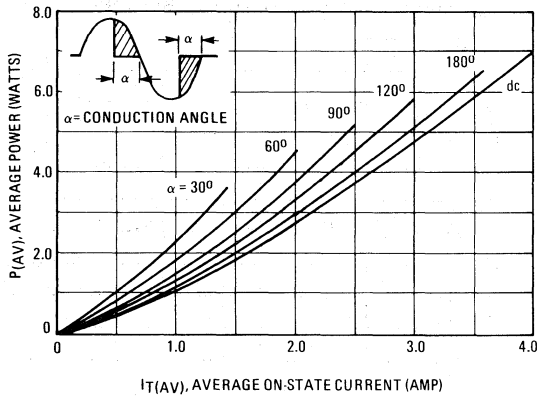


FIGURE 4 – POWER DISSIPATION

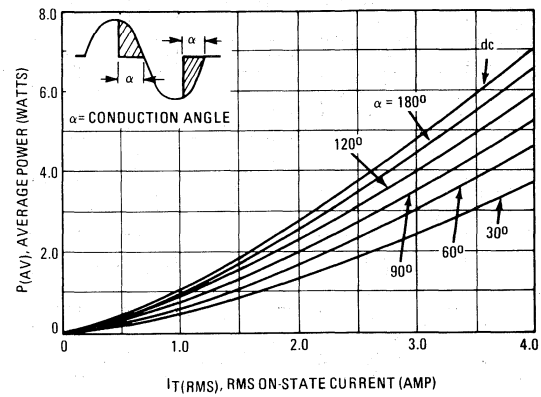


FIGURE 5 – TYPICAL GATE-TRIGGER VOLTAGE

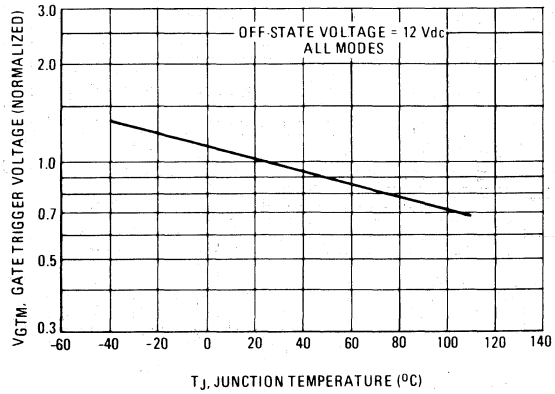


FIGURE 6 – TYPICAL GATE-TRIGGER CURRENT

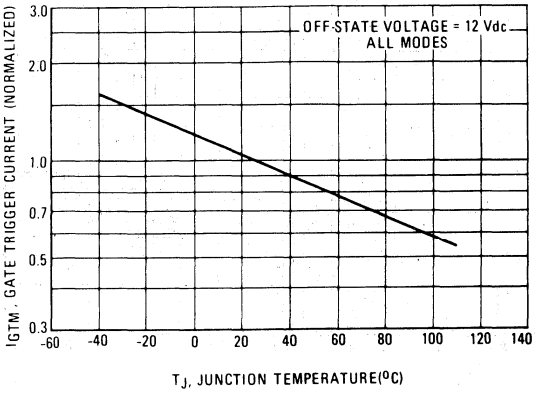


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FIGURE 7 – MAXIMUM ON-STATE CHARACTERISTICS

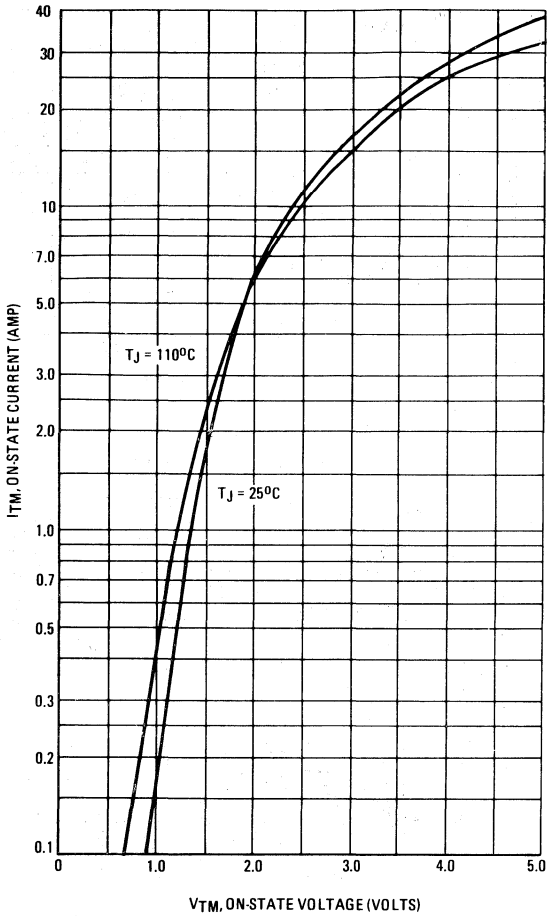


FIGURE 8 – TYPICAL HOLDING CURRENT

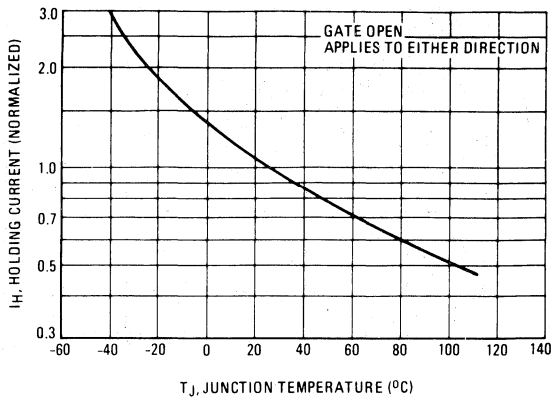


FIGURE 9 – MAXIMUM ALLOWABLE SURGE CURRENT

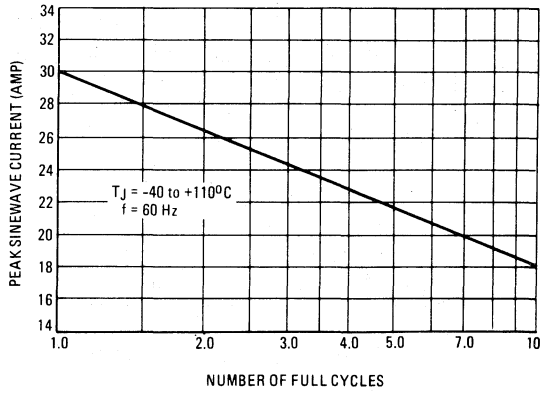
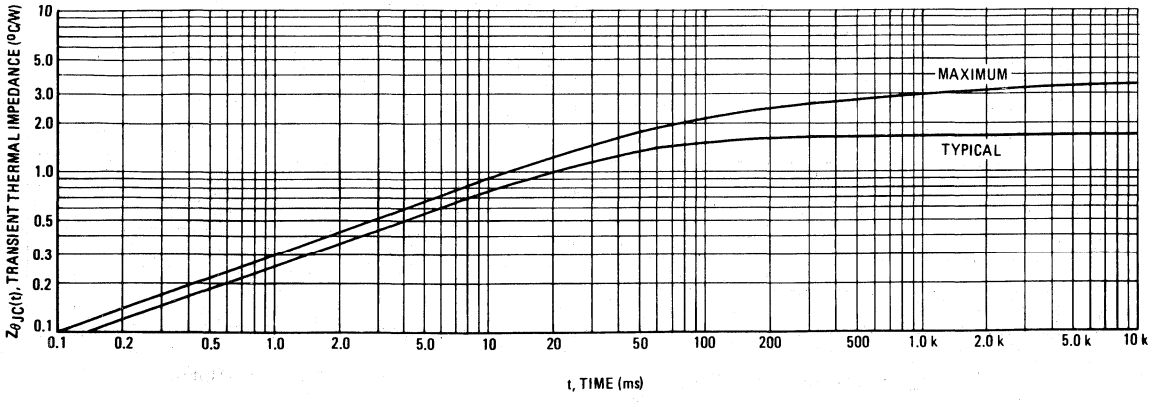


FIGURE 10 – THERMAL RESPONSE



2N6145–47
(See 2N5571)

Triacs

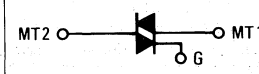
Silicon Bidirectional Triode Thyristors

2N6157 thru 2N6165

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire
- Isolated Stud for Ease of Assembly
- Gate Triggering Guaranteed In All 4 Quadrants

TRIACs
30 AMPERES RMS
200 thru 600 VOLTS



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −65 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open *Peak Principal Voltage 2N6157, 2N6160, 2N6163 2N6158, 2N6161, 2N6164 2N6159, 2N6162, 2N6165	V_{DRM}	200 400 600	Volts
*Peak Gate Voltage	V_{GM}	10	Volts
*RMS On-State Current (T_C = −65 to +85°C) (T_C = +100°C) Full Sine Wave, 50 to 60 Hz	$I_{T(RMS)}$	30 20	Amps
*Peak Non-Repetitive Surge Current (One Full Cycle of surge current at 60 Hz, preceded and followed by a 30 A RMS current, T_C = 85°C)	I_{TSM}	250	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	260	A^2s
*Peak Gate Power (T_J = +80°C, Pulse Width = 2 μs)	P_{GM}	20	Watts
*Average Gate Power (T_J = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
*Peak Gate Current	I_{GM}	2	Amps
*Operating Junction Temperature Range	T_J	−65 to +125	°C
*Storage Temperature Range	T_{stg}	−65 to +150	°C
*Stud Torque 2N6160 thru 2N6165	—	30	in. lb.

CASE 174-04
(TO-203AA)
STYLE 3
2N6157-59

CASE 263-04
STYLE 2
2N6160-62

CASE 311-02
STYLE 2
2N6163-65

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	1	°C/W

*Indicates JEDEC Registered Data.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
*Peak On-State Voltage (Either Direction) (I_{TM} = 42 A Peak, Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.5	2	Volts
Gate Trigger Current (Continuous dc), Note 1 (Main Terminal Voltage = 12 Vdc, R_L = 50 Ohms)	I_{GT}				mA
MT2(+), G(+)		—	15	60	
MT2(+), G(−)		—	20	70	
MT2(−), G(−)		—	20	70	
MT2(−), G(+)		—	30	100	
*MT2(+), G(+); MT2(−), G(−) T_C = −65°C		—	—	200	
*MT2(+), G(−); MT2(−), G(+) T_C = −65°C		—	—	250	
Gate Trigger Voltage (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 50 Ohms)	V_{GT}				Volts
MT2(+), G(+)		—	0.8	2	
MT2(+), G(−)		—	0.7	2.1	
MT2(−), G(−)		—	0.85	2.1	
MT2(−), G(+)		—	1.1	2.5	
*All Quadrants, T_C = −65°C		—	—	3.4	
*Main Terminal Voltage = Rated V_{DRM}, R_L = 10 k ohms, T_J = +125°C		0.2	—	—	
Holding Current (Main Terminal Voltage = 12 Vdc, Gate Open) (Initiating Current = 500 mA)	I_H				mA
MT2(+)		—	8	70	
MT2(−)		—	10	80	
*Either Direction, T_C = −65°C		—	—	200	
*Turn-On Time (Main Terminal Voltage = Rated V_{DRM}, I_{TM} = 42 A, Gate Source Voltage = 12 V, R_S = 50 Ohms, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1	2	μs
Blocking Voltage Application Rate at Commutation, f = 60 Hz, T_C = 85°C On-State Conditions: (I_{TM} = 42 A, Pulse Width = 4 ms, di/dt = 17.5 A/ms) Off-State Conditions: (Main Terminal Voltage = Rated V_{DRM} (200 μs min), Gate Source Voltage = 0 V, R_S = 50 Ω)	dv/dt(c)	—	5	—	V/μs

*Indicates JEDEC Registered Data.
Note 1. All voltage polarities referenced to main terminal 1.

FIGURE 1 — RMS CURRENT DERATING

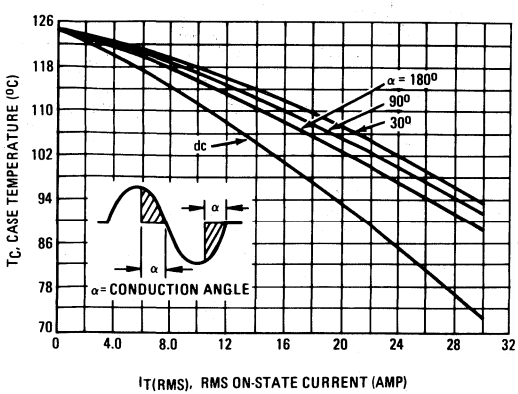


FIGURE 2 — POWER DISSIPATION

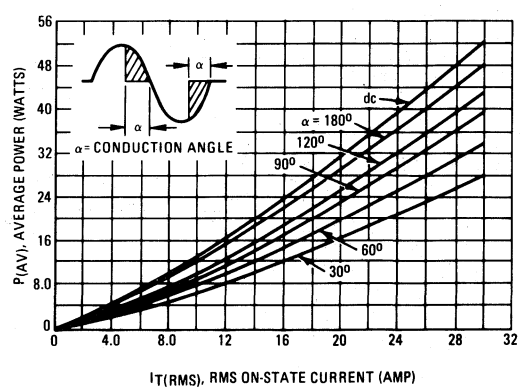


FIGURE 3 – TYPICAL GATE TRIGGER VOLTAGE

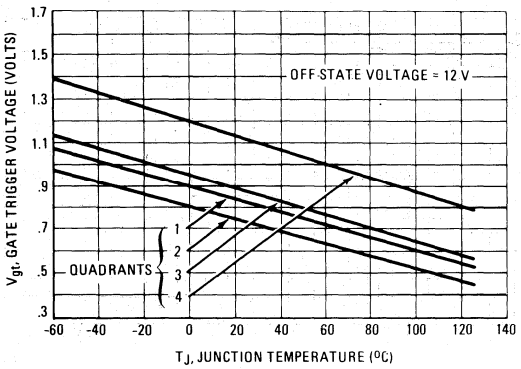


FIGURE 4 – TYPICAL GATE TRIGGER CURRENT

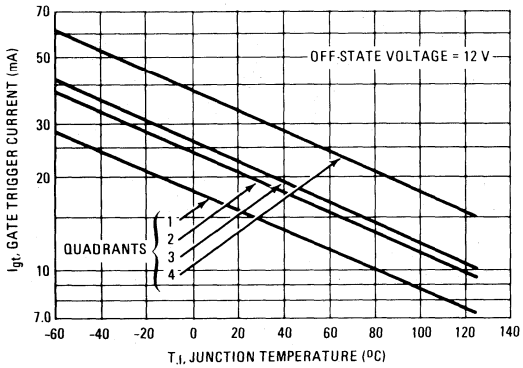


FIGURE 5 – MAXIMUM ON-STATE CHARACTERISTICS

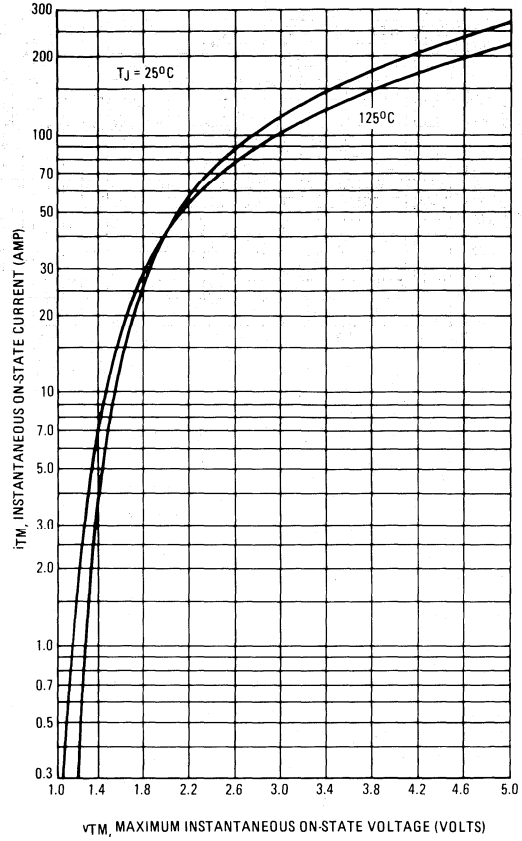


FIGURE 6 – TYPICAL HOLDING CURRENT

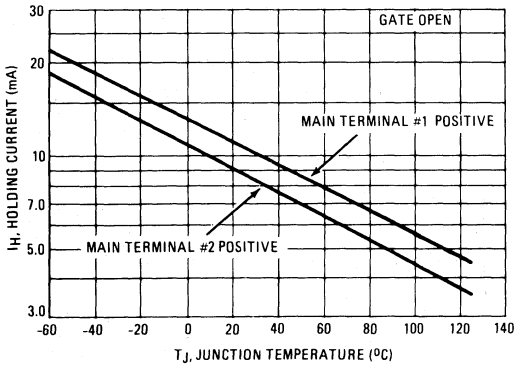


FIGURE 7 – MAXIMUM ALLOWABLE SURGE CURRENT

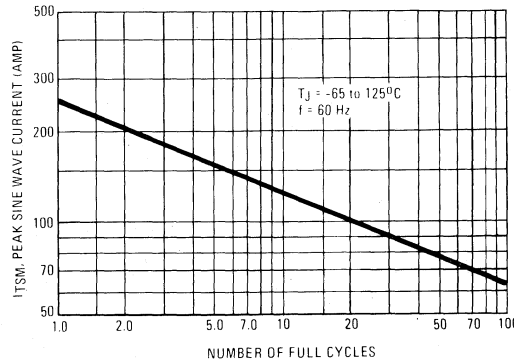


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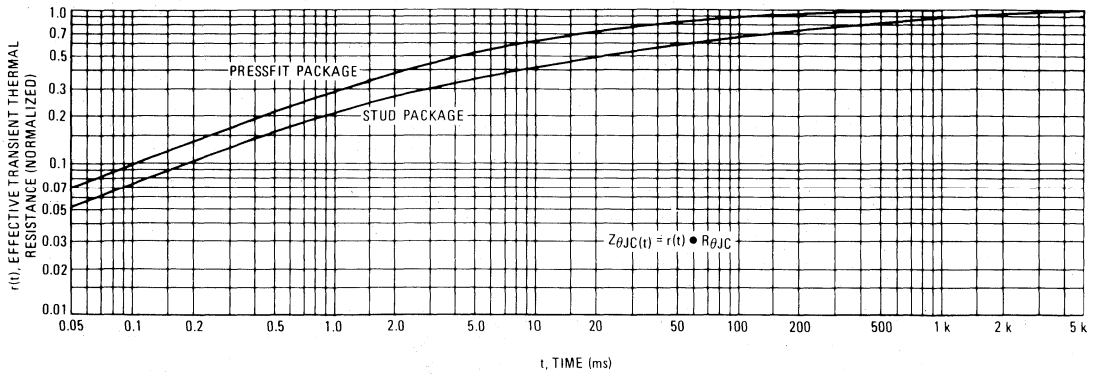
FIGURE 8 – TYPICAL THERMAL RESPONSE



3

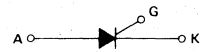
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

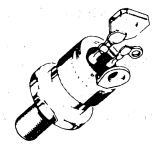
2N6167 thru 2N6170

SCRs
20 AMPERES RMS
100 thru 600 VOLTS

. . . designed for industrial and consumer applications such as power supplies; battery chargers; temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 240 Amps
- Rugged Construction in Isolated Stud Package





CASE 311-02
STYLE 1

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Repetitive Forward and Reverse Blocking Voltage, Note 1 (T_J = −40°C to +100°C) 2N6167	V_{DRM} or V_{RRM}	100	Volts
2N6168		200	
2N6169		400	
2N6170		600	
*Non-Repetitive Peak Reverse Blocking Voltage (t ≤ 5 ms) 2N6167	V_{RSM}	150	Volts
2N6168		250	
2N6169		450	
2N6170		650	
*Average Forward Current (T_C = −40 to +65°C)	$I_{T(AV)}$	13	Amps
(+85°C)		6.5	
*Peak Surge Current (One cycle, 60 Hz) (T_C = +65°C)	I_{TSM}	240	Amps
(1.5 ms pulse @ T_J = 100°C) Preceded and followed by no current or Voltage		560	
Circuit Fusing (t = 8.3 ms)	I^2t	240	A^2s
*Peak Gate Power	P_{GM}	5	Watts
*Average Gate Power	$P_{G(AV)}$	0.5	Watt

*Indicates JEDEC Registered Data. (cont.)

Note 1. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode. Devices should not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
*Peak Forward Gate Current	I_{GFM}	2	Amps
*Operating Junction Temperature Range	T_J	−40 to +100	°C
*Storage Temperature Range	T_{stg}	−40 to +150	°C
*Stud Torque	—	30	in. lb.

***THERMAL CHARACTERISTICS**

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.5	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open, T_C = 100°C)	I_{DRM}, I_{RRM}				mA
2N6167		—	1	2	
2N6168		—	1	2.5	
2N6169		—	1	3	
2N6170		—	1	4	
(Rated V_{DRM} or V_{RRM}, gate open, T_C = 25°C) All Devices		—	—	10	μA
*Peak Forward "On" Voltage (I_{TM} = 41 A Peak)	V_{TM}	—	1.5	1.7	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 24 Ω) *T_C = −40°C	I_{GT}	—	—	75	mA
T_C = 25°C		—	2.1	40	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 24 Ω) *T_C = −40°C	V_{GT}	—	0.8	2.5	Volts
T_C = 25°C		—	0.63	1.6	
Holding Current (V_D = 12 V, gate open, I_T = 200 mA) *T_C = −40°C	I_H	—	—	90	mA
T_C = 25°C		—	3.5	50	
*Turn-On Time (t_d + t_r) (I_{TM} = 41 Adc, V_D = Rated V_{DRM}, I_{GT} = 200 mAdc, Rise Time ≤ 0.05 μs, Pulse Width = 10 μs)	t_{on}	—	—	1	μs
Turn-Off Time (I_{TM} = 10 A, I_R = 10 A)	t_{off}	—	25	—	μs
(I_{TM} = 10 A, I_R = 10 A, T_J = 100°C)		—	40	—	
Forward Voltage Application Rate (T_J = 100°C, V_D = Rated V_{DRM})	dv/dt	—	50	—	V/μs

*Indicates JEDEC Registered Data.

FIGURE 1 — AVERAGE CURRENT DERATING

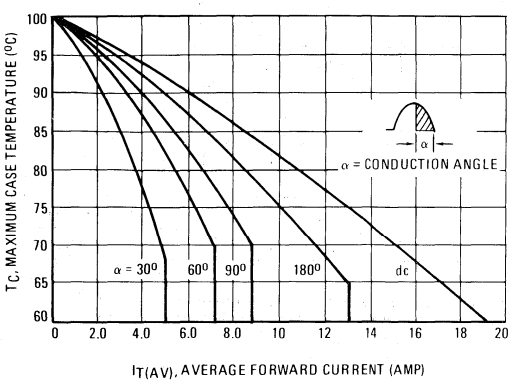


FIGURE 2 — POWER DISSIPATION

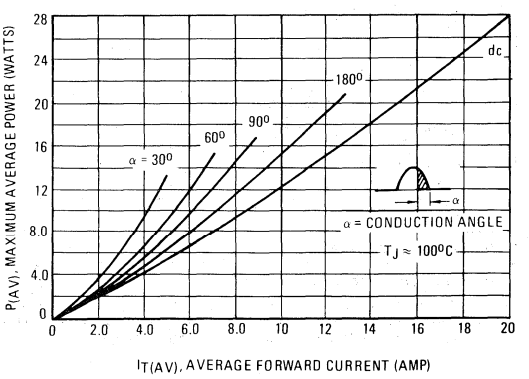


FIGURE 3 – MAXIMUM ON-STATE CHARACTERISTICS

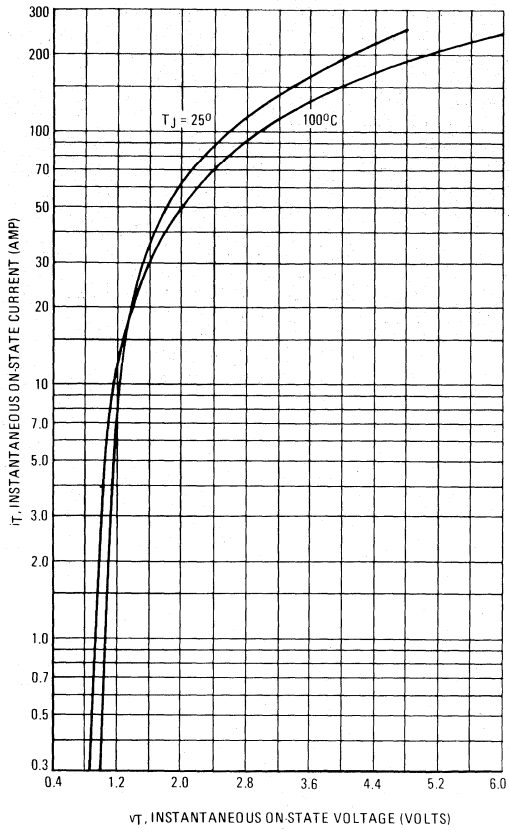


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

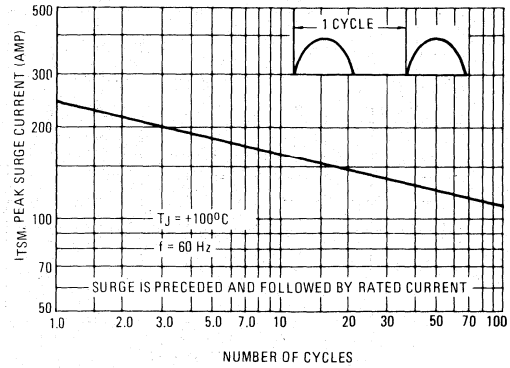


FIGURE 5 – CHARACTERISTICS AND SYMBOLS

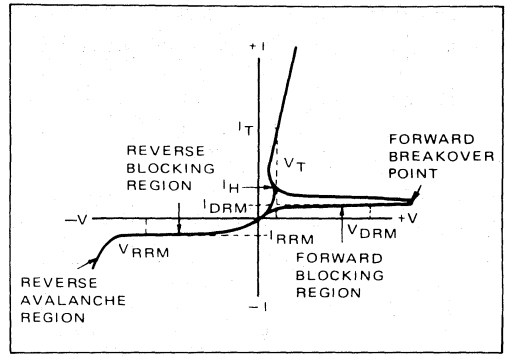


FIGURE 6 – THERMAL RESPONSE

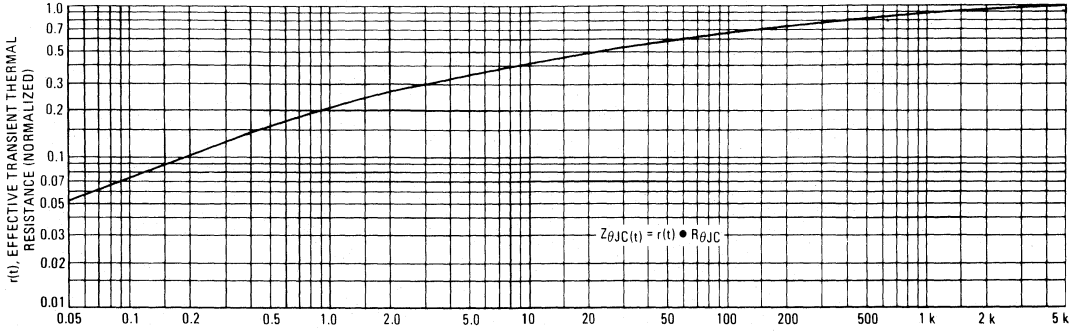


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FIGURE 7 – TYPICAL GATE TRIGGER CURRENT

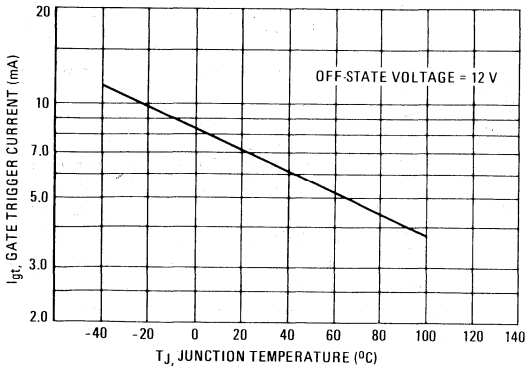


FIGURE 8 – TYPICAL GATE TRIGGER VOLTAGE

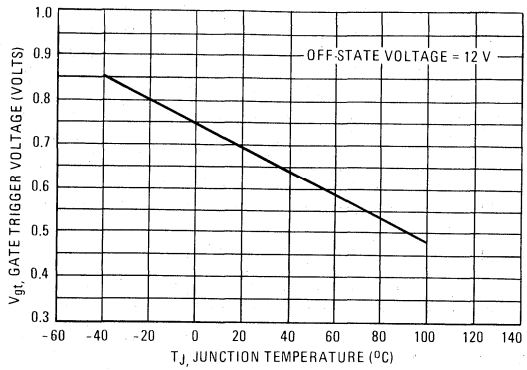
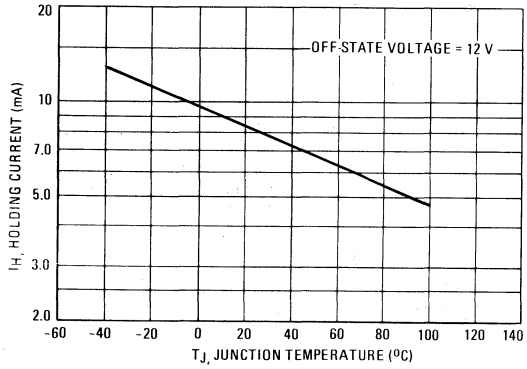


FIGURE 9 – TYPICAL HOLDING CURRENT



3

2N6171–74
(See 2N3870)

Silicon Controlled Rectifiers

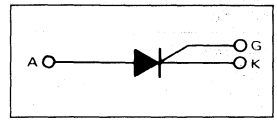
Reverse Blocking Triode Thyristors

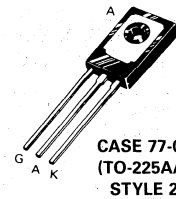
2N6237 thru 2N6241

SCRs
4 AMPERES RMS
50 thru 600 VOLTS

. . . PNPN devices designed for high volume consumer applications such as temperature, light, and speed control; process and remote control, and warning systems where reliability of operation is important.

- Passivated Surface for Reliability and Uniformity
- Power Rated at Economical Prices
- Practical Level Triggering and Holding Characteristics
- Flat, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Recommended Electrical Replacement for C106





CASE 77-07
(TO-225AA)
STYLE 2

MAXIMUM RATINGS (T_C = 110°C unless otherwise noted.)

Rating		Symbol	Value	Unit
*Repetitive Peak Forward and Reverse Blocking Voltage, Note 1 (1/2 Sine Wave) (R_{GK} = 1000 ohms, T_C = −40 to +110°C)	2N6237 2N6238 2N6239 2N6240 2N6241	V_{DRM} or V_{RRM}	50 100 200 400 600	Volts
*Non-Repetitive Peak Reverse Blocking Voltage (1/2 Sine Wave, R_{GK} = 1000 ohms, T_C = −40 to +110°C)	2N6237 2N6238 2N6239 2N6240 2N6241	V_{RSM}	100 150 250 450 650	Volts
*Average On-State Current (T_C = −40 to +90°C) (T_C = +100°C)		$I_{T(AV)}$	2.6 1.6	Amps
*Surge On-State Current (1/2 Sine Wave, 60 Hz, T_C = +90°C) (1/2 Sine Wave, 1.5 ms, T_C = +90°C)		I_{TSM}	25 35	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	2.6	A^2s
*Peak Gate Power (Pulse Width = 10 μs, T_C = 90°C)		P_{GM}	0.5	Watt

*Indicates JEDEC Registered Data.

(continued)

Note 1. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode. Devices should not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued (T_C = 110°C unless otherwise noted.)

Rating	Symbol	Value	Unit
*Average Gate Power (t = 8.3 ms, T_C = 90°C)	$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current	I_{GM}	0.2	Amp
Peak Reverse Gate Voltage	V_{RGM}	6	Volts
*Operating Junction Temperature Range	T_J	−40 to +110	°C
*Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque, Note 1	—	6	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Min	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	—	3	°C/W
Thermal Resistance Junction to Ambient	$R_{\theta JA}$	—	75	°C/W

*Indicates JEDEC Registered Data.

ELECTRICAL CHARACTERISTICS (T_C = 25°C and R_{GK} = 1000 ohms unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C T_C = 110°C	I_{DRM}, I_{RRM}	— —	— —	10 200	μA μA
*Peak Forward "On" Voltage (I_{TM} = 8.2 A Peak, Pulse Width = 1 to 2 ms, 2% Duty Cycle)	V_{TM}	—	—	2.2	Volts
Gate Trigger Current (Continuous dc), Note 2 (V_{AK} = 12 Vdc, R_L = 24 Ohms) *(V_{AK} = 12 Vdc, R_L = 24 Ohms, T_C = −40°C)	I_{GT}	— —	— —	200 500	μA
Gate Trigger Voltage (Continuous dc) (Source Voltage = 12 V, R_S = 50 Ohms) *(V_{AK} = 12 Vdc, R_L = 24 Ohms, T_C = −40°C)	V_{GT}	—	—	1	Volts
Gate Non-Trigger Voltage (V_{AK} = Rated V_{DRM}, R_L = 100 Ohms, T_C = 110°C)	V_{GD}	0.2	—	—	Volts
Holding Current (V_{AK} = 12 Vdc, I_{GT} = 2 mA) T_C = 25°C *(Initiating On-State Current = 200 mA) T_C = −40°C	I_H	— —	— —	5 10	mA
*Total Turn-On Time (Source Voltage = 12 V, R_S = 6 k Ohms) (I_{TM} = 8.2 A, I_{GT} = 2 mA, Rated V_{DRM}) (Rise Time = 20 ns, Pulse Width = 10 μs)	t_{gt}	—	2	—	μs
Forward Voltage Application Rate (V_D = Rated V_{DRM}, T_C = 110°C)	dv/dt	—	10	—	V/μs

*Indicates JEDEC Registered Data.

Notes: 1. Torque rating applies with use of compression washer (B52200F006 or equivalent). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Anode lead and heatsink contact pad are common. (See AN-209 B)

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +200°C. For optimum results an activated flux (oxide removing) is recommended.

2. Measurement does not include R_{GK} current.

3

FIGURE 1 – MAXIMUM CASE TEMPERATURE

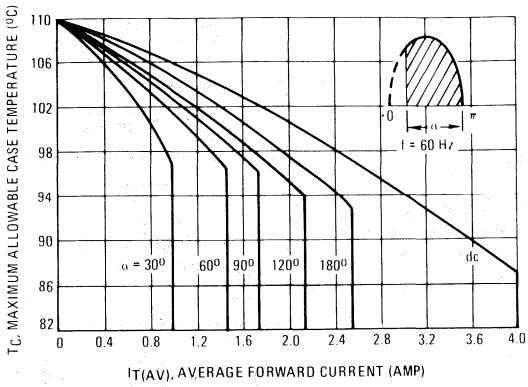
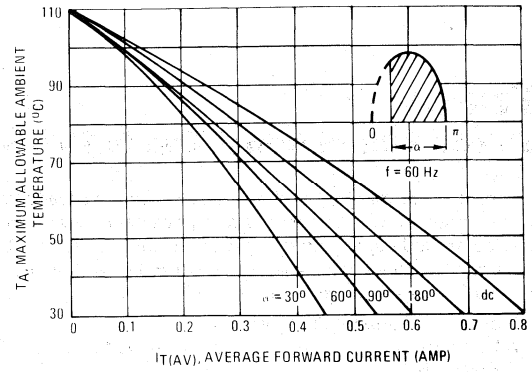


FIGURE 2 – MAXIMUM AMBIENT TEMPERATURE



3

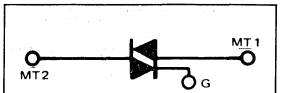
Triacs
Silicon Bidirectional Triode Thyristors

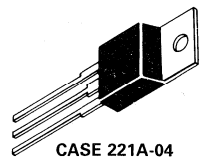
2N6342 thru 2N6349

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Two Modes (2N6342, 2N6343, 2N6344, 2N6345) or Four Modes (2N6346, 2N6347, 2N6348, 2N6349)
- For 400 Hz Operation, Consult Factory
- 12 Ampere Devices Available as 2N6342A thru 2N6349A





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −40 to +100°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	2N6342, 2N6346 2N6343, 2N6347 2N6344, 2N6348 2N6345, 2N6349	V_{DRM}	 200 400 600 800	Volts
*RMS On-State Current Full Cycle Sine Wave 50 to 60 Hz	(T_C = +80°C) (T_C = +90°C)	$I_{T(RMS)}$	8 4	Amps
*Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +80°C) Preceded and followed by Rated Current		I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	40	A^2s
*Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)		P_{GM}	20	Watts
*Average Gate Power (T_C = +80°C, t = 8.3 ms)		$P_{G(AV)}$	0.5	Watt
*Peak Gate Current		I_{GM}	2	Amps
*Peak Gate Voltage		V_{GM}	10	Volts
*Operating Junction Temperature Range		T_J	−40 to +125	°C
*Storage Temperature Range		T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 100°C		—	—	2	mA
*Peak On-State Voltage (I_{TM} = 11 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.55	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms) (Minimum Gate Pulse Width = 2 μs)	I_{GT}				mA
MT2(+), G(+) All Types		—	12	50	
MT2(+), G(−) 2N6346 thru 49		—	12	75	
MT2(−), G(−) All Types		—	20	50	
MT2(−), G(+) 2N6346 thru 49		—	35	75	
*MT2(+), G(+); MT2(−), G(−) T_C = −40°C All Types		—	—	100	
*MT2(+), G(−); MT2(−), G(+) T_C = −40°C 2N6346 thru 49		—	—	125	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms) (Minimum Gate Pulse Width = 2 μs)	V_{GT}				Volts
MT2(+), G(+) All Types		—	0.9	2	
MT2(+), G(−) 2N6346 thru 49		—	0.9	2.5	
MT2(−), G(−) All Types		—	1.1	2	
MT2(−), G(+) 2N6346 thru 49		—	1.4	2.5	
*MT2(+), G(+); MT2(−), G(−) T_C = −40°C All Types		—	—	2.5	
*MT2(+), G(−); MT2(−), G(+) T_C = −40°C 2N6346 thru 49		—	—	3	
(V_D = Rated V_{DRM}, R_L = 10 k Ohms, T_J = 100°C)					
*MT2(+), G(+); MT2(−), G(−) All Types		0.2	—	—	
*MT2(+), G(−); MT2(−), G(−) 2N6346 thru 49		0.2	—	—	
*Holding Current (V_D = 12 Vdc, Gate Open) T_C = 25°C	I_H	—	6	40	mA
(I_T = 200 mA) *T_C = −40°C		—	—	75	
*Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 11 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1.5	2	μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 11 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

*Indicates JEDEC Registered Data.

FIGURE 1 — RMS CURRENT DERATING

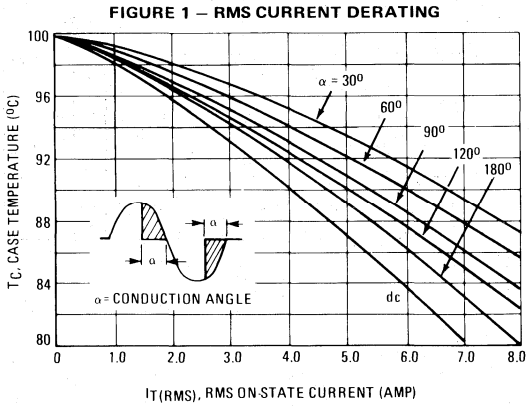


FIGURE 2 — ON-STATE POWER DISSIPATION

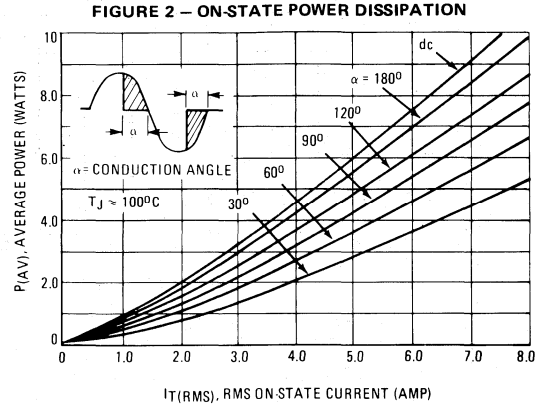


FIGURE 3 – TYPICAL GATE TRIGGER VOLTAGE

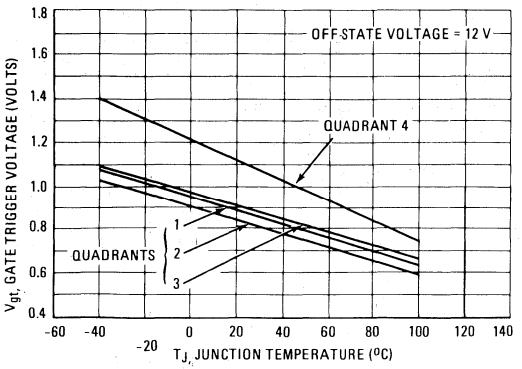


FIGURE 4 – TYPICAL GATE TRIGGER CURRENT

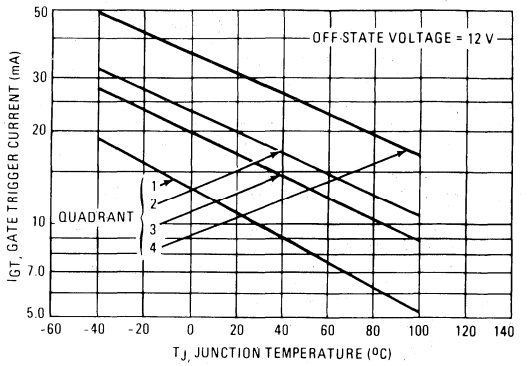


FIGURE 5 – ON-STATE CHARACTERISTICS

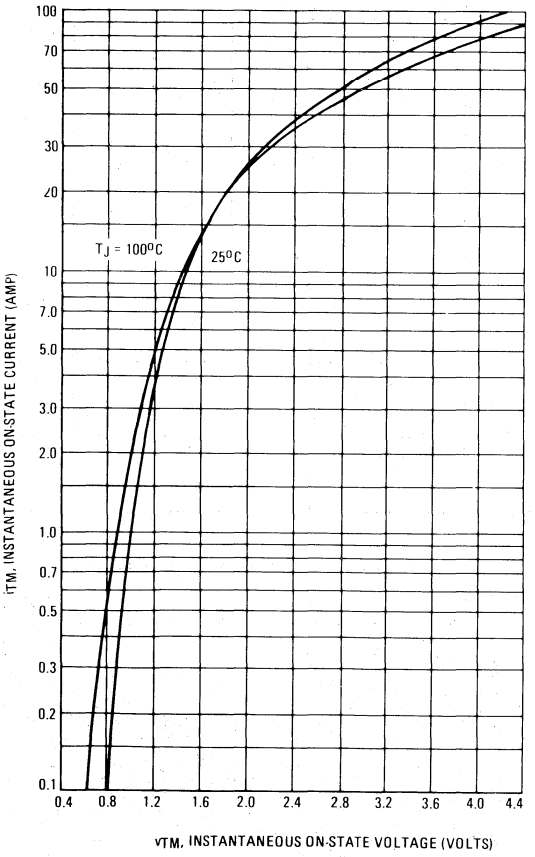


FIGURE 6 – TYPICAL HOLDING CURRENT

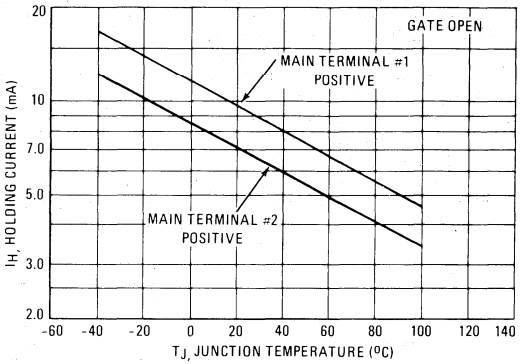


FIGURE 7 – MAXIMUM NON-REPETITIVE SURGE CURRENT

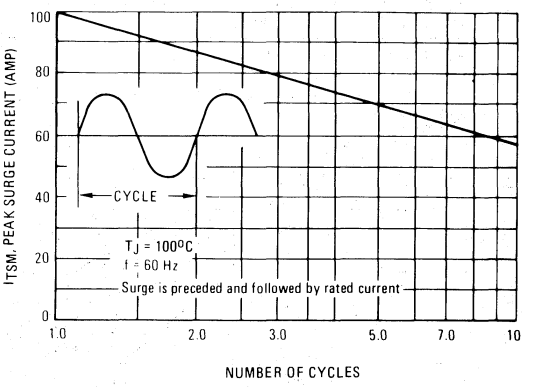


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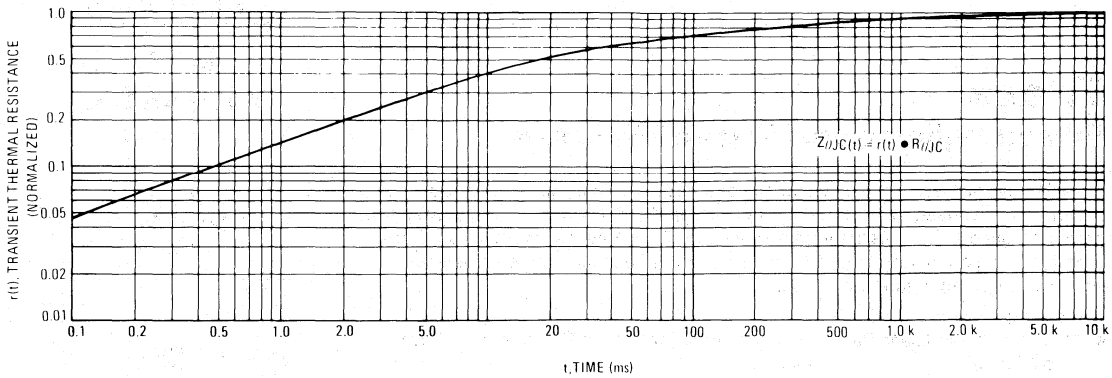
FIGURE 8 – TYPICAL THERMAL RESPONSE



3

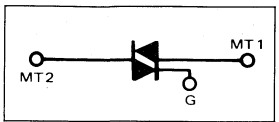
Triacs
Silicon Bidirectional Triode Thyristors

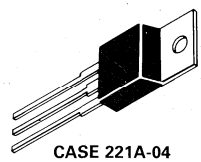
2N6342A thru 2N6349A

TRIACs
12 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Two Modes (2N6342A, 2N6343A, 2N6344A, 2N6345A) or Four Modes (2N6346A, 2N6347A, 2N6348A, 2N6349A)
- For 400 Hz Operation, Consult Factory
- 8 Ampere Devices Available as 2N6342 thru 2N6349





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
*Peak Repetitive Off-State Voltage (T_J = −40 to +110°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	2N6342A, 2N6346A 2N6343A, 2N6347A 2N6344A, 2N6348A 2N6345A, 2N6349A	V_{DRM}	 200 400 600 800	Volts
*RMS On-State Current (Full Cycle Sine Wave 50 to 60 Hz)	(T_C = +80°C) (T_C = +95°C)	$I_{T(RMS)}$	12 6	Amps
*Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +80°C) Preceded and Followed by Rated Current		I_{TSM}	120	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	59	A^2s
*Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)		P_{GM}	20	Watts
*Average Gate Power (T_C = +80°C, t = 8.3 ms)		$P_{G(AV)}$	0.5	Watt
*Peak Gate Current		I_{GM}	2	Amps
*Peak Gate Voltage		V_{GM}	±10	Volts
*Operating Junction Temperature Range		T_J	−40 to +125	°C
*Storage Temperature Range		T_{stg}	−40 to +150	°C

*Indicates JEDEC Registered Data.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
*Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	2	mA
*Peak On-State Voltage (Either Direction) (I_{TM} = 17 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.75	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	I_{GT}				mA
MT2(+), G(+) All Types		—	6	50	
MT2(+), G(−) 2N6346A thru 2N6349A		—	6	75	
MT2(−), G(−) All Types		—	10	50	
MT2(−), G(+) 2N6346A thru 2N6349A		—	25	75	
*MT2(+), G(+); MT2(−), G(−) T_C = −40°C All Types		—	—	100	
*MT2(+), G(−); MT2(−), G(+) T_C = −40°C 2N6346A thru 2N6349A		—	—	125	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 100 ohms)	V_{GT}				Volts
MT2(+), G(+) All Types		—	0.9	2	
MT2(+), G(−) 2N6346A thru 2N6349A		—	0.9	2.5	
MT2(−), G(−) All Types		—	1.1	2	
MT2(−), G(+) 2N6346A thru 2N6349A		—	1.4	2.5	
*MT2(+), G(+); MT2(−), G(−) T_C = −40°C All Types		—	—	2.5	
*MT2(+), G(−); MT2(−), G(+) T_C = −40°C 2N6346A thru 2N6349A		—	—	3	
(V_D = Rated V_{DRM}, R_L = 10 k ohms, T_J = 100°C)					
*MT2(+), G(+); MT2(−), G(−) All Types		0.2	—	—	
*MT2(+), G(−); MT2(−), G(−) 2N6346A thru 2N6349A		0.2	—	—	
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open) T_C = 25°C	I_H	—	6	40	mA
(I_T = 200 mA) *T_C = −40°C		—	—	75	
*Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 17 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1.5	2	μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 17 A, Commutating di/dt = 6.5 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

*Indicates JEDEC Registered Data.

3

FIGURE 1 — RMS CURRENT DERATING

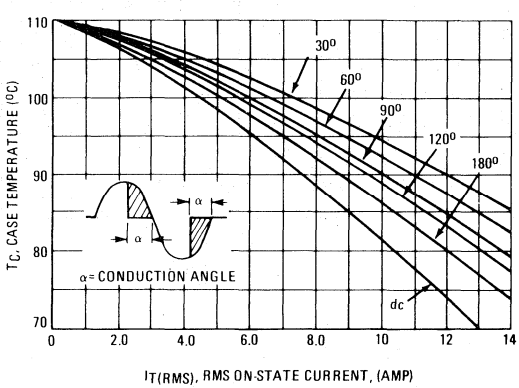


FIGURE 2 — ON-STATE POWER DISSIPATION

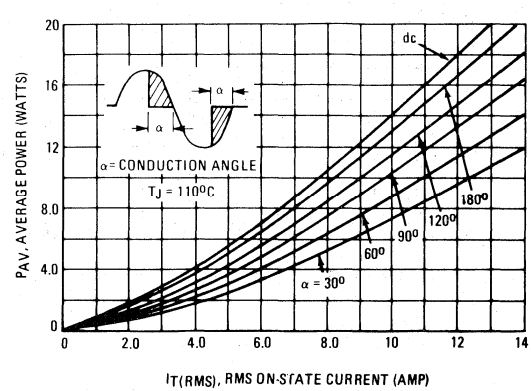


FIGURE 3 – TYPICAL GATE TRIGGER VOLTAGE

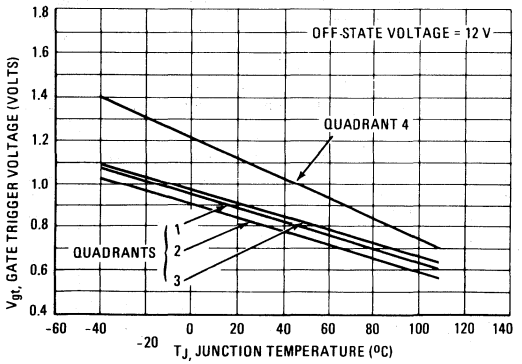


FIGURE 4 – TYPICAL GATE TRIGGER CURRENT

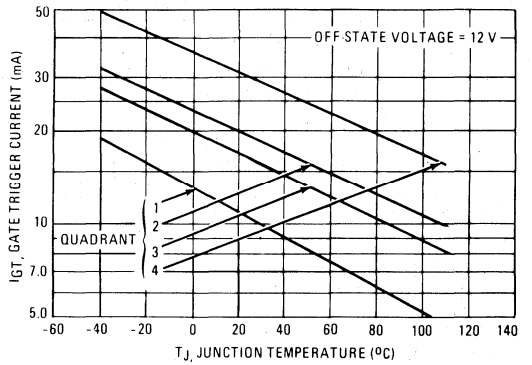


FIGURE 5 – ON-STATE CHARACTERISTICS

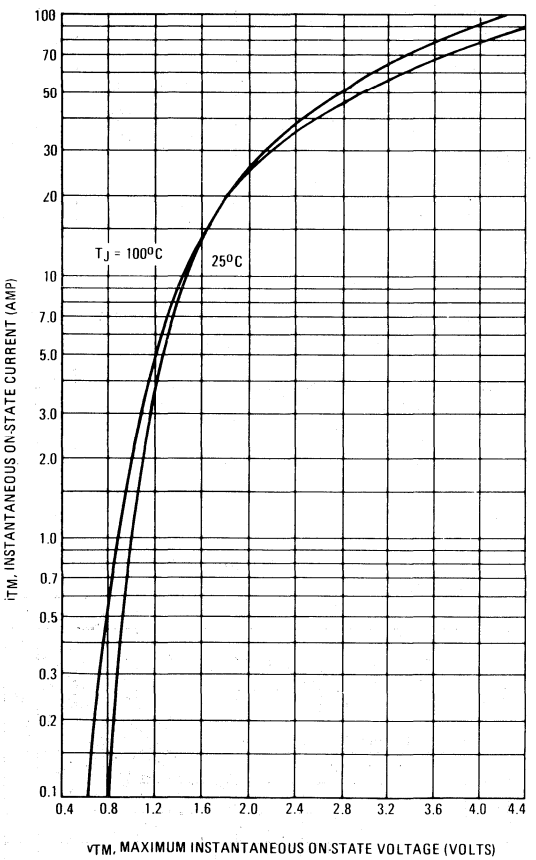


FIGURE 6 – TYPICAL HOLDING CURRENT

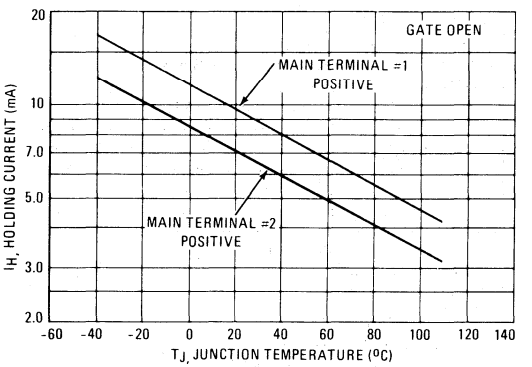


FIGURE 7 – MAXIMUM NON-REPETITIVE SURGE CURRENT

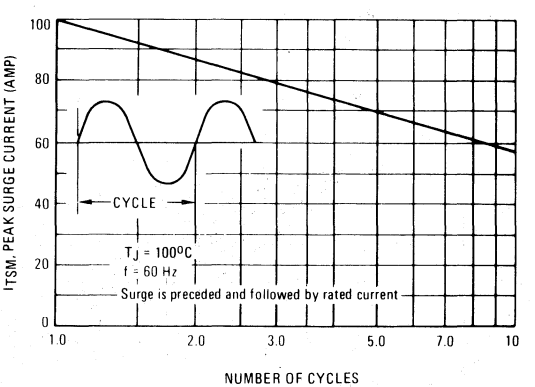


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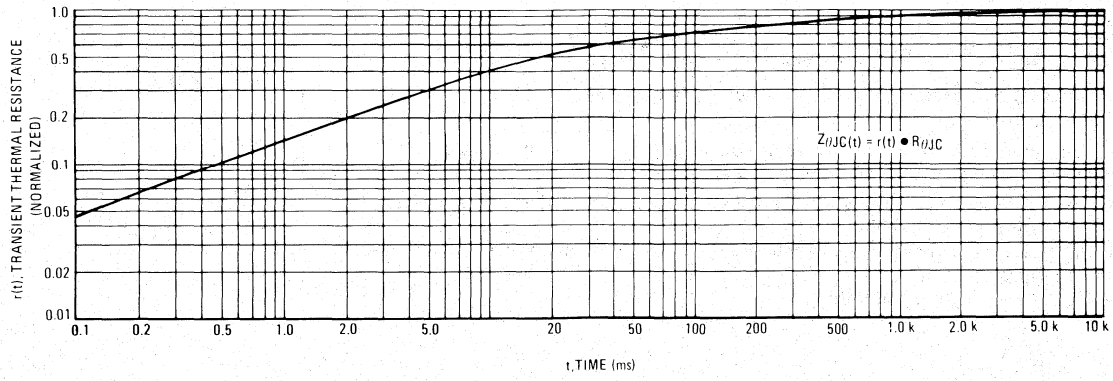
FIGURE 8 – TYPICAL THERMAL RESPONSE



3

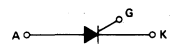
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

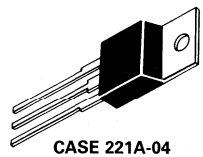
2N6394 thru 2N6399

SCRs
12 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies.

- Glass Passivated Junctions with Center Gate Geometry for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts





CASE 221A-04
(TO-220AB)
STYLE 3

3

***MAXIMUM RATINGS**

Rating		Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage (T_J = −40 to 125°C)	2N6394	V_{RRM} or V_{DRM}	50	Volts
	2N6395		100	
	2N6396		200	
	2N6397		400	
	2N6398		600	
	2N6399		800	
RMS On-State Current (All Conduction Angles)	T_C = 90°C	$I_{T(RMS)}$	12	Amps
Peak Non-Repetitive Surge Current (1/2 cycle, Sine Wave, 60 Hz, T_J = 125°C)		I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	40	A^2s
Forward Peak Power		P_{GM}	20	Watts
Forward Average Gate Power		$P_{G(AV)}$	0.5	Watt

*Indicates JEDEC Registered Data.

(cont.)

***MAXIMUM RATINGS — continued**

Rating	Symbol	Value	Unit
Forward Peak Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
*Forward "On" Voltage (I_{TM} = 24 A Peak)	V_{TM}	—	1.7	2.2	Volts
*Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	I_{GT}	—	5	30	mA
*Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms) (V_D = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)	V_{GT} V_{GD}	— 0.2	0.7 —	1.5 —	Volts
*Holding Current (V_D = 12 Vdc)	I_H	—	6	40	mA
Turn-On Time (I_{TM} = 12 A, I_{GT} = 40 mAdc, V_D = Rated V_{DRM})	t_{gt}	—	1	2	μs
Turn-Off Time (V_D = Rated V_{DRM}) (I_{TM} = 12 A, I_R = 12 A) (I_{TM} = 12 A, I_R = 12 A, T_J = 125°C)	t_q	— —	15 35	— —	μs
Critical Rate-of-Rise of Off-State Voltage Exponential (V_D = Rated V_{DRM}, T_J = 125°C)	dv/dt	—	50	—	V/μs

*Indicates JEDEC Registered Data.

3

FIGURE 1 – CURRENT DERATING

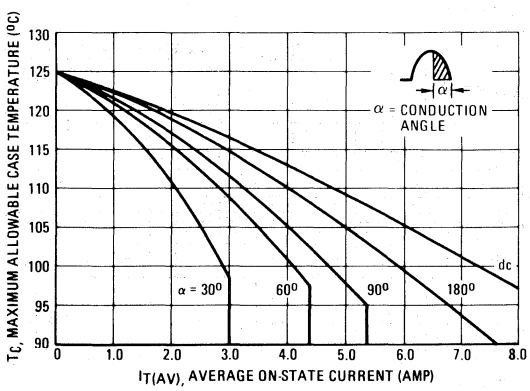


FIGURE 2 – MAXIMUM ON-STATE POWER DISSIPATION

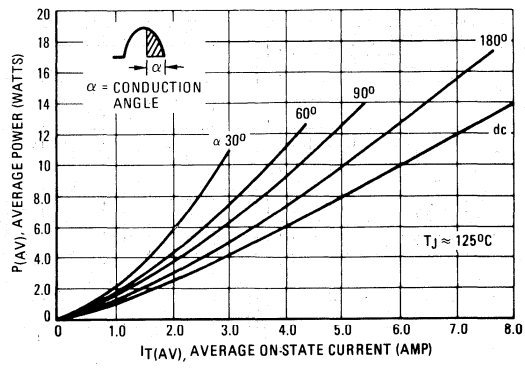


FIGURE 3 – ON-STATE CHARACTERISTICS

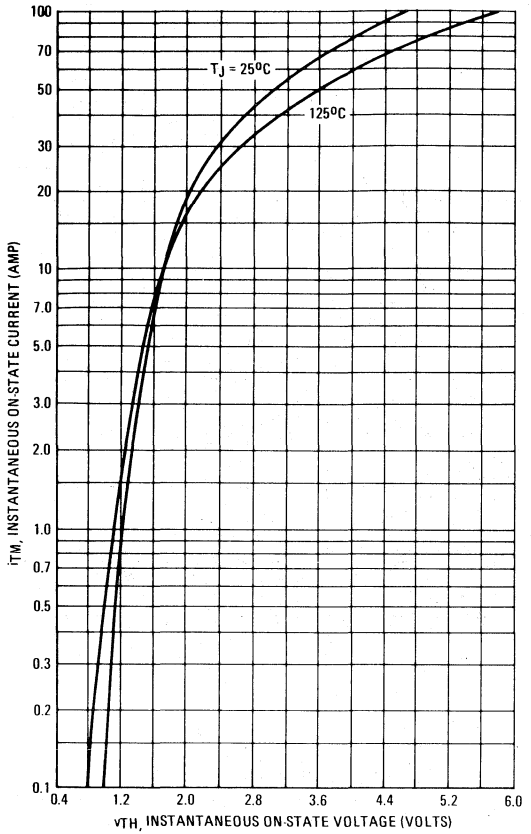


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

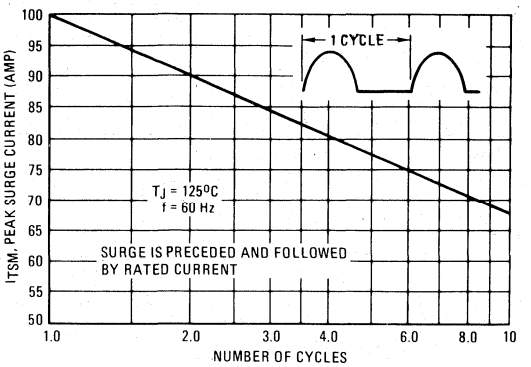
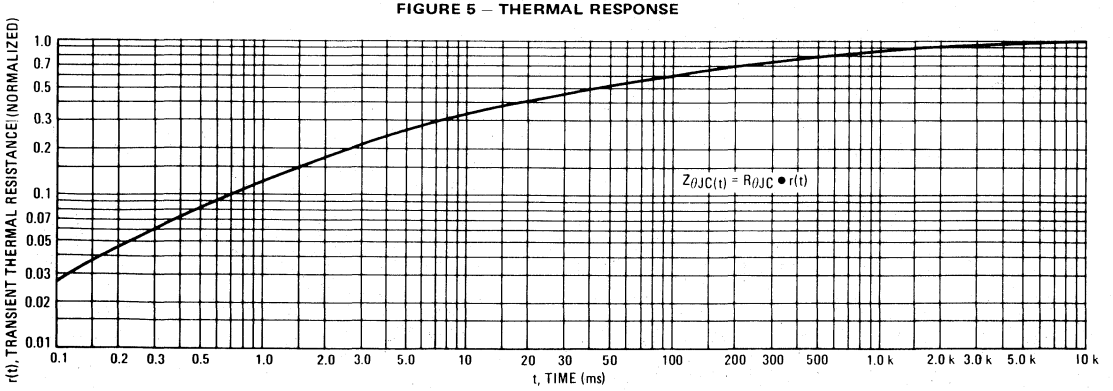


FIGURE 5 – THERMAL RESPONSE



3

TYPICAL CHARACTERISTICS

FIGURE 6 – PULSE TRIGGER CURRENT

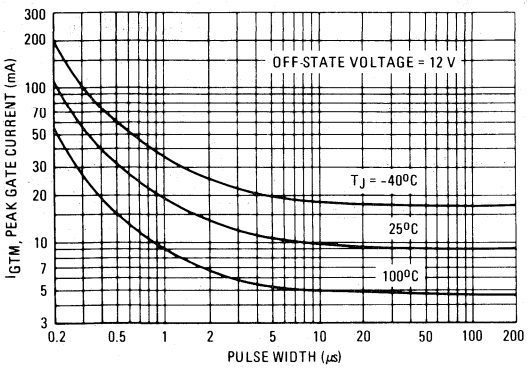


FIGURE 7 – GATE TRIGGER CURRENT

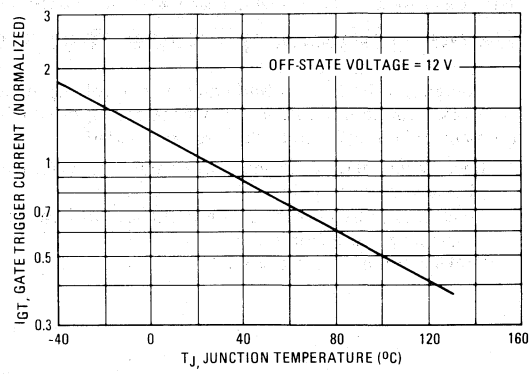


FIGURE 8 – GATE TRIGGER VOLTAGE

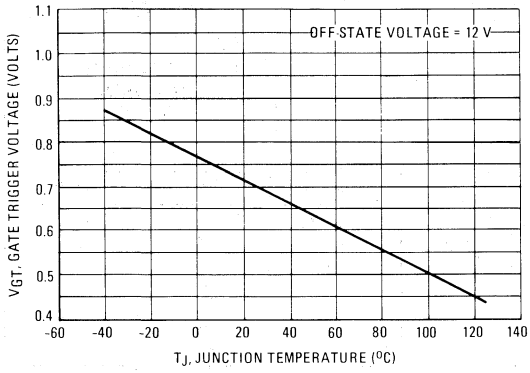
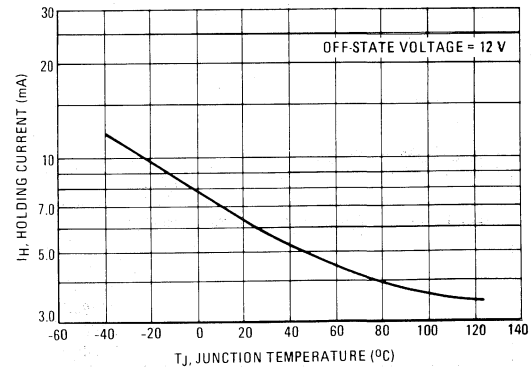


FIGURE 9 – HOLDING CURRENT



3

Silicon Controlled Rectifiers

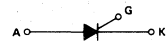
Reverse Blocking Triode Thyristors

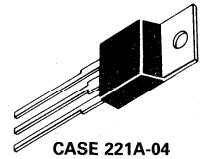
2N6400 thru 2N6405

SCRs
16 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions with Center Gate Geometry for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts





CASE 221A-04
(TO-220AB)
STYLE 3

3

***MAXIMUM RATINGS**

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Voltage	V_{RRM} or V_{DRM}		Volts
2N6400		50	
2N6401		100	
2N6402		200	
2N6403		400	
2N6404		600	
2N6405		800	
RMS On-State Current, T_C = 90°C	$I_{T(RMS)}$	16	Amps
Average On-State Current	$I_{T(AV)}$	10	Amps
Peak Non-Repetitive Forward Surge Current (1/2 cycle, Sine Wave, 60 Hz, T_J = 125°C)	I_{TSM}	160	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	145	A^2s
Forward Peak Gate Power	P_{GM}	20	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt

*Indicates JEDEC Registered Data.

(cont.)

***MAXIMUM RATINGS — continued**

Rating	Symbol	Value	Unit
Forward Peak Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.5	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM})	T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
*Peak On-State Voltage (I_{TM} = 32 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)		V_{TM}	—	—	1.7	Volts
*Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 50 Ohms)		I_{GT}	—	5	30	mA
*Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 50 Ohms) (V_D = Rated V_{DRM}, R_L = 50 Ohms)	T_C = 25°C T_C = −40°C T_C = +125°C	V_{GT}	— — 0.2	0.7 — —	1.5 2.5 —	Volts
*Holding Current (V_D = 12 Vdc)	T_C = 25°C *T_C = −40°C	I_H	— —	6 —	40 60	mA
Turn-On Time (I_{TM} = 16 A, I_{GT} = 40 mAdc, V_D = Rated V_{DRM})		t_{gt}	—	1	—	μs
Turn-Off Time (I_{TM} = 16 A, I_R = 16 A, V_D = Rated V_{DRM})	T_C = 25°C T_J = +125°C	t_q	— —	15 35	— —	μs
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform)	T_J = +125°C	dv/dt	—	50	—	V/μs

*Indicates JEDEC Registered Data.

FIGURE 1 – AVERAGE CURRENT DERATING

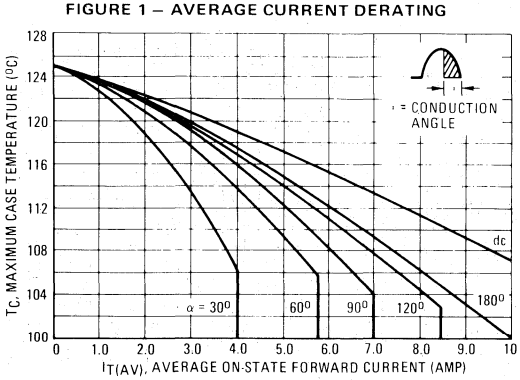


FIGURE 2 – MAXIMUM ON-STATE POWER DISSIPATION

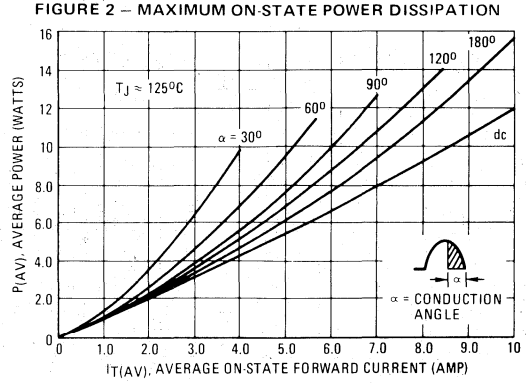


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FIGURE 3 – ON-STATE CHARACTERISTICS

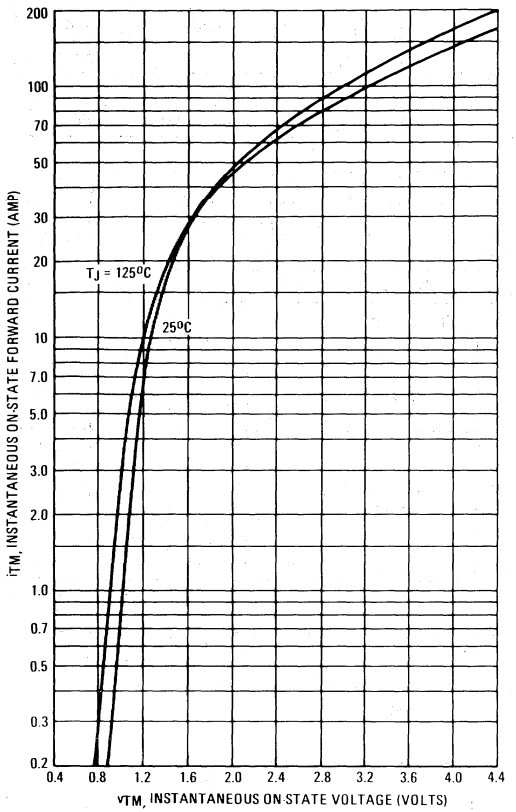


FIGURE 4 – MAXIMUM NON-REPETITIVE SURGE CURRENT

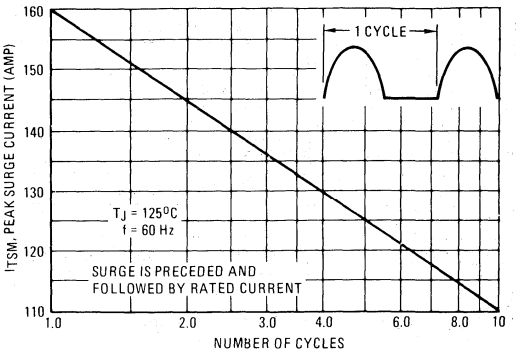
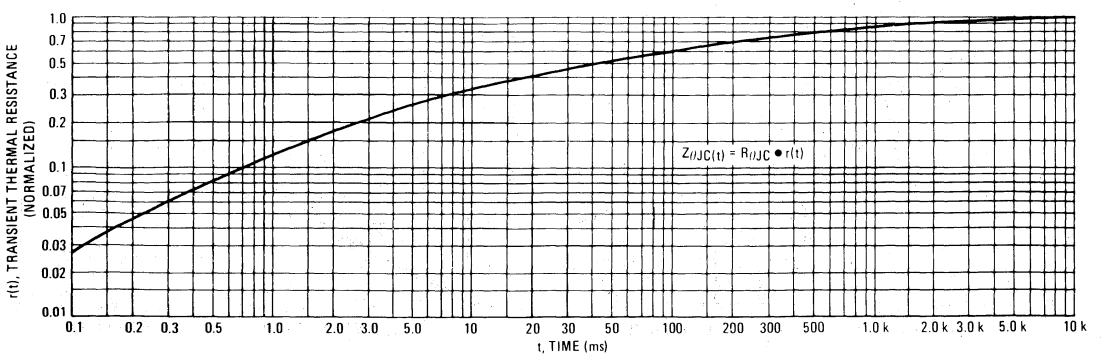


FIGURE 5 – THERMAL RESPONSE



TYPICAL TRIGGER CHARACTERISTICS

FIGURE 6 – PULSE TRIGGER CURRENT

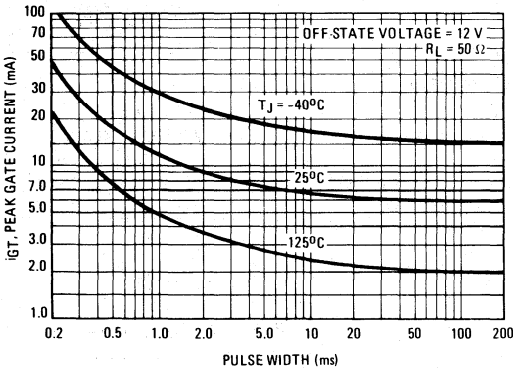


FIGURE 7 – GATE TRIGGER CURRENT

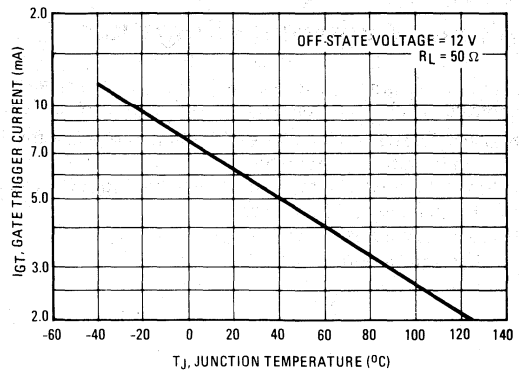


FIGURE 8 – GATE TRIGGER VOLTAGE

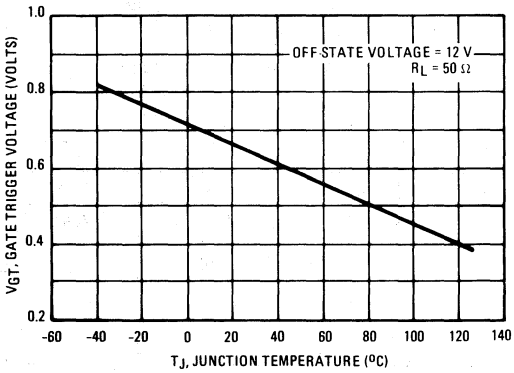
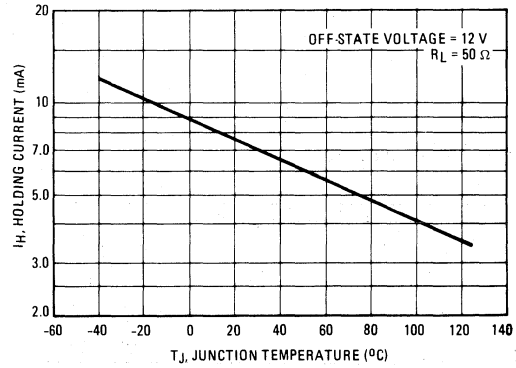


FIGURE 9 – HOLDING CURRENT



3

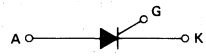
Thyristors
Silicon Controlled Rectifiers

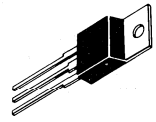
2N6504 thru 2N6509

SCRs
25 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supply crowbar circuits.

- Glass Passivated Junctions with Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Constructed for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts
- 300 A Surge Current Capability





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
*Peak Reverse Blocking Voltage, Note 1	V_{RRM}		Volts
2N6504		50	
2N6505		100	
2N6506		200	
2N6507		400	
2N6508		600	
2N6509		800	
Forward Current (T_C = 85°C) (All Conduction Angles)	$I_{T(RMS)}$	25	Amps
	$I_{T(AV)}$	16	
Peak Non-Repetitive Surge Current — 8.3 ms (1/2 Cycle, Sine Wave)	I_{TSM}	300	Amps
1.5 ms		350	

*Indicates JEDEC Registered Data.

(cont.)

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Forward Peak Gate Power	P_{GM}	20	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt
Forward Peak Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

***THERMAL CHARACTERISTICS**

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.5	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
*Peak Forward Blocking Voltage (T_J = 125°C)		V_{DRM}				Volts
	2N6504		50	—	—	
	2N6505		100	—	—	
	2N6506		200	—	—	
	2N6507		400	—	—	
	2N6508		600	—	—	
	2N6509		800	—	—	
*Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM})	T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
	T_J = 125°C		—	—	2	mA
*Forward "On" Voltage, Note 1 (I_{TM} = 50 A)		V_{TM}	—	—	1.8	Volts
*Gate Trigger Current (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	—	40	mA
	T_C = −40°C		—	25	75	
*Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms, T_C = −40°C)		V_{GT}	—	1	1.5	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)		V_{GD}	0.2	—	—	Volts
*Holding Current (Anode Voltage = 12 Vdc, T_C = −40°C)		I_H	—	35	40	mA
*Turn-On Time (I_{TM} = 25 A, I_{GT} = 50 mAdc)		t_{gt}	—	1.5	2	μs
Turn-Off Time (V_{DRM} = rated voltage)		t_q				μs
(I_{TM} = 25 A, I_R = 25 A)			—	15	—	
(I_{TM} = 25 A, I_R = 25 A, T_J = 125°C)			—	35	—	
Critical Rate of Rise of Off-State Voltage (Gate Open, Rated V_{DRM}, Exponential Waveform)		dv/dt	—	50	—	V/μs

*Indicates JEDEC Registered Data.
Note 1. Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.

3

FIGURE 1 — AVERAGE CURRENT DERATING

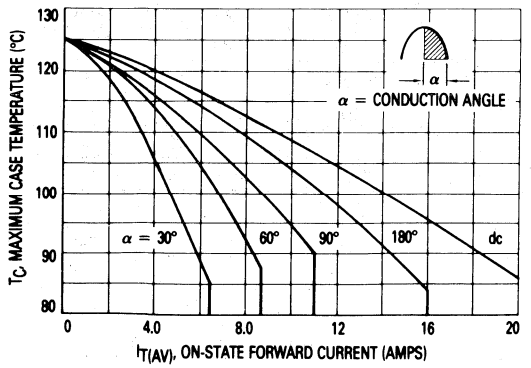


FIGURE 2 — MAXIMUM ON-STATE POWER DISSIPATION

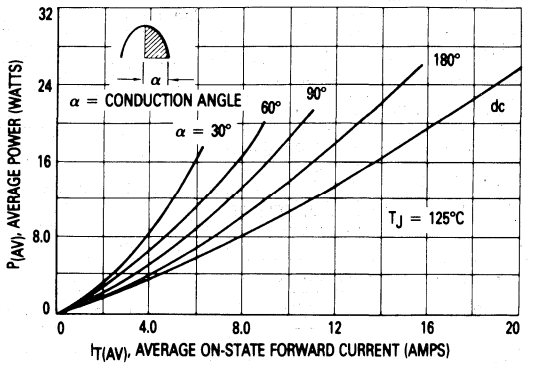


FIGURE 3 — MAXIMUM FORWARD VOLTAGE

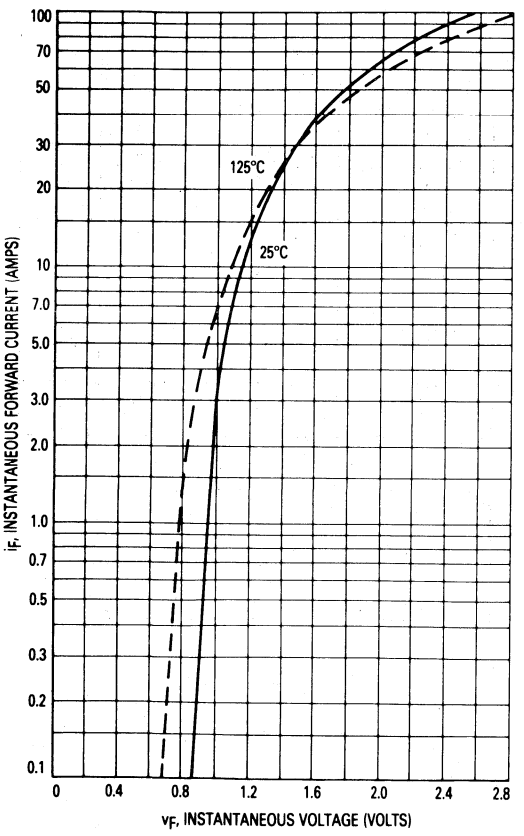


FIGURE 4 — MAXIMUM NON-REPETITIVE SURGE CURRENT

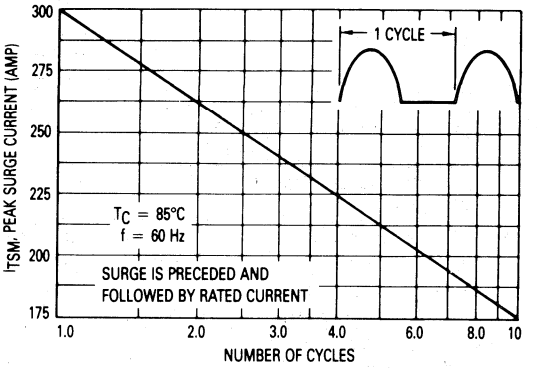


FIGURE 5 — CHARACTERISTICS AND SYMBOLS

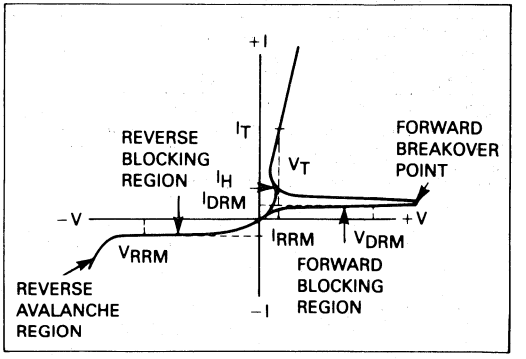


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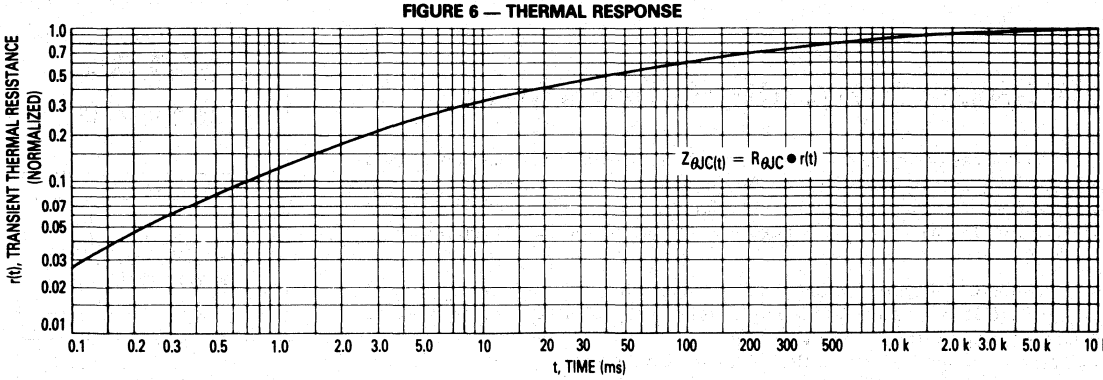
FIGURE 6 — THERMAL RESPONSE



TYPICAL TRIGGER CHARACTERISTICS

FIGURE 7 — GATE TRIGGER CURRENT

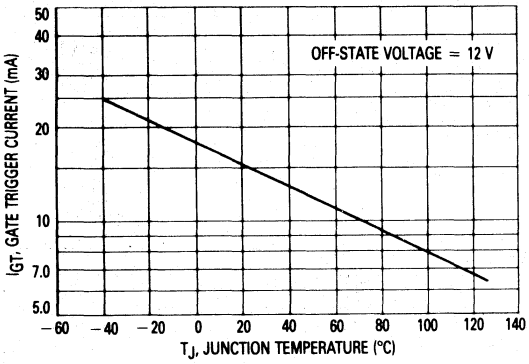


FIGURE 8 — GATE TRIGGER VOLTAGE

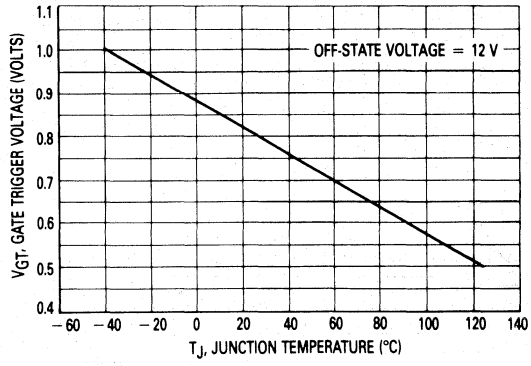
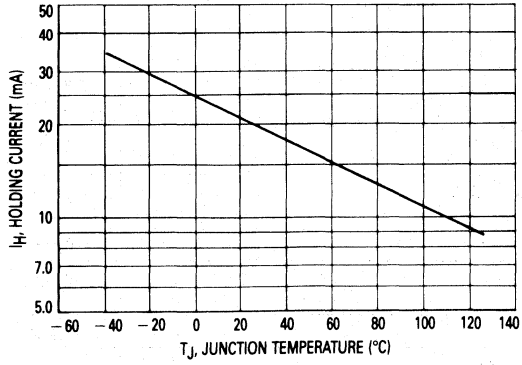


FIGURE 9 — HOLDING CURRENT



3

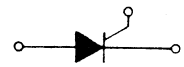
Silicon Controlled Rectifiers

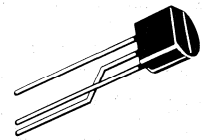
BRX44 thru 49

SCRs
0.8 AMPERE RMS
30 TO 400 VOLTS

PNPN devices designed for high volume, line-powered consumer applications such as relay and lamp drivers, small motor controls, gate drivers for larger thyristors, and sensing and detection circuits. Supplied in an inexpensive TO-226AA (TO-92) package which is readily adaptable for use in automatic insertion equipment.

- Sensitive Gate Trigger Current — 200 μA Maximum
- Low Reverse and Forward Blocking Current — 100 μA Maximum, T_C = 125°C
- Low Holding Current — 5 mA Maximum
- Glass-Passivated Surface for Reliability and Uniformity





CASE 29-04
(TO-226AA)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage	V_{RRM}		Volts
BRX44		30	
BRX45		60	
BRX46		100	
BRX47		200	
BRX48		300	
BRX49		400	
Forward Current RMS (All Conduction Angles)	$I_{T(RMS)}$	0.8	Amp
Peak Forward Surge Current, T_A = 25°C (1/2 Cycle, Sine Wave, 60 Hz)	I_{TSM}	8	Amps
Circuit Fusing Considerations, T_A = 25°C (t = 8.3 ms)	I^2t	0.15	A^2s
Peak Gate Power — Forward, T_A = 25°C	P_{GM}	0.1	Watt
Peak Gate Current Forward, T_A = 25°C (300 μs, 120 PPS)	I_{GFM}	1	Amp
Peak Gate Voltage — Reverse	V_{GRM}	5	Volts
Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Lead Solder Temperature (<1.5 mm from case, 10 s max.)		+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS

Characteristic		Symbol	Min	Max	Unit
Peak Forward Blocking Voltage (T_C = 125°C)		V_{RRM}			Volts
	BRX44		30	—	
	BRX45		60	—	
	BRX46		100	—	
	BRX47		200	—	
	BRX48		300	—	
	BRX49		400	—	
Peak Forward Blocking Current (Rated V_{DRM} @ T_C = 125°C)		I_{DRM}	—	100	μA
Peak Reverse Blocking Current (Rated V_{RRM} @ T_C = 125°C)		I_{RRM}	—	100	μA
Forward "On" Voltage (Note 1) (I_{TM} = 1 A Peak @ T_A = 25°C)		V_{TM}	—	1.7	Volts
Gate Trigger Current (Continuous dc) (Note 2) (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	200	μA
Gate Trigger Voltage (Continuous dc)	T_C = 25°C	V_{GT}	—	0.8	Volts
(Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = −40°C		—	1.2	
(Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms)	T_C = 125°C		0.1	—	
Holding Current	T_C = 25°C	I_H	—	5	mA
(Anode Voltage = 7 Vdc, initiating current = 20 mA)	T_C = −40°C		—	10	

NOTE:
1. Forward current applied for 1 ms maximum duration, duty cycle ≤ 1%.
2. R_{GK} current is not included in measurement.

3

FIGURE 1 — CURRENT DERATING (REFERENCE: CASE TEMPERATURE)

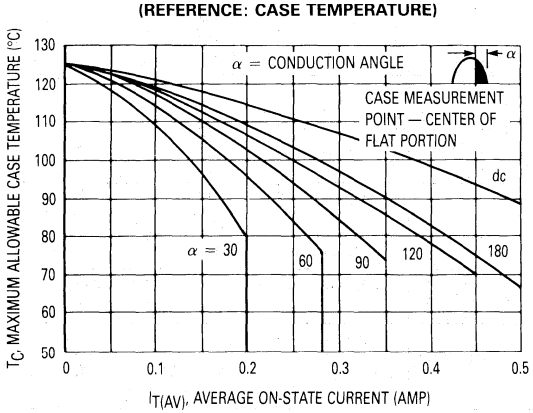
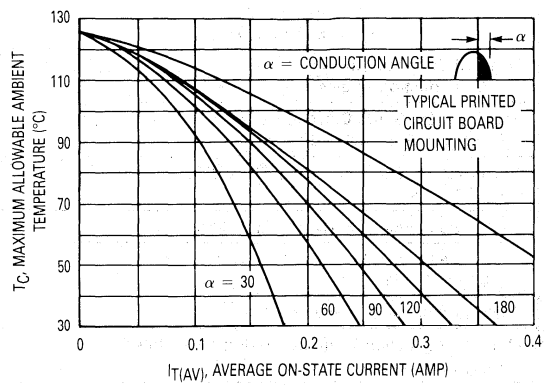


FIGURE 2 — CURRENT DERATING (REFERENCE: AMBIENT TEMPERATURE)



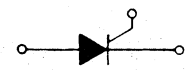
Silicon Controlled Rectifiers

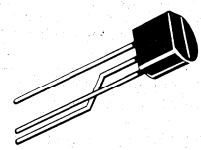
BRY55-30 thru 600

PNPN devices designed for high volume, line-powered consumer applications such as relay and lamp drivers, small motor controls, gate drivers for larger thyristors, and sensing and detection circuits. Supplied in an inexpensive TO-226AA (TO-92) package which is readily adaptable for use in automatic insertion equipment.

- Sensitive Gate Trigger Current — 200 μA Maximum
- Low Reverse and Forward Blocking Current — 100 μA Maximum, T_C = 125°C
- Low Holding Current — 5 mA Maximum
- Glass-Passivated Surface for Reliability and Uniformity

SCRs
0.8 AMPERE RMS
30 TO 600 VOLTS





CASE 29-04
(TO-226AA)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage Marking: BRY55-1 . . . BRY55-30 -2 . . . BRY55-60 -3 . . . BRY55-100 -4 . . . BRY55-200 -5 . . . BRY55-300 -6 . . . BRY55-400 -7 . . . BRY55-500 -8 . . . BRY55-600	V_{RRM}	 30 60 100 200 300 400 500 600	Volts
Forward Current RMS (All Conduction Angles)	$I_{T(RMS)}$	0.8	Amp
Peak Forward Surge Current, T_A = 25°C (1/2 Cycle, Sine Wave, 60 Hz)	I_{TSM}	8	Amps
Circuit Fusing Considerations, T_A = 25°C (t = 8.3 ms)	I^2t	0.15	A^2s
Peak Gate Power — Forward, T_A = 25°C	P_{GM}	0.1	Watt
Peak Gate Current Forward, T_A = 25°C (300 μs, 120 PPS)	I_{GFM}	1	Amp
Peak Gate Voltage — Reverse	V_{GRM}	5	Volts
Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Lead Solder Temperature (<1.5 mm from case, 10 s max.)		+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS

Characteristic		Symbol	Min	Max	Unit
Peak Forward Blocking Voltage (T_C = 125°C)		V_{RRM}			Volts
	BRY55-30		30	—	
	BRY55-60		60	—	
	BRY55-100		100	—	
	BRY55-200		200	—	
	BRY55-300		300	—	
	BRY55-400		400	—	
	BRY55-500		500	—	
	BRY55-600		600	—	
Peak Forward Blocking Current (Rated V_{DRM} @ T_C = 125°C)		I_{DRM}	—	100	μA
Peak Reverse Blocking Current (Rated V_{RRM} @ T_C = 125°C)		I_{RRM}	—	100	μA
Forward "On" Voltage (Note 1) (I_{TM} = 1 A Peak @ T_A = 25°C)		V_{TM}	—	1.7	Volts
Gate Trigger Current (Continuous dc) (Note 2) (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	200	μA
Gate Trigger Voltage (Continuous dc)	T_C = 25°C	V_{GT}	—	0.8	Volts
(Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = −40°C		—	1.2	
(Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms)	T_C = 125°C		0.1	—	
Holding Current	T_C = 25°C	I_H	—	5	mA
(Anode Voltage = 7 Vdc, initiating current = 20 mA)	T_C = −40°C		—	10	

NOTE:
1. Forward current applied for 1 ms maximum duration, duty cycle ≤ 1%.
2. R_{GK} current is not included in measurement.
3. MARKING: BRY55-30 = BRY55-1
   BRY55-60 = BRY55-2
   BRY55-100 = BRY55-3
   BRY55-200 = BRY55-4
   BRY55-300 = BRY55-5
   BRY55-400 = BRY55-6
   BRY55-500 = BRY55-7
   BRY55-600 = BRY55-8

FIGURE 1 — CURRENT DERATING (REFERENCE: CASE TEMPERATURE)

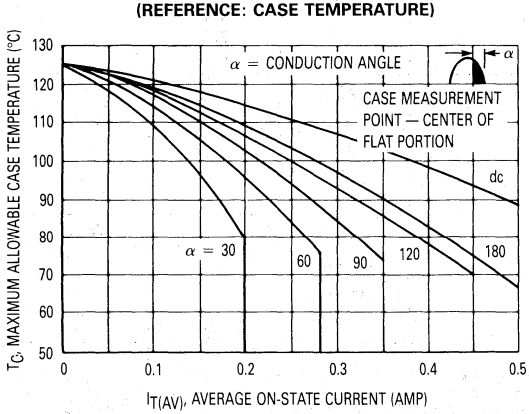
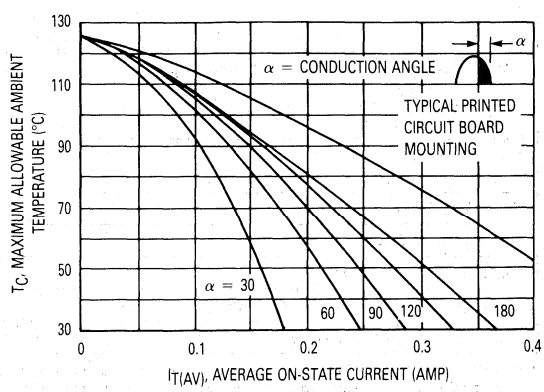


FIGURE 2 — CURRENT DERATING (REFERENCE: AMBIENT TEMPERATURE)



Silicon Controlled Rectifiers

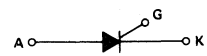
Reverse Blocking Triode Thyristors

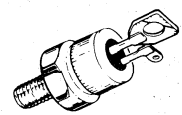
C35 Series

SCRs
35 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire for Greater Parameter Uniformity and Stability
- Blocking Voltage to 800 Volts





CASE 263-04
STYLE 1

3

MAXIMUM RATINGS (T_J = 125°C unless otherwise noted.)

Rating		Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage, Note 1 (T_C = −65 to +125°C)	C35F	V_{DRM} or V_{RRM}	50	Volts
	C35A		100	
	C35B		200	
	C35D		400	
	C35M		600	
	C35N		800	
Non-Repetitive Peak Reverse Voltage (T_C = −65 to +125°C, V < 5 ms)	C35F	V_{RSM}	75	Volts
	C35A		150	
	C35B		300	
	C35D		500	
	C35M		720	
	C35N		960	
RMS On-State Current (All Conduction Angles)		$I_{T(RMS)}$	35	Amps
Peak Non-Repetitive Surge Current (One cycle, 60 Hz)		I_{TSM}	225	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	210	A^2s
Peak Gate Power		P_{GM}	5	Watts
Average Gate Power		$P_{G(AV)}$	0.5	Watt
Peak Reverse Gate Voltage		V_{GRM}	5	Volts
Operating Junction Temperature Range		T_J	−65 to +125	°C
Storage Temperature Range		T_{stg}	−65 to +150	°C

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage. (cont.)

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.7	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
*Peak Forward Blocking Current		I_{DRM}				mA
(V_D = Rated V_{DRM} @ T_C = +125°C)	C35F,A	or	—	—	13	
	C35B	I_{RRM}	—	—	12	
	C35D		—	—	6	
	C35M		—	—	5	
	C35N		—	—	4	
(V_D = Rated V_{DRM} @ T_C = 125°C)	All Devices		—	—	10	μA
Peak Reverse Blocking Current		$I_{DRM(AV)}$				mA
(V_R = Rated V_{RRM} @ T_C = +125°C)	C35F,A	or	—	—	6.5	
	C35B	$I_{RRM(AV)}$	—	—	6	
	C35D		—	—	4	
	C35M		—	—	2.5	
	C35N		—	—	2	
(V_R = Rated V_{RRM} @ T_C = 125°C)	All Devices		—	—	10	μA
Peak On-State Voltage (I_{TM} = 50.3 A peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)		V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc)		I_{GT}				mA
(V_D = 12 Vdc, R_L = 50 Ω)			—	6	40	
(V_D = 12 Vdc, R_L = 50 Ω, T_C = −65°C)			—	—	80	
Gate Trigger Voltage (Continuous dc)		V_{GT}				Volts
(V_D = 12 Vdc, R_L = 50 Ω, T_C = −65°C to +125°C)			—	—	3	
(V_D = Rated V_{DRM}, R_L = 1000 Ω, T_C = 125°C)			0.25	—	—	
Holding Current (V_D = 24 Vdc, Gate Supply = 10 V, 20 Ω, 45 μs minimum pulse width, I_T = 0.5 A)		I_H	—	—	100	mA
Critical Rate of Rise of Forward Blocking Voltage		dv/dt				V/μs
(V_D = Rated V_{DRM}, T_C = +125°C)	C35F,M,N		10	—	—	
	C35A,B		20	—	—	
	C35D		25	—	—	

3

FIGURE 1 – CURRENT DERATING (HALF-WAVE RECTIFIED SINE WAVE)

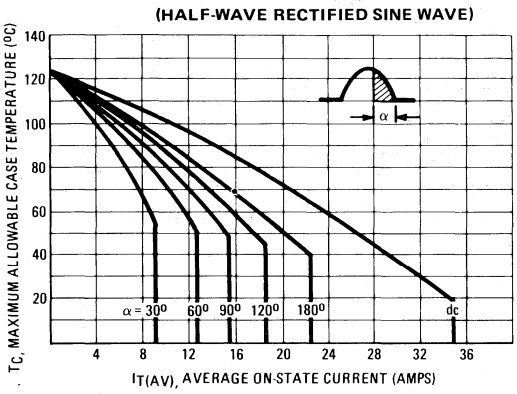
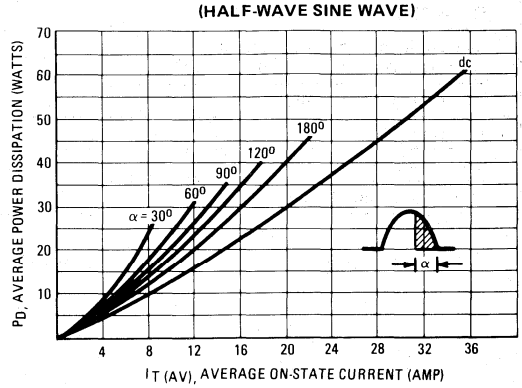


FIGURE 2 – POWER DISSIPATION (HALF-WAVE SINE WAVE)



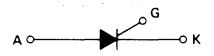
Silicon Controlled Rectifier
Reverse Blocking Triode Thyristors

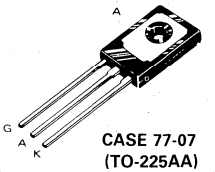
C106 Series

SCRs
4 AMPERES RMS
50 thru 600 VOLTS

. . . Glassivated PNPN devices designed for high volume consumer applications such as temperature, light, and speed control; process and remote control, and warning systems where reliability of operation is important.

- Glassivated Surface for Reliability and Uniformity
- Power Rated at Economical Prices
- Practical Level Triggering and Holding Characteristics
- Flat, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability





CASE 77-07
(TO-225AA)
STYLE 2

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage (R_{GK} = 1 kΩ) (T_C = −40° to 110°C)	C106F C106A C106B C106D C106M	V_{DRM} or V_{RRM}	50 100 200 400 600	Volts
RMS Forward Current (All Conduction Angles)		$I_{T(RMS)}$	4	Amps
Average Forward Current (T_A = 30°C)		$I_{T(AV)}$	2.55	Amps
Peak Non-Repetitive Surge Current (1/2 Cycle, 60 Hz, T_J = −40 to +110°C)		I_{TSM}	20	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	1.65	A^2s
Peak Gate Power		P_{GM}	0.5	Watt
Average Gate Power		$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current		I_{GFM}	0.2	Amp

(cont.)

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Peak Reverse Gate Voltage	V_{GRM}	6	Volts
Operating Junction Temperature Range	T_J	−40 to +110	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque, Note 1	—	6	in. lb.

Note 1. Torque rating applies with use of compression washer (B52200F006). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Anode lead and heatsink contact pad are common. (See AN-290 B)

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +200°C. For optimum results, an activated flux (oxide removing) is recommended.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	75	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, R_{GK} = 1000 Ohms) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	100	μA
Forward "On" Voltage (I_{FM} = 1 A Peak)	V_{TM}	—	—	2.2	Volts
Gate Trigger Current (Continuous dc)	I_{GT}				μA
(V_{AK} = 6 Vdc, R_L = 100 Ohms)		—	30	200	
(V_{AK} = 6 Vdc, R_L = 100 Ohms, T_C = −40°C)		—	75	500	
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volts
(V_{AK} = 6 Vdc, R_L = 100 Ohms, R_{GK} = 1000 Ohms) T_J = 25°C		0.4	—	0.8	
(V_{AK} = Rated V_{DRM}, R_L = 3000 Ohms, R_{GK} = 1000 Ohms, T_J = 110°C) T_J = −40°C		0.5	—	1	
		0.2	—	—	
Holding Current (V_D = 12 Vdc, R_{GK} = 1000 Ohms) T_J = 25°C	I_{HX}	0.3	—	3	mA
T_J = −40°C		0.4	—	6	
T_J = +110°C		0.14	—	2	
Forward Voltage Application Rate (T_J = 110°C, R_{GK} = 1000 Ohms, V_D = Rated V_{DRM})	dv/dt	—	8	—	V/μs
Turn-On Time	t_{gt}	—	1.2	—	μs
Turn-Off Time	t_q	—	40	—	μs

3

FIGURE 1 – AVERAGE CURRENT DERATING

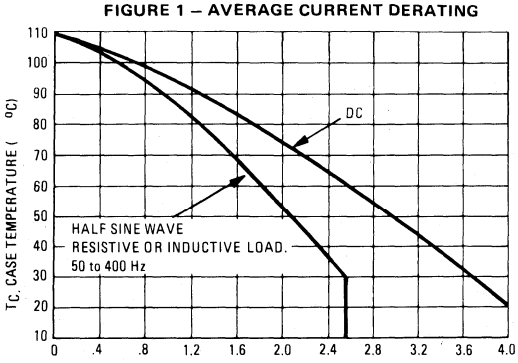
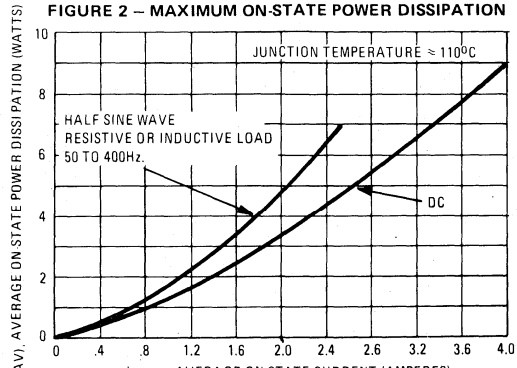
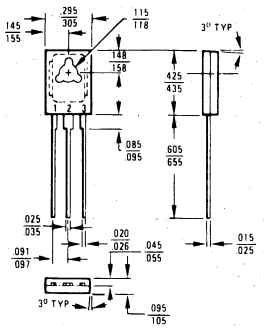


FIGURE 2 – MAXIMUM ON-STATE POWER DISSIPATION

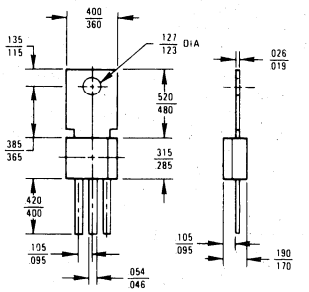


Package Interchangeability

The dimensional diagrams below compare the critical dimensions of the Motorola C-106 package with competitive devices. It has been demonstrated that the smaller dimensions of the Motorola package make it compatible in most lead-mount and chassis-mount applications. The user is advised to compare all critical dimensions for mounting compatibility.



Motorola C-106 Package



Competitive C-106 Package

3

Silicon Controlled Rectifiers

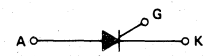
Reverse Blocking Triode Thyristors

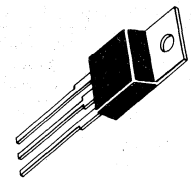
C122( )1 Series

SCRs
8 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts
- Different Leadform Configurations, Suffix (2) thru (6) available, see Leadform Options (Section 4) for Information





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1		V_{DRM}		Volts
Repetitive Peak Reverse Voltage	C122F1	V_{RRM}	50	
	C122A1		100	
	C122B1		200	
	C122D1		400	
	C122M1		600	
	C122N1		800	
Non-Repetitive Peak Reverse Voltage		V_{RSM}		Volts
	C122F1		75	
	C122A1		200	
	C122B1		300	
	C122D1		500	
	C122M1		700	
	C122N1		800	
Forward Current RMS (All Conduction Angles)	$T_C \leq 75°C$	$I_{T(RMS)}$	8	Amps
Peak Forward Surge Current (1/2 Cycle, Sine Wave, 60 Hz)		I_{TSM}	90	Amps
Circuit Fusing Considerations (t = 8.3 ms)		I^2t	34	A^2s

Note 1. V_{DRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. (cont.) Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Forward Peak Gate Power (t = 10 μs)	P_{GM}	5	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt
Forward Peak Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.8	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C T_C = 100°C	I_{DRM}, I_{RRM}	— —	— —	10 0.5	μA mA
Peak On-State Voltage, Note 1 (I_{TM} = 16 A Peak, T_C = 25°C)	V_{TM}	—	—	1.83	Volts
Gate Trigger Current (Continuous dc) (V_D = 6 V, R_L = 91 Ohms, T_C = 25°C) (V_D = 6 V, R_L = 45 Ohms, T_C = −40°C)	I_{GT}	— —	— —	25 40	mA
Gate Trigger Voltage (Continuous dc) (V_D = 6 V, R_L = 91 Ohms, T_C = 25°C) (V_D = 6 V, R_L = 45 Ohms, T_C = −40°C) (V_D = Rated V_{DRM}, R_L = 1000 Ohms, T_C = 100°C)	V_{GT}	— — 0.2	— — —	1.5 2 —	Volts
Holding Current (V_D = 24 Vdc, I_T = 0.5 A, 0.1 to 10 ms Pulse, Gate Trigger Source = 7 V, 20 Ohms) T_C = 25°C T_C = −40°C	I_H	— —	— —	30 60	mA
Turn-Off Time (V_D = Rated V_{DRM}) (I_{TM} = 8 A, I_R = 8 A)	t_q	—	50	—	μs
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Linear, T_C = 100°C)	dv/dt	—	50	—	V/μs

Note 1. Pulse Test: Pulse Width = 1 ms, Duty Cycle ≤ 2%.

3

FIGURE 1 – CURRENT DERATING (HALF-WAVE)

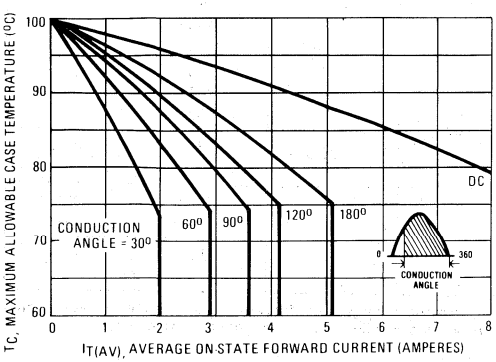


FIGURE 2 – CURRENT DERATING (FULL-WAVE)

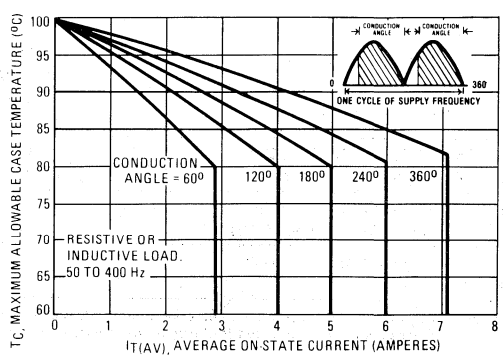


FIGURE 3 – MAXIMUM POWER DISSIPATION (HALF-WAVE)

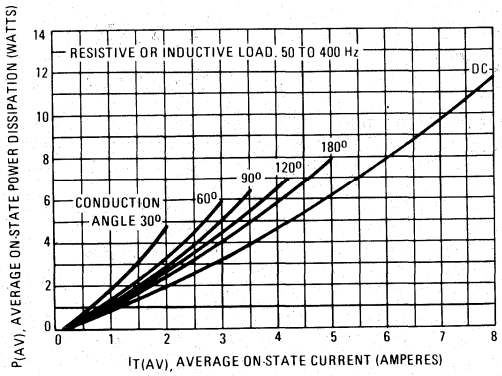
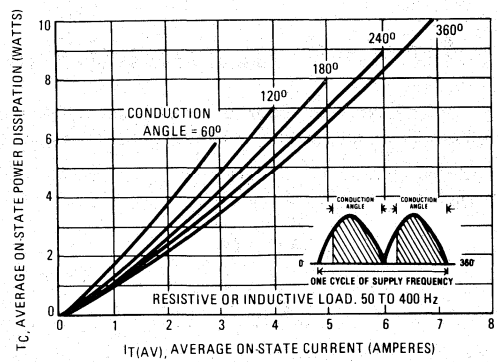


FIGURE 4 – MAXIMUM POWER DISSIPATION (FULL-WAVE)



3

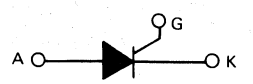
Silicon Controlled Rectifiers

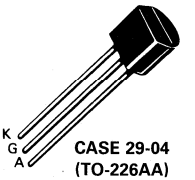
C205 Series

. . . designed and tested for repetitive peak operation required for CD ignition, fuel ignitors, flash circuits, motor controls and low-power switching applications.

- 150 Amperes for 2 μs Safe Area
- High dv/dt
- Very Low VF at High Current
- Low-Cost TO-226AA (TO-92)

SCRs
1.2 AMPERES RMS
30 thru 400 VOLTS





CASE 29-04
(TO-226AA)
STYLE 10

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 Repetitive Peak Reverse Voltage	V_{RRM} V_{DRM}		Volts
C205Y		30	
C205YY		60	
C205A		100	
C205B		200	
C205D		400	
Forward Current RMS (All Conduction Angles)	$I_{T(RMS)}$	1.2	Amps
Peak Forward Surge Current (1/2 Cycle, Sine Wave, 60 Hz)	I_{TSM}	10	Amps
Forward Peak Gate Power T_A = 25°C	P_{GM}	0.5	Watts
Forward Average Gate Power T_A = 25°C	$P_{G(AV)}$	0.1	Watt
Forward Peak Gate Current T_A = 25°C	I_{GM}	0.2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	R_θ	75	°C/W

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted. R_{GK} = 1000 Ohms)

Characteristic		Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Current (Rated V_{DRM})	T_C = 25°C T_C = 125°C	I_{DRM}	— —	— —	10 200	μA
Peak Reverse Blocking Current (Rated V_{RRM}	T_C = 25°C T_C = 125°C	I_{RRM}	— —	— —	10 200	μA
Peak On-State Voltage, Note 1 (I_{TM} = 1 A Peak, T_C = 25°C)		V_{TM}	—	—	1.6	Volts
Gate Trigger Current (Continuous dc) (V_D = 6 V, R_L = 100 Ohms, T_C = 25°C T_C = −40°C		I_{GT}	— —	— —	200 500	μA
Gate Trigger Voltage (Continuous dc) (V_D = 7 V, R_L = 100 Ohms, T_C = 25°C)		V_{GT}	—	—	0.8	Volts
Holding Current Anode Voltage = 12 Vdc	T_C = 25°C T_C = −40°C	I_H	— —	— —	5 10	mA
Turn-Off Time (V_{DRM} = Rated Voltage) T_J = +125°C		tq	—	15	—	μs
Forward Voltage Application Rate (T_C = 100°C)		dv/dt	—	20	—	V/μs

Note 1. Pulse Test: Pulse Width = 1 ms, Duty Cycle ≤ 2%.

3

Silicon Controlled Rectifiers

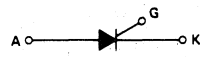
Reverse Blocking Triode Thyristors

C228
C228( )3
MCR229
Series

SCRs
35 AMPERES RMS
100 thru 600 VOLTS

. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 300 Amps
- Low Forward "On" Voltage −1.2 V (Typ) @ I_{TM} = 35 Amps
- Practical Level Triggering and Holding Characteristics — 10 mA (Typ) @ T_C = 25°C
- Rugged Construction in Either Pressfit, Stud, or Isolated Stud Packages
- Glass Passivated Junctions for Maximum Reliability



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C)	V_{DRM} and V_{RRM}		Volts
C228A, C228A3, MCR229A		100	
C228B, C228B3, MCR229B		200	
C228D, C228D3, MCR229D		400	
C228M, C228M3, MCR229M		600	
Non-Repetitive Reverse Voltage (T_J = −40 to +125°C)	V_{RSM}		Volts
C228A, C228A3, MCR229A		150	
C228B, C228B3, MCR229B		300	
C228D, C228D3, MCR229D		500	
C228M, C228M3, MCR229M		720	
Forward Current RMS	$I_{T(RMS)}$	35	Amps
Peak Surge Current (One Cycle, 60 Hz, T_C = −40 to +125°C)	I_{TSM}	300	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	370	A^2s
Peak Gate Power	P_{GM}	5	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Peak Forward Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Stud Torque	—	30	in. lb.

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode.

CASE 174-04
(TO-203AA)
STYLE 3
MCR229 Series

CASE 263-04
STYLE 1
C228 Series

CASE 311-02
STYLE 1
C228( )3 Series

3

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case C228 and MAC229 Series C228( )3 Series	$R_{\theta JC}$	 1.7 1.85	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C T_C = 100°C	I_{DRM}, I_{RRM}	 — —	 — —	 10 3	 μA mA
Forward "On" Voltage (I_{TM} = 100 A Peak)	V_{TM}	—	—	1.9	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 80 Ohms, T_C = 25°C) (V_D = 6 Vdc, R_L = 50 Ohms, T_C = −40°C)	I_{GT}	 — —	 — —	 40 80	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 80 Ohms, T_C = 25°C) (V_D = 6 Vdc, R_L = 80 Ohms, T_C = −40°C)	V_{GT}	 — —	 — —	 2.5 3	Volts
Gate Trigger Voltage (Rated V_{DRM}, R_L = 1000 Ohms, T_C = +125°C)	V_{GT}	0.2	—	—	Volts
Holding Current T_C = 25°C (Anode Voltage = 24 V, gate open) T_C = −40°C	I_H	— —	— —	75 150	mA
Turn-On Time (t_d + t_r) (I_{TM} = 35 Adc, I_{GT} = 40 mAdc)	t_{on}	—	1	—	μs
Turn-Off Time (I_{TM} = 10 A, I_R = 10 A) (I_{TM} = 10 A, I_R = 10 A, T_C = 100°C)	t_{off}	 — —	 20 35	 — —	μs
Forward Voltage Application Rate (T_C = 100°C)	dv/dt	—	50	—	V/μs

3

FIGURE 1 – CURRENT DERATING (HALF-WAVE RECTIFIED SINE WAVE)

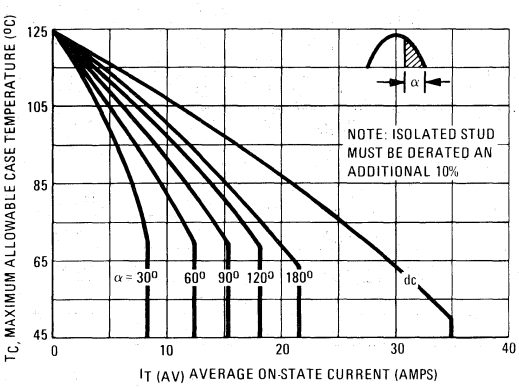
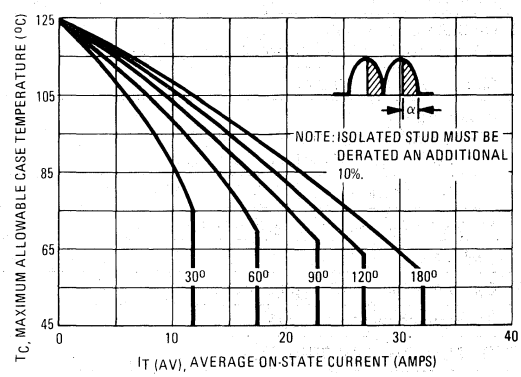


FIGURE 2 – CURRENT DERATING (FULL-WAVE RECTIFIED SINE WAVE)



C232, 233
(See MCR230)

Triacs

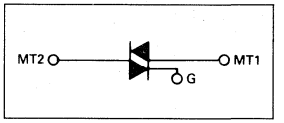
Silicon Bidirectional Triode Thyristors

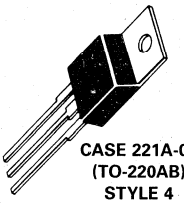
MAC15 Series
MAC15A Series

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC15 Series) or Four Modes (MAC15A Series)

TRIACs
15 AMPERES RMS
200 thru 800 VOLTS





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Repetitive Off-State Voltage (T_J = −40 to 125°C)	MAC15-4, MAC15A4 MAC15-6, MAC15A6 MAC15-8, MAC15A8 MAC15-10, MAC15A10	V_{DRM}	200 400 600 800	Volts
Peak Gate Voltage		V_{GM}	10	Volts
On-State Current RMS Full Cycle Sine Wave 50 to 60 Hz (T_C = +90°C)		$I_{T(RMS)}$	15	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	93	A^2s
Peak Surge Current (One Full Cycle, 60 Hz, T_C = +80°C) Preceded and followed by rated current)		I_{TSM}	150	Amps
Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)		P_{GM}	20	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)		$P_{G(AV)}$	0.5	Watt
Peak Gate Current		I_{GM}	2	Amps
Operating Junction Temperature Range		T_J	−40 to +125	°C
Storage Temperature Range		T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and either polarity of MT2 to MT1 Voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Peak On-State Voltage (I_{TM} = 21 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.6	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	I_{GT}				mA
MT2(+), G(+)		—	—	50	
MT2(+), G(−)		—	—	50	
MT2(−), G(−)		—	—	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	75	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
(V_D = Rated V_{DRM}, R_L = 10 k Ohms, T_J = 110°C)					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open) (I_T = 200 mA)	I_H	—	6	40	mA
Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 17 A) (I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 21 A, Commutating di/dt = 8 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

3

FIGURE 1 – RMS CURRENT DERATING

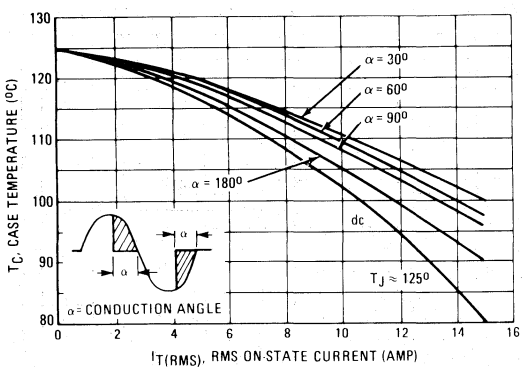


FIGURE 2 – ON-STATE POWER DISSIPATION

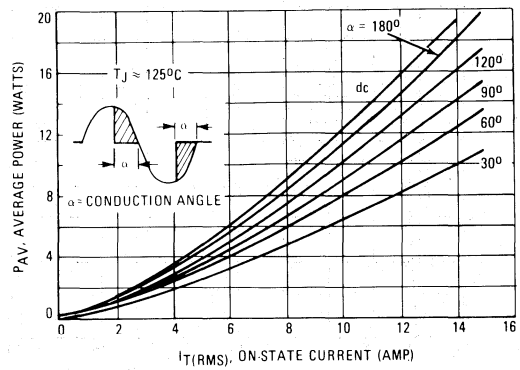


FIGURE 3 – TYPICAL GATE TRIGGER VOLTAGE

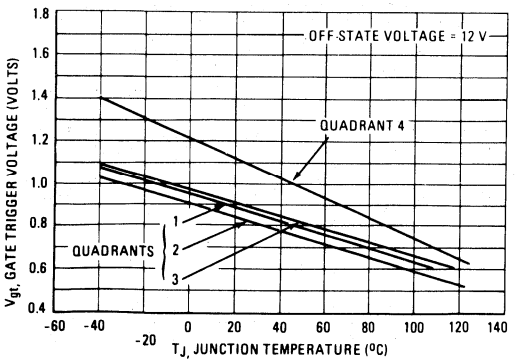


FIGURE 4 – TYPICAL GATE TRIGGER CURRENT

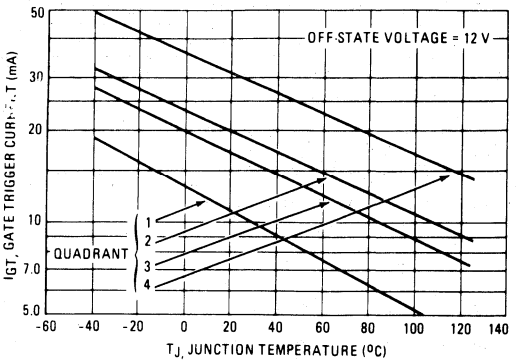


FIGURE 5 – ON-STATE CHARACTERISTICS

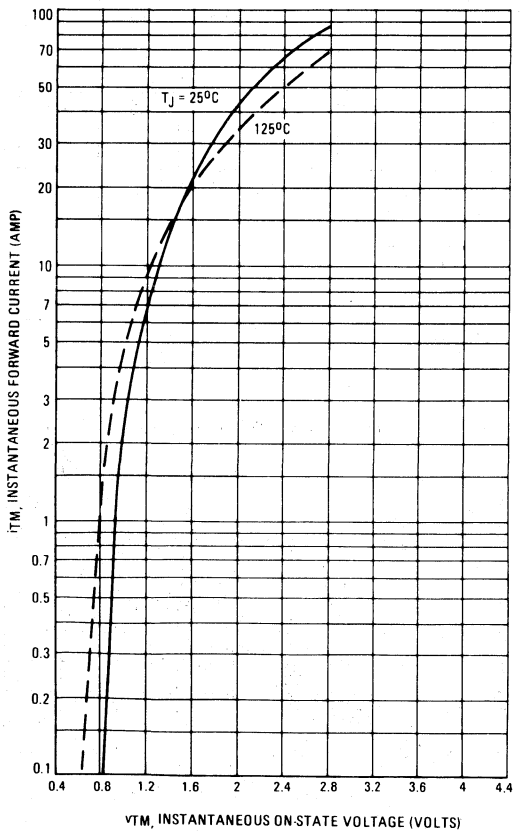


FIGURE 6 – TYPICAL HOLDING CURRENT

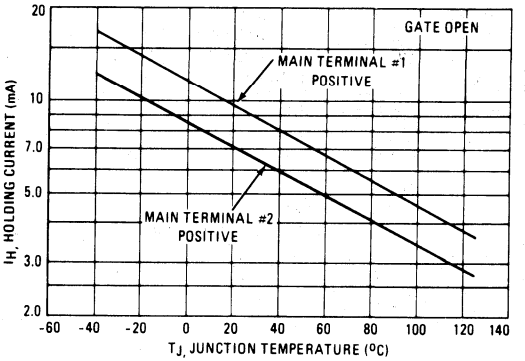


FIGURE 7 – MAXIMUM NON-REPETITIVE SURGE CURRENT

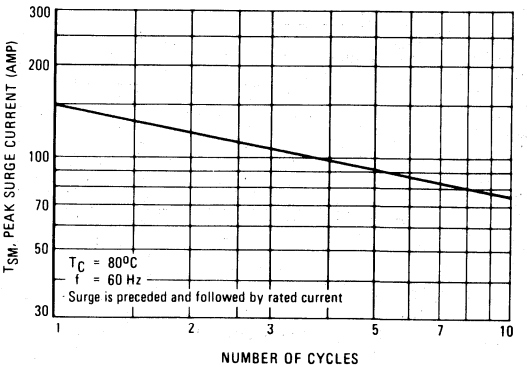
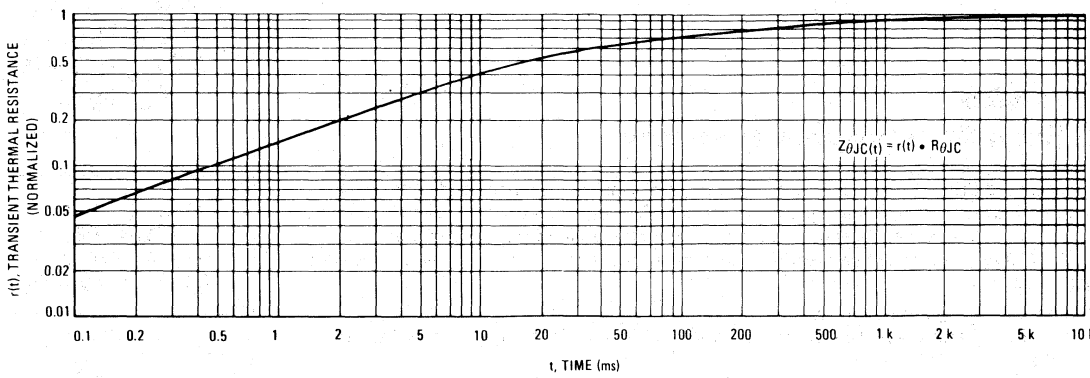


FIGURE 8 – THERMAL RESPONSE



3

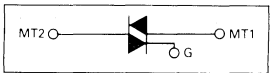
Triacs
Silicon Bidirectional Thyristors

MAC15FP Series
MAC15AFP Series

ISOLATED TRIACs
THYRISTORS
15 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC15FP Series) or Four Modes (MAC15AFP Series)



CASE 221C-02
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC15-4FP, MAC15A4FP MAC15-6FP, MAC15A6FP MAC15-8FP, MAC15A8FP MAC15-10FP, MAC15A10FP	V_{DRM}	 200 400 600 800	Volts
On-State RMS Current (T_C = +80°C), Note 2 Full Cycle Sine Wave 50 to 60 Hz (T_C = +95°C)	$I_{T(RMS)}$	15 12	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +80°C) preceded and followed by rated current	I_{TSM}	150	Amps
Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current	I_{GM}	2	Amps
Peak Gate Voltage	V_{GM}	10	Volts
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

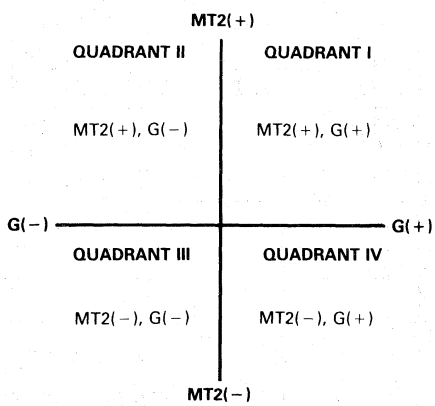
THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) T_J = 25°C	I_{DRM}	—	—	10	μA
Rated V_{DRM} @ T_J = +125°C, Gate Open		—	—	2	mA
Peak On-State Voltage (Either Direction) I_{TM} = 21 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.3	1.6	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	I_{GT}				mA
MT2(+), G(+)		—	—	50	
MT2(+), G(−)		—	—	50	
MT2(−), G(−)		—	—	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	75	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +110°C					
MT2(+), G(+); MT2(+), G(-); MT2(-), G(—)		0.2	—	—	
MT2(-), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 200 mA	I_H	—	6	40	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 17 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 21 A, Commutating di/dt = 8 A/ms, Gate Unenergized, T_C = +80°C	dv/dt(c)	—	5	—	V/μs

QUADRANT DEFINITIONS



Trigger devices are recommended for gating on Triacs. They provide:

1. Consistent predictable turn-on points.
2. Simplified circuitry.
3. Fast turn-on time for cooler, more efficient and reliable operation.

ELECTRICAL CHARACTERISTICS of RECOMMENDED BIDIRECTIONAL SWITCHES

Usage	General	
Part Number	MBS4991	MBS4992
V_S	6–10 V	7.5–9 V
I_S	350 μA Max	120 μA Max
V_{S1}–V_{S2}	0.5 V Max	0.2 V Max
Temperature Coefficient	0.02%/°C Typ	

1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

TYPICAL CHARACTERISTICS

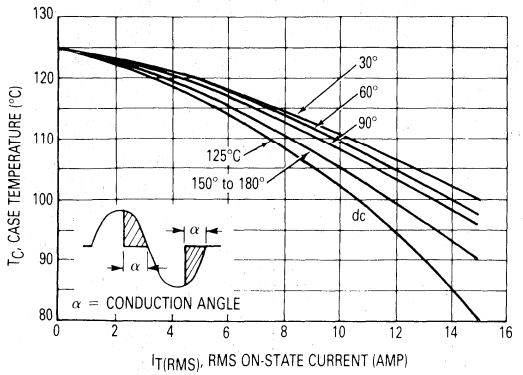


Figure 1. RMS Current Derating

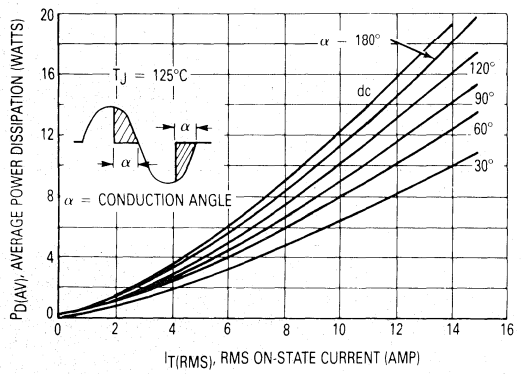


Figure 2. On-State Power Dissipation

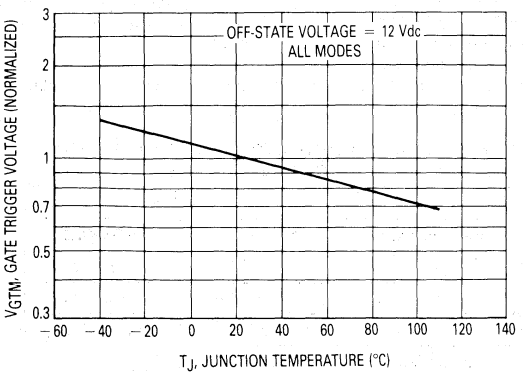


Figure 3. Typical Gate Trigger Voltage

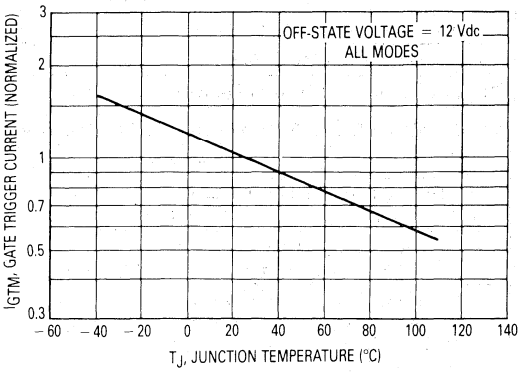


Figure 4. Typical Gate Trigger Current

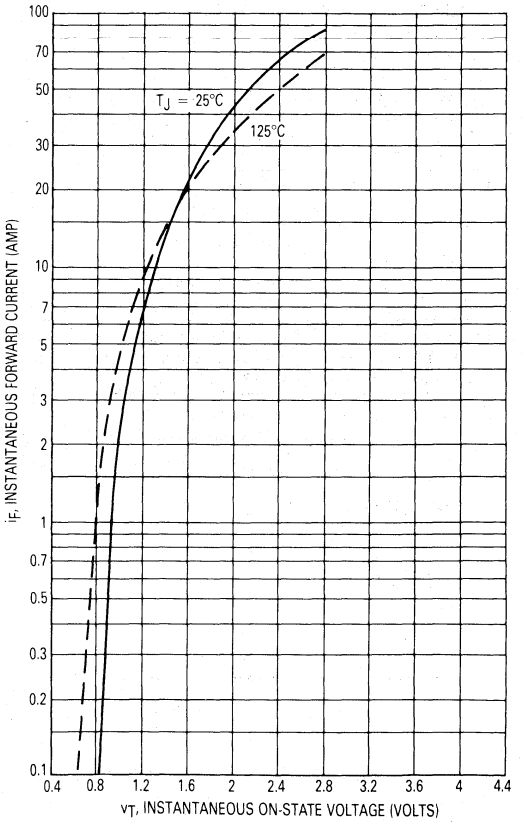


Figure 5. Maximum On-State Characteristics

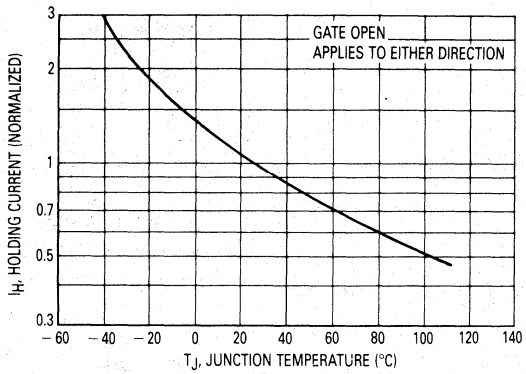


Figure 6. Typical Holding Current

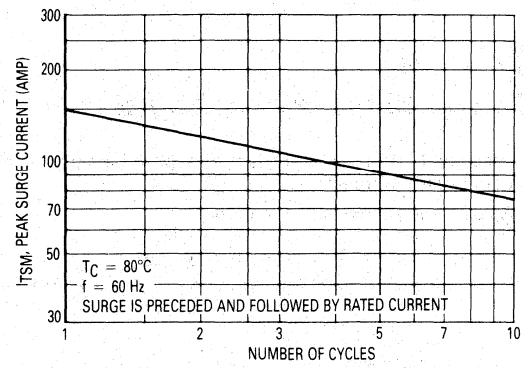


Figure 7. Maximum Nonrepetitive Surge Current

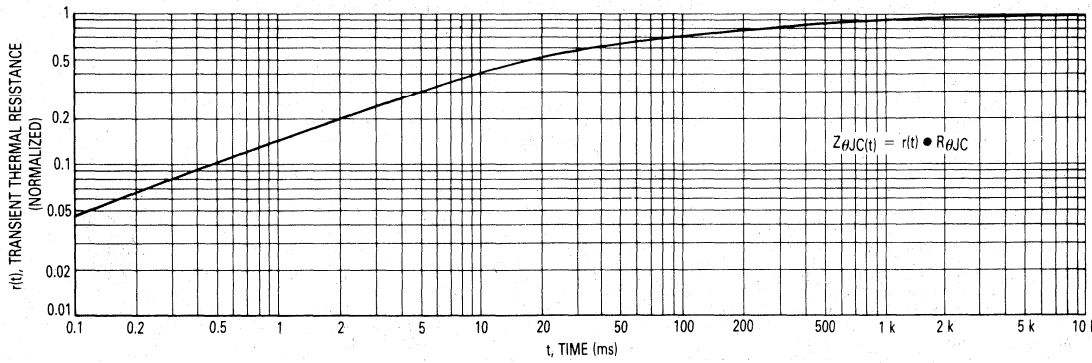


Figure 8. Thermal Response

3

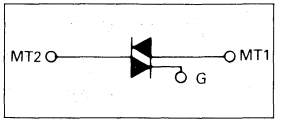
Triacs
Silicon Bidirectional Thyristors

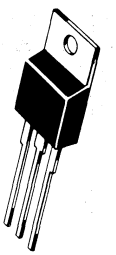
MAC16 Series

TRIACs
15 AMPERES RMS
200 thru 800 VOLTS

. . . designed for full-wave ac control applications primarily in industrial environments needing noise immunity.

- Guaranteed High Noise Immunity
  dv/dt(s) — 500 V/μs Min @ T_C = 25°C
- High Blocking Voltage — V_{DRM} to 800 V
- Photo Glass Passivated Junction for Improved Power Cycling Capability and Reliability





CASE 221A-04
(TO-220AB)
STYLE 4

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to 125°C) MAC16-4	V_{DRM}	200	Volts
-6		400	
-8		600	
-10		800	
Peak Gate Voltage	V_{GM}	±10	Volts
On-State Current RMS Full Cycle Sine Wave 50 to 60 Hz (T_C = +90°C)	$I_{T(RMS)}$	15	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	93	A^2s
Peak Surge Current (One Full Cycle, 60 Hz, T_C = +80°C) Preceded and followed by rated current	I_{TSM}	150	Amps
Peak Gate Power (T_C = +80°C, Pulse Width = 2.0 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current	I_{GM}	2.0	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.0	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C, and either polarity of MT2 to MT1 Voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2.0	mA
Peak On-State Voltage (I_{TM} = 21 A Peak; Pulse Width ≤ 2.0 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.3	1.6	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	I_{GT}				mA
MT2(+), Gate(+)		—	—	100	
MT2(+), Gate(−)		—	—	100	
MT2(−), Gate(−)		—	—	100	

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

(continued)

ELECTRICAL CHARACTERISTICS — continued (T_C = +25°C, and either polarity of MT2 to MT1 Voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volts
(V_D = 12 Vdc, R_L = 100 Ohms)					
MT2(+), G(+)		—	0.9	2.0	
MT2(+), G(−)		—	0.9	2.0	
MT2(−), G(−)		—	1.1	2.0	
(V_D = Rated V_{DRM}, R_L = 10 k Ohms, T_J = +110°C)					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
Holding Current (Either Direction)	I_H	—	—	100	mA
(V_D = 12 Vdc, Gate Open)					
(I_T = 200 mA)					
Critical Rate of Rise of Off-State Voltage	dv/dt(s)				V/μs
(V_D = Rated V_{DRM}, Exponential Voltage Rise, Gate Open)					
T_J = 25°C		500	—	—	
T_J = 125°C		200	—	—	

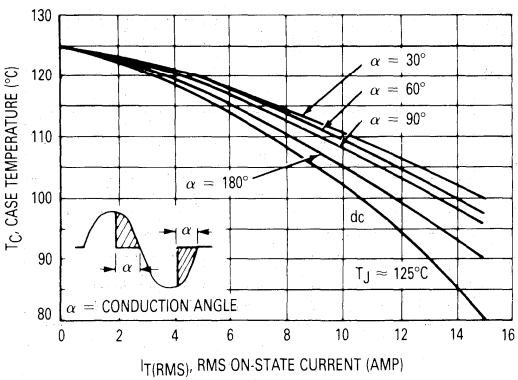


Figure 1. RMS Current Derating

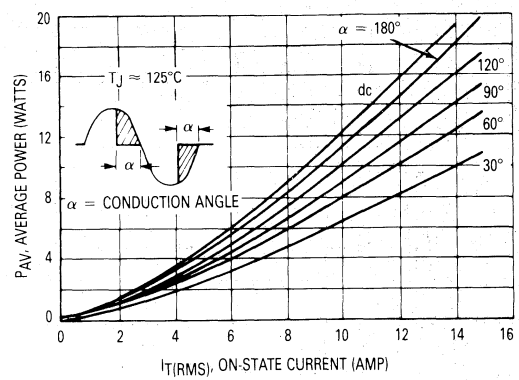


Figure 2. On-State Power Dissipation

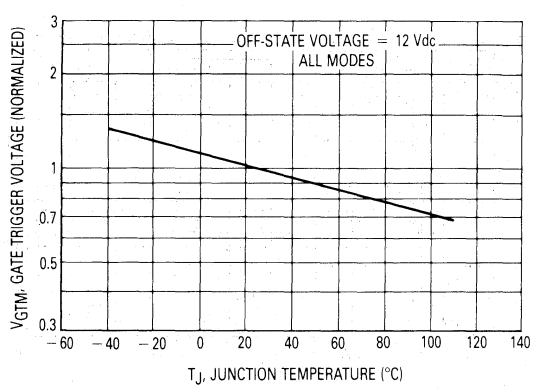


Figure 3. Typical Gate Trigger Voltage

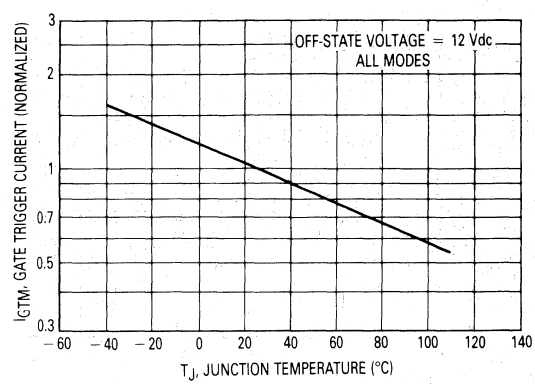


Figure 4. Typical Gate Trigger Current

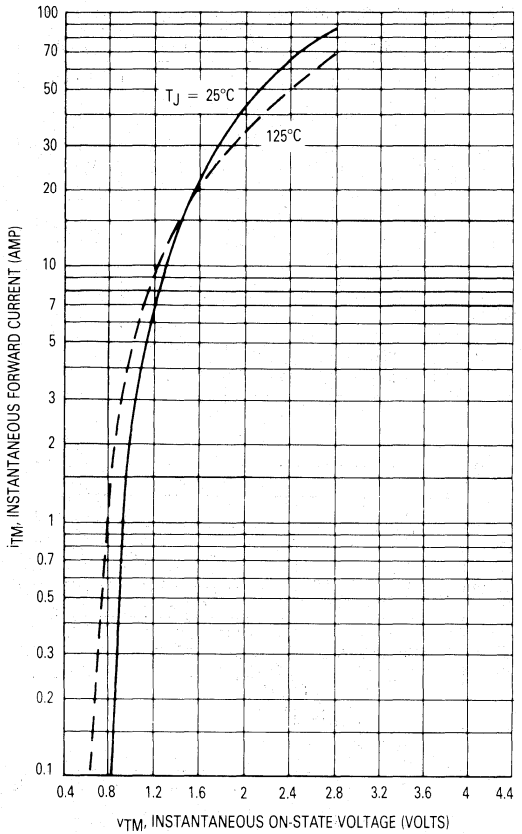


Figure 5. On-State Characteristics

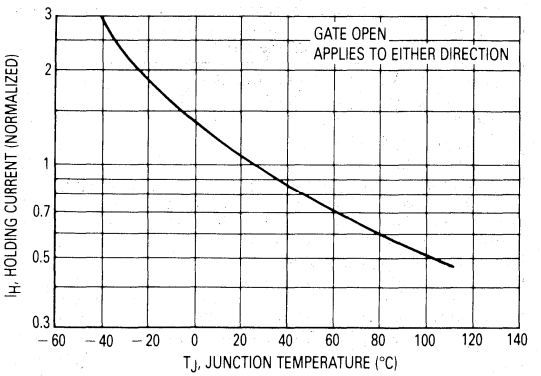


Figure 6. Typical Holding Current

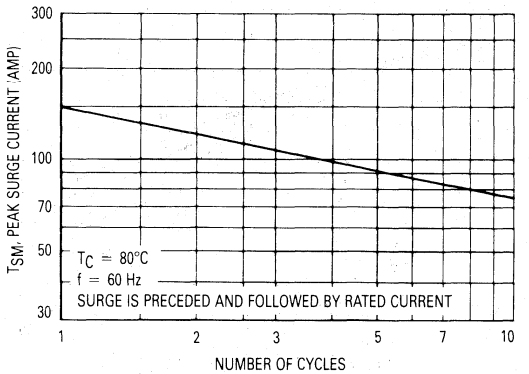


Figure 7. Maximum Non-Repetitive Surge Current

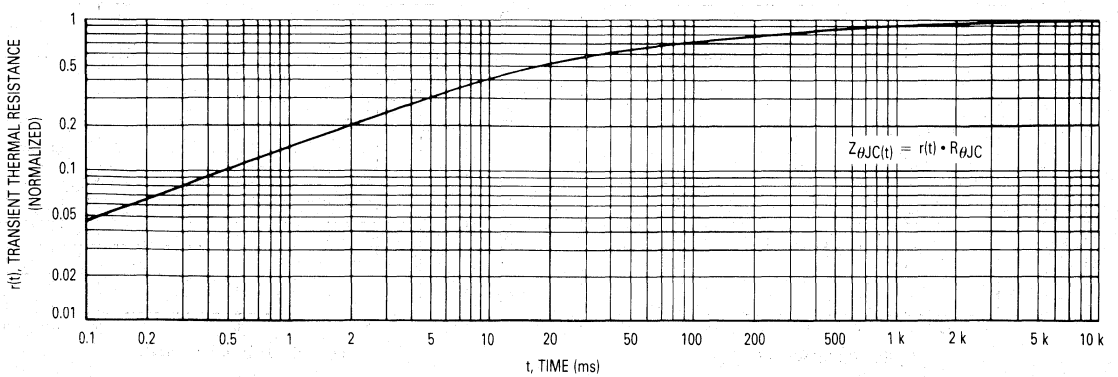


Figure 8. Thermal Response

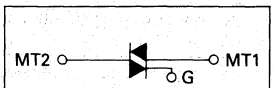
Triacs
Silicon Bidirectional Triode Thyristors

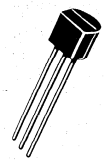
MAC97,A Series

TRIACs
0.6 AMPERE RMS
200 thru 600 VOLTS

. . . designed for use in solid state relays, MPU interface, TTL logic and any other light industrial or consumer application. Supplied in an inexpensive TO-226AA (TO-92) package which is readily adaptable for use in automatic insertion equipment.

- One-Piece, Injection-Molded Unibloc Package
- Sensitive Gate Triggering in Four Trigger Modes for all possible Combinations of Trigger Sources, and Especially Suitable for Circuits that Source Gate Drives.
- All Diffused and Glassivated Junctions for Maximum Uniformity of Parameters and Reliability
- Available in TO-5 or TO-206AA (TO-18) Leadforms





CASE 29-04
(TO-226AA)
STYLE 12

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage (T_J = −40 to +110°C) Note 1 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC97- 4 / or / MAC97A- 6 / 8	V_{DRM}	200 400 600	Volts
On-State RMS Current (Full Cycle Sine Wave 50 to 60 Hz, T_C = +50°C)	$I_{T(RMS)}$	0.6	Amp
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_A = 25°C)	I_{TSM}	8	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	0.26	A^2s
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	5	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	5	Watts
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.1	Watt
Peak Gate Current (t ≤ 2 μs)	I_{GM}	1	Amp
Operating Junction Temperature Range	T_J	−40 to +110	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	8	0.1	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 0.85 A Peak; Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.9	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	I_{GT}	See Table 1			mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 100 Ohms)	V_{GT}				Volts
MT2(+), G(+) All Types		—	—	2	
MT2(+), G(−) All Types		—	—	2	
MT2(−), G(−) All Types		—	—	2	
MT2(−), G(+) All Types		—	—	2.5	
(V_D = Rated V_{DRM}, R_L = 10 k ohms, T_J = 110°C) MT2(+), G(+); MT2(−), G(−); MT2(+), G(−) All Types		0.1	—	—	
MT2(−), G(+) All Types		0.1	—	—	
Holding Current (V_D = 12 Vdc, I_{TM} = 200 mA, Gate Open)	I_H	—	—	10	mA
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 1 A pk, I_G = 25 mA)	t_{gt}	—	2	—	μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 0.84 μA pk) (Commutating di/dt = 0.32 A/ms, Gate Unenergized, T_C = 50°C)	dv/dt(c)	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM} Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

3

QUADRANT DEFINITIONS

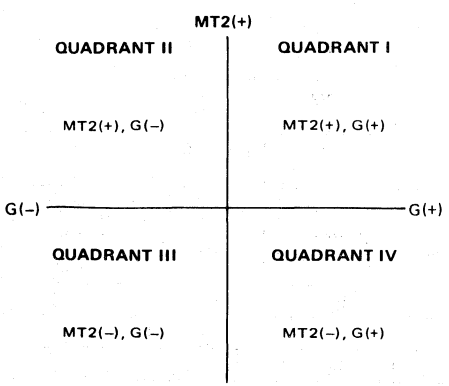


TABLE 1 — MAXIMUM GATE TRIGGER CURRENTS (V_D = 12 V, R_L = 100 Ω)

Quadrant and Polarity	MAC Series 97	MAC Series 97A	Unit
I MT2(+), G(+)	10	5	mA
II MT2(+), G(−)	10	5	mA
III MT2(−), G(−)	10	5	mA
IV MT2(−), G(+)	10	7	mA

FIGURE 1 — RMS CURRENT DERATING
(Reference: Case Temperature)

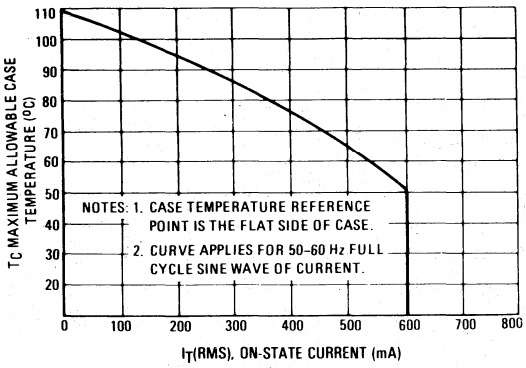


FIGURE 2 — ON-STATE CHARACTERISTICS

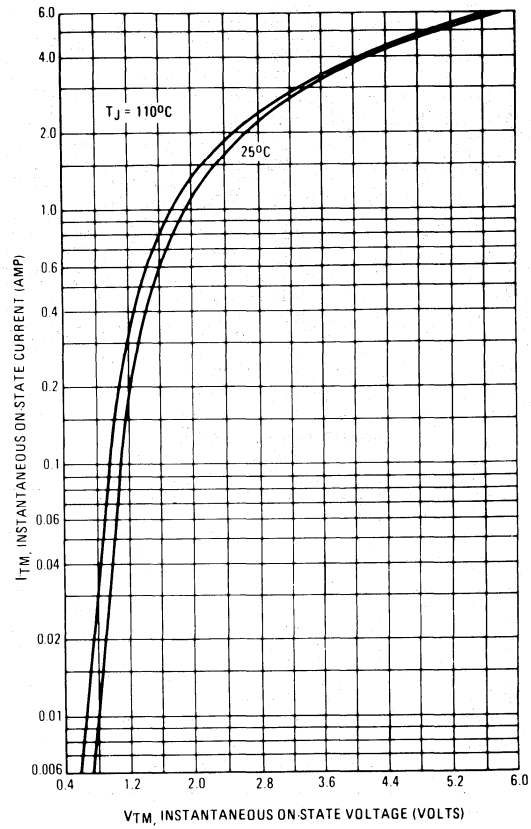


FIGURE 3 — RMS CURRENT DERATING
(Reference: Ambient Temperature)

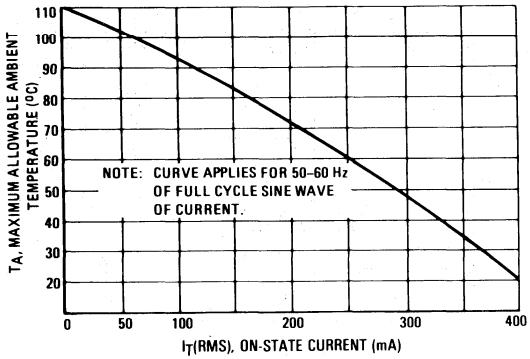


FIGURE 4 — ON-STATE POWER DISSIPATION

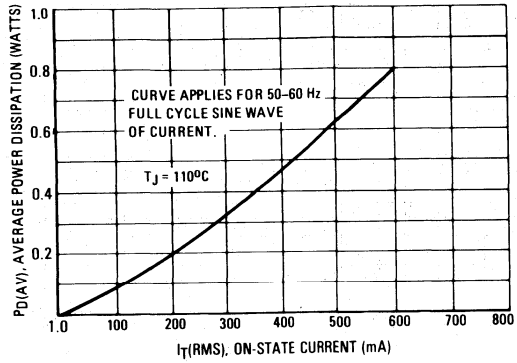


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FIGURE 5 – NORMALIZED GATE TRIGGER CURRENT

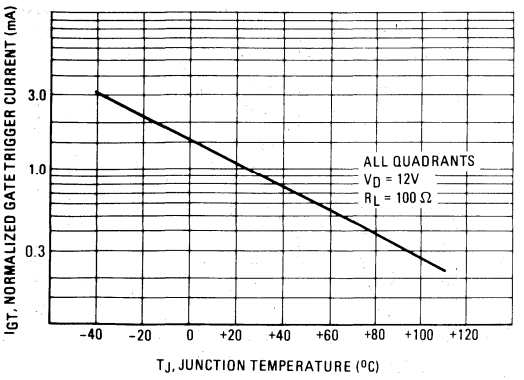


FIGURE 6 – NORMALIZED GATE TRIGGER VOLTAGE

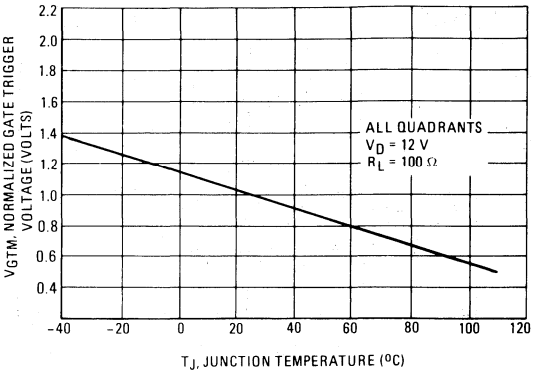


FIGURE 7 – NORMALIZED HOLD CURRENT

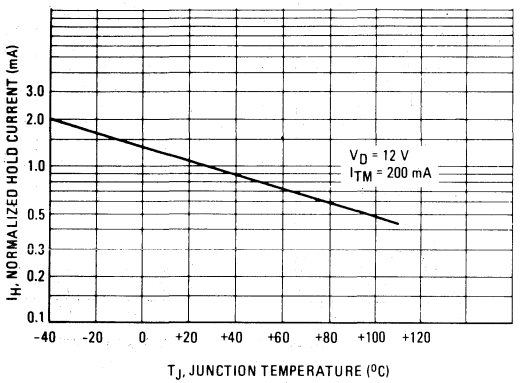


FIGURE 8 – MAXIMUM ALLOWABLE SURGE CURRENT

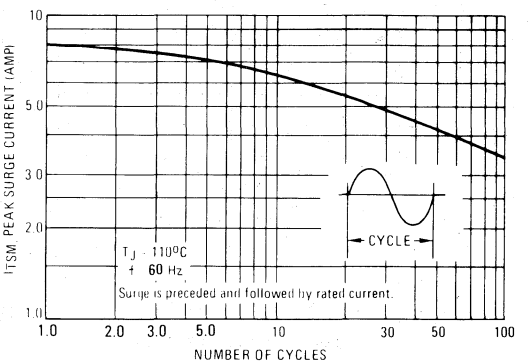
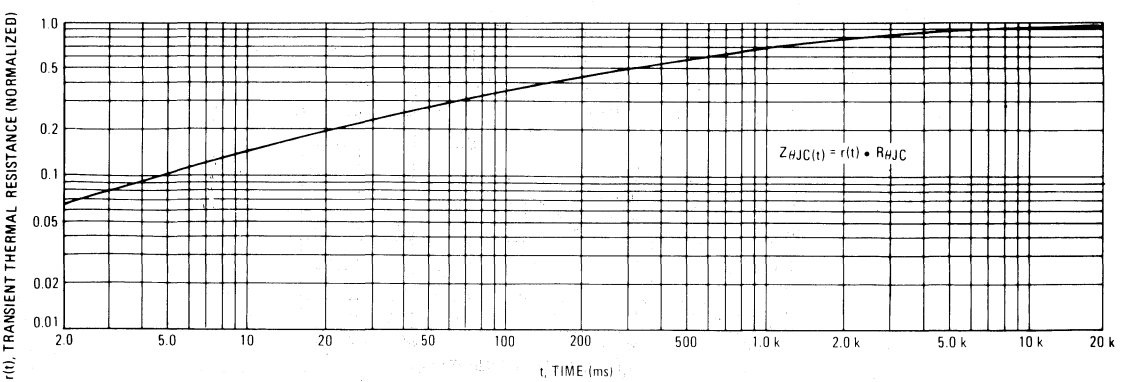


FIGURE 9 – THERMAL RESPONSE



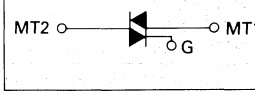
Triacs
Silicon Bidirectional Thyristors

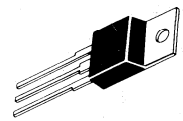
MAC210 Series
MAC210A Series

TRIACs
10 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC210 Series) or Four Modes (MAC210A Series)





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	MAC210-4, MAC210A4	V_{DRM}	200	Volts
	MAC210-6, MAC210A6		400	
	MAC210-8, MAC210A8		600	
	MAC210-10, MAC210A10		800	
On-State Current RMS (T_C = +70°C) Full Cycle Sine Wave 50 to 60 Hz		$I_{T(RMS)}$	10	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +70°C) Preceded and followed by Rated Current		I_{TSM}	100	Amps
Circuit Fusing Considerations (t = 8.3 ms)		I^2t	40	A^2s
Peak Gate Power (T_C = +70°C, Pulse Width = 10 μs)		P_{GM}	20	Watts
Average Gate Power (T_C = +70°C, t = 8.3 ms)		$P_{G(AV)}$	0.35	Watt
Peak Gate Current (T_C = +70°C, Pulse Width = 10 μs)		I_{GM}	2	Amps
Operating Junction Temperature Range		T_J	−40 to +125	°C
Storage Temperature Range		T_{stg}	−40 to +125	°C

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = +125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 14 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.2	1.65	Volts
Gate Trigger Current (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms)	I_{GT}				mA
MT2(+), G(+)		—	12	50	
MT2(+), G(−)		—	12	50	
MT2(−), G(−)		—	20	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	35	75	
Gate Trigger Voltage (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms)	V_{GT}				volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
(Main Terminal Voltage = Rated V_{DRM}, R_L = 10 k ohms, T_J = +125°C)					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) (Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 500 mA, T_C = +25°C)	I_H	—	6	50	mA
Turn-On Time (Rated V_{DRM}, I_{TM} = 14 A) (I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage (Rated V_{DRM}, I_{TM} = 14 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 70°C)	dv/dt(c)	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = +70°C)	dv/dt	—	100	—	V/μs

3

FIGURE 1 — CURRENT DERATING

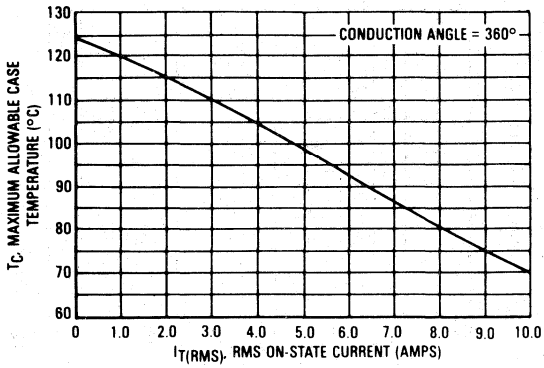


FIGURE 2 — POWER DISSIPATION

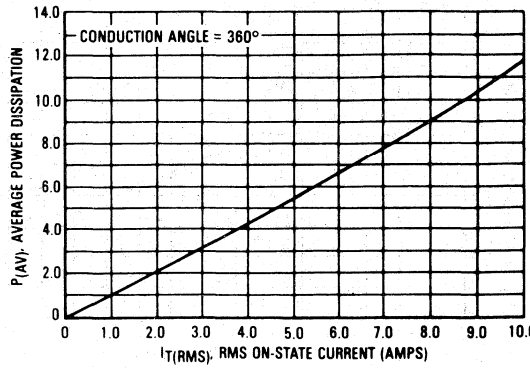


FIGURE 3 — MAXIMUM ON-STATE CHARACTERISTICS

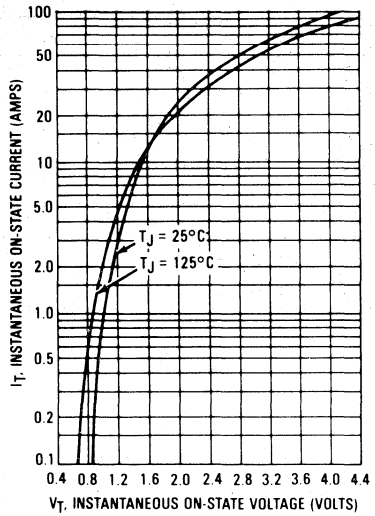


FIGURE 4 — MAXIMUM NON-REPETITIVE SURGE CURRENT

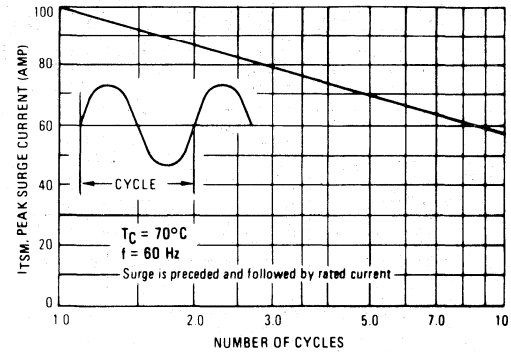


FIGURE 5 — TYPICAL GATE TRIGGER VOLTAGE

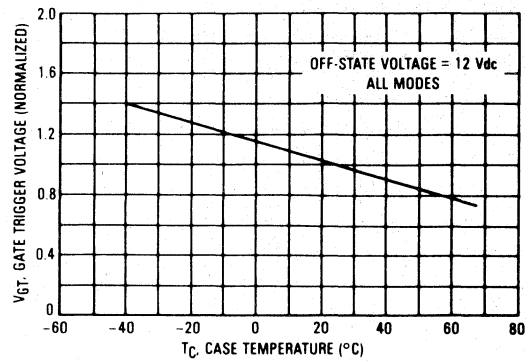


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FIGURE 6 — TYPICAL GATE TRIGGER CURRENT

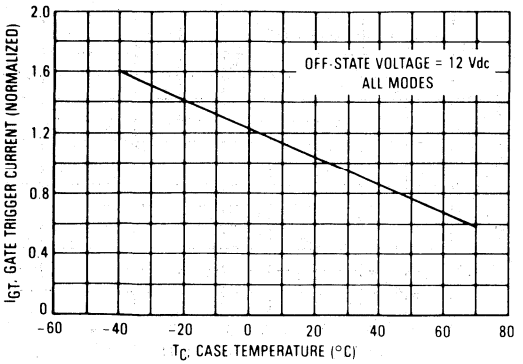


FIGURE 7 — TYPICAL HOLDING CURRENT

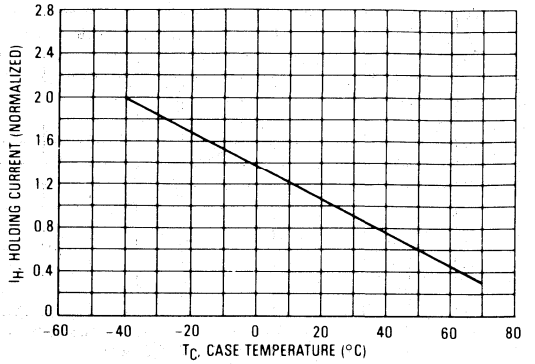
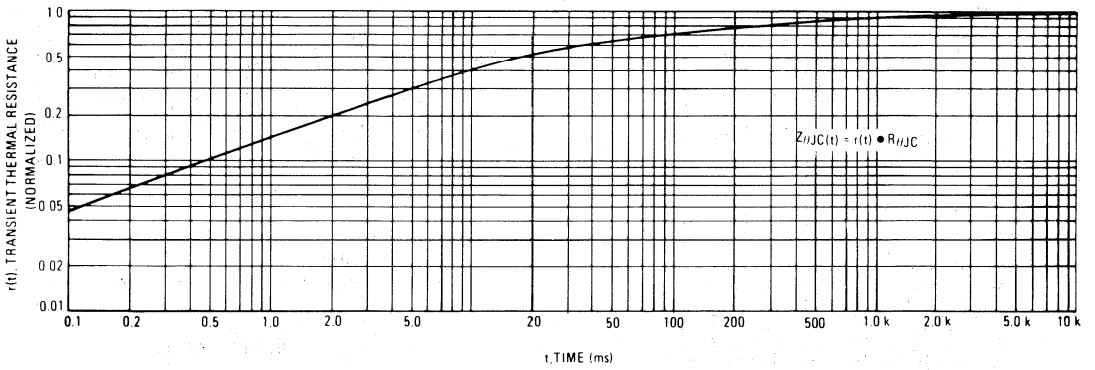


FIGURE 8 — THERMAL RESPONSE



3

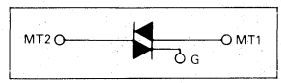
Triacs
Silicon Bidirectional Thyristors

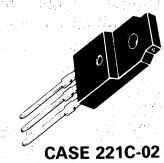
MAC210FP Series
MAC210AFP Series

ISOLATED TRIACs THYRISTORS
10 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC210FP Series) or Four Modes (MAC210AFP Series)





CASE 221C-02
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC210-4FP, MAC210A4FP MAC210-6FP, MAC210A6FP MAC210-8FP, MAC210A8FP MAC210-10FP, MAC210A10FP	V_{DRM}	 200 400 600 800	Volts
On-State RMS Current (T_C = +70°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	10	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +70°C) preceded and followed by rated current	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +70°C, Pulse Width = 10 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +70°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Current (T_C = +70°C, Pulse Width = 10 μs)	I_{GM}	2	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) I_{TM} = 14 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.2	1.65	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms Minimum Gate Pulse Width = 2 μs	I_{GT}				mA
MT2(+), G(+)		—	12	50	
MT2(+), G(−)		—	12	50	
MT2(−), G(−)		—	20	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	35	75	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms Minimum Gate Pulse Width = 2 μs	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C					
MT2(+), G(+); MT2(+), G(-); MT2(-), G(-)		0.2	—	—	
MT2(-), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 500 mA, T_C = +25°C	I_H	—	6	50	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 14 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 14 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = +70°C	$dv/dt_{(c)}$	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = +70°C)	dv/dt	—	100	—	V/μs

3

TYPICAL CHARACTERISTICS

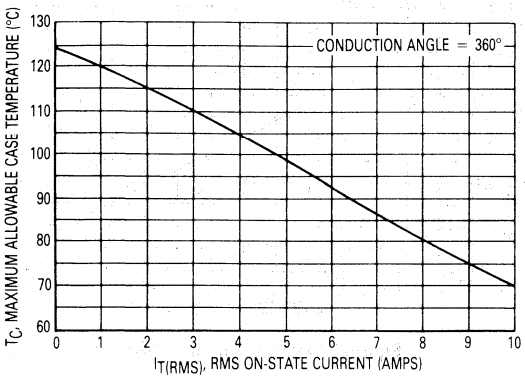


Figure 1. Current Derating

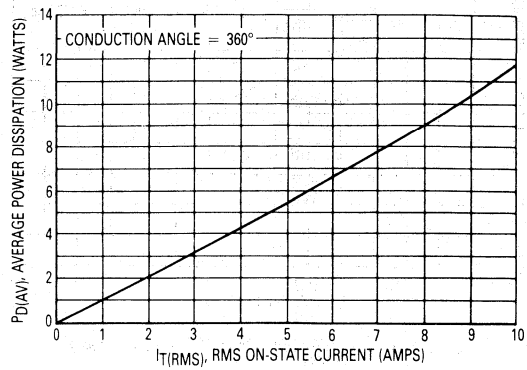


Figure 2. Power Dissipation

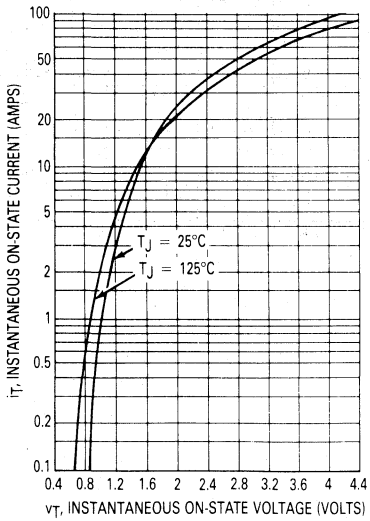


Figure 3. Maximum On-State Characteristics

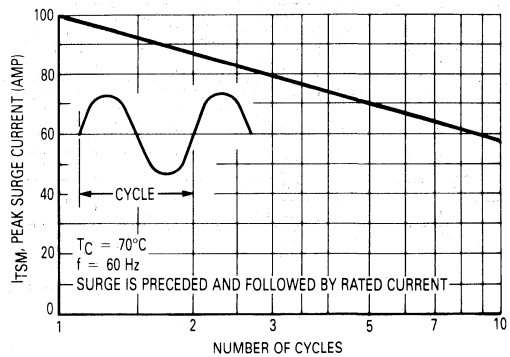


Figure 4. Maximum Nonrepetitive Surge Current

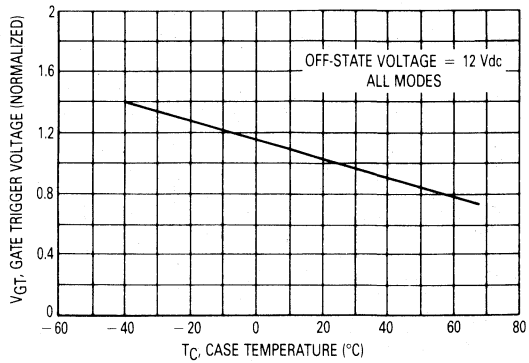


Figure 5. Typical Gate Trigger Voltage

3

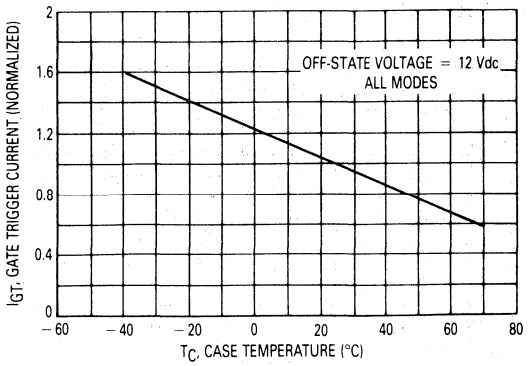


Figure 6. Typical Gate Trigger Current

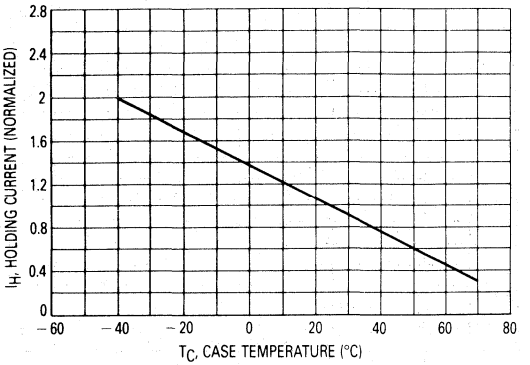


Figure 7. Typical Holding Current

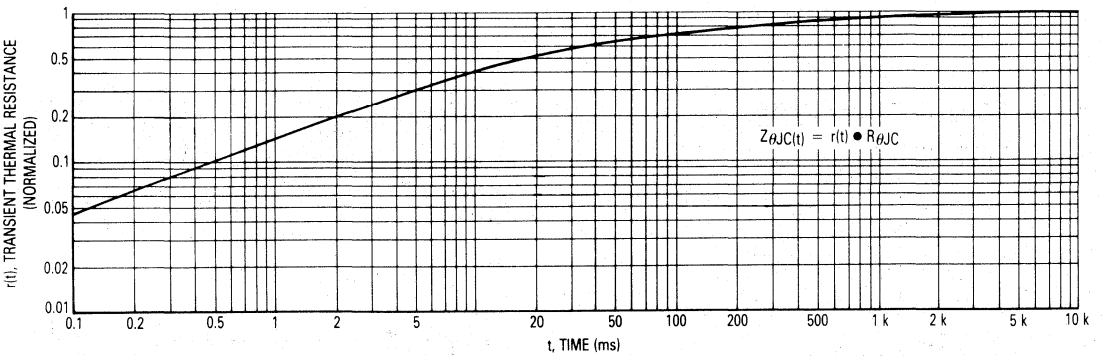


Figure 8. Thermal Response

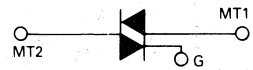
Triacs
Silicon Bidirectional Thyristors

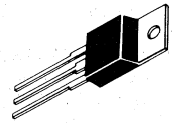
MAC212 Series
MAC212A Series

TRIACs
12 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC212 Series) or Four Modes (MAC212A Series)





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC212-4, MAC212A4 MAC212-6, MAC212A6 MAC212-8, MAC212A8 MAC212-10, MAC212A10	V_{DRM}	 200 400 600 800	Volts
On-State Current RMS (T_C = +85°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	12	Amp
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +85°C) preceded and followed by Rated Current	I_{TSM}	100	Amp
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +85°C, Pulse Width = 10 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +85°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Current (T_C = +85°C, Pulse Width = 10 μs)	I_{GM}	2	Amp
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.1	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) I_{TM} = 17 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.3	1.75	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	I_{GT}				mA
MT2(+), G(+)		—	12	50	
MT2(+), G(−)		—	12	50	
MT2(−), G(−)		—	20	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	35	75	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 500 mA, T_C = +25°C	I_H	—	6	50	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 17 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 17 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = +85°C	$dv/dt_{(c)}$	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = +85°C)	dv/dt	—	100	—	V/μs

3

FIGURE 1 — CURRENT DERATING

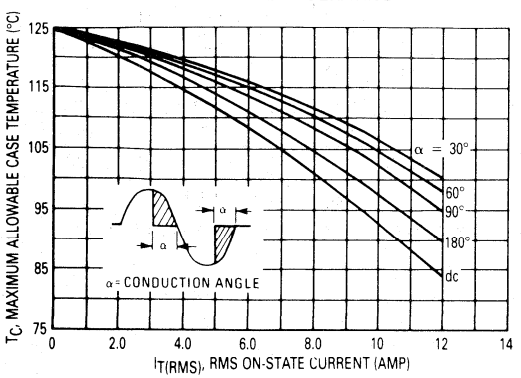


FIGURE 2 — POWER DISSIPATION

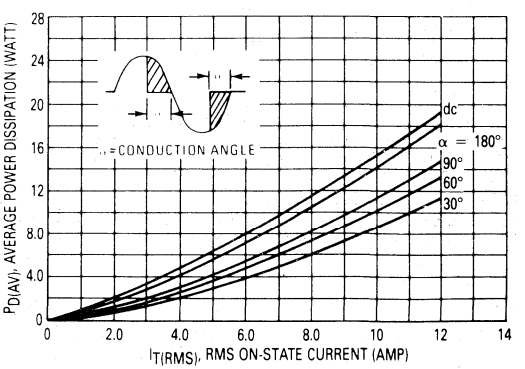


FIGURE 3 — MAXIMUM ON-STATE CHARACTERISTICS

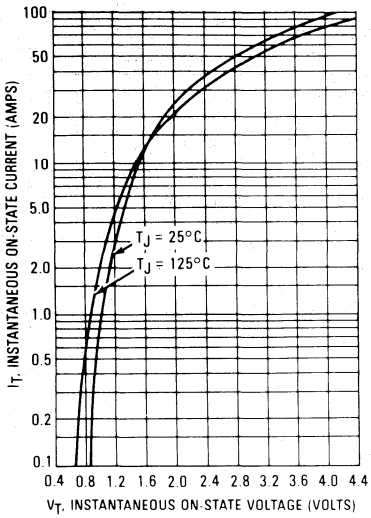


FIGURE 4 — MAXIMUM NON-REPETITIVE SURGE CURRENT

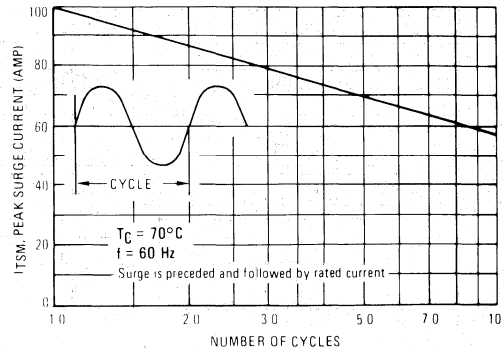


FIGURE 5 — TYPICAL GATE TRIGGER VOLTAGE

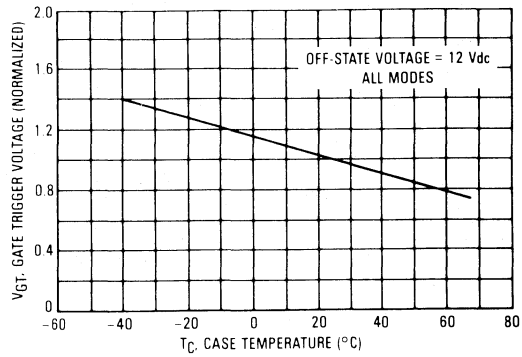


FIGURE 6 — TYPICAL GATE TRIGGER CURRENT

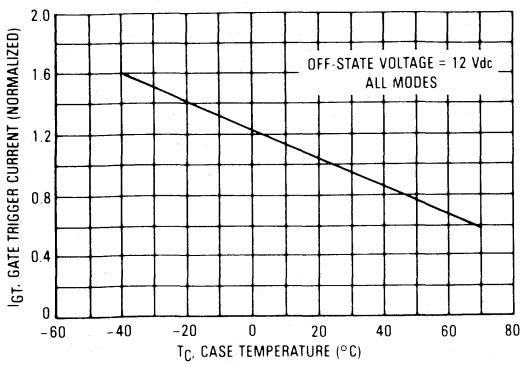


FIGURE 7 — TYPICAL HOLDING CURRENT

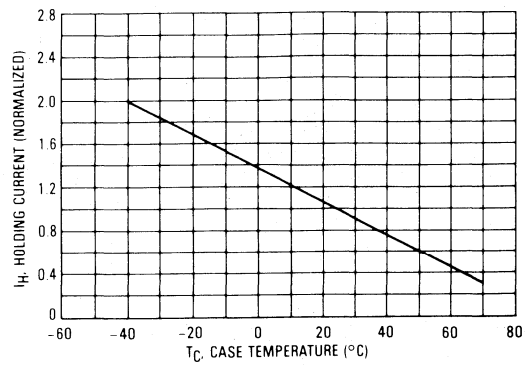


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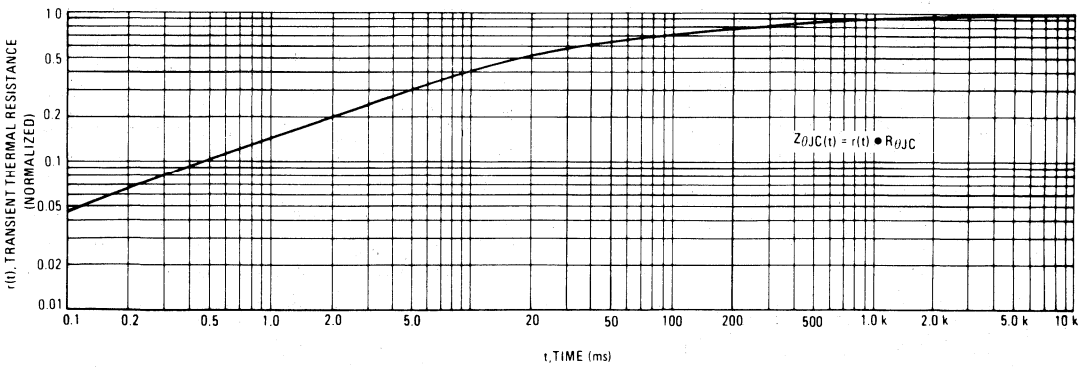
FIGURE 8 — THERMAL RESPONSE



3

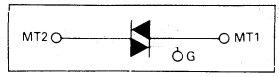
Triacs
Silicon Bidirectional Thyristors

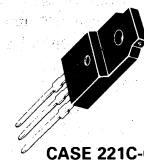
MAC212FP Series
MAC212AFP Series

ISOLATED TRIACs
THYRISTORS
12 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC212FP Series) or Four Modes (MAC212AFP Series)





CASE 221C-02
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC212-4FP, MAC212A4FP MAC212-6FP, MAC212A6FP MAC212-8FP, MAC212A8FP MAC212-10FP, MAC212A10FP	V_{DRM}	 200 400 600 800	Volts
On-State RMS Current (T_C = +85°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	12	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +85°C) preceded and followed by rated current	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +85°C, Pulse Width = 10 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +85°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Current (T_C = +85°C, Pulse Width = 10 μs)	I_{GM}	2	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.1	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) I_{TM} = 17 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.3	1.75	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms Minimum Gate Pulse Width = 2 μs	I_{GT}				mA
MT2(+), G(+)		—	12	50	
MT2(+), G(−)		—	12	50	
MT2(−), G(−)		—	20	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	35	75	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms Minimum Gate Pulse Width = 2 μs	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C					
MT2(+), G(+); MT2(+), G(-); MT2(-), G(-)		0.2	—	—	
MT2(-), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 500 mA, T_C = +25°C	I_H	—	6	50	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 17 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 17 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = +85°C	$dv/dt_{(c)}$	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = +85°C)	dv/dt	—	100	—	V/μs

3

TYPICAL CHARACTERISTICS

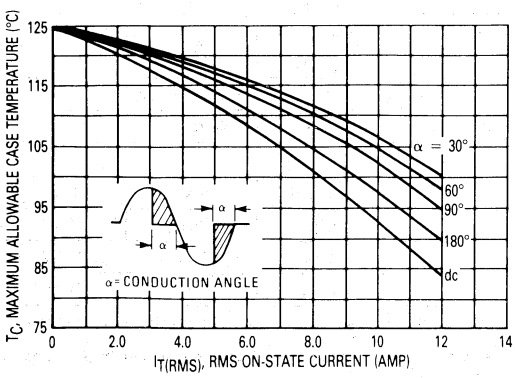


Figure 1. Current Derating

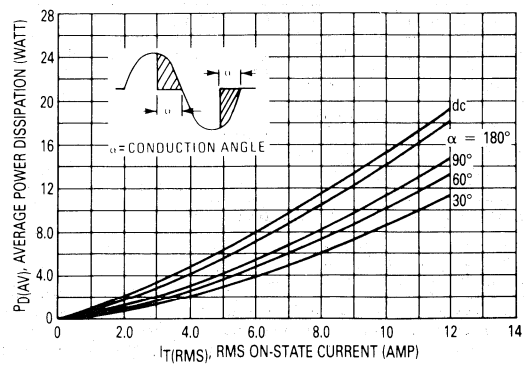


Figure 2. Power Dissipation

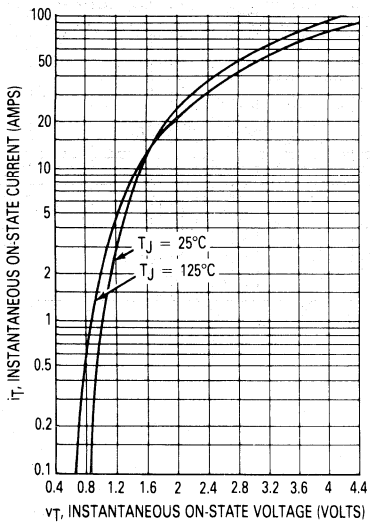


Figure 3. Maximum On-State Characteristics

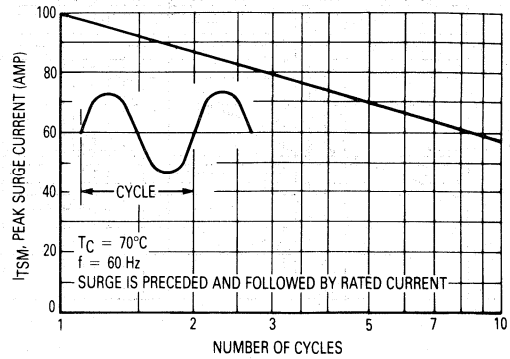


Figure 4. Maximum Nonrepetitive Surge Current

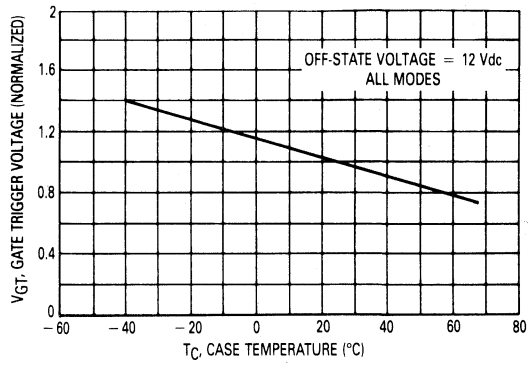


Figure 5. Typical Gate Trigger Voltage

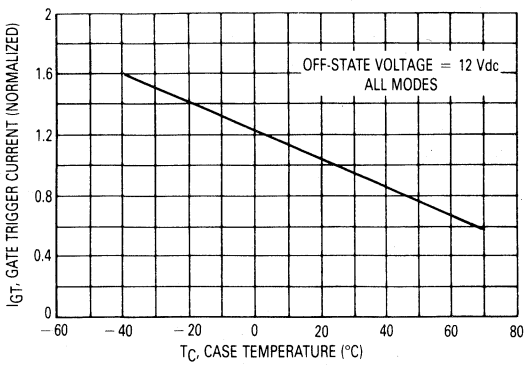


Figure 6. Typical Gate Trigger Current

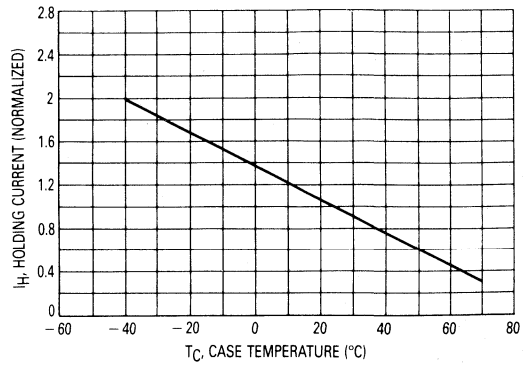


Figure 7. Typical Holding Current

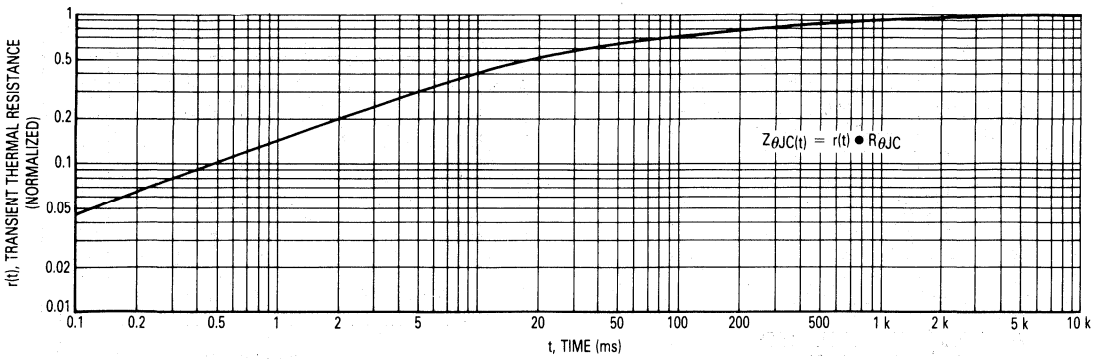


Figure 8. Thermal Response

3

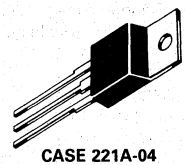
Triacs
Silicon Bidirectional Thyristors

MAC213 Series

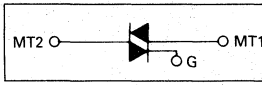
TRIACs
12 AMPERES RMS
200 thru 800 VOLTS

. . . designed for full-wave ac control applications primarily in industrial environments needing noise immunity.

- Guaranteed High Commutation Voltage
  dv/dt — 500 V/μs Min @ T_C = 25°C
- High Blocking Voltage — V_{DRM} to 800 V
- Photo Glass Passivated Junction for Improved Power Cycling Capability and Reliability



CASE 221A-04
(TO-220AB)
STYLE 4



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C)	V_{DRM}		Volts
MAC213-4		200	
-6		400	
-8		600	
-10		800	
Peak Gate Voltage	V_{GM}	10	Volts
On-State Current RMS (T_C = +85°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	12	Amp
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +85°C) preceded and followed by Rated Current	I_{TSM}	100	Amp
Circuit Fusing Considerations (T_C = +85°C, t = 1.0 to 8.3 ms)	I^2t	41	A^2s
Peak Gate Power (T_C = +85°C, Pulse Width = 10 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +85°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Current (T_C = +85°C, Pulse Width = 10 μs)	I_{GM}	2.0	Amp
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.1	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2.0	mA
Peak On-State Voltage (Either Direction) I_{TM} = 17 A Peak; Pulse Width ≤ 2.0 ms, Duty Cycle ≤ 2.0%	V_{TM}	—	1.3	1.75	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	I_{GT}				mA
MT2(+), G(+)		—	—	100	
MT2(+), G(−)		—	—	100	
MT2(−), G(−)		—	—	100	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	—	2.0	
MT2(+), G(−)		—	—	2.0	
MT2(−), G(−)		—	—	2.0	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 200 mA, T_C = +25°C	I_H	—	—	100	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 17 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2.0 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage V_D = Rated V_{DRM}, Exponential Voltage Rise, Gate Open	dv/dt(s)				V/μs
T_J = 25°C		500	—	—	
T_J = +125°C		200	—	—	

3

TYPICAL CHARACTERISTICS

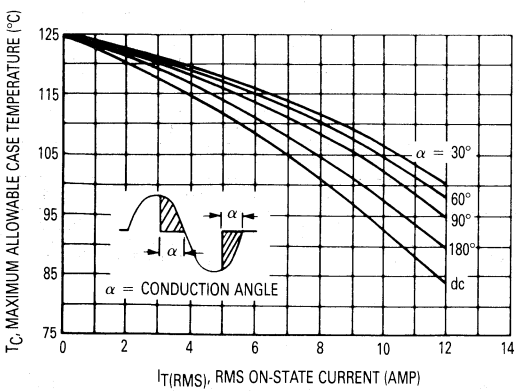


Figure 1. Current Derating

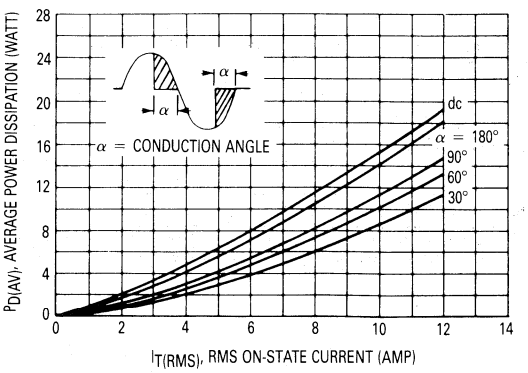


Figure 2. Power Dissipation

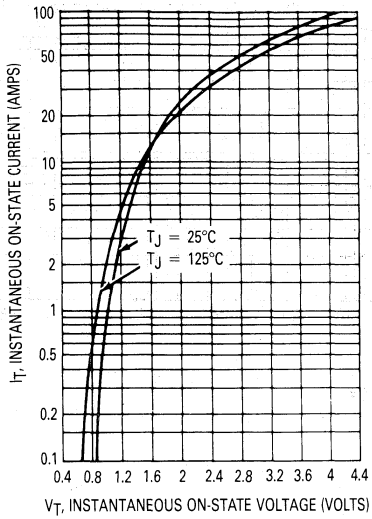


Figure 3. Maximum On-State Characteristics

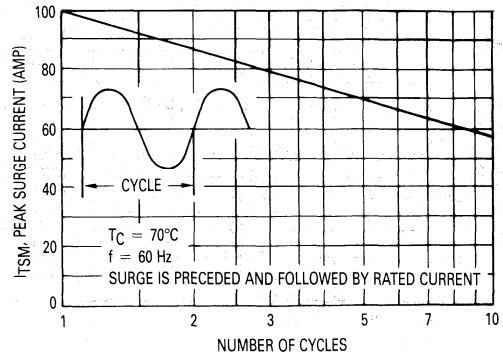


Figure 4. Maximum Non-Repetitive Surge Current

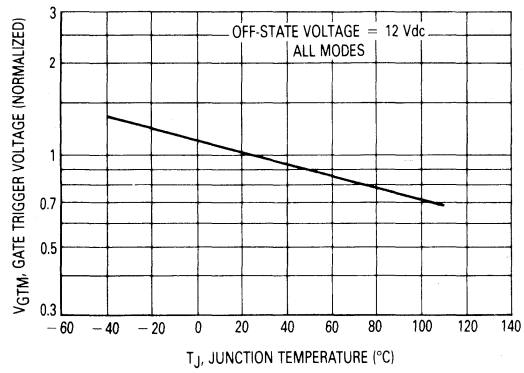


Figure 5. Typical Gate Trigger Voltage

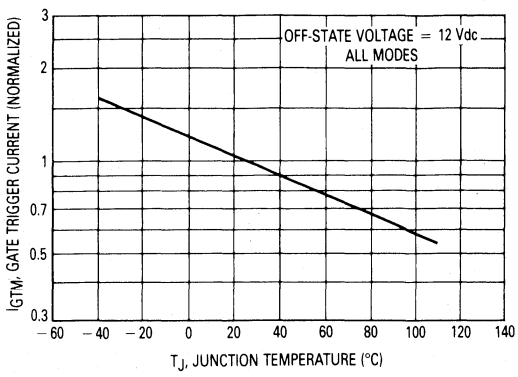


Figure 6. Typical Gate Trigger Current

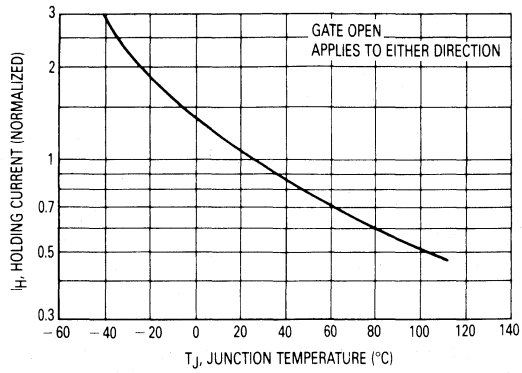


Figure 7. Typical Holding Current

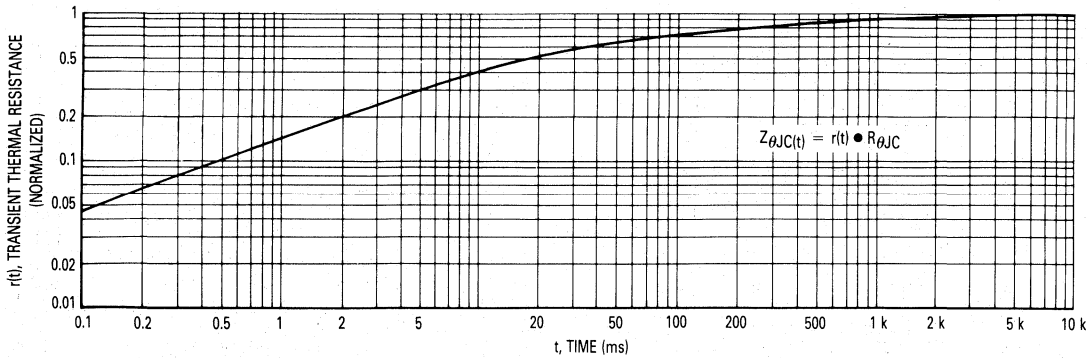


Figure 8. Thermal Response

3

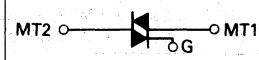
Triacs
Silicon Bidirectional Thyristors

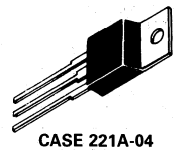
MAC218 Series
MAC218A Series

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Blocking Voltage to 800 Volts
- Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- TO-220 Construction Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC218 Series) or Four Modes (MAC218A Series)





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 Gate Open MAC218-4, MAC218A4 MAC218-6, MAC218A6 MAC218-8, MAC218A8 MAC218-10, MAC218A10	V_{DRM}	 200 400 600 800	Volts
On-State Current RMS (Conduction Angle = 360°, T_C = +80°C)	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = 80°C, preceded and followed by rated current)	I_{TSM}	100	Amps
Fusing Current (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)	P_{GM}	16	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Trigger Current (Pulse Width = 1 μs)	I_{GTM}	4	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 11.3 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle < 2%)	V_{TM}	—	1.7	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 12Ω)	I_{GT}				mA
Trigger Mode					
MT2(+), Gate(+); MT2(+), Gate(−); MT2(−), Gate(−)		—	—	50	
MT2(−), Gate(+) "A" SUFFIX ONLY		—	—	75	
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volts
Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms					
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) (V_D = 24 Vdc, Gate Open, Initiating Current = 200 mA)	I_H	—	—	50	mA
Critical Rate of Rise of Commutating Off-State Voltage (Rated V_{DROM}, $I_{T(RMS)}$ = 6 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_J = 125°C)	dv/dt	—	100	—	V/μs

3

FIGURE 1 — CURRENT DERATING

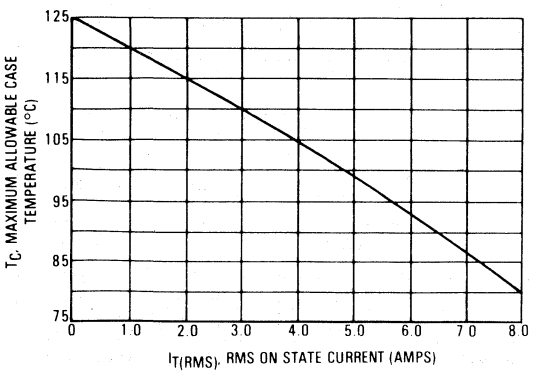


FIGURE 2 — POWER DISSIPATION

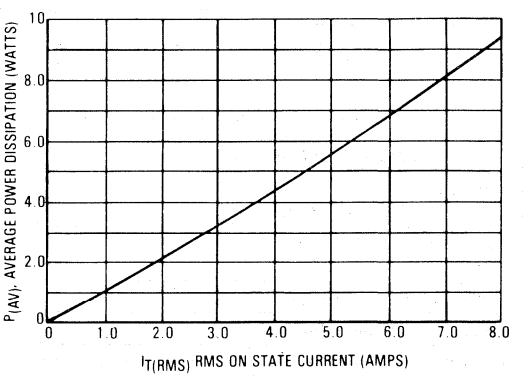


FIGURE 3 — NORMALIZED GATE TRIGGER CURRENT

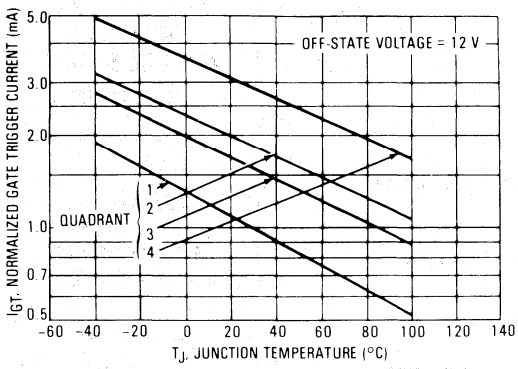


FIGURE 4 — NORMALIZED GATE TRIGGER VOLTAGE

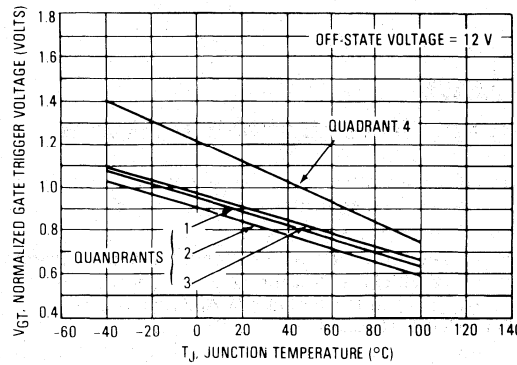
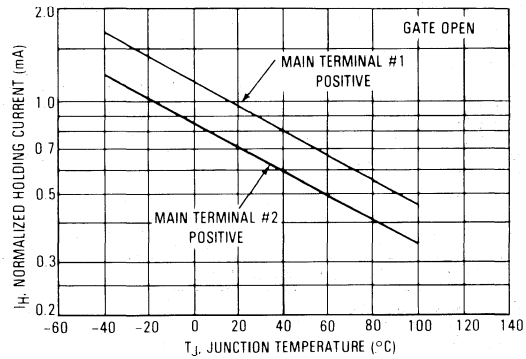


FIGURE 5 — NORMALIZED HOLDING CURRENT



3

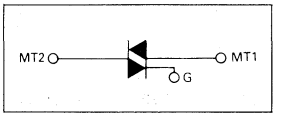
Triacs
Silicon Bidirectional Thyristors

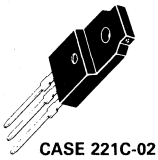
MAC218FP Series
MAC218AFP Series

ISOLATED TRIACs THYRISTORS
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Blocking Voltage to 800 Volts
- Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Isolated TO-220 Type Package for Ease of Mounting
- Gate Triggering in Three Modes (MAC218FP Series) or Four Modes (MAC218AFP Series)





CASE 221C-02
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC218-4FP, MAC218A4FP MAC218-6FP, MAC218A6FP MAC218-8FP, MAC218A8FP MAC218-10FP, MAC218A10FP	V_{DRM}	 200 400 600 800	Volts
On-State RMS Current (T_C = +80°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	8	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz) preceded and followed by rated current, T_C = 80°C	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)	P_{GM}	16	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Current (Pulse Width = 1 μs)	I_{GM}	4	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Off-State Current (Either Direction) (Rated V_{DROM} @ T_J = 125°C, Gate Open)	I_{DROM}	—	—	2	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 11.3 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle < 2%)	V_{TM}	—	1.7	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 12 Ω) Trigger Mode	I_{GT}				mA
MT2(+), G(+)		—	—	50	
MT2(+), G(−)		—	—	50	
MT2(−), G(−)		—	—	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	75	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C					
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
MT2(-), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) (V_D = 24 Vdc, Gate Open, Initiating Current = 200 mA)	I_H	—	—	50	mA
Critical Rate of Rise of Commutating Off-State Voltage (Rated V_{DROM}, $I_{T(RMS)}$ = 6 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	$dv/dt_{(c)}$	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open, T_J = 125°C)	dv/dt	—	100	—	V/μs

3

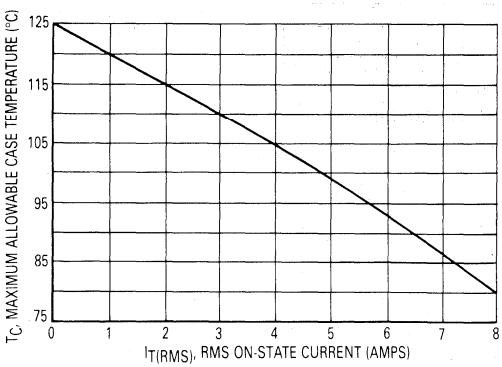


Figure 1. Current Derating

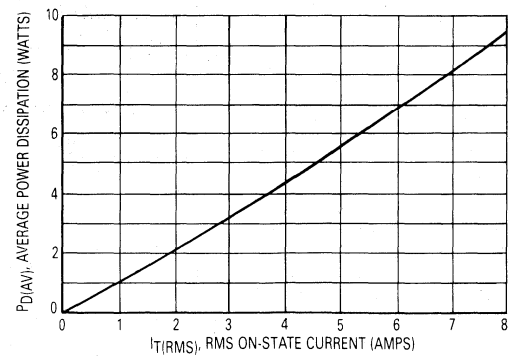


Figure 2. Power Dissipation

TYPICAL CHARACTERISTICS

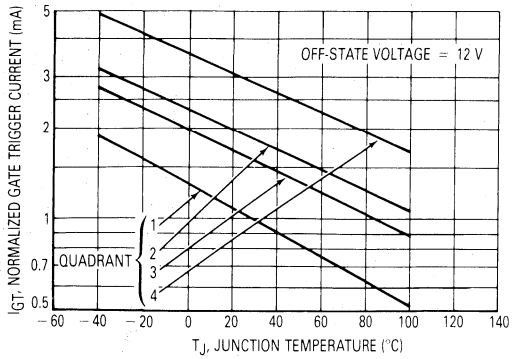


Figure 3. Normalized Gate Trigger Current

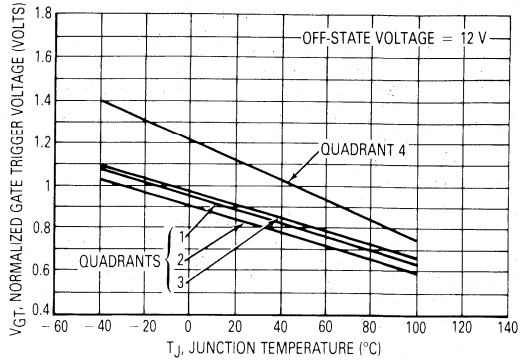


Figure 4. Normalized Gate Trigger Voltage

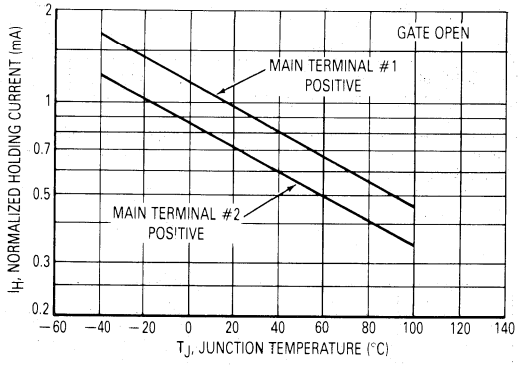


Figure 5. Normalized Holding Current

3

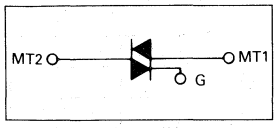
Triacs
Silicon Bidirectional Thyristors

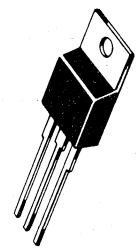
MAC219 Series

TRIACs
8.0 AMPERES RMS
200 thru 800 VOLTS

. . . designed for full-wave ac control applications primarily in industrial environments needing noise immunity.

- Guaranteed High Noise Immunity
  dv/dt(s) — 500 V/μs Min @ T_C = 25°C
- High Blocking Voltage — V_{DRM} to 800 V
- Photo Glass Passivated Junction for Improved Power Cycling Capability and Reliability





CASE 221A-04
(TO-220AB)
STYLE 4

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to 125°C) MAC219-4	V_{DRM}	200	Volts
-6		400	
-8		600	
-10		800	
Peak Gate Voltage	V_{GM}	±10	Volts
On-State Current RMS (Conduction Angle = 360°, T_C = +80°C)	$I_{T(RMS)}$	8.0	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	100	Amps
Fusing Current (t = 8.3 ms)	I^2t	35	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 2.0 μs)	P_{GM}	16	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Trigger Current (Pulse Width = 1.0 μs)	I_{GTM}	4.0	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2.0	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 11.3 A Peak; Pulse Width = ≤2.0 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.7	2.0	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 12 Ω) Trigger Mode MT2(+), Gate(+); MT2(+), Gate (−); MT2(−), Gate(−)	I_{GT}	—	—	100	mA
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	0.9	2.0	
MT2(+), G(−)		—	0.9	2.0	
MT2(−), G(−)		—	1.1	2.0	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

(continued)

ELECTRICAL CHARACTERISTICS — continued (T_C = +25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open, Initiating Current = 200 mA)	I_H	—	—	100	mA
Rate of Rise of Commutation Voltage (Rated V_{DROM}, $I_{T(RMS)}$ = 6.0 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5.0	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DROM}, Exponential Voltage Rise, Gate Open)	dv/dt(s)				V/μs
T_J = 25°C		500	—	—	
T_J = 125°C		200	—	—	

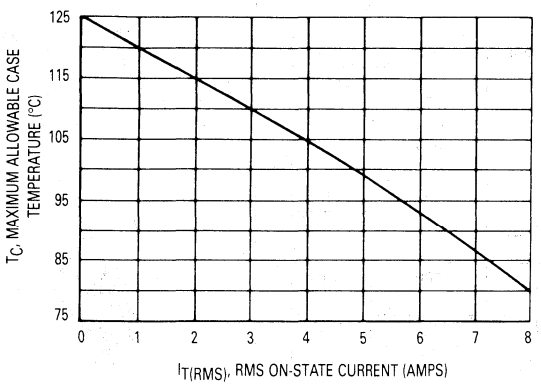


Figure 1. Current Derating

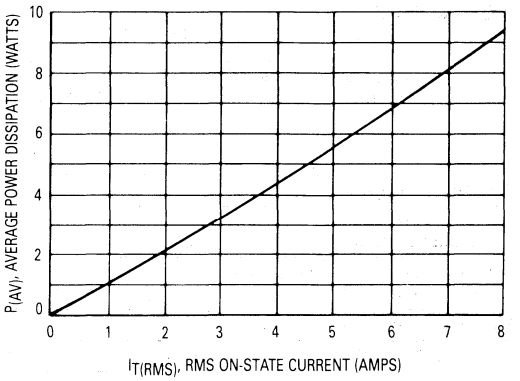


Figure 2. Power Dissipation

3

Triacs

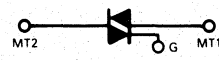
Silicon Bidirectional Triode Thyristors

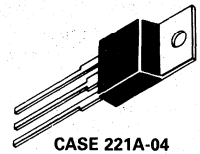
MAC223 Series
MAC223A Series

TRIACs
25 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications such as lighting sysjtems, heater controls, motor controls and power supplies; or wherever full-wave silicon-gate-controlled devices are needed.

- Off-State Voltages to 800 Volts
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged Thermowatt Construction for Thermal Resistance and High Heat Dissipation
- Gate Triggering Guaranteed in Three Modes (MAC223 Series) or Four Modes (MAC223A Series)





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage (T_J = −40 to 125°C), Note 1 (1/2 Sine Wave 50 to 60 Hz, Gate Open)	V_{DRM}		Volts
MAC223-4, MAC223A4		200	
MAC223-6, MAC223A6		400	
MAC223-8, MAC223A8		600	
MAC223-10, MAC223A10		800	
On-State RMS Current (T_C = 80°C) (Full Cycle Sine Wave 50 to 60 Hz)	$I_{T(RMS)}$	25	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = 80°C, preceded and followed by rated current)	I_{TSM}	250	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	260	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 125	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque	—	8	in. lb.

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

(cont.)

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Note 1) (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Peak On-State Voltage (I_{TM} = 35 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.4	1.85	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}				mA
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		—	20	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	30	75	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	V_{GT}				Volts
MT2(+), G(+); MT2(−), G(−); MT(+), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.3	2.5	
(V_D = Rated V_{DRM}, T_J = 125°C, R_L = 10 k) MT(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	0.4	—	
MT2(−), G(+) "A" SUFFIX ONLY		0.2	0.4	—	
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	10	50	mA
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 35 A Peak, I_G = 200 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 125°C)	dv/dt	—	40	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 35 A Peak, Commutating di/dt = 13.4 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

3

FIGURE 1 — RMS CURRENT DERATING

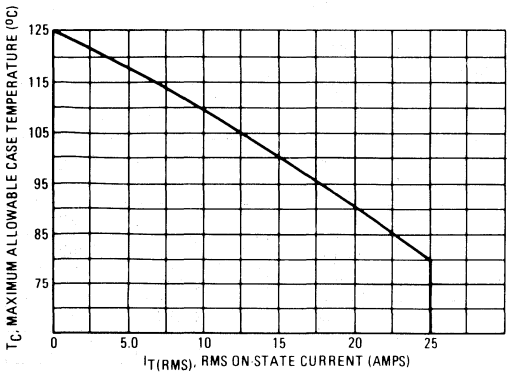


FIGURE 2 — ON-STATE POWER DISSIPATION

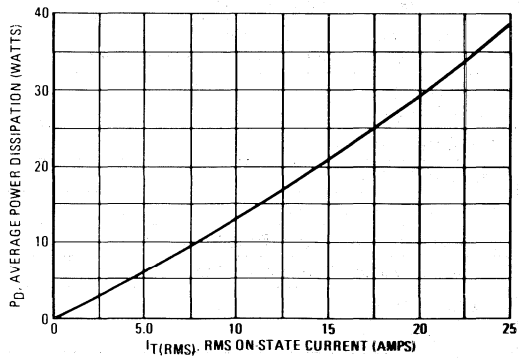


FIGURE 3 — GATE TRIGGER CURRENT

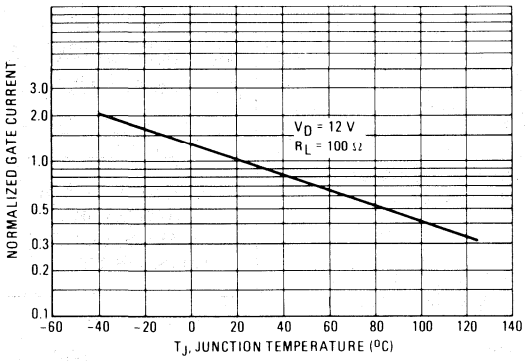


FIGURE 4 — GATE TRIGGER VOLTAGE

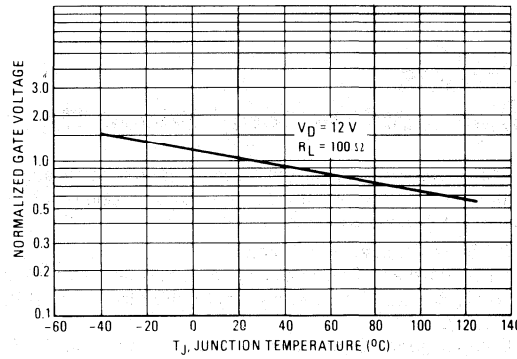


FIGURE 5 — HOLD CURRENT

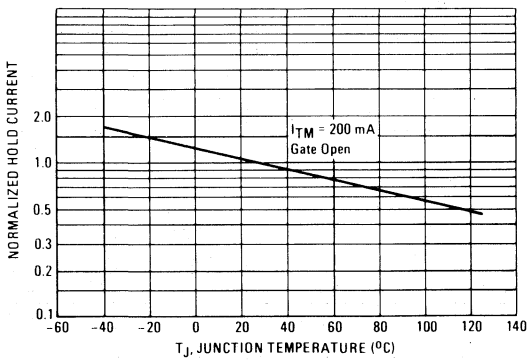
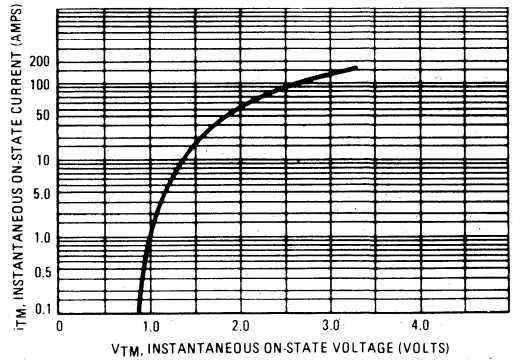


FIGURE 6 — ON-STATE CHARACTERISTICS



3

Triacs

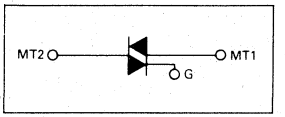
Silicon Bidirectional Triode Thyristors

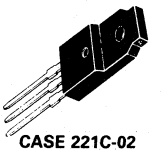
MAC223FP Series
MAC223AFP Series

ISOLATED TRIACs THYRISTORS
25 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as lighting systems, heater controls, motor controls and power supplies; or wherever full-wave silicon-gate-controlled devices are needed.

- Off-State Voltages to 800 Volts
- All Diffused and Glass Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged Thermowatt Construction for Thermal Resistance and High Heat Dissipation
- Gate Triggering Guaranteed in Three Modes (MAC223FP Series) or Four Modes (MAC223AFP Series)





CASE 221C-02
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
MAC223-4FP, MAC223A4FP		200	
MAC223-6FP, MAC223A6FP		400	
MAC223-8FP, MAC223A8FP		600	
MAC223-10FP, MAC223A10FP		800	
On-State RMS Current (T_C = +80°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	25	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = 80°C, preceded and followed by rated current)	I_{TSM}	250	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	260	A^2s
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current (t ≤ 2 μs)	I_{GM}	2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque	—	8	in/lb

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.2	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2	°C/W
Thermal Resistance, Case to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current, Note 1 T_J = 25°C (V_D = Rated V_{DRM}, T_J = 125°C	I_{DRM}	— —	— —	10 2	μA mA
Peak On-State Voltage (I_{TM} = 35 A Peak; Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.4	1.85	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−), MT2(+), G(−) MT2(−), G(+) "A" SUFFIX ONLY	I_{GT}	 — —	 20 30	 50 75	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−), MT2(+), G(−) MT2(−), G(+) "A" SUFFIX ONLY (V_D = Rated V_{DRM}, T_J = 125°C, R_L = 10 k) MT2(+), G(+); MT2(-), G(-), MT2(+), G(-) MT2(-), G(+) "A" SUFFIX ONLY	V_{GT}	 — — 0.2 0.2	 1.1 1.3 0.4 0.4	 2 2.5 — —	Volts
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	10	50	mA
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 35 A Peak, I_G = 200 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 125°C)	dv/dt	—	40	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 35 A Peak, Commutating di/dt = 13.4 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

NOTE 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

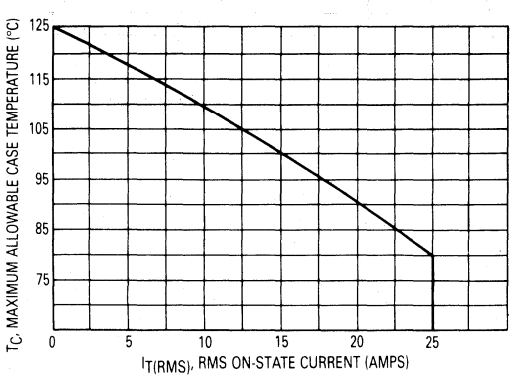


Figure 1. RMS Current Derating

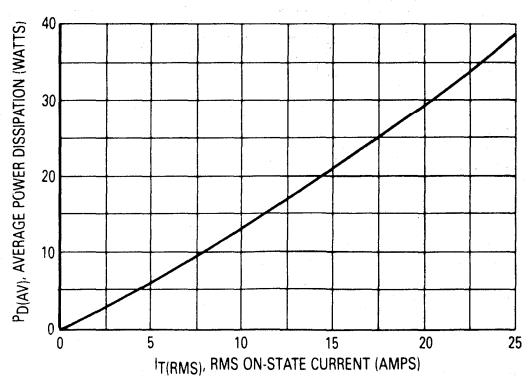


Figure 2. On-State Power Dissipation

TYPICAL CHARACTERISTICS

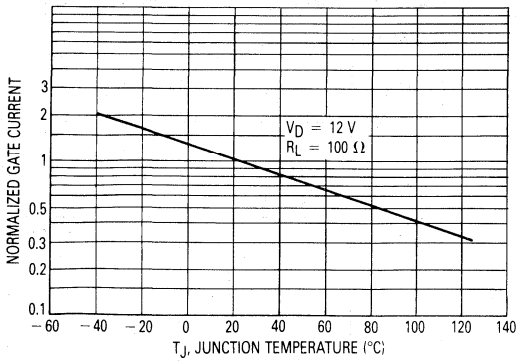


Figure 3. Gate Trigger Current

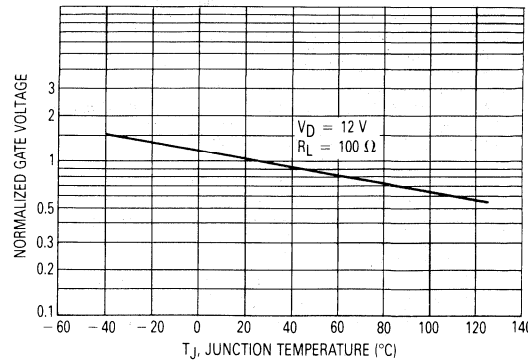


Figure 4. Gate Trigger Voltage

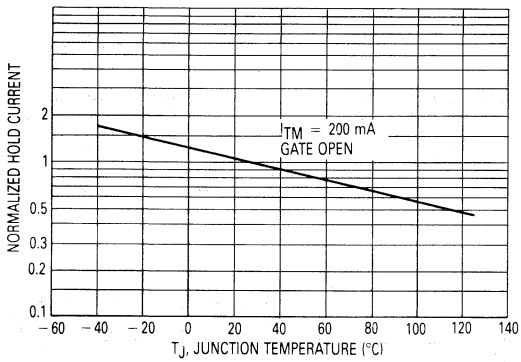


Figure 5. Hold Current

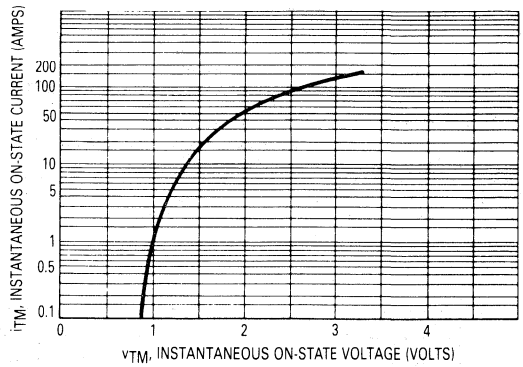


Figure 6. On-State Characteristics

3

Triacs

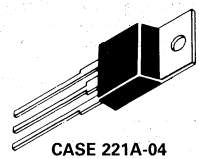
Silicon Bidirectional 40 Amperes RMS Triode Thyristors

MAC224 Series
MAC224A Series

. . . designed primarily for full-wave ac control applications such as lighting systems, heater controls, motor controls and power supplies.

- Blocking Voltage to 800 Volts
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Gate Triggering Guaranteed in Three Modes (MAC224 Series) or Four Modes (MAC224A Series)

TRIACs
40 AMPERES RMS
200 thru 800 VOLTS



CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage (T_J = −40 to 125°C), Note 1 (1/2 Sine Wave 50 to 60 Hz, Gate Open)	V_{DRM}		Volts
MAC224-4, MAC224A4		200	
MAC224-6, MAC224A6		400	
MAC224-8, MAC224A8		600	
MAC224-10, MAC224A10		800	
On-State RMS Current (T_C = 75°C), Note 2 (Full Cycle Sine Wave 50 to 60 Hz)	$I_{T(RMS)}$	40	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_J = 125°C)	I_{TSM}	350	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	500	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = 75°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 125	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque	—	8	in. lb.

Notes: 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage. (cont.)
2. This device is rated for use in applications subject to high surge conditions. Care must be taken to insure proper heat sinking when the device is to be used at high sustained currents. (See Figure 1 for maximum case temperatures.)

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
Peak On-State Voltage (I_{TM} = 56 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.4	1.85	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(+), G(−); MT2(+), G(−) MT2(−), G(+) "A" SUFFIX ONLY	I_{GT}	— —	25 40	50 75	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−); MT(+), G(−) MT2(−), G(+) "A" SUFFIX ONLY	V_{GT}	— —	1.1 1.3	2 2.5	Volts
Gate Non-Trigger Voltage (V_D = Rated V_{DRM}, T_J = 125°C, R_L = 10 k) MT2(+), G(+); MT2(−), G(−); MT(+), G(−) MT2(−), G(+)	V_{GD}	0.2 0.2	— —	— —	Volts
Holding Current (V_D = 12 Vdc, Gate Open)	I_H	—	30	75	mA
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 56 A Peak, I_G = 200 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 125°C)	dv/dt	—	50	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 56 A Peak, Commutating di/dt = 13.4 A/ms, Gate Unenergized, T_C = 75°C)	dv/dt(c)	—	5	—	V/μs

3

FIGURE 1 — RMS CURRENT DERATING

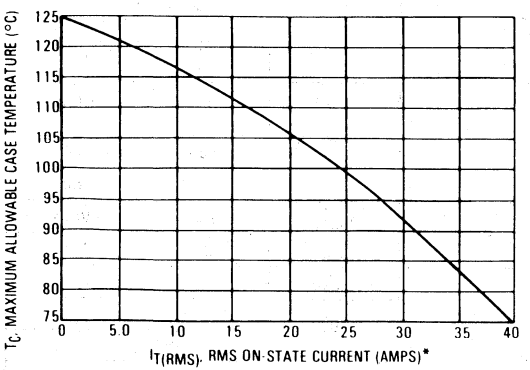
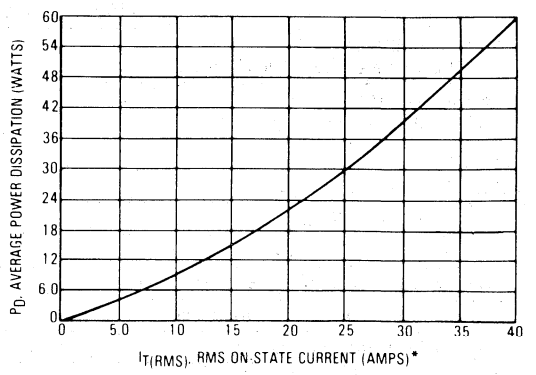


FIGURE 2 — ON-STATE POWER DISSIPATION



*This device is rated for use in applications subject to high surge conditions. Care must be taken to insure proper heat sinking when the device is to be used at high sustained currents.

FIGURE 3 — GATE TRIGGER CURRENT

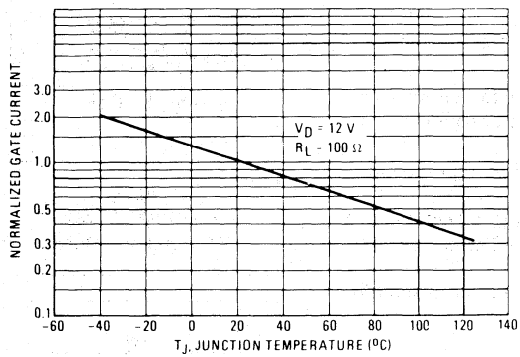


FIGURE 4 — GATE TRIGGER VOLTAGE

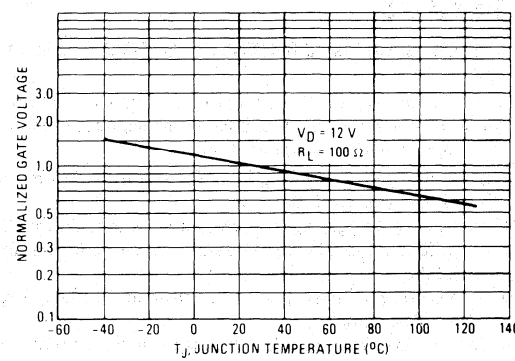


FIGURE 5 — HOLDING CURRENT

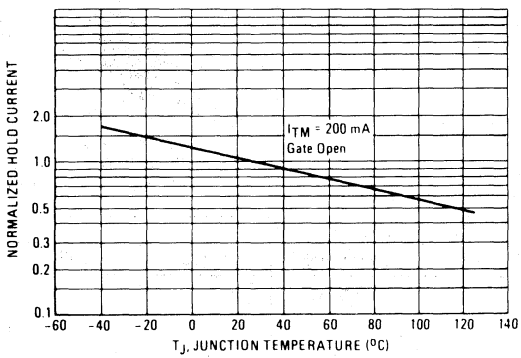


FIGURE 6 — ON-STATE CHARACTERISTICS

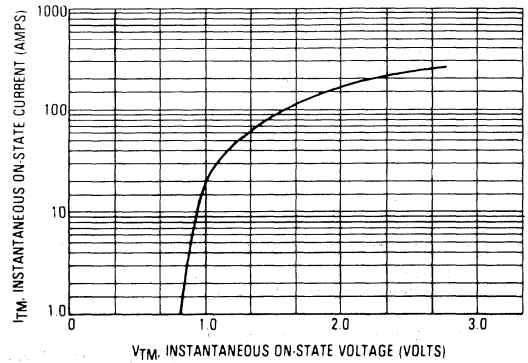
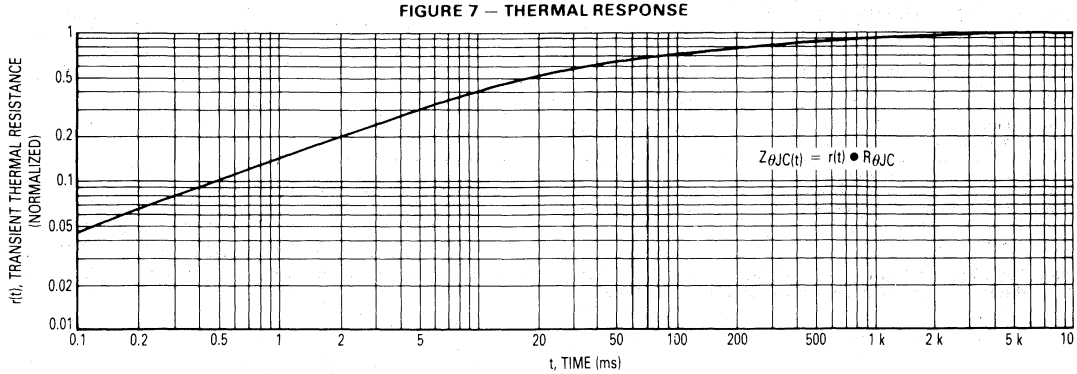


FIGURE 7 — THERMAL RESPONSE



3

Triacs

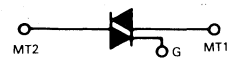
Silicon Bidirectional Triode Thyristors

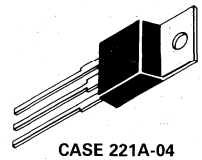
MAC228 Series
MAC228A Series

. . . designed primarily for industrial and consumer applications for full wave control of ac loads such as appliance controls, heater controls, motor controls, and other power switching applications.

- Sensitive Gate Triggering in 3 Modes for AC Triggering on Sinking Current Sources (MAC228 Series)
- Four Mode Triggering for Drive Circuits that Source Current (MAC228A Series)
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal resistance and High Heat Dissipation
- Center Gate Geometry for Uniform Current Spreading

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to 110°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
MAC228-4, MAC228A4		200	
MAC228-6, MAC228A6		400	
MAC228-8, MAC228A8		600	
MAC228-10, MAC228A10		800	
On-State RMS Current (T_C = 80°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle 60 Hz, T_J = 110°C)	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

(continued)

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 110	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque		8	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 110°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
Peak On-State Voltage (I_{TM} = 11 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.8	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−) MT2(−), G(+) "A" Suffix Only	I_{GT}	 — —	 — —	 5 10	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−) MT2(−), G(+) "A" Suffix Only (V_D = Rated V_{DRM}, T_C = 110°C, R_L = 10 k) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−) MT2(−), G(+) "A" Suffix Only	V_{GT}	 — — 0.2 0.2	 — — — —	 2 2.5 — —	Volts
Holding Current (V_D = 12 Vdc, I_{TM} = 200 mA, Gate Open)	I_H	—	—	15	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 16 A Peak, I_G = 30 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 11 A Peak, Commutating di/dt = 5.8 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

FIGURE 1 — RMS CURRENT DERATING

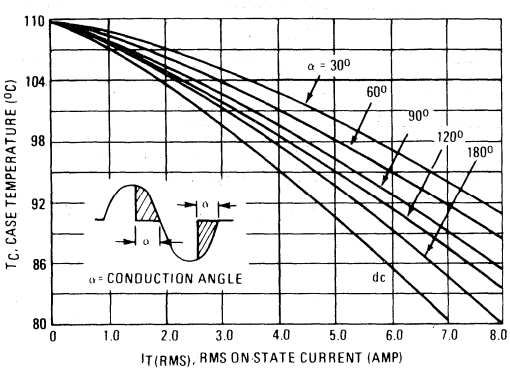
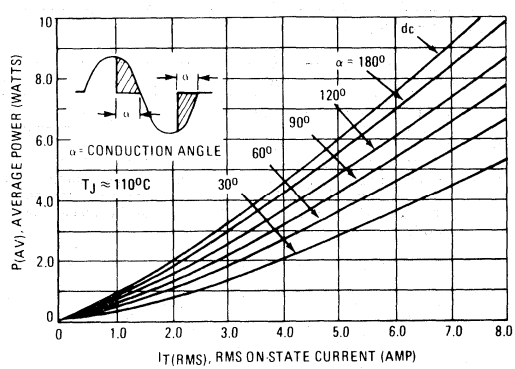


FIGURE 2 — ON-STATE POWER DISSIPATION



3

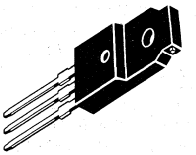
Triacs
Silicon Bidirectional Triode Thyristors

MAC228FP Series
MAC228AFP Series

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and consumer applications for full wave control of ac loads such as appliance controls, heater controls, motor controls, and other power switching applications.

- Four Mode Triggering for Drive Circuits that Source Current
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal resistance and High Heat Dissipation
- Center Gate Geometry for Uniform Current Spreading



CASE 221C-02
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to 110°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
MAC228-4FP, MAC228A4FP		200	
MAC228-6FP, MAC228A6FP		400	
MAC228-8FP, MAC228A8FP		600	
MAC228-10FP, MAC228A10FP		800	
On-State RMS Current (T_C = 80°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle 60 Hz, T_J = 110°C)	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage. (continued)
2. The case temperature reference point for all TC measurements is a point on the center lead of the package as close as possible to the plastic body.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 110	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque		8	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	2	mA
Peak On-State Voltage (I_{TM} = 11 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.8	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}				mA
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	5	
MT2(−), G(+) "A" Suffix Only		—	—	10	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	V_{GT}				Volts
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	2	
MT2(−), G(+) "A" Suffix Only		—	—	2.5	
(V_D = Rated V_{DRM}, T_C = 110°C, R_L = 10 k) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		0.2	—	—	
MT2(−), G(+) "A" Suffix Only		0.2	—	—	
Holding Current (V_D = 12 Vdc, I_{TM} = 200 mA, Gate Open)	I_H	—	—	15	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 16 A Peak, I_G = 30 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 11 A Peak, Commutating di/dt = 5.8 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

FIGURE 1 — RMS CURRENT DERATING

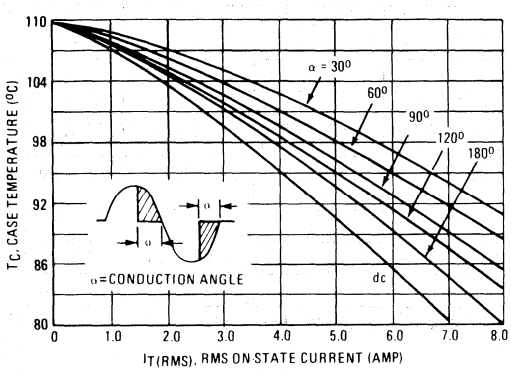
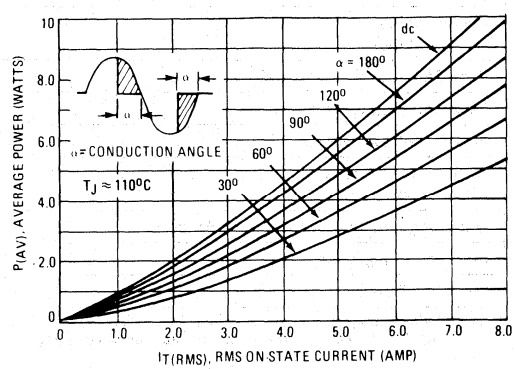


FIGURE 2 — ON-STATE POWER DISSIPATION



3

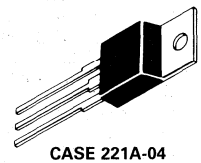
Triacs
Silicon Bidirectional Triode Thyristors

MAC229 Series
MAC229A Series

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and consumer applications for full wave control of ac loads such as appliance controls, heater controls, motor controls, and other power switching applications.

- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal resistance and High Heat Dissipation
- Center Gate Geometry for Uniform Current Spreading
- Gate Triggering Guaranteed in Three Modes (MAC229 Series) or Four Modes (MAC229A Series)



CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to 110°C) 1/2 Sine ave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
MAC229-4, MAC229A4		200	
MAC229-6, MAC229A6		400	
MAC229-8, MAC229A8		600	
MAC229-10, MAC229A10		800	
On-State RMS Current (T_C = 80°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle 60 Hz, T_J = 110°C)	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 110	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque		8	in. lb.

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage. (cont.)

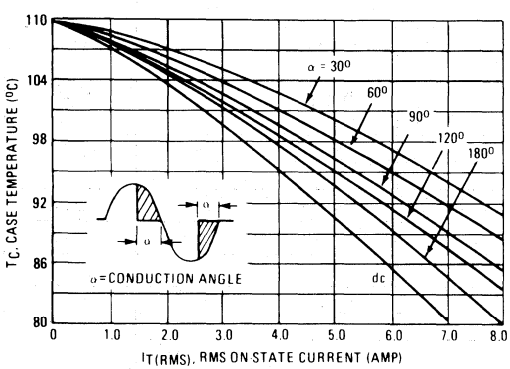
THERMAL CHARACTERISTICS

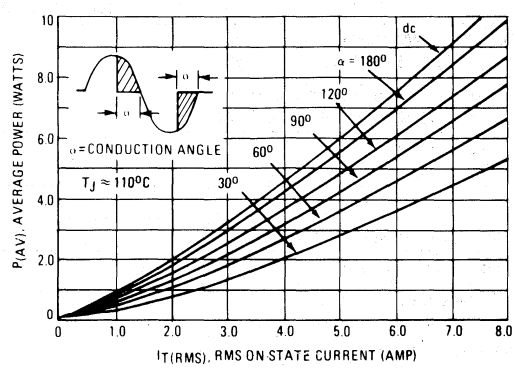
Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current, Note 1 (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	2	mA
Peak On-State Voltage (I_{TM} = 11 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.8	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}				mA
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	10	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	20	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	V_{GT}				Volts
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	2.5	
(V_D = Rated V_{DRM}, T_C = 110°C, R_L = 10 k)					
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−); All types		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY MAC229 series		0.2	—	—	
Holding Current (V_D = 12 Vdc, I_{TM} = 200 mA, Gate Open)	I_H	—	—	15	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 16 A Peak, I_G = 30 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 11 A Peak, Commutating di/dt = 5.8 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.





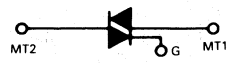
Triacs
Silicon Bidirectional Triode Thyristors

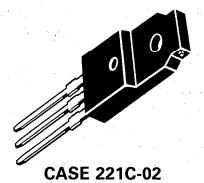
MAC229FP Series
MAC229AFP Series

TRIACs
8 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and consumer applications for full wave control of ac loads such as appliance controls, heater controls, motor controls, and other power switching applications.

- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal resistance and High Heat Dissipation
- Center Gate Geometry for Uniform Current Spreading
- Gate Triggering Guaranteed in Three Modes (MAC229FP Series) or Four Modes (MAC229AFP Series)





CASE 221C-02
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to 110°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open	V_{DRM}		Volts
MAC229-4FP, MAC229A4FP		200	
MAC229-6FP, MAC229A6FP		400	
MAC229-8FP, MAC229A8FP		600	
MAC229-10FP, MAC229A10FP		800	
On-State RMS Current (T_C = 80°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle 60 Hz, T_J = 110°C)	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Current (t ≤ 2 μs)	I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20	Watts
AVerage Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range	T_J	−40 to 110	°C
Storage Temperature Range	T_{stg}	−40 to 150	°C
Mounting Torque		8	in. lb.

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage. (continued)
2. The case temperature reference point for all TC measurements is a point on the center lead of the package as close as possible to the plastic body.

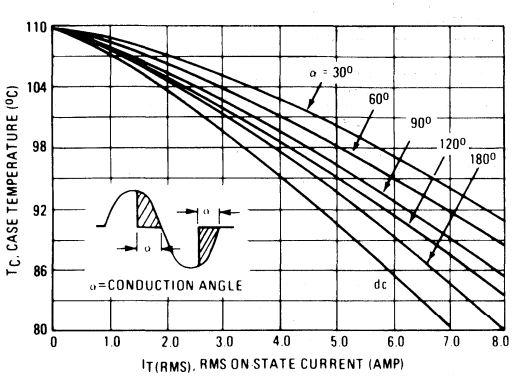
THERMAL CHARACTERISTICS

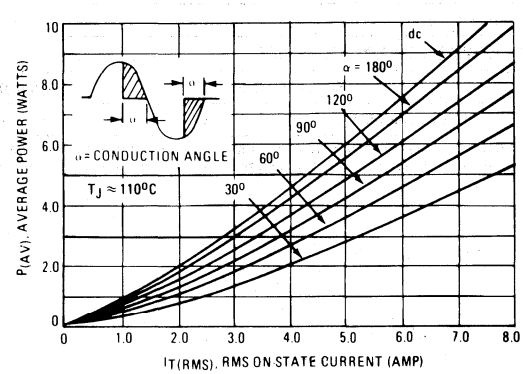
Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current, Note 1 (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	2	mA
Peak On-State Voltage (I_{TM} = 11 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.8	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}				mA
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	10	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	20	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	V_{GT}				Volts
MT2(+), G(+); MT2(+), G(−); MT2(−), G(−)		—	—	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	2.5	
(V_D = Rated V_{DRM}, T_C = 110°C, R_L = 10 k) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−); All types		0.2	—	—	
MT2(−), G(+) "A" SUFFIX ONLY MAC229 series		0.2	—	—	
Holding Current (V_D = 12 Vdc, I_{TM} = 200 mA, Gate Open)	I_H	—	—	15	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 16 A Peak, I_G = 30 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 11 A Peak, Commutating di/dt = 5.8 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.





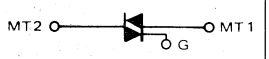
Triacs
Silicon Bidirectional Thyristors

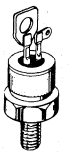
. . . designed primarily for industrial and military applications for the fullwave control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Pressfit and Stud Packages
- Gate Triggering Guaranteed In All 4 Quadrants

MAC245

TRIAC
10 AMPERES RMS
200 thru 800 VOLTS





CASE 263-04
STYLE 2
MAC245
STUD

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage (1) T_C = −40°C to +100°C MAC245B MAC245D MAC245M MAC245N	V_{DRM}	 200 400 600 800	Volts
RMS On-State Current	$I_{T(RMS)}$	10	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	100	Amps
Circuit Fusing Considerations t = 1 ms t = 8.3 ms	I^2t	 20 41.5	A^2s
Peak Gate Power	P_{GM}	10	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

(1) Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Off-State Current (1) Rated V_{DRM} = Peak Off-State Voltage, Gate Open-Circuited T_C = +25°C T_C = +115°C	I_{DRM}	 — —	 — —	 0.1 0.5	mA
Peak On-State Voltage (1) I_{TM} = 14 A Peak, Pulse Width = 1 ms, Duty Cycle ≤ 2%. T_C = +25°C	V_{TM}	—	—	1.65	Volts
Critical Rate of Rise of Off-State Voltage (1) Rated V_{DRM}, Gate Open-Circuited, Exponential Waveform T_C = +100°C	dv/dt	—	50	—	Volts/μs
Critical Rate-of-Rise of Commutating Off-State Voltage (1) $I_{T(RMS)}$ = Rated RMS On-State Current V_{DRM} = Rated Peak Off-State Voltage, Gate Open-Circuited, Commutating di/dt = 5.4 A/ms T_C = +78.5°C	dv/dt(C)	4	—	—	Volts/μs
DC Gate Trigger Current (2) V_D = 12 Vdc Trigger Mode MT2(+), Gate(+), R_L = 100 Ohms MT2(−), Gate(−), R_L = 100 Ohms MT2(+), Gate(−), R_L = 50 Ohms MT2(+), Gate(+), R_L = 50 Ohms, T_C = −40°C MT2(−), Gate(−), R_L = 50 Ohms, T_C = −40°C MT2(+), Gate(−), R_L = 25 Ohms, T_C = −40°C	I_{GT}	 — — — — — —	 — — — — — —	 50 50 50 80 80 80	mAdc
DC Gate Trigger Voltage (2) V_D = 12 Vdc Trigger Mode MT2(+), Gate(+), R_L = 100 Ohms MT2(−), Gate(−), R_L = 100 Ohms MT2(+), Gate(−), R_L = 50 Ohms MT2(+), Gate(+), R_L = 50 Ohms, T_C = −40°C MT2(−), Gate(−), R_L = 50 Ohms, T_C = −40°C MT2(+), Gate(−), R_L = 25 Ohms, T_C = −40°C MT2(+), Gate(+), R_L = 1000 Ohms, T_C = +100°C (2,3) MT2(−), Gate(−), R_L = 1000 Ohms, T_C = +100°C (2,3) MT2(+), Gate(−), R_L = 1000 Ohms, T_C = +100°C (2,3) MT2(−), Gate(+), R_L = 1000 Ohms, T_C = +100°C (2,3)	V_{GT}	 — — — — — — 0.2 0.2 0.2 0.2	 — — — — — — — — — —	 2.5 2.5 2.5 3.5 3.5 3.5 — — — —	Vdc
Holding Current (1) Main Terminal Voltage = 24 Vdc, Peak Initiating Current = 0.5 A, Pulse Width = 0.1 to 10 ms Gate Trigger Source = 7 V, 20 Ohms T_C = +25°C T_C = −40°C	I_H	 — —	 — —	 50 100	mAdc
Latching Current (2) Main Terminal Source Voltage = 24 Vdc, Gate Trigger Source = 15 V, 100 Ohms, Pulse Width = 50 μs, Rise and Fall Times Maximum = 5 μs Trigger Mode MT2(+), Gate(+) MT2(−), Gate(−) MT2(+), Gate(−) MT2(+), Gate(+), T_C = −40°C MT2(−), Gate(−), T_C = −40°C MT2(+), Gate(−), T_C = −40°C	I_L	 — — — — — —	 — — — — — —	 100 100 200 200 200 400	mAdc

NOTES:
1. Values apply for either polarity of Main Terminal 2 characteristics referenced to Main Terminal 1.
2. Main Terminal 1 is the reference terminal.
3. With V_D equal to rated off-state voltage.

3

Triacs

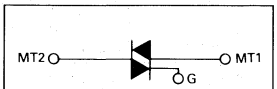
Bidirectional Triode Thyristors

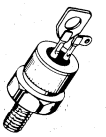
MAC250
MAC250( )3

TRIACs
15 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Pressfit, Stud and Isolated Stud Packages
- Gate Triggering Guaranteed In All 3 Quadrants





CASE 263-04
STYLE 1
MAC250
STUD

CASE 311-02
STYLE 2
MAC250( )3
ISOLATED STUD

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage MAC250B, MAC250B3 MAC250D, MAC250D3 MAC250M, MAC250M3 MAC250N	V_{DRM}	 200 400 600 800	Volts
RMS On-State Current	$I_{T(RMS)}$	15	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	100	Amps
Circuit Fusing Considerations t = 1 ms t = 8.3 ms	I^2t	 20 41.5	A^2s
Peak Gate Power	P_{GM}	10	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Peak Gate Power (Pulse Width = 10 μs)	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +115	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case MAC250, SC251 MAC250( )3	$R_{\theta JC}$	 2 2.3	°C/W

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted. Values apply for either polarity of Main Terminal 2 Characteristics referenced to Main Terminal 1.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = +115°C		—	—	0.5	mA
Peak On-State Voltage (I_{TM} = 21 A, Pulse Width = 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.65	Volts
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Gate Open-Circuited, Exponential Waveform) T_C = +115°C	dv/dt	100	—	—	V/μs
Critical Rate-of-Rise of Commutating Off-State Voltage, Note 1 ($I_{T(RMS)}$ = Rated RMS On-State Current, V_D = V_{DRM}) (Gate Open-Circuited, Commutating di/dt = 8 A/ms)	dv/dt(c)				V/μs
MAC250 T_C = +84°C		4	—	—	
MAC250( )3 T_C = +78°C		4	—	—	
DC Gate Trigger Current (Continuous dc) (V_D = 12 Vdc)	I_{GT}				mAdc
MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms		—	—	50	
MT2(+), G(−); R_L = 50 Ohms		—	—	50	
DC Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, T_C = −40°C)	I_{GT}				mAdc
MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms		—	—	80	
MT2(+), G(−); R_L = 25 Ohms		—	—	80	
DC Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc)	V_{GT}				Vdc
MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms		—	—	2.5	
MT2(+), G(−); R_L = 50 Ohms		—	—	2.5	
DC Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, T_C = −40°C)	V_{GT}				Vdc
MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms		—	—	3.5	
MT2(+), G(−); R_L = 25 Ohms		—	—	3.5	
DC Gate Non-Trigger Voltage (V_D = Rated V_{DRM}, R_L = 1K Ohms, T_C = 115°C) All Trigger Modes	V_{GD}	0.20	—	—	Vdc
Holding Current (V_D = 24 Vdc, Peak Initiating Current = 0.5 A, Pulse Width = 0.1 to 10 ms, Gate Trigger) (Source = 7 V, 20 Ohms) T_C = +25°C	I_H	—	—	50	mAdc
T_C = −40°C		—	—	100	
Latching Current (V_D = 24 Vdc, Gate Trigger Source = 15 V, 100 Ohms, Pulse Width = 50 μs, 5 μs Maximum Rise and Fall Times)	I_L				mAdc
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−) T_C = 25°C		—	—	100	
MT2(+), G(+); MT2(−), G(−); MT2(+), G(−) T_C = −40°C		—	—	200	

FIGURE 1 – CURRENT DERATING

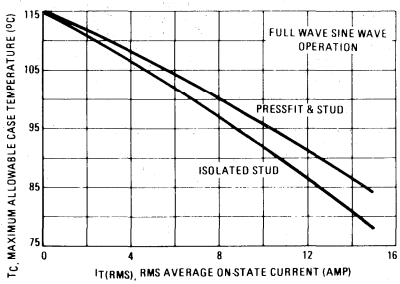
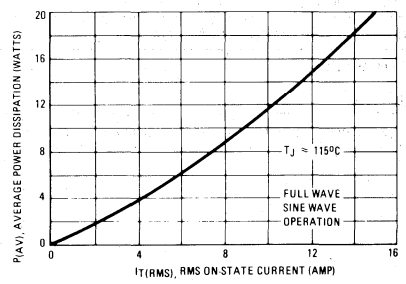


FIGURE 2 – MAXIMUM ON-STATE POWER DISSIPATION



3

Triacs

Silicon Bidirectional Triode Thyristors

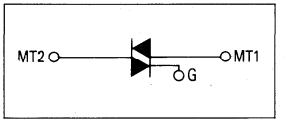
MAC261
(See SC260)

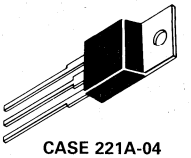
MAC310 Series
MAC310A Series

TRIACs
10 AMPERES RMS
200 thru 600 VOLTS

. . . designed primarily for industrial and consumer applications for full wave control of ac loads such as appliance controls, heater controls, motor controls, and other power switching applications.

- Sensitive Gate Triggering in Three Trigger Modes for AC Triggering on Sinking Current Sources (MAC310 Series)
- Four Mode Triggering (10 mA) for Drive Circuits that Source Current (MAC310A Series)
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation
- Center Gate Geometry for Uniform Current Spreading





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Repetitive Off-State Voltage (T_J = −40 to 110°C) Note 1. 1/2 Sine Wave 50 to 60 Hz, Gate Open	MAC310-4, MAC310A4 MAC310-6, MAC310A6 MAC310-8, MAC310A8	V_{DRM}	 200 400 600	Volts
On-State RMS Current (T_C = 80°C) Full Cycle Sine Wave 50 to 60 Hz		$I_{T(RMS)}$	10	Amps
Peak Non-Repetitive Surge Current (One Full Cycle 60 Hz, T_J = 110°C)		I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	40	A^2s
Peak Gate Current (t ≤ 2 μs)		I_{GM}	±2	Amps
Peak Gate Voltage (t ≤ 2 μs)		V_{GM}	±10	Volts
Peak Gate Power (t ≤ 2 μs)		P_{GM}	20	Watts
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)		$P_{G(AV)}$	0.5	Watts
Operating Junction Temperature Range		T_J	−40 to 110	°C
Storage Temperature Range		T_{stg}	−40 to 150	°C
Mounting Torque		—	8	in-lb

Note 1. Ratings apply for open gate conditions. Devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

(continued)

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Note 1) T_J = 25°C (V_D = Rated V_{DRM}, T_J = 110°C)	I_{DRM}	— —	— —	10 2	mA
Peak On-State Voltage (I_{TM} = 14 A Peak, Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−) MT2(−), G(+) "A" Suffix Only	I_{GT}	 — —	 — —	 5 10	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(+), G(−); MT2(−), G(−) MT2(−), G(+) "A" Suffix Only (V_D = Rated V_{DRM}, T_C = 110°C, R_L = 10 k) All Trigger Modes	V_{GT}	 — — 0.2	 — — —	 2 2.5 —	Volts
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	—	15	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 14 A Peak, I_G = 30 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 110°C)	dv/dt	—	25	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 14 A Peak, Commutating di/dt = 14 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs

3

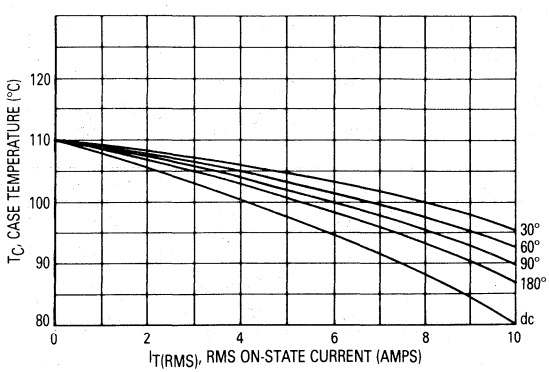


Figure 1. RMS Current Derating

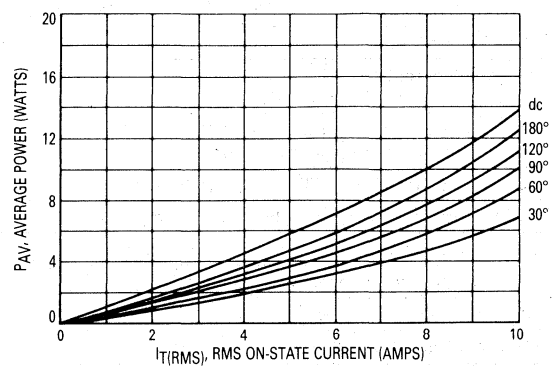


Figure 2. On-State Power Dissipation

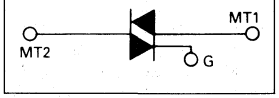
Triacs
Silicon Bidirectional Thyristors

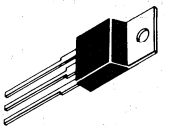
MAC320 Series
MAC320A Series

TRIACs
20 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC320 Series) or Four Modes (MAC320A Series)





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC320-4, MAC320A4 MAC320-6, MAC320A6 MAC320-8, MAC320A8 MAC320-10, MAC320A10	V_{DRM}	 200 400 600 800	Volts
Peak Gate Voltage	V_{GM}	10	Volts
On-State Current RMS (T_C = +75°C) (Full Cycle, Sine Wave, 50 to 60 Hz)	$I_{T(RMS)}$	20	Amp
Peak Surge Current (One Full Cycle, 60 Hz, T_C = +75°C) preceded and followed by rated current	I_{TSM}	150	Amp
Peak Gate Power (T_C = +75°C, Pulse Width = 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +75°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current	I_{GM}	2	Amp
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.8	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open T_J = 25°C T_J = +125°C	I_{DRM}	 — —	 — —	 10 2	 μA mA
Peak On-State Voltage (Either Direction) I_{TM} = 28 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.4	1.7	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms MT2 (+), G(+); MT2 (+), G(−); MT2 (−), G(−) MT2 (−), G(+) "A" SUFFIX ONLY	I_{GT}	 — —	 — —	 50 75	mA
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms MT2 (+), G(+); MT2 (+), G(−); MT2 (−), G(−) MT2 (−), G(+) "A" SUFFIX ONLY Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +110°C MT2 (+), G(+); MT2 (−), G(−); MT2 (+), G(−); MT2 (−), G(+) "A" SUFFIX ONLY	V_{GT}	 — — 0.2 0.2	 0.9 1.4 — —	 2 2.5 — —	Volts
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 200 mA	I_H	—	6	40	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 28 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 28 A, Commutating di/dt = 8 A/ms, Gate Unenergized, T_C = +75°C	dv/dt(C)	—	5	—	V/μs

3

FIGURE 1 — RMS CURRENT DERATING

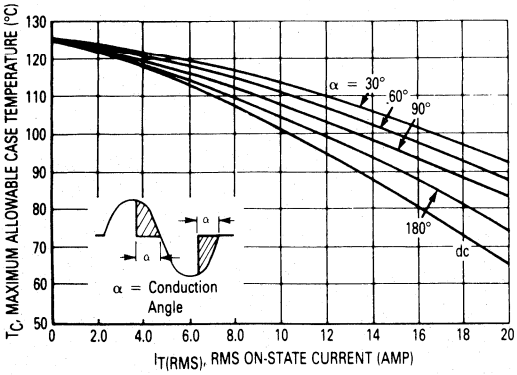


FIGURE 2 — ON-STATE POWER DISSIPATION

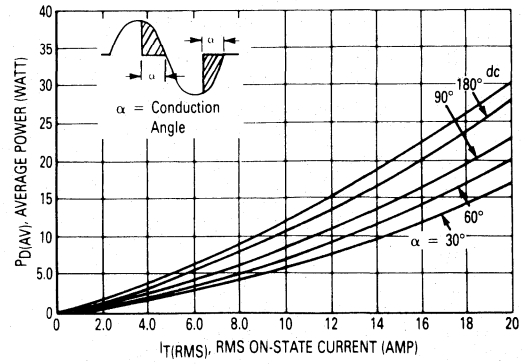


FIGURE 3 — TYPICAL GATE TRIGGER VOLTAGE

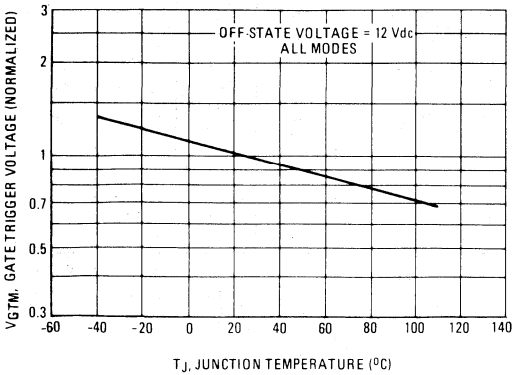


FIGURE 4 — TYPICAL GATE TRIGGER CURRENT

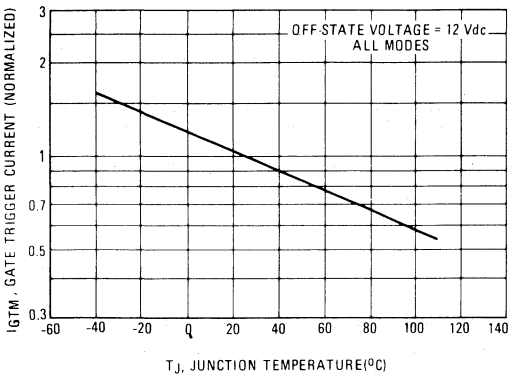


FIGURE 5 — MAXIMUM ON-STATE CHARACTERISTICS

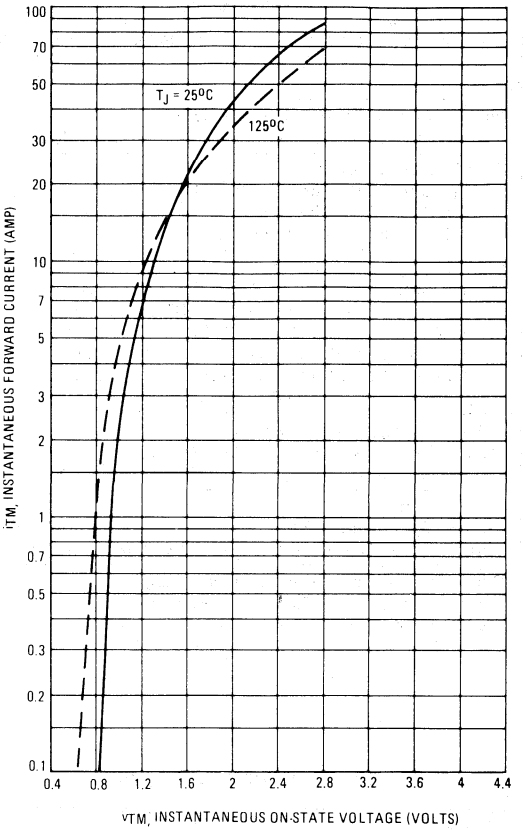


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FIGURE 6 — TYPICAL HOLDING CURRENT

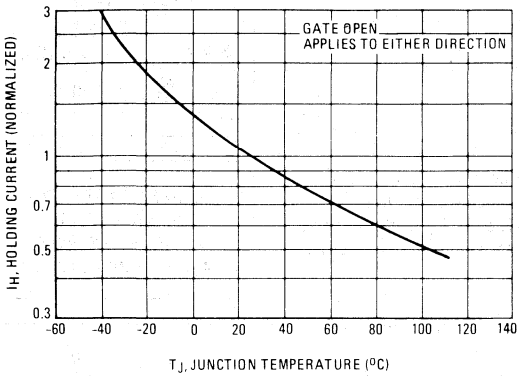


FIGURE 7 — MAXIMUM ON-REPETITIVE SURGE CURRENT

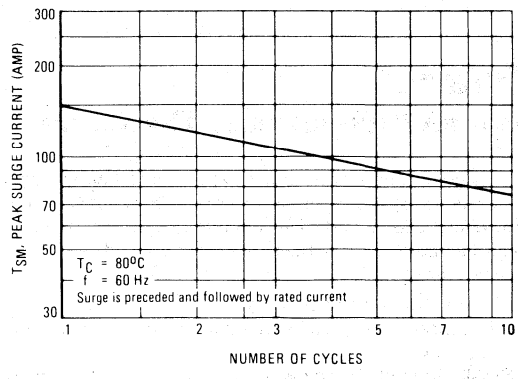
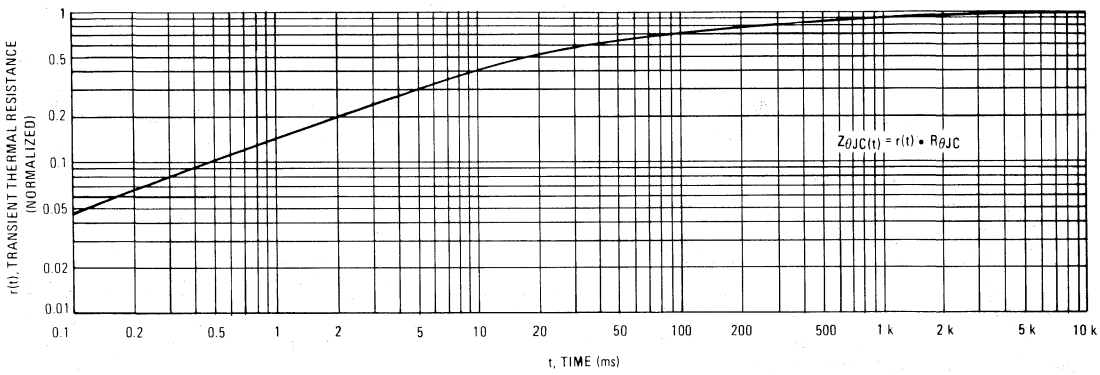


FIGURE 8 — THERMAL RESPONSE



3

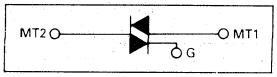
Triacs
Silicon Bidirectional Thyristors

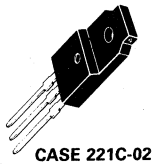
MAC320FP Series
MAC320AFP Series

ISOLATED TRIACs
THYRISTORS
20 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Gate Triggering Guaranteed in Three Modes (MAC320FP Series) or Four Modes (MAC320AFP Series)





CASE 221C-02
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open MAC320-4FP, MAC320A4FP MAC320-6FP, MAC320A6FP MAC320-8FP, MAC320A8FP MAC320-10FP, MAC320A10FP	V_{DRM}	 200 400 600 800	Volts
Peak Gate Voltage	V_{GM}	10	Volts
On-State RMS Current (T_C = +75°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	20	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +75°C) preceded and followed by rated current	I_{TSM}	150	Amps
Peak Gate Power (T_C = +75°C, Pulse Width = 2 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +75°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current	I_{GM}	2	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.8	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2	mA
Peak On-State Voltage (Either Direction) I_{TM} = 28 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%	V_{TM}	—	1.4	1.7	Volts
Peak Gate Trigger Current	I_{GT}				mA
Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms					
Minimum Gate Pulse Width = 2 μs					
MT2(+), G(+)		—	—	50	
MT2(+), G(−)		—	—	50	
MT2(−), G(−)		—	—	50	
MT2(−), G(+) "A" SUFFIX ONLY		—	—	75	
Peak Gate Trigger Voltage	V_{GT}				Volts
Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms					
Minimum Gate Pulse Width = 2 μs					
MT2(+), G(+)		—	0.9	2	
MT2(+), G(−)		—	0.9	2	
MT2(−), G(−)		—	1.1	2	
MT2(−), G(+) "A" SUFFIX ONLY		—	1.4	2.5	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +110°C					
MT2(+), G(+); MT2(+), G(-); MT2(-), G(-)		0.2	—	—	
MT2(-), G(+) "A" SUFFIX ONLY		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 200 mA	I_H	—	6	40	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 28 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DRM}, I_{TM} = 28 A, Commutating di/dt = 8 A/ms, Gate Unenergized, T_C = +75°C	dv/dt(c)	—	5	—	V/μs

3

TYPICAL CHARACTERISTICS

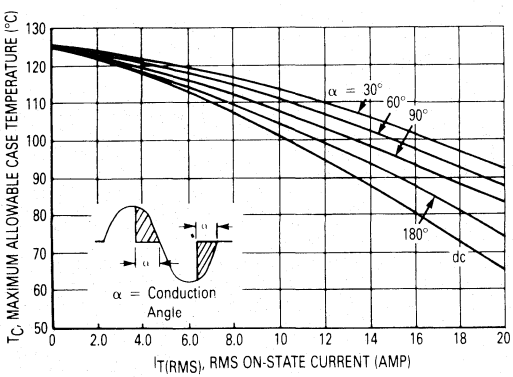


Figure 1. RMS Current Derating

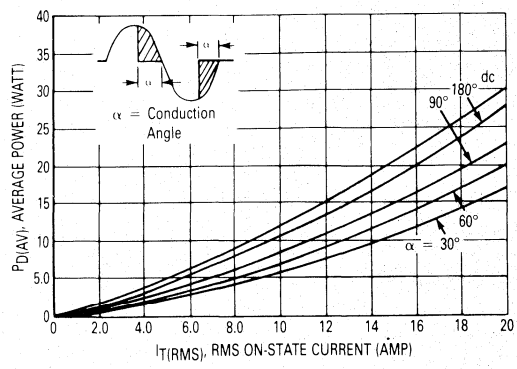


Figure 2. On-State Power Dissipation

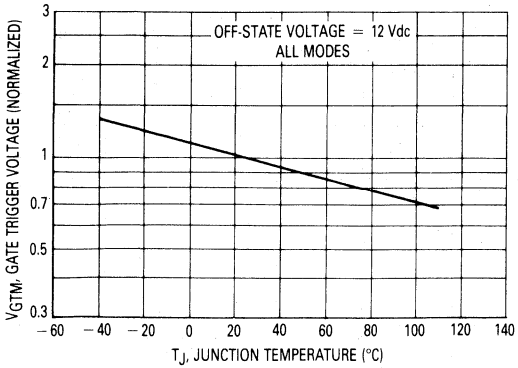


Figure 3. Typical Gate Trigger Voltage

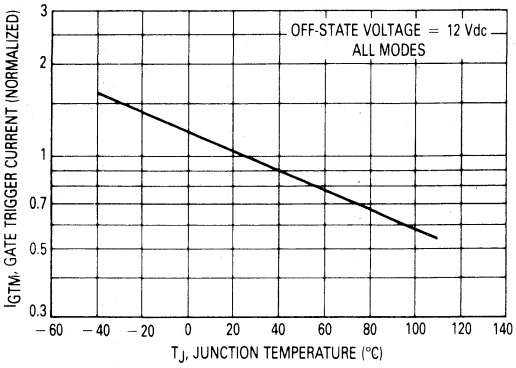


Figure 4. Typical Gate Trigger Current

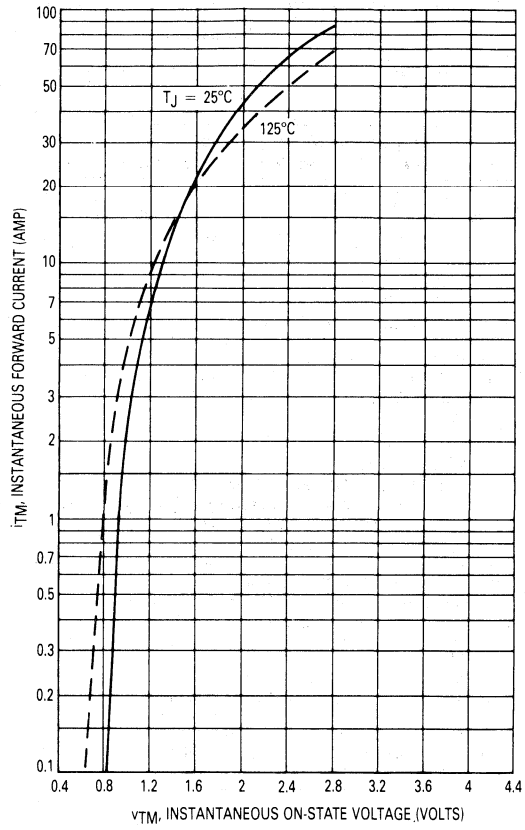


Figure 5. Maximum On-State Characteristics

3

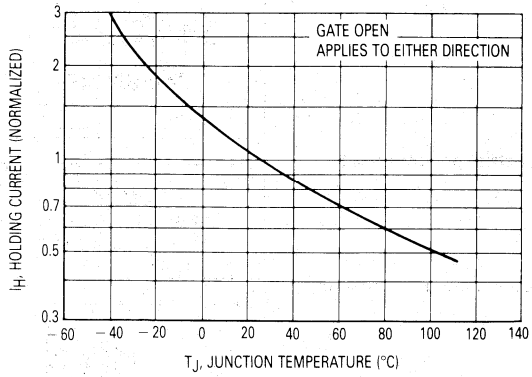


Figure 6. Typical Holding Current

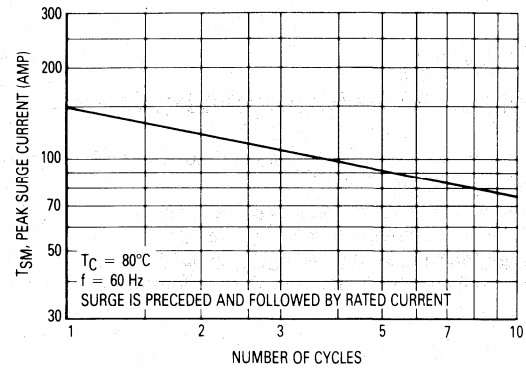


Figure 7. Maximum Nonrepetitive Surge Current

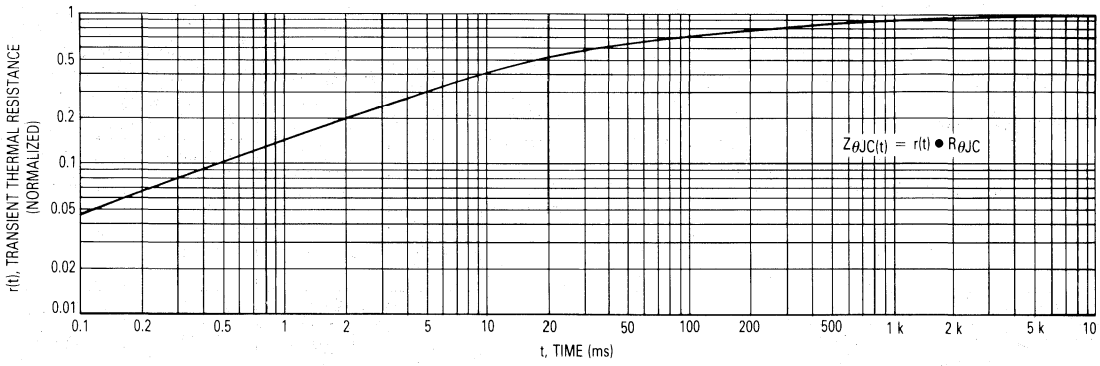


Figure 8. Thermal Response

3

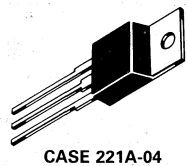
Triacs
Silicon Bidirectional Thyristors

MAC321 Series

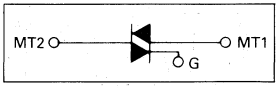
TRIACs
20 AMPERES RMS
200 thru 800 VOLTS

. . . designed for full-wave ac control applications primarily in industrial environments needing noise immunity.

- Guaranteed High Commutation Voltage
  dv/dt — 500 V/μs Min @ T_C = 25°C
- High Blocking Voltage — V_{DRM} to 800 V
- Photo Glass Passivated Junction for Improved Power Cycling Capability and Reliability



CASE 221A-04
(TO-220AB)
STYLE 4



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C)	V_{DRM}		Volts
MAC321-4		200	
-6		400	
-8		600	
-10		800	
Peak Gate Voltage	V_{GM}	10	Volts
On-State Current RMS (T_C = +75°C) Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	20	Amp
Peak Surge Current (One Full Cycle, 60 Hz, T_C = +75°C) preceded and followed by Rated Current	I_{TSM}	150	Amp
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	93	A^2s
Peak Gate Power (T_C = +75°C, Pulse Width = 2.0 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +75°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current	I_{GM}	2.0	Amp
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.8	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM}				
T_J = 25°C		—	—	10	μA
T_J = +125°C		—	—	2.0	mA
Peak On-State Voltage (Either Direction) I_{TM} = 28 A Peak; Pulse Width ≤ 2.0 ms, Duty Cycle ≤ 2.0%	V_{TM}	—	1.4	1.7	Volts
Gate Trigger Current (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	I_{GT}				mA
MT2(+), G(+)		—	—	100	
MT2(+), G(−)		—	—	100	
MT2(−), G(−)		—	—	100	
Gate Trigger Voltage (Continuous dc) Main Terminal Voltage = 12 Vdc, R_L = 100 Ohms	V_{GT}				Volts
MT2(+), G(+)		—	—	2.0	
MT2(+), G(−)		—	—	2.0	
MT2(−), G(−)		—	—	2.0	
Main Terminal Voltage = Rated V_{DRM}, R_L = 10 kΩ, T_J = +125°C MT2(+), G(+); MT2(−), G(−); MT2(+), G(−)		0.2	—	—	
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 200 mA	I_H	—	—	100	mA
Turn-On Time Rated V_{DRM}, I_{TM} = 28 A, I_{GT} = 120 mA, Rise Time = 0.1 μs, Pulse Width = 2.0 μs	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage V_D = Rated V_{DRM}, Exponential Voltage Rise, Gate Open	dv/dt(s)				V/μs
T_J = 25°C		500	—	—	
T_J = +125°C		200	—	—	

3

TYPICAL CHARACTERISTICS

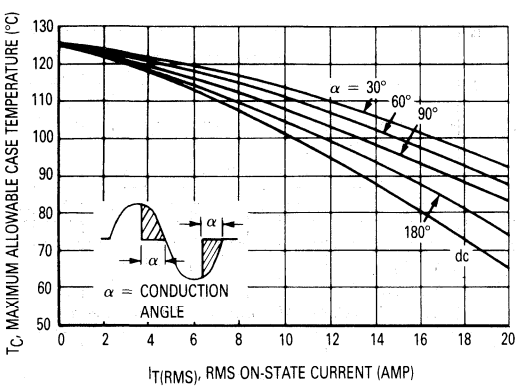


Figure 1. RMS Current Derating

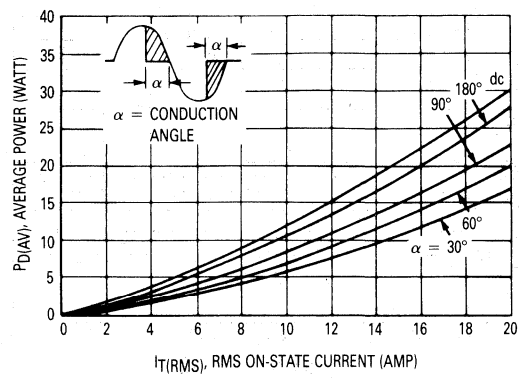


Figure 2. On-State Power Dissipation

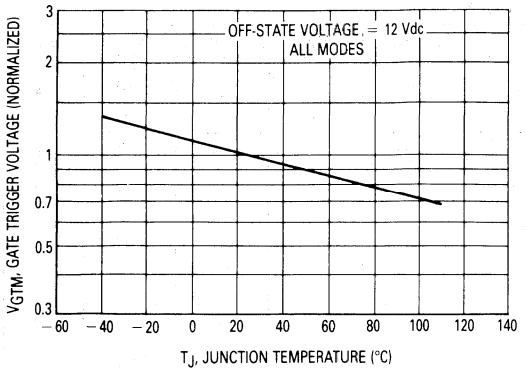


Figure 3. Typical Gate Trigger Voltage

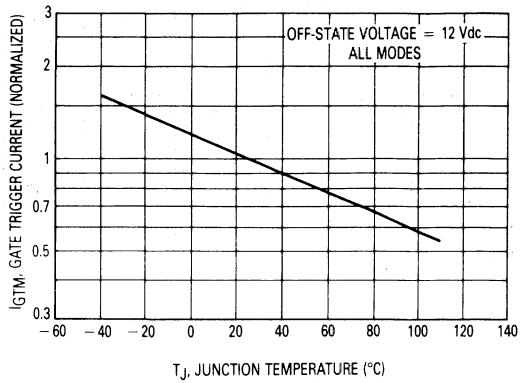


Figure 4. Typical Gate Trigger Current

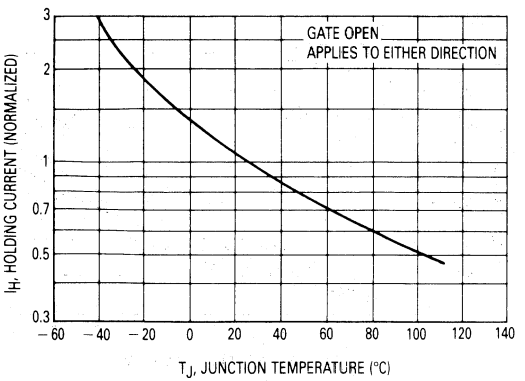


Figure 6. Typical Holding Current

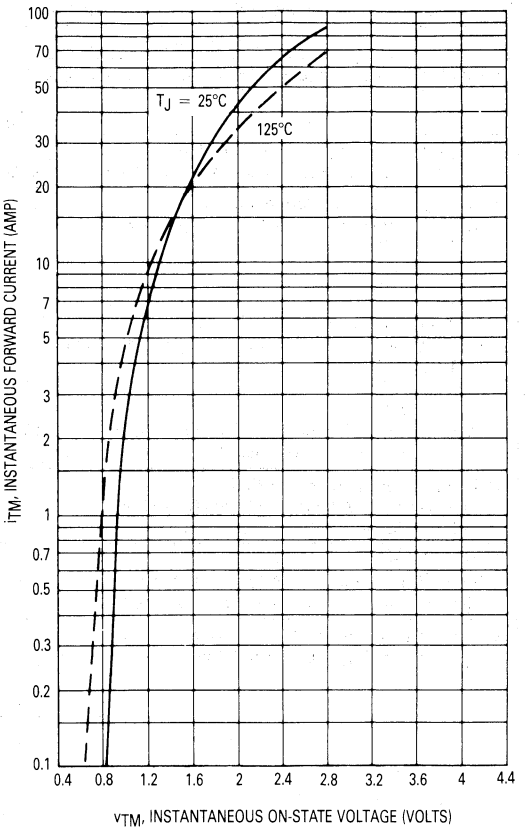


Figure 5. Maximum On-State Characteristics

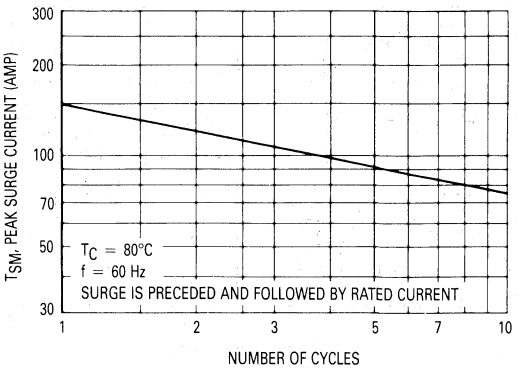


Figure 7. Maximum On-Repetitive Surge Current

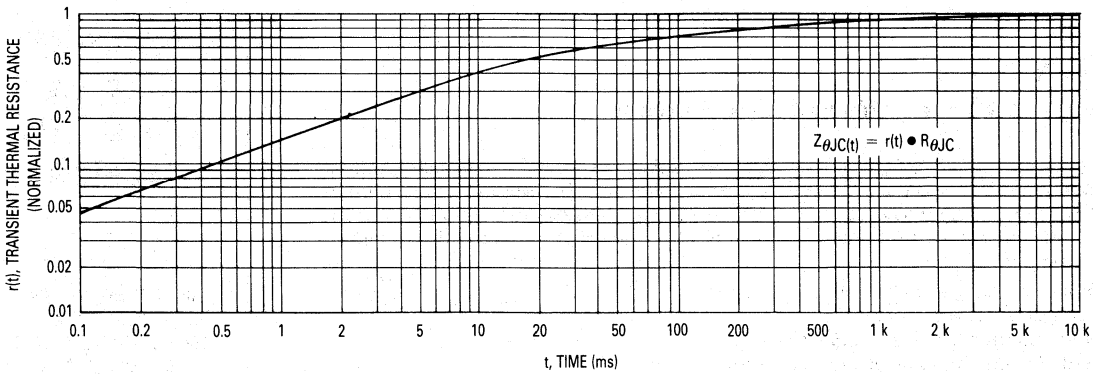


Figure 8. Thermal Response

3

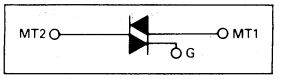
Triacs
Silicon Bidirectional Thyristors

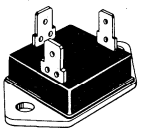
MAC625 Series
MAC635 Series

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 600 Volts
- TO-3 Isolated Mounting
- High Isolation Voltage — 2.5 kV rms
- High Commutating dv/dt — 6 V/μs Min
- UL Recognized — File # E69369

ISOLATED TRIACs
THYRISTORS
25/35 AMPERES RMS
200–600 VOLTS





CASE 383-01
STYLE 1

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Repetitive Peak Off-State Voltage (1) (T_J = −40 to +125°C) 1/2 Sine Wave 50 to 60 Hz, Gate Open		V_{DRM}		Volts
	MAC625-4/635-4 MAC625-6/635-6 MAC625-8/635-8		200 400 600	
Peak Gate Voltage		V_{GM}	10	Volts
On-State RMS Current (T_C = +75°C) Full Cycle Sine Wave 50 to 60 Hz (2) (T_C = 58°C)	 MAC625 Series MAC635 Series	$I_{T(RMS)}$	 25 35	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +75°C) preceded and followed by rated current	MAC625 Series MAC635 Series	I_{TSM}	250 330	Amps
Peak Gate Power (T_C = +75°C, Pulse Width = 2 μs)		P_{GM}	10	Watts
Average Gate Power (T_C = +75°C, t = 8.3 ms)		$P_{G(AV)}$	1	Watt
Peak Gate Current		I_{GM}	3	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)		$V_{(ISO)}$	2500	Volts
Operating Junction Temperature		T_J	−40 to +125	°C
Storage Temperature Range		T_{stg}	−40 to +125	°C
Circuit Fusing (t = 8.3 ms)	MAC625 Series MAC635 Series	I^2t	260 350	A^2s

(1) Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
(2) The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.5	°C/W

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Either Direction) Rated V_{DRM}, Gate Open	I_{DRM} I_{RRM}	—	—	10	μA
Peak On-State Voltage (Either Direction) I_{TM} = 1.4 $I_{T(RMS)}$	V_{TM}	—	—	1.4	Volts
Peak Gate Trigger Current Main Terminal Voltage = 6 Vdc, I_T = 1 A	I_{GT}				mA
MT2(+), G(+)		—	—	50	
MT2(+), G(−)		—	—	50	
MT2(−), G(−)		—	—	50	
MT2(−), G(+)		—	—	—	
Peak Gate Trigger Voltage Main Terminal Voltage = 6 Vdc, I_T = 1 A Minimum Gate Pulse Width = 2 μs	V_{GT}				Volts
MT2(+), G(+)		—	—	3	
MT2(+), G(−)		—	—	3	
MT2(−), G(−)		—	—	3	
MT2(−), G(+)		—	—	—	
Gate Non-Trigger Voltage Main Terminal Voltage = 1/2 Rated V_{DRM}, R_L = 10 kΩ, T_J = 125°C	V_{GD}	0.2	—	—	Volts
Holding Current (Either Direction)	I_H	—	30	—	mA
Turn-On Time V_D = 1/2 Rated V_{DRM}, I_{TM} = Rated I_T, I_{GT} = 100 mA	t_{gt}	—	—	10	μs
Critical Rate of Rise of Commutation Voltage V_D = 2/3 Rated V_{DRM}, Commutating di/dt = 15 A/ms T_J = 125°C	dv/dt(c)	6	—	—	V/μs
dv/dt @ V_D = 2/3 V_{DRM}, T_J = 125°C	dv/dt	100	—	—	V/μs

3

TYPICAL CHARACTERISTICS

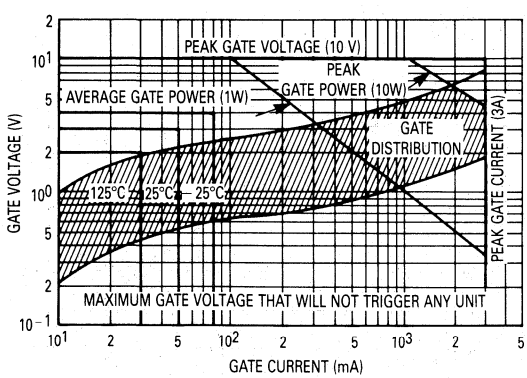


Figure 1. Gate Characteristics

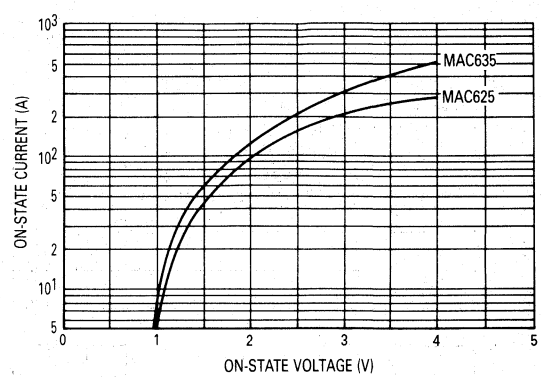


Figure 2. On-State Voltage

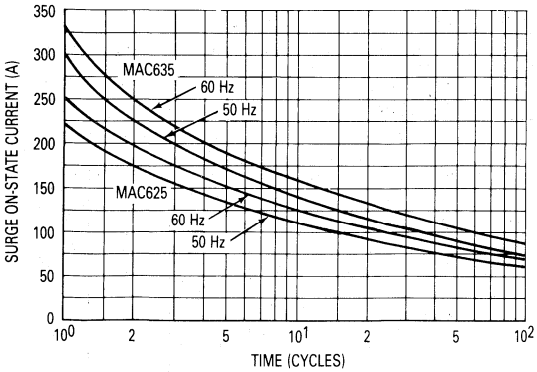


Figure 3. Surge On-State Current Rating (Non-Repetitive)

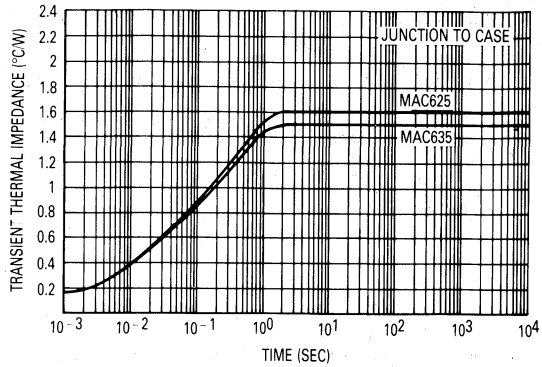


Figure 4. Transient Thermal Impedance

RMS On-State Current versus Maximum Power Dissipation

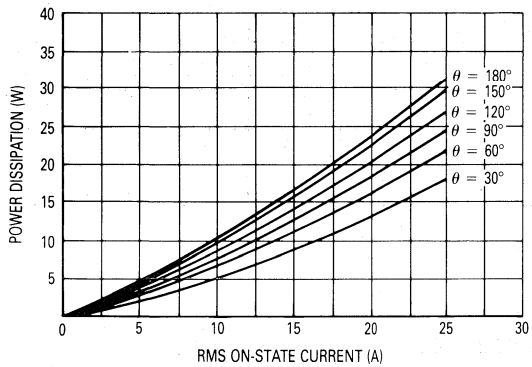


Figure 5. MAC625

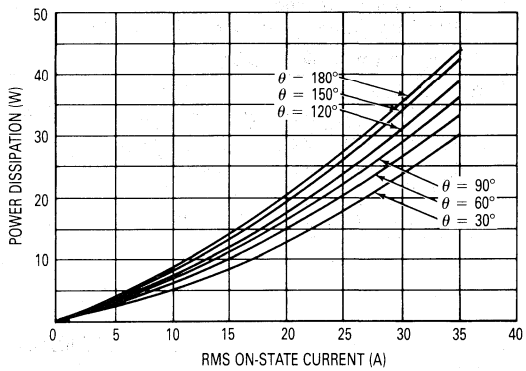


Figure 6. MAC635

RMS On-State Current versus Allowable Case Temperature

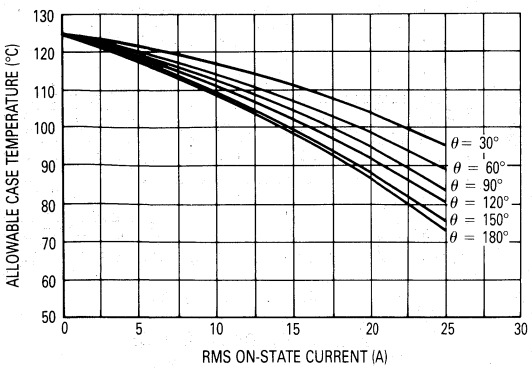


Figure 7. MAC625

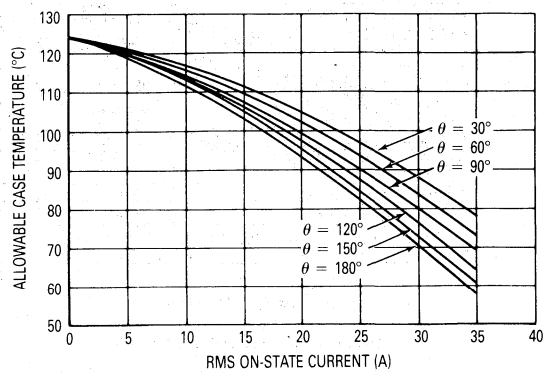


Figure 8. MAC635

Ambient Temperature versus RMS On-State Current

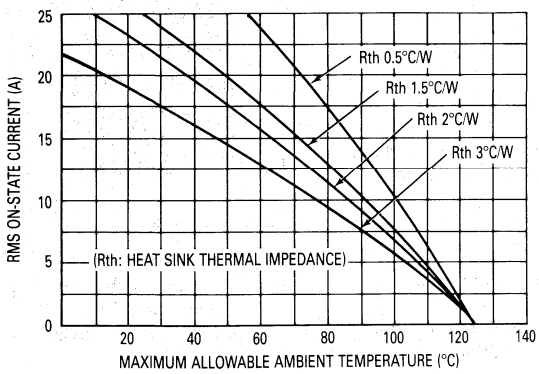


Figure 9. MAC625

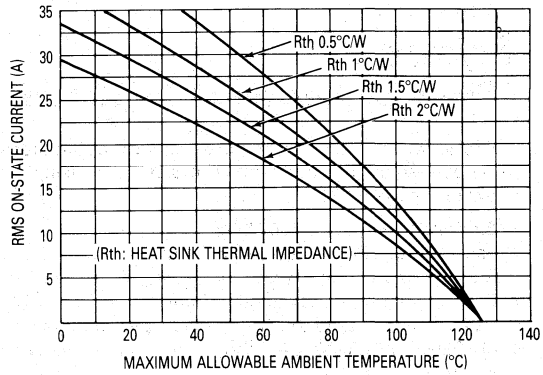


Figure 10. MAC635

3

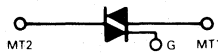
Triacs
Silicon Bidirectional Thyristors

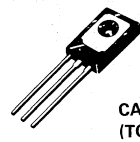
. . . designed for full-wave ac power control applications, and specifically designed to be used in conjunction with MOC30XX opto couplers in circuits similar to that shown on page 3-221.

- Blocking Voltages to 400 Volts
- Load Current Controlled Up to 40 A
- Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Gate Triggering Guaranteed in Four Modes
- Designed for Use with MOC Series Optoisolators Having Triac Driver Outputs
- MAC3010/MAC3030 Are Recommended For Use With MOC3010/MOC3030 Optoisolators
- MAC3020/MAC3040 Are Recommended For Use With MOC3020/MOC3040 Optoisolators

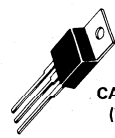
MAC3010
MAC3020
MAC3030
MAC3040
Series

TRIACs
4, 8, 15, 25 and 40
AMPERES RMS
250 thru 400 VOLTS

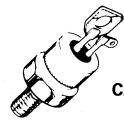




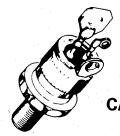
CASE 77-07
(TO-225AA)
STYLE 5
-4



CASE 221A-04
(TO-220AB)
STYLE 4
-8, -15, -25



CASE 263-04
STYLE 2
-40



CASE 311-02
STYLE 2
-40I

3

MAXIMUM RATINGS

Rating	Symbol	Current Ratings					Unit
		-4	-8	-15	-25	-40 -40I	
On-State RMS Current (see Figure 1) (Full Cycle Sine Wave 50 to 60 Hz)	$I_{T(RMS)}$	4	8	15	25	40	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_J = 110°C)	I_{TSM}	30	80	150	250	300	Amps
Circuit Fusing Considerations (T_J = −40 to +110°C, t = 8.3 ms)	I^2t	3.6	26	90	260	370	A^2s
Peak Gate Voltage (t ≤ 2 μs)	V_{GM}	±5	±10	±10	±10	±10	Volts
Peak Gate Power (t ≤ 2 μs)	P_{GM}	10	20	20	20	20	Watts
Average Gate Power (T_C = 80°C, t ≤ 8.3 ms)	$P_{G(AV)}$	0.5	0.5	0.5	0.5	0.5	Watts
Peak Gate Current (t ≤ 2 μs)	I_{GM}	11	12	12	12	12	Amps
Operating Junction Temperature Range	T_J	*	−40 to +125			*	°C
Storage Temperature Range	T_{stg}	−40 to +150					°C
Mounting Torque	—	6	8	8	8	30	in. lb.
MAC3010/MAC3030, Note 1 MAC3020/MAC3040	V_{DRM}	250 400	250 400	250 400	250 400	250 400	Volts

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

—4 CURRENT RATING

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Blocking Current (Note 1) (V_D = 250 V, T_J = 110°C)	I_{DRM}	—	—	2.0	mA
Peak On-State Voltage (Either Direction) (I_{TM} = 6 A Peak; Pulse Width ≤ 2.0 ms, Duty Cycle ≤ 2.0%)	V_{TM}	—	—	2.0	Volts
Gate Trigger Current, Continuous dc (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−)	I_{GT}	—	—	3.0	mA
Gate Trigger Voltage, Continuous dc (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−)	V_{GT}	—	—	2.0	Volts
(V_D = 250 V, R_L = 10 k Ω, T_J = 110°C) MT2(+), G(+); MT1(−), G(−)		0.2	—	—	
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	—	40	mA
Gate Controlled Turn-On Time (V_D = 250 V, I_{TM} = 6 A pk, I_G = 100 mA)	tgt	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage (V_D = 250 V, I_{TM} = 6 A pk, Commutating di/dt = 3.1 A/ms, Gate Unenergized, T_C = 85°C)	dv/dt(C)	—	5.0	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = 250 V, Exponential Waveform, T_C = 110°C)	dv/dt	—	20	—	V/μs

Note 1: Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking voltage such that the voltage applied exceeds the rated blocking voltage.

3

FIGURE 1 — CURRENT DERATING AND POWER DISSIPATION REFERENCE: CASE TEMPERATURE

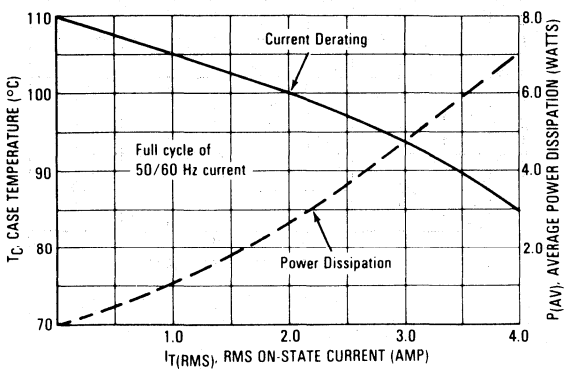
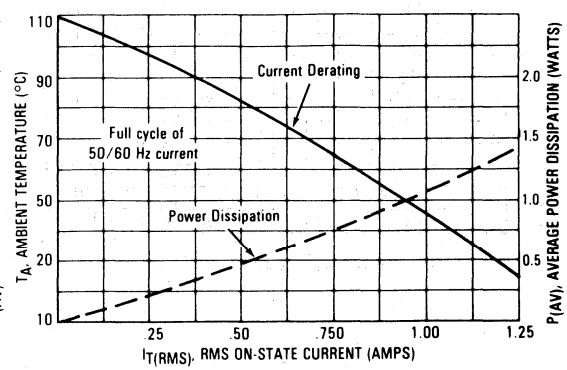


FIGURE 2 — CURRENT DERATING AND POWER DISSIPATION REFERENCE: AMBIENT TEMPERATURE



-8, -15, -25 CURRENT RATINGS

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Peak On-State Voltage ($I_{TM} = \sqrt{2}\, I_{T(RMS)}$ A Peak; Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}				Volts
MAC3030-8		—	—	1.6	
MAC3030-15		—	—	1.6	
MAC3030-25		—	—	1.85	
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ohms) MT2(+), G(+); MT2(−), G(−) All Types	I_{GT}	—	—	40	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ohms)	V_{GT}				Volts
MT2(+), G(+); MT2(−), G(−) All Types		—	—	2	
(T_J = 125°C, R_L = 10 k Ohms) MT2(+), G(+); MT1(−), G(−) All Types		0.2	—	—	
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	—	40	mA
Gate Controlled Turn-On Time ($I_{TM} = 2\, I_{T(RMS)}$ A Peak, I_G = 100 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage ($I_{TM} = 2\, I_{T(RMS)}$ A Peak, Commutating di/dt = 0.52 $I_{T(RMS)}$ A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (Exponential Waveform, T_C = 125°C)	dv/dt	—	40	—	V/μs

3

FIGURE 2 — RMS CURRENT DERATING

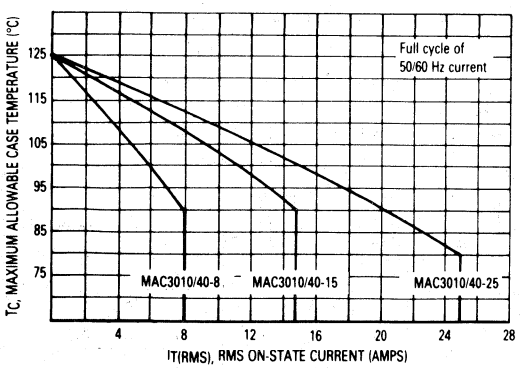
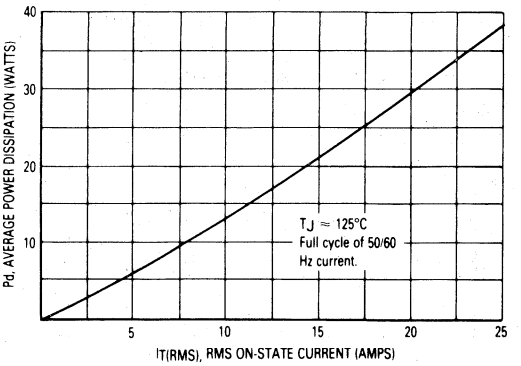


FIGURE 3 — ON-STATE POWER DISSIPATION



-40, -40I CURRENT RATINGS

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 110°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
Peak On-State Voltage (Either Direction) (I_{TM} = 56 A Peak; Pulse Width ≤ 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.85	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−)	I_{GT}	—	—	40	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) MT2(+), G(+); MT2(−), G(−) (R_L = 10 kΩ, T_J = 110°C) MT2(+), G(+); MT1(−), G(−)	V_{GT}	— 0.2	— —	2 —	Volts
Holding Current (V_D = 12 V, I_{TM} = 200 mA, Gate Open)	I_H	—	—	50	mA
Gate Controlled Turn-On Time (I_{TM} = 56 A pk, I_G = 200 mA)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Commutation Voltage (I_{TM} = 56 A pk, Commutating di/dt = 22 A/ms, Gate Unenergized, T_C = 60°C)	dv/dt(c)	—	5	—	V/μs
Critical Rate of Rise of Off-State Voltage (Exponential Waveform, T_C = 110°C)	dv/dt	—	30	—	V/μs

3

FIGURE 4 — RMS CURRENT DERATING

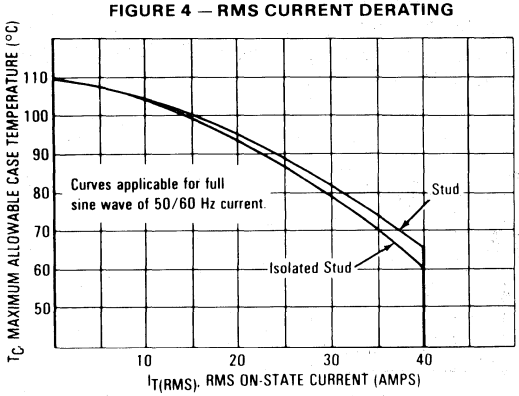
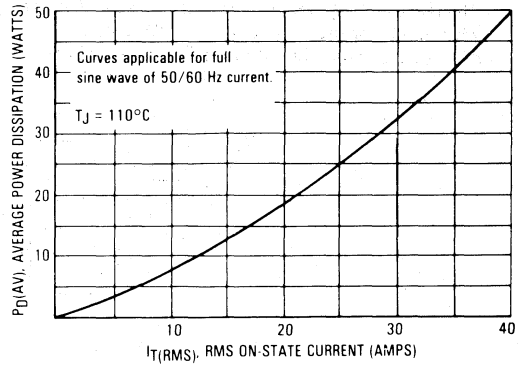


FIGURE 5 — ON-STATE POWER DISSIPATION



USING THE MOC OPTO COUPLERS AND MAC TRIAC SERIES DEVICES

The MOCXXXX Opto Coupler can be used as a triac driver with MACXXXX-X by selecting R_C to limit the surge current thru the coupler and yet supply enough gate drive to the triac to guarantee complete turn on. The maximum surge current rating of the coupler (I_{TSM}) determines the minimum value of R_C:

$$R_C\,(\text{min}) = \frac{V_{in}(pk)}{I_{TSM}\,(\text{coupler})}$$

For high line 110 Vac nominal voltage: $V_{in}(pk) = 187$ V.

$$R_C\,(\text{min}) = \frac{187\text{ V}}{1.2\text{ A}} = 155.8\text{ ohms}$$

In practice, this would be a 180 ohm resistor.

The maximum gate drive required determines the maximum value of R_C:

$$R_C\,(\text{max}) = \frac{V_{IH}-V_{TM}}{I_{GT}(\text{triac})}$$

Where V_{IH} is the inhibit voltage of the coupler and V_{TM} is the on-state voltage of the triac in the coupler.

For the MOC3040 and MAC3040 – 25 V_{IH} = 40 V. V_{TM} = 3.0 V, and I_{GT} = 40 mA.

$$R_C\,(\text{max}) = \frac{40\text{ V}-3.0\text{ V}}{40\text{ mA}} = 930\text{ ohms}$$

In practice, the gate is driven two or three times I_{GT} to guarantee complete turn on. R_C (max) would be 460 ohms or 310 ohms.

The line voltage at turn on is:

$$V_{\text{Line at turn on}} = R_C \cdot I_{GT} + V_{TM}(\text{coupler}) + V_{GT}(\text{triac})$$

For the above example V_{GT}(triac) = 2.0 V, I_{GT} = 80 mA, R_C = 210 ohms.

$$V_{\text{Line at turn on}} = (210)(0.08\text{ A}) + 3.0\text{V} + 2.0\text{V} = 22\text{ V}$$

Resistive Loads

Resistive heating elements and incandescent lamps are typical loads for the triac. Cold incandescent lamps can draw 5–6 times their hot RMS value on start up. The triac must be specified to sustain the repetitive surge (I_{TSM}). In practice, the RMS value is chosen at two times actual so the surge rating of the triac will be very high.

Inductive Loads

Motors, solenoids, and magnets are typical problem loads for the triac and coupler. Since the triac turns off as the current approaches zero, but the inductive voltage is still high, it appears to the triac as a rise in applied voltage. If this rate of rise in voltage exceeds the dv/dt commutating of the triac or the dv/dt static of the coupler, the triac will turn back on.

Snubber Network

When the dv/dt of the circuit exceeds the capability of the coupler or triac, a R_SC_S network is placed across the main terminals of the triac. In most applications the snubber used for the triac will also protect the coupler. The R_S also limits the energy from the C_S destroying the gate region on the first use of the triac.

Since the power factor of the board (cosine of the I-V phase shift) is not always known, a typical design can be a starting point for scope verification.

For power factor = 0.1, 110 V nominal line.

$$V_{\text{turn off voltage}} = V_{pk}\sin\phi \approx V_{pk} \approx 187\text{ V}$$

Setting the dv/dt $_C$ (triac) equal to the circuit $V_{Turn\ off}$ over the snubber time constant and solving for R_S:

$$dv/dt\ C\,(\text{triac}) = \frac{V_{Turn\ off}}{R_SC_S}$$

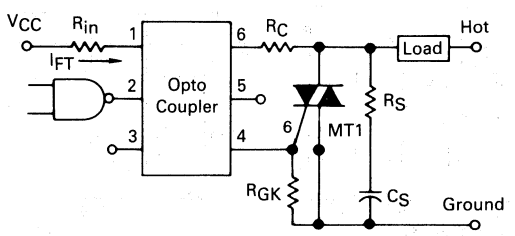
$$R_S = \frac{V_{Turn\ off}}{dv/dt\,(c)\,C_S}$$

For MAC3030-25 dv/dt (c) = 5.0 V/μs, and choosing C_S = 0.1 μF

$$R_S = \frac{187\text{ V}}{(5.0\text{ V}/\mu\text{s})(0.1\ \mu\text{F})}$$

$$R_S = 374\text{ ohms}$$

In practice, R_S is selected empirically. For more details see AN1048 (Section 1).



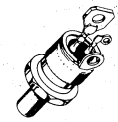
Triacs
Silicon Bidirectional Triode Thyristors

MAC4120 Series

TRIACs
15 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and military applications for full wave control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Isolated Stud Package
- Gate Triggering Guaranteed In All 4 Quadrants



CASE 311-02
STYLE 2

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −65 to +100°C) Gate Open	V_{DRM}		Volts
MAC4120 B		200	
D		400	
M		600	
N		800	
RMS On-State Current (Conduction Angle = 360°) T_C = +75°C	$I_{T(RMS)}$	15	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (Pulse Width = 1 μs)	P_{GM}	16	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Peak Gate Trigger Current (Pulse Width = 1 μs)	I_{GTM}	4	Amps
Operating Case Junction Temperature Range	T_C	−65 to +100	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.1	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C, either polarity of MT2 to MT1 voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	2	mA
Peak On-State Voltage (I_T = 21 A Peak)	V_{TM}	—	1.4	1.8	Volt
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 30 Ohms)	I_{GT}				mA
$V_{MT2(+)}$, $V_{G(+)}$; $V_{MT2(-)}$, $V_{G(-)}$		—	—	50	
$V_{MT2(+)}$, $V_{G(-)}$; $V_{MT2(-)}$, $V_{G(+)}$		—	—	80	
$V_{MT2(+)}$, $V_{G(+)}$; $V_{MT2(-)}$, $V_{G(-)}$, T_C = −65°C		—	—	150	
$V_{MT2(+)}$, $V_{G(-)}$; $V_{MT2(-)}$, $V_{G(+)}$, T_C = −65°C		—	—	200	
Gate Trigger Voltage (Continuous dc) (All Quadrants) (V_D = 12 Vdc, R_L = 30 Ohms)	V_{GT}				Volts
T_C = 25°C		—	—	2.5	
T_C = −65°C		—	—	4	
(V_D = Rated V_{DROM}, R_L = 125 Ohms, T_C = 100°C)		0.2	—	—	
Holding Current (V_D = 12 Vdc, Gate Open) (I_T = 500 mA) T_C = 25°C	I_H	—	—	75	mA
T_C = −65°C		—	—	300	
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 25 A Peak, I_{GT} = 160 mA, Rise Time = 0.1 μs)	t_{gt}	—	1.6	2.5	μs
Critical Rate of Rise of Commutation Voltage (Rated V_{DRM}, $I_{T(RMS)}$ = 15 A, Cummutating di/dt = 8 A/ms, Gate Unenergized, T_C = 75°C)	dv/dt(c)	2	10	—	V/μs
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 100°C)	dv/dt				V/μs
MAC4120 B		30	150	—	
D		20	100	—	
M		10	75	—	
N		10	—	—	

Note 1. All Voltage polarities referenced to main terminal 1.

3

FIGURE 1 – CURRENT DERATING

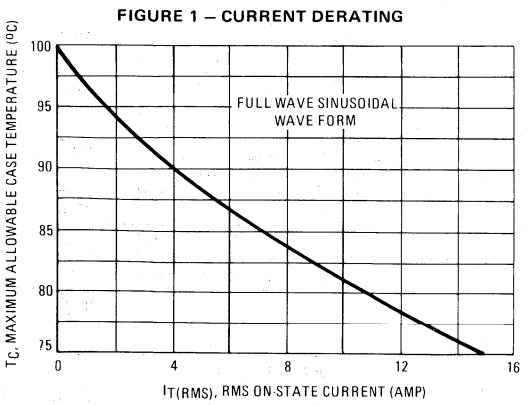
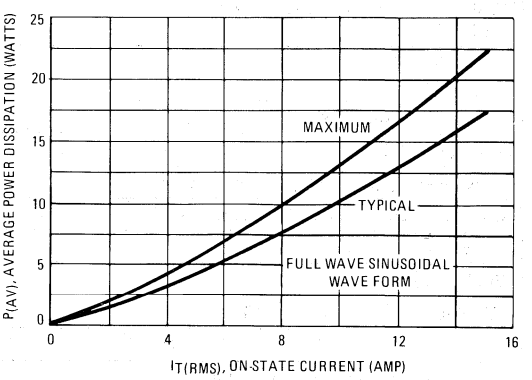


FIGURE 2 – POWER DISSIPATION



Triacs
Silicon Bidirectional Triode Thyristors

MAC4121 Series (See 2N5567)

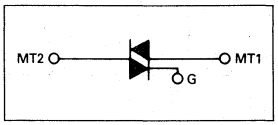
MAC5441–43 (See 2N5444)

MAC6400 T6410 T6420 Series

TRIACs
40 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire
- Press Fit Stud — MAC6400
  Stud — T6410
  Isolated Stud — T6420
- Gate Triggering Guaranteed in All 4 Quadrants



CASE 263-04
STYLE 2
T6410
STUD

CASE 174-04
(TO-203AA)
STYLE 3
MAC6400
PRESS FIT

CASE 311-02
STYLE 2
T6420
ISOLATED STUD

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −65 to +110°c) Gate Open MAC6400B, T6410B, T6420B MAC6400D, T6410D, T6420D MAC6400M, T6410M, T6420M MAC6400N, T6410N, T6420N	V_{DRM}	 200 400 600 800	Volts
On-State Current RMS (Conduction Angle = 360°) T_C (Pressfit) = 70°C T_C (Stud) = 65°C	$I_{T(RMS)}$	40	Amps
Peak Surge Current (Non-Repetitive) (One Full Cycle, 60 Hz)	I_{TSM}	300	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	375	A^2s
Peak Gate Power (Pulse Width = 10 μs)	P_{GM}	40	Watts
Average Gate Power	$P_{G(AV)}$	0.75	Watt
Peak Gate Current (Pulse Width = 1 μs)	I_{GTM}	12	Amps
Operating Temperature Range	T_C	−65 to +110	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case Pressfit Stud Isolated Stud	$R_{\theta JC}$	0.8 0.9 1	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

3

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	4	mA
Maximum On-State Voltage (Either Direction) (I_T = 100 A Peak)	V_{TM}	—	1.5	2	Volts
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 30 Ohms	I_{GT}				mA
$V_{MT2(+)}$, $V_{G(+)}$		—	15	50	
$V_{MT2(+)}$, $V_{G(-)}$		—	30	80	
$V_{MT2(-)}$, $V_{G(-)}$		—	20	50	
$V_{MT2(-)}$, $V_{G(+)}$		—	40	80	
$V_{MT2(+)}$, $V_{G(+)}$, $V_{MT2(-)}$, $V_{G(-)}$, T_C = −65°C		—	—	125	
$V_{MT2(+)}$, $V_{G(-)}$, $V_{MT2(-)}$, $V_{G(+)}$, T_C = −65°C		—	—	240	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 30 Ohms, T_C = 25°C	V_{GT}	—	1.35	2.5	Volts
T_C = −65°C		—	—	3.4	
(V_D = Rated V_{DRM}, R_L = 125 Ohms, T_C = 110°C)		0.2	—	—	
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open) (Initiating Current = 500 mA) T_C = 25°C	I_{HO}	—	25	60	mA
T_C = −65°C		—	—	100	
Gate Controlled Turn-On Time (Rated V_{DRM}, I_T = 60 A, I_{GT} = 200 mA, Rise Time = 0.1 μs)	t_{gt}	—	1.7	3	μs
Critical Rate of Rise of Commutation Voltage, On-State Conditions (di/dt = 22 A/ms, Gate Unenergized, V_D = Rated V_{DROM}, $I_{T(RMS)}$ = 40 A, T_C (Pressfit) = 70°C) T_C (Stud) = 65°C	dv/dt(c)	—	5	—	V/μs

Note 1. All voltage polarities referenced to main terminal 1.

3

FIGURE 1 – ON-STATE POWER DISSIPATION

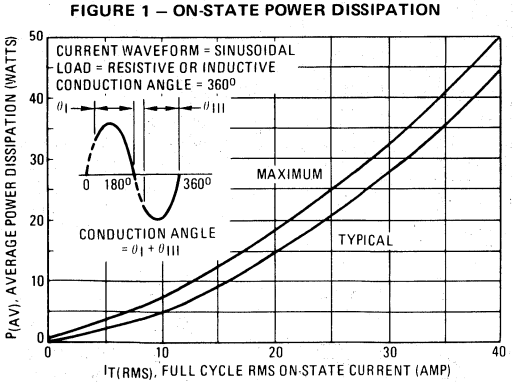
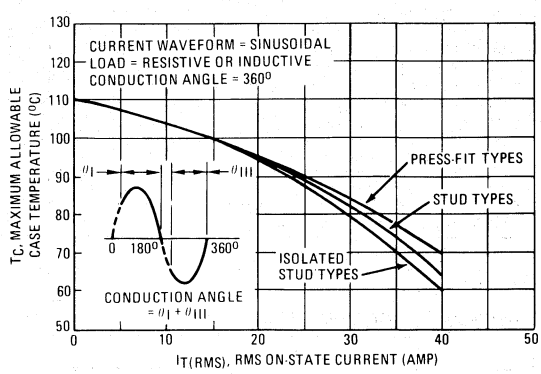


FIGURE 2 – RMS CURRENT DERATING



Triacs

Silicon Bidirectional Triode Thyristors

. . . designed primarily for industrial and military applications for full wave control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems.

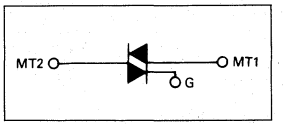
- Glass Passivated Junctions and Center Gate Fire
- Press Fit, Stud, Isolated Stud Packages
- Gate Triggering Guaranteed In All 4 Modes

MAC5569–70
(See 2N5567)

MAC5573–74
(See 2N5571)

MAC6401
T6411
T6421
Series

TRIACs
30 AMPERES RMS
200 thru 800 VOLTS



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −65 to +100°C) Gate Open	V_{DRM}		Volts
MAC6401B, T6411B, T6421B		200	
MAC6401D, T6411D, T6421D		400	
MAC6401M, T6411M, T6421M		600	
MAC6401N, T6411N, T6421N		800	
On-State Current RMS (Conduction Angle = 360°) $T_C \leq +65°C$	$I_{T(RMS)}$	30	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	300	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	375	A^2s
Peak Gate Power (Pulse Width = 1 μs)	P_{GM}	40	Watts
Average Gate Power	$P_{G(AV)}$	0.75	Watt
Peak Gate Current Pulse Width ≤ 1 μs)	I_{GTM}	2	Amps
Operating Case Temperature Range	T_C	−65 to +100	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case Pressfit	$R_{\theta JC}$	0.8	°C/W
Stud		0.9	
Isolated Stud		1	

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

CASE 174-04
(TO-203AA)
STYLE 3
MAC6401
PRESS FIT

CASE 263-04
STYLE 2
T6411
STUD

CASE 311-02
STYLE 2
T6421
ISOLATED STUD

3

ELECTRICAL CHARACTERISTICS (T_C = 25°C, and Either Polarity of MT2 to MT1, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	4	mA
Maximum On-State Voltage (Either Direction) (I_T = 100 A Peak)	V_{TM}	—	2.1	2.5	Volts
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 30 Ohms	I_{GT}				mA
$V_{MT2(+)}$, $V_{G(+)}$; $V_{MT2(-)}$, $V_{G(-)}$		—	20	50	
$V_{MT2(+)}$, $V_{G(-)}$; $V_{MT2(-)}$, $V_{G(+)}$		—	35	80	
Gate Trigger Voltage (Continuous dc) (All Trigger Modes) (V_D = 12 Vdc, R_L = 30 Ohms)	V_{GT}	—	1.35	2.5	Volts
(V_D = Rated V_{DRM}, R_L = 125 Ohms, T_C = 100°C)		0.2	—	—	
Holding Current (V_D = 12 Vdc, Gate Open) (I_T = 150 mA)	I_{HO}	—	—	60	mA
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 45 A, I_{GT} = 200 mA, Rise Time = 0.1 μs)	t_{gt}	—	1.7	3	μs
Critical Rate of Rise of Commutation Voltage, On-State Conditions (di/dt = 16 A/ms, Gate Unenergized, V_D = Rated V_{DRM}, $I_{T(RMS)}$ = 30 A, T_C = Rated Value from Figure 1)	dv/dt(c)	3	20	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Rise, T_C = 100°C)	dv/dt				V/μs
MAC6401B, T6411B, T6421B		40	—	—	
MAC6401D, T6411D, T6421D		25	—	—	
MAC6401M, T6411M, T6421M		20	—	—	
MAC6401N, T6411N, T6421N		20	—	—	

3

Note 1. All voltage polarities referenced to main terminal 1.

FIGURE 1 – CURRENT DERATING

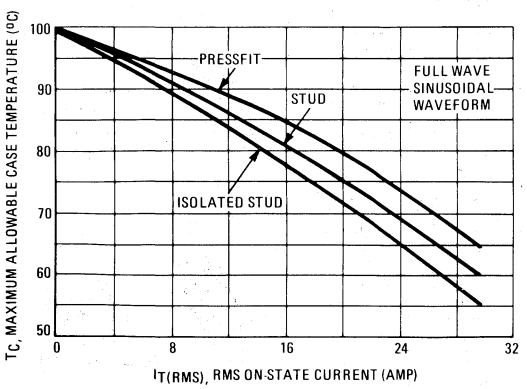
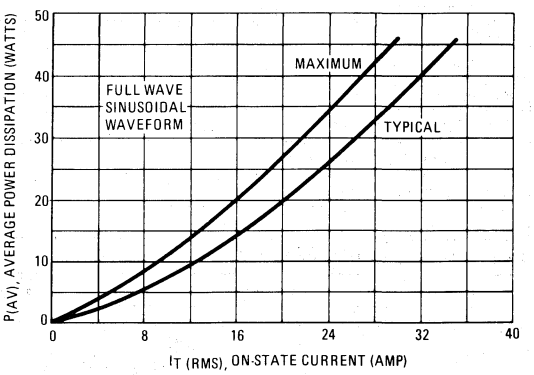


FIGURE 2 – POWER DISSIPATION



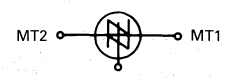
Silicon Bidirectional Switches
Diode Thyristors

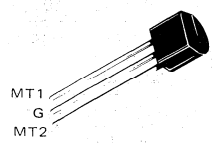
MBS4991
MBS4992
MBS4993

SBS
(PLASTIC)

. . . designed for full-wave triggering in Triac phase control circuits, half-wave SCR triggering application and as voltage level detectors. Supplied in an inexpensive plastic TO-226AA package for high-volume requirements, this low-cost plastic package is readily adaptable for use in automatic insertion equipment.

- Low Switching Voltage — 8 Volts Typical
- Uniform Characteristics in Each Direction
- Low On-State Voltage — 1.7 Volts Maximum
- Low Off-State Current — 0.1 μA Maximum
- Low Temperature Coefficient — 0.02 %/°C Typical





CASE 29-04
(TO-226AA)
STYLE 12

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Power Dissipation	P_D	500	mW
DC Forward Current	I_F	200	mA
DC Gate Current (off-state only)	I_G(off)	5	mA
Repetitive Peak Forward Current (1% Duty Cycle, 10 μs Pulse Width, T_A = 100°C)	I_{FM}(rep)	2	Amps
Non-Repetitive Forward Current (10 μs Pulse Width, T_A = 25°C)	I_{FM}(nonrep)	6	Amps
Operating Junction Temperature Range	T_J	−55 to +125	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
Switching Voltage	MBS4991 MBS4992, MBS4993	V_S	6 7.5	8 8	10 9	Vdc
Switching Current	MBS4991 MBS4992 MBS4993	I_S	— —	175 90 175	500 120 250	μAdc
Switching Voltage Differential (See Figure 10)	MBS4991 MBS4992, MBS4993	$\lvert V_{S1}-V_{S2}\rvert$	— —	0.3 0.1	0.5 0.2	Vdc
Gate Trigger Current (V_F = 5 Vdc, R_L = 1 k ohm)	MBS4992 MBS4993	I_{GF}	— —	— —	100 500	μAdc
Holding Current	MBS4991 MBS4992 MBS4993	I_H	— — —	0.7 0.2 0.3	1.5 0.5 0.75	mAdc
Off-State Blocking Current (V_F = 5 Vdc, T_A = 25°C) (V_F = 5 Vdc, T_A = 85°C) (V_F = 5 Vdc, T_A = 25°C) (V_F = 5 Vdc, T_A = 100°C)	 MBS4991 MBS4991 MBS4992, MBS4993 MBS4992, MBS4993	I_B	 — — — —	 0.08 2 0.08 6	 1 10 0.1 10	μAdc
Forward On-State Voltage (I_F = 175 mAdc) (I_F = 200 mAdc)	 MBS4991 MBS4992, MBS4993	V_F	 — —	 1.4 1.5	 1.7 1.7	Vdc
Peak Output Voltage (C_C = 0.1 μF, R_L = 20 ohms, (Figure 7)		V_O	3.5	4.8	—	Vdc
Turn-On Time (Figure 8)		t_{on}	—	1	—	μs
Turn-Off Time (Figure 9)		t_{off}	—	30	—	μs
Temperature Coefficient of Switching Voltage (−50 to +125°C)		T_C	—	+0.02	—	%/°C
Switching Current Differential (See Figure 10)		$I_{S1}-I_{S2}$	—	—	100	μA

3

TYPICAL ELECTRICAL CHARACTERISTICS

FIGURE 1 – SWITCHING VOLTAGE versus TEMPERATURE

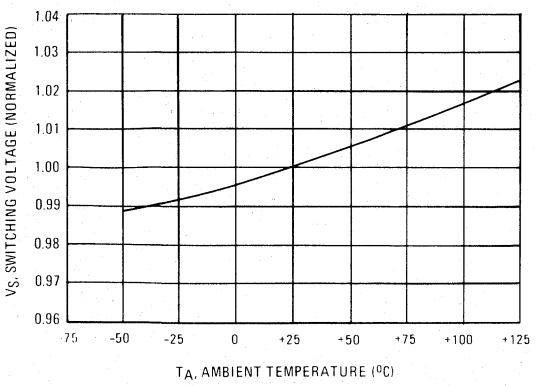


FIGURE 2 – SWITCHING CURRENT versus TEMPERATURE

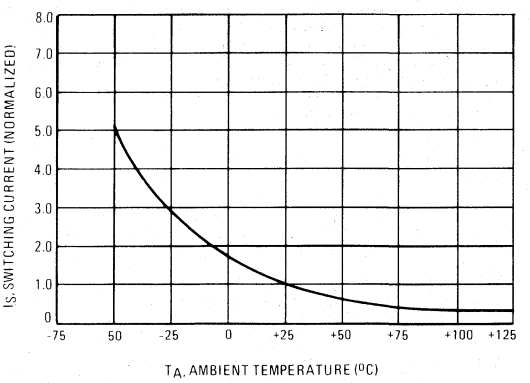


FIGURE 3 – HOLDING CURRENT versus TEMPERATURE

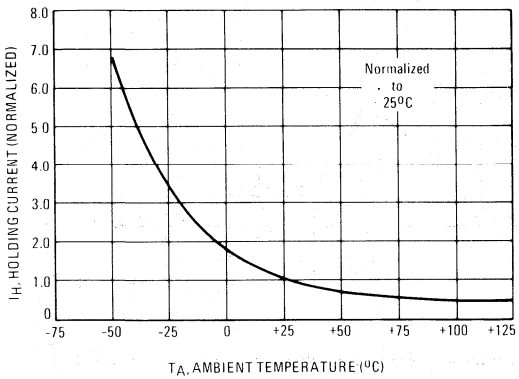


FIGURE 4 – OFF-STATE BLOCKING CURRENT versus TEMPERATURE

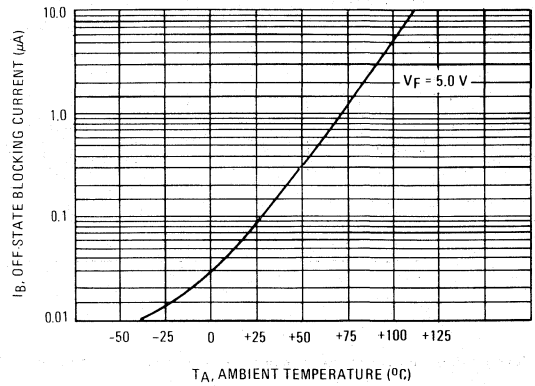


FIGURE 5 – ON-STATE VOLTAGE versus FORWARD CURRENT

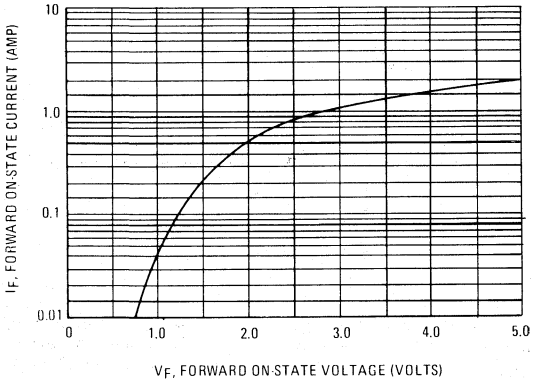


FIGURE 6 – PEAK OUTPUT VOLTAGE (FUNCTION OF R_L AND C_c)

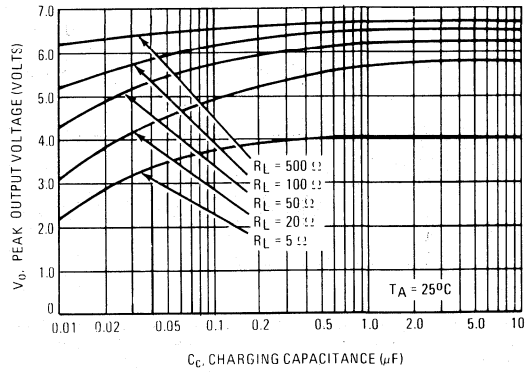


FIGURE 7 – PEAK OUTPUT VOLTAGE TEST CIRCUIT

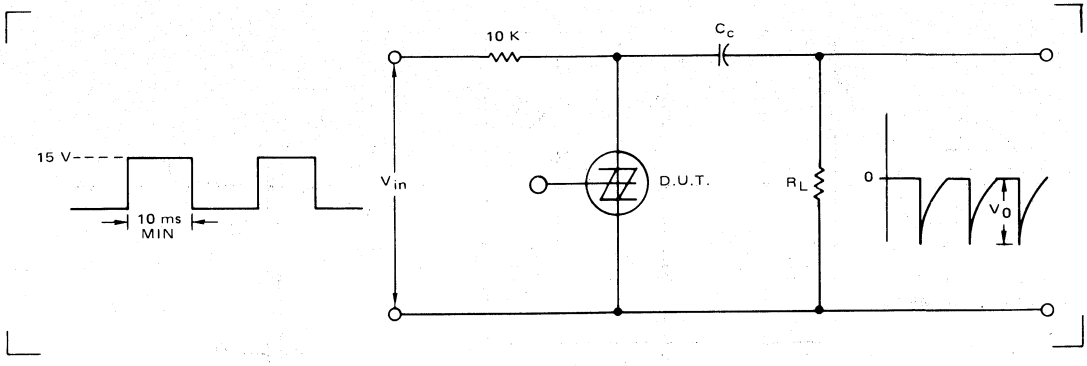
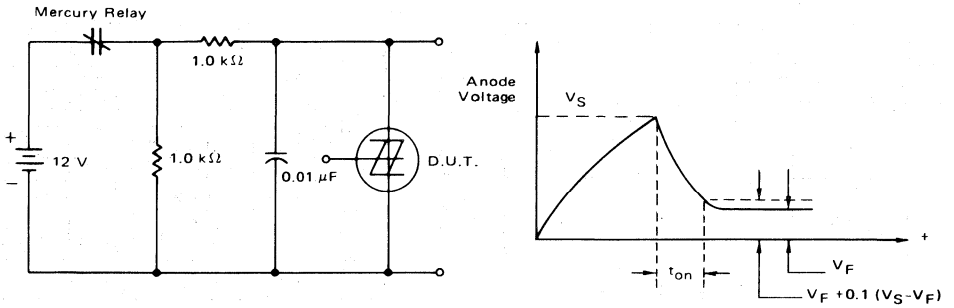
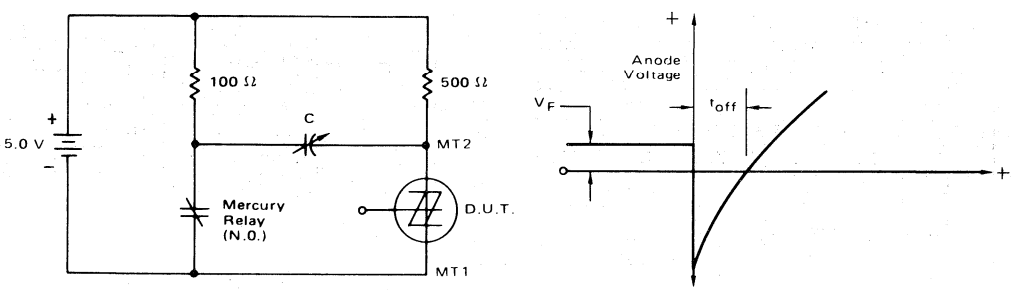


FIGURE 8 – TURN-ON TIME TEST CIRCUIT



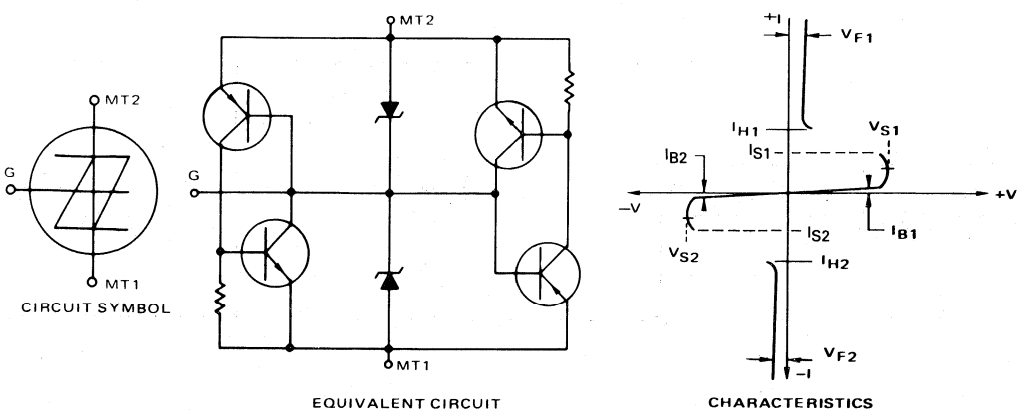
Turn-on time is measured from the time V_S is achieved to the time when the anode voltage drops to within 90% of the difference between V_S and V_F.

FIGURE 9 – TURN-OFF TIME TEST CIRCUIT



With the SBS in conduction and the relay contacts open, close the contacts to cause anode A2 to be driven negative. Decrease C until the SBS just remains off when anode A2 becomes positive. The turn off time, t_{off}, is the time from initial contact closure and until anode A2 voltage reaches zero volts.

FIGURE 10 – DEVICE EQUIVALENT CIRCUIT, CHARACTERISTICS AND SYMBOLS



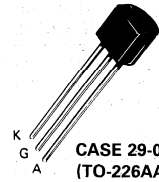
Silicon Controlled Rectifiers

MCR22-2 thru 8

SCRs
1.5 AMPERES RMS
50 thru 600 VOLTS

. . . designed and tested for repetitive peak operation required for CD ignition, fuel ignitors, flash circuits, motor controls and low-power switching applications.

- 150 Amperes for 2 μs Safe Area
- High dv/dt
- Very Low VF at High Current
- Low-Cost TO-226AA (TO-92)



CASE 29-04
(TO-226AA)
STYLE 10

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Reverse Blocking Voltage	V_{RRM}		Volts
MCR22-2		50	
MCR22-3		100	
MCR22-4		200	
MCR22-6		400	
MCR22-8		600	
On-State Current RMS (All Conduction Angles)	$I_{T(RMS)}$	1.5	Amps
Peak Nonrepetitive Surge Current, T_A = 25°C (1/2 Cycle, Sine Wave, 60 Hz)	I_{TSM}	15	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	0.9	A^2s
Peak Gate Power, T_A = 25°C	P_{GM}	0.5	Watt
Average Gate Power, T_A = 25°C	$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current, T_A = 25°C (300 μs, 120 PPS)	I_{FGM}	0.2	Amp
Peak Reverse Gate Voltage	V_{RGM}	5	Volts
Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Lead Solder Temperature (Lead Length ≥ 1/16″ from case, 10 s Max)	—	+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	50	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	160	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted. R_{GK} = 1000 Ohms.)

Characteristic		Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (T_C = 125°C)		V_{DRM}				Volts
	MCR22-2		50	—	—	
	MCR22-3		100	—	—	
	MCR22-4		200	—	—	
	MCR22-6		400	—	—	
	MCR22-8		600	—	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM})	T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
	T_C = 125°C		—	—	200	μA
Forward "On" Voltage (I_{TM} = 1 A peak)		V_{TM}	—	1.2	1.7	Volts
Gate Trigger Current (Continuous dc), Note 1 (Anode Voltage = 6 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	30	200	μA
	T_C = −40°C		—	—	500	
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	V_{GT}	—	—	0.8	Volts
	T_C = −40°C		—	—	1.2	
(Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms)	T_C = 125°C	V_{GD}	0.1	—	—	
Holding Current (Anode Voltage = 12 Vdc)	T_C = 25°C	I_H	—	2	5	mA
	T_C = −40°C		—	—	10	
Forward Voltage Application Rate (T_C = 125°C)		dv/dt	—	25	—	V/μs
Turn-On Time		t_{on}	—	1.2	—	μs
Turn-Off Time		t_{off}	—	40	—	μs

Note 1. R_{GK} Current Not Included in Measurement.

3

CURRENT DERATING

FIGURE 1 — MAXIMUM CASE TEMPERATURE

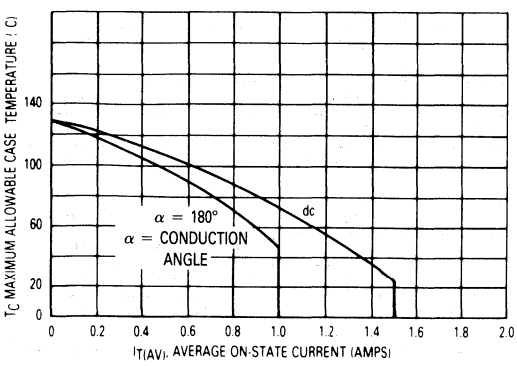


FIGURE 2 — MAXIMUM AMBIENT TEMPERATURE

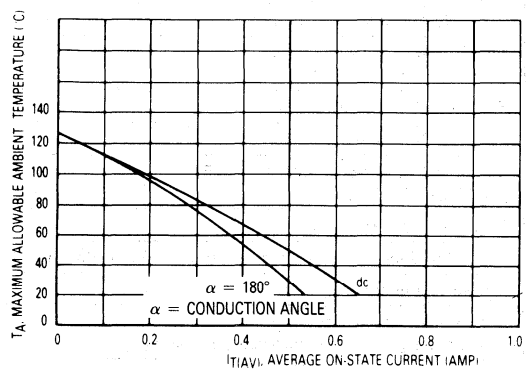


FIGURE 3 — TYPICAL FORWARD VOLTAGE

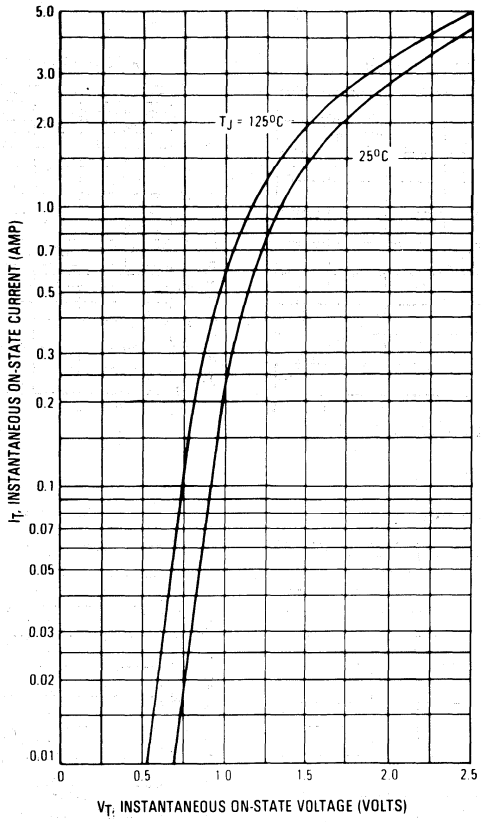
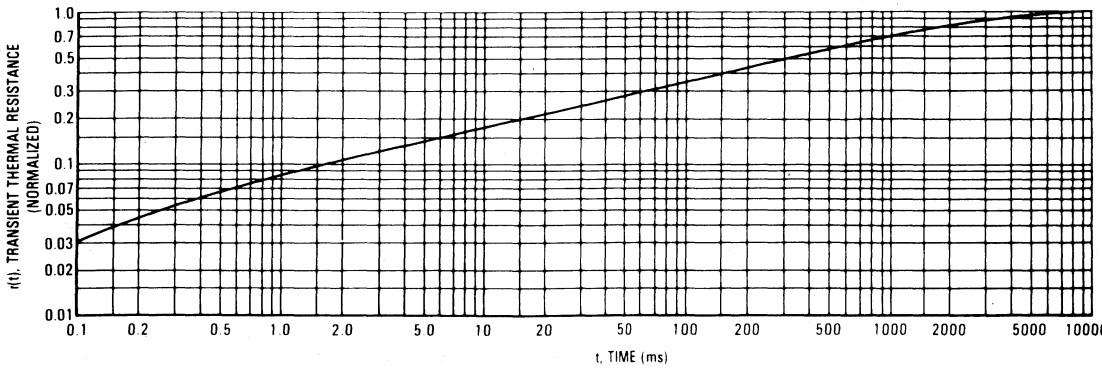


FIGURE 4 — THERMAL RESPONSE



TYPICAL CHARACTERISTICS

FIGURE 5 — GATE TRIGGER VOLTAGE

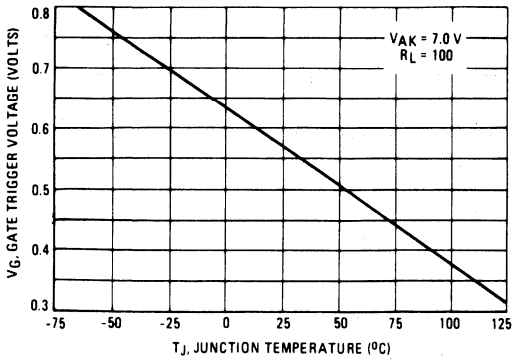


FIGURE 6 — TYPICAL GATE TRIGGER CURRENT

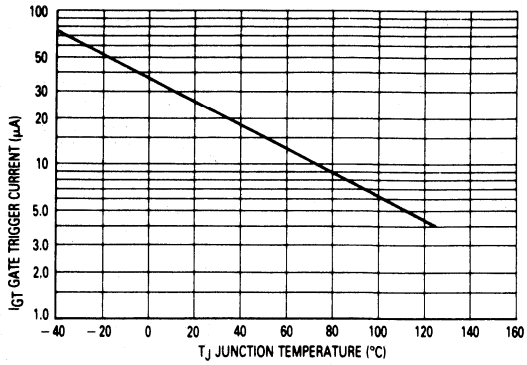


FIGURE 7 — HOLDING CURRENT

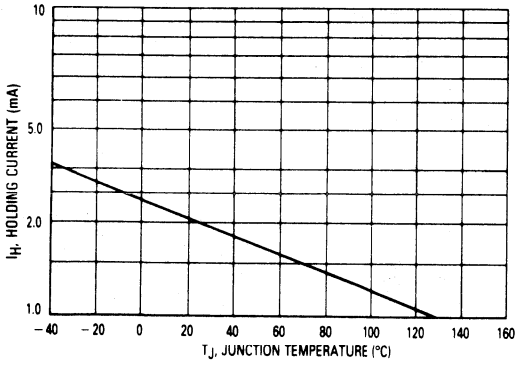
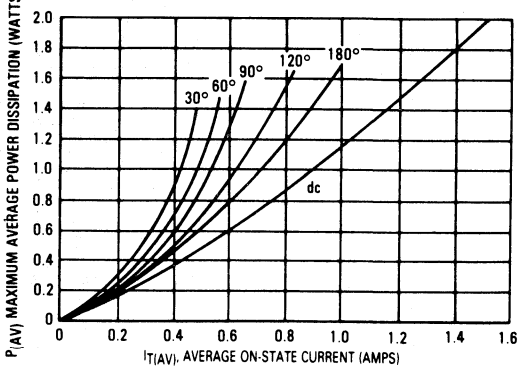


FIGURE 8 — POWER DISSIPATION



3

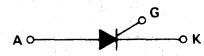
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

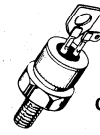
MCR63-( )A Series
MCR64 Series
MCR65 Series

SCRs
55 AMPERES RMS
50 thru 800 VOLTS

. . . designed for industrial and consumer applications such as power supplies; battery chargers; temperature, motor, light, and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 550 Amps
- Rugged Construction in Either Pressfit, Stud, or Isolated Stud
- Glass Passivated Junctions for Maximum Reliability





CASE 263-04
STYLE 1
MCR64 Series

CASE 174-04
(TO-203AA)
STYLE 1
MCR63-( )A Series

CASE 311-02
STYLE 1
MCR65 Series

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage, Note 1		V_{DRM} or V_{RRM}		Volts
MCR63-( )A	2		50	
MCR64-	3		100	
MCR65-	4		200	
	6		400	
	8		600	
	10		800	
Non-Repetitive Peak Reverse Blocking Voltage (t ≤ 5 ms)		V_{RSM}		Volts
MCR63-( )A	2		75	
MCR64-	3		150	
MCR65-	4		300	
	6		500	
	8		700	
	10		900	
Forward Current RMS		$I_{T(RMS)}$	55	Amps
Peak Surge Current (One cycle, 60 Hz, T_J = −40 to +125°C)		I_{TSM}	550	Amps
Circuit Fusing Considerations (t = 8.3 ms)		I^2t	1255	A^2s
Peak Gate Power		P_{GFM}	20	Watts
Average Gate Power (Pulse Width ≤ 2 μs)		$P_{GF(AV)}$	0.5	Watt
Peak Forward Gate Current		I_{GFM}	2	Amps
Peak Gate Voltage — Forward		V_{GFM}	10	Volts
Reverse		V_{GRM}	10	
Operating Junction Temperature Range		T_J	−40 to +125	°C
Storage Temperature Range		T_{stg}	−40 to +150	°C
Stud Torque		—	30	in. lb.

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode.

3

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$		°C/W
Pressfit and Stud		1	
Isolated Stud		1.1	

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	10	μA
T_J = 125°C		—	2	mA
Forward "On" Voltage (I_{TM} = 175 A Peak)	V_{TM}	—	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 50 Ω) T_C = 25°C	I_{GT}	—	40	mA
T_C = −40°C		—	75	
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 50 Ω) T_C = 25°C	V_{GT}	—	3	Volts
T_C = −40°C		—	3.5	
(V_D = Rated V_{DRM}, R_L = 1 kΩ, T_J = 125°C)		0.2	—	
Holding Current (V_D = 12 V, R_L = 50 Ω, Gate Open)	I_H	—	60	mA
Forward Voltage Application Rate (T_J = 125°C, V_D = Rated V_{DRM})	dv/dt	50	—	V/μs

3

FIGURE 1 — AVERAGE CURRENT DERATING

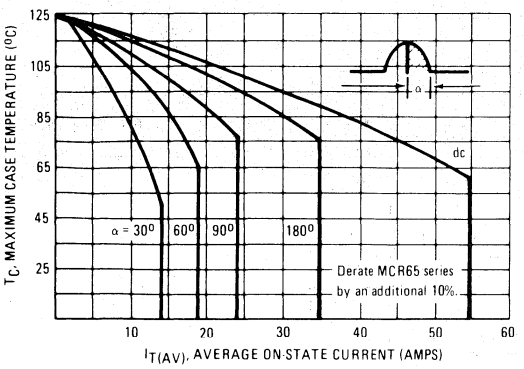
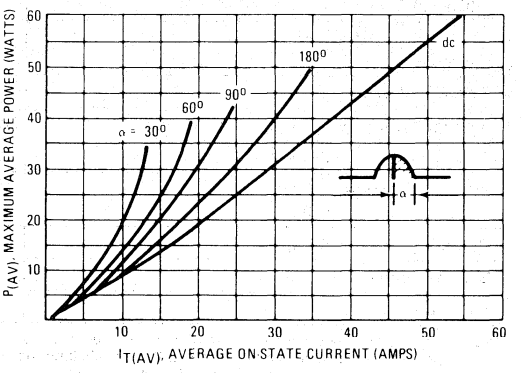


FIGURE 2 — POWER DISSIPATION



Silicon Controlled Rectifiers

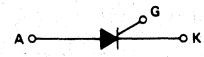
Reverse Blocking Triode Thyristors

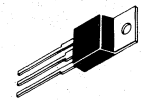
MCR68 Series
MCR69 Series

SCRs
12 and 25 AMPERES RMS
50 thru 400 VOLTS

. . . designed for overvoltage protection in crowbar circuits.

- Glass-Passivated Junctions for Greater Parameter Stability and Reliability
- Center-Gate Geometry for Uniform Current Spreading Enabling High Discharge Current
- Small Rugged, Thermowatt or Metal Packages Constructed for Low Thermal Resistance for Maximum Power Dissipation and Durability
- High Capacitor Discharge Current
  300 Amps (MCR68)
  750 Amps (MCR69)





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value MCR68	Value MCR69	Unit
Repetitive Peak Forward or Reverse Blocking Voltage, Note 1 (T_J = −40 to +125°C)	V_{DRM} or V_{RRM}			Volts
MCR68, 69 -2		50		
MCR68, 69 -3		100		
MCR68, 69 -6		400		
Peak Discharge Current, Note 2	I_{TM}	300	750	Amps
On-State Current (T_C = 85°C) (1/2 Cycle, Sine Wave, 60 Hz)	$I_{T(RMS)}$	12	25	Amps
	$I_{T(AV)}$	8	16	
Peak Non-Repetitive Surge Current (1/2 Cycle, Sine Wave, 60 Hz, T_J = 125°C)	I_{TSM}	100	300	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	375	A^2s
Critical Rate-of-Rise of Current (Note 3)	di/dt	75	100	A/μs
Peak Gate Current (t ≤ 2 μs)	I_{GM}	2		Amps
Peak Gate Power (t ≤ 2 μs)	P_{GM}	20		Watts
Average Gate Power	$P_{G(AV)}$	0.5		Watt
Operating Junction Temperature Range	T_J	−40 to +125		°C
Storage Temperature Range	T_{stg}	−40 to +150		°C
Mounting Torque	—	8		in. lb.

THERMAL CHARACTERISTICS

Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	1.5	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60		°C/W

Notes: 1. V_{DRM} for all types can be applied on a continuous basis over the operating junction temperature range without recurring damage. Ratings apply for open or shorted gate conditions or negative voltage on the gate. Devices should not be tested for blocking voltages such that the supply voltage exceeds the rating of the device.
2. Ratings apply for t_w = 1 ms. See Figure 1 for I_{TM} capability for various duration of an exponentially decaying current waveform, t_w is defined as 5 time constants of an exponentially decaying current pulse.
3. Test Conditions: I_G = 150 mA, V_D = Rated V_{DRM}, I_{TM} = Rated Value, T_J = 125°C.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Forward On-State Voltage	V_{TM}				Volts
(I_{TM} = 24 Amps), Note 1 MCR68		—	—	2.2	
(I_{TM} = 50 Amps), Note 1 MCR69		—	—	1.8	
(I_{TM} = 300 Amps, t_w = 1 ms), Note 2 MCR68		—	6	—	
(I_{TM} = 750 Amps, t_w = 1 ms), Note 2 MCR69		—	6	—	
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}	2	7	30	mA
Gate Trigger Voltage (Continuous dc)	V_{GT}				volts
(V_D = 12 Volts, R_L = 100 Ω)		—	0.65	1.5	
(V_D = Rated V_{DRM}, R_L = 1 kΩ, T_J = 125°C)		0.2	0.40	—	
Holding Current (I_{TM} = 100 mA, Gate-Open)	I_H	3	15	50	mA
Latching Current (V_D = 12 Vdc, I_G = 150 mA, t_r ≤ 50 μs)	I_L	—	—	60	mA
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Gate Open, Exponential Waveform, T_J = 125°C)	dv/dt	10	—	—	V/μs
Gate Controlled Turn-On Time, Note 3 (V_D = Rated V_{DRM}, I_G = 150 mA)	t_{gt}				μs
(I_{TM} = 24 Amps, peak) MCR68		—	1	—	
(I_{TM} = 50 Amps, peak) MCR69		—	1	—	

Notes: 1. Pulse duration ≤ 300 μs, duty cycle ≤ 2%.
2. Ratings apply for t_w = 1 ms. See Figure 1 for I_{TM} capability for various durations of an exponentially decyaing current waveform. t_w is defined as 5 time constants of an exponentially decaying current pulse.
3. The gate controlled turn-on time in a crowbar circuit will be influenced by the circuit inductance.

3

FIGURE 1 – PEAK CAPACITOR DISCHARGE CURRENT

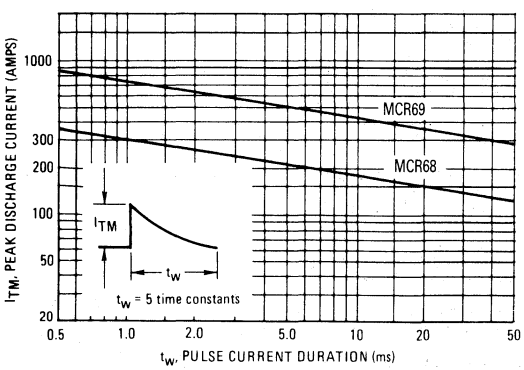


FIGURE 2 – PEAK CAPACITOR DISCHARGE CURRENT DERATING

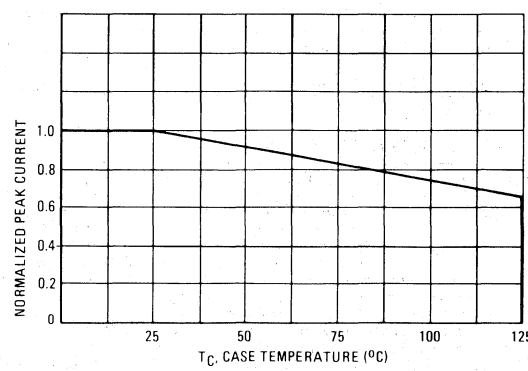


FIGURE 3 — CURRENT DERATING
MCR68

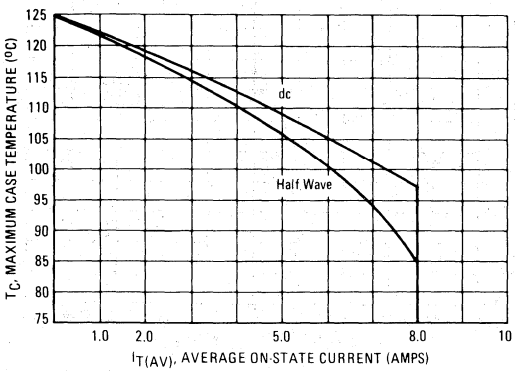


FIGURE 4 – CURRENT DERATING
MCR69

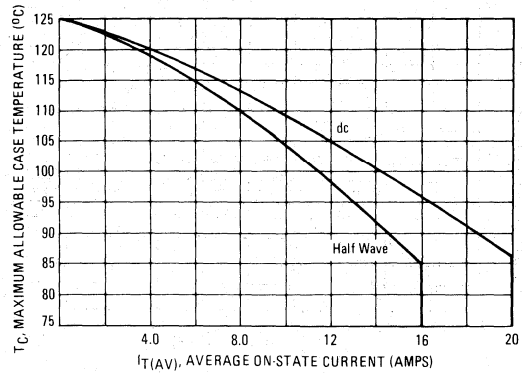


FIGURE 5 — MAXIMUM POWER DISSIPATION
MCR68

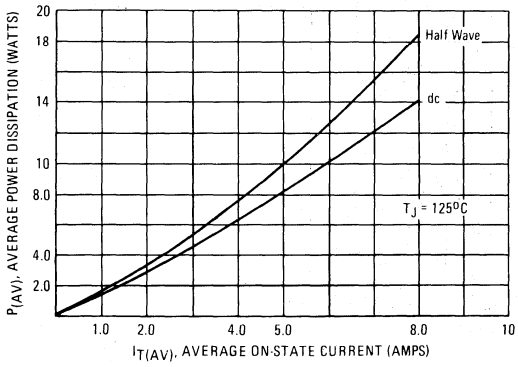


FIGURE 6 – MAXIMUM POWER DISSIPATION
MCR69

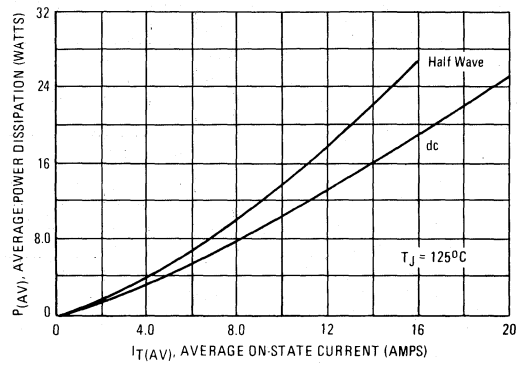


FIGURE 7 – THERMAL RESPONSE

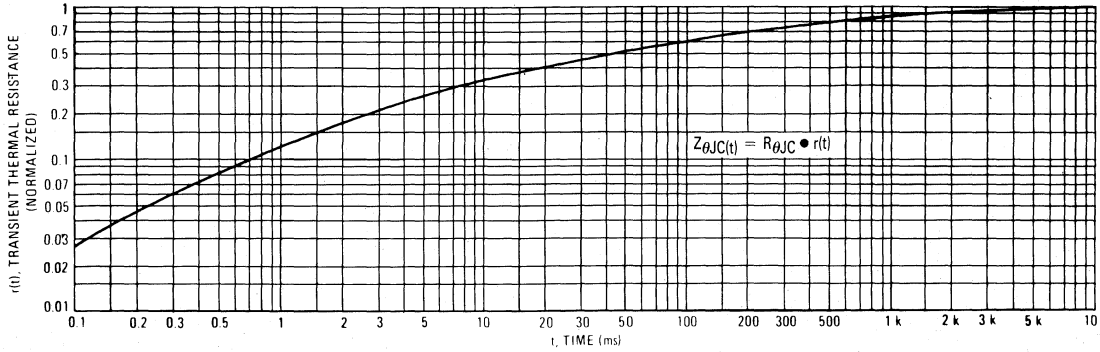


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FIGURE 8 – GATE TRIGGER CURRENT

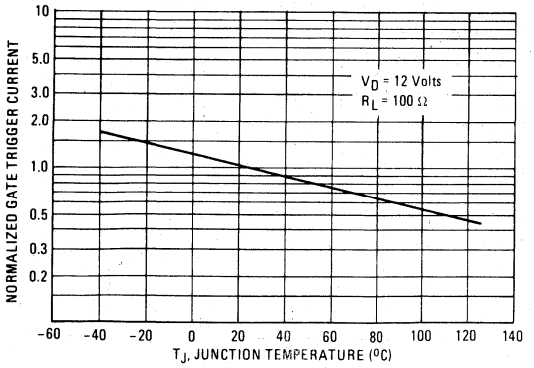


FIGURE 9 – GATE TRIGGER VOLTAGE

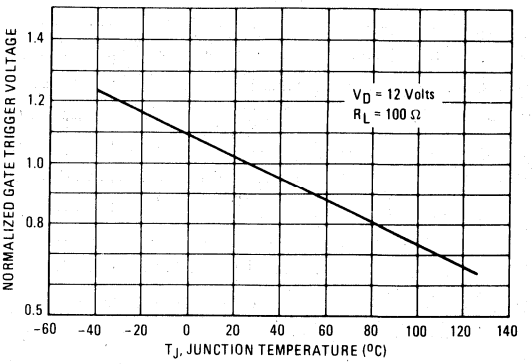
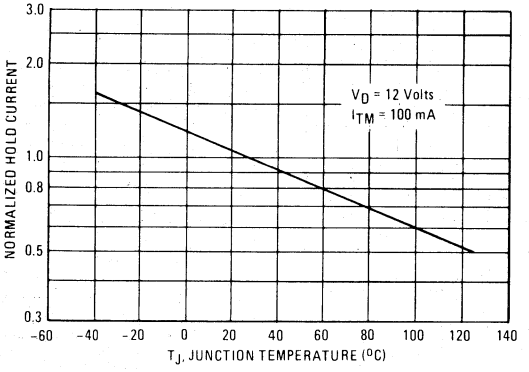


FIGURE 10 – HOLDING CURRENT



3

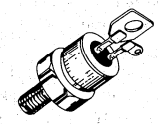
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

MCR70-( )A Series
MCR71 Series

SCRs
35 and 55 AMPERES RMS
50 thru 400 VOLTS

. . . designed for overvoltage protection in crowbar circuits.

- Glass-Passivated Junctions for Greater Parameter Uniformity and Stability
- Center-Gate Geometry for Uniform Current Spreading Enabling High Peak Current Capability
- High Capacitor Discharge Current Capability
  850 Amps (MCR70-( )A Series)
  1700 Amps (MCR71 Series)
- Hermetically-Sealed Metal Packages



CASE 263-04
STYLE 1

3

MAXIMUM RATINGS

Rating	Symbol	Value MCR70-( )A	Value MCR71	Unit
Repetitive Peak Forward or Reverse Blocking Voltage, Note 1 MCR70-( )A 2 / MCR71 3 / 6	V_{DRM} or V_{RRM}	50 / 100 / 400		Volts
Peak Discharge Current, Note 2	I_{TM}	850	1700	Amps
On-State Current ($T_C \leq 75°C$)	$I_{T(RMS)}$ / $I_{T(AV)}$	35 / 22	55 / 35	Amps
Peak Non-Repetitive Surge Current (1/2 Cycle, Sine Wave, 60 Hz, T_J = 125°C)	I_{TSM}	350	550	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	510	1255	A^2s
Critical Rate-of-Rise of Current, Note 3	di/dt	100	200	A/μS
Forward Peak Gate Power (t ≤ 20 μs)	P_{GM}	20		Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5		Watts
Forward Peak Gate Current (t ≤ 20 μs)	I_{GM}	2		Amps
Operating Junction Temperature Range	T_J	−40 to +125		°C
Storage Temperature Range	T_{stg}	−40 to +150		°C
Mounting Torque	—	30		in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1	°C/W

Notes: 1. The rated voltage can be applied over the rated operating junction temperatures without incurring damage. Ratings apply for shorted-open or shorted-gate conditions or negative voltage on the gate. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltages.
2. Rating is for t_w = 1 ms. See Figure 1 for I_{TM} limits of an exponentially decaying current pulse of various durations.
3. Test Conditions: I_G = 150 mA, V_D = Rated V_{DRM}, I_{TM} = Rated Value, T_J = 125°C.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
On-State Voltage, Note 1	V_{TM}				Volts
(I_{TM} = 70 A) MCR70 series		—	1.5	1.85	
(I_{TM} = 175 A) MCR71 series		—	1.7	2.1	
(I_{TM} = 850 A, t_w = 1 ms) Note 2 MCR70 series		—	6	—	
(I_{TM} = 1700 A, t_w = 1 ms) Note 2 MCR71 series		—	7	—	
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ω)	I_{GT}	2	10	30	mA
Gate Trigger Voltage (Continuous dc)	V_{GT}				volts
(V_D = 12 Volts, R_L = 100 Ω)		—	1	1.5	
(V_D = Rated V_{DRM}, R_L = 1 kΩ, T_J = 125°C)		0.2	—	—	
Holding Current (I_{TM} = 0.5 A, Gate-Open)	I_H	3	15	50	mA
Latching Current (V_D = 12 Vdc, I_G = 150 mA, t_r ≤ 50 μs)	I_L	—	30	60	mA
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Gate Open, Exponential Waveform, T_C = 125°C)	dv/dt	10	—	—	V/μs
Turn-On Time, Note 3 (V_D = Rated V_{DRM}, I_G = 150 mA)	t_{on}				μs
(I_{TM} = 70 Amps, peak) MCR70 series		—	1	—	
(I_{TM} = 110 Amps, peak) MCR71 series		—	1.2	—	

Notes: 1. Duty Cycle ≤ 1%, Pulse Width ≤ 300 μs.
2. Characteristic applies for t_w = 1 ms. t_w is defined as 5 time constants of an exponentially decaying current pulse.
3. The gate controlled turn-on time in a crowbar circuit will be influenced by the circuit inductance.

3

FIGURE 1 – PEAK CAPACITOR DISCHARGE CURRENT

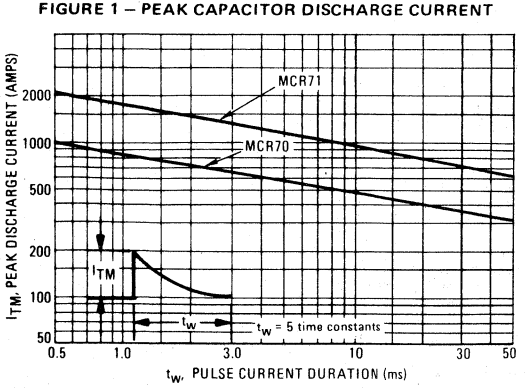


FIGURE 2 – PEAK CAPACITOR DISCHARGE CURRENT DERATING

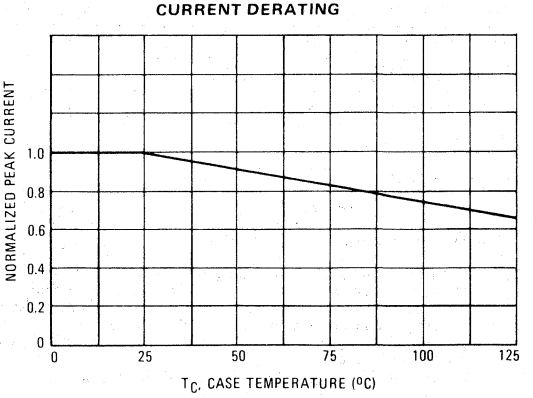


FIGURE 3 – AVERAGE CURRENT DERATING
MCR70

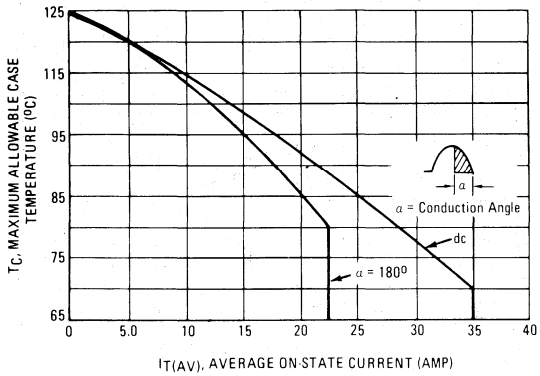


FIGURE 4 – POWER DISSIPATION
MCR70

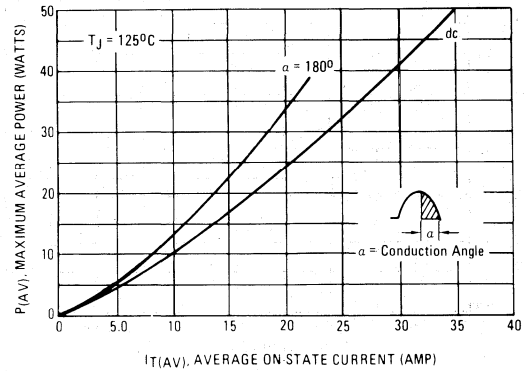


FIGURE 5 – CURRENT DERATING
MCR71

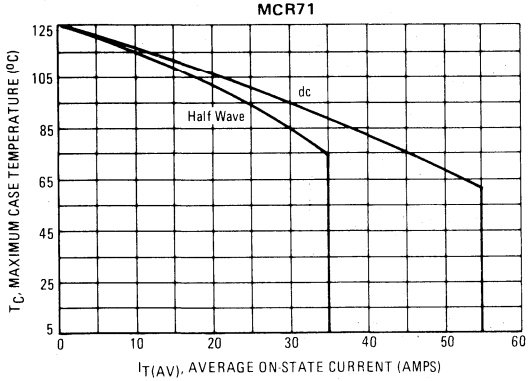


FIGURE 6 – POWER DISSIPATION
MCR71

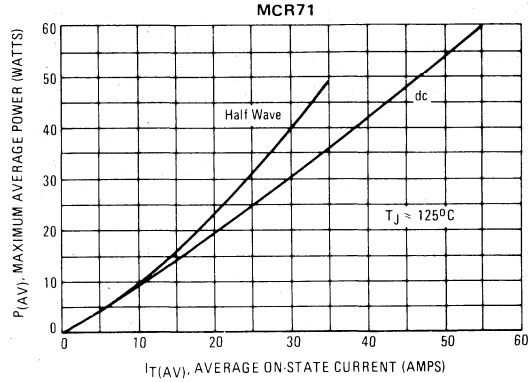


FIGURE 7 – TYPICAL THERMAL RESPONSE

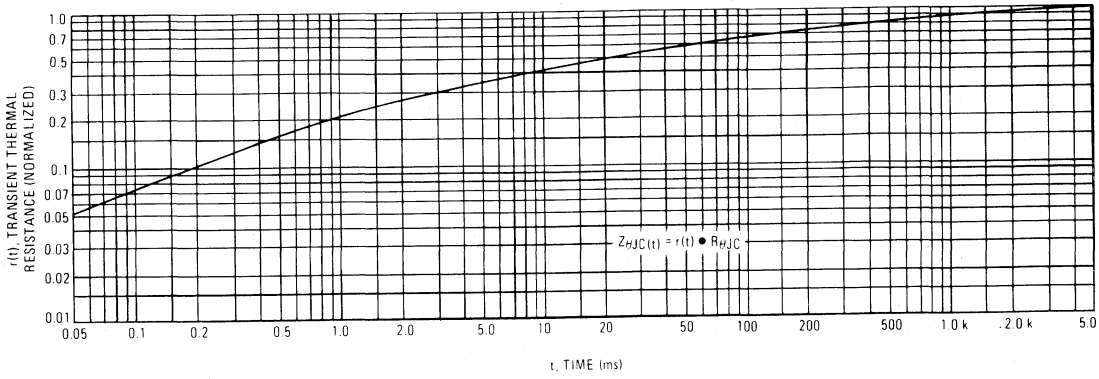


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FIGURE 8 – GATE TRIGGER CURRENT

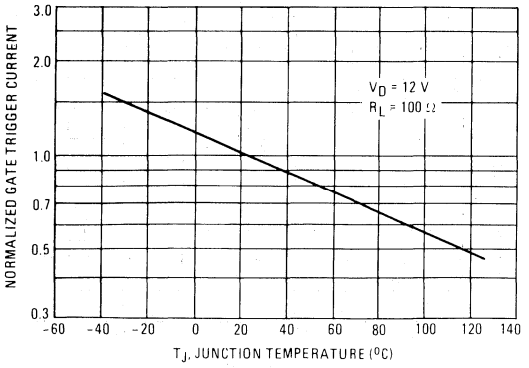


FIGURE 9 – GATE TRIGGER VOLTAGE

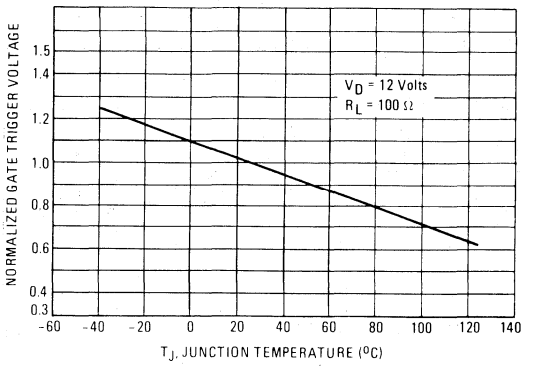
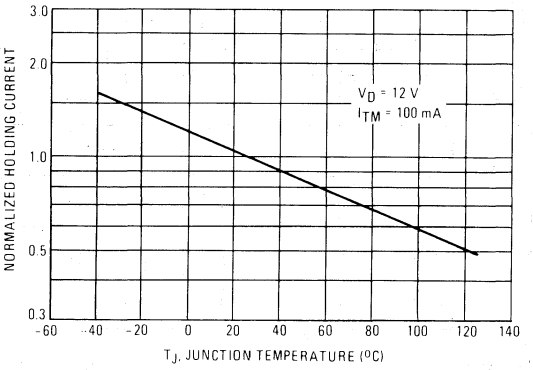


FIGURE 10 – HOLDING CURRENT



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Silicon Controlled Rectifiers

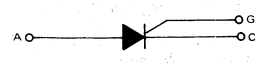
Reverse Blocking Triode Thyristors

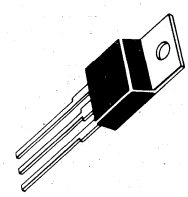
MCR72 Series

SCRs
8 AMPERES RMS
50 thru 800 VOLTS

. . . designed for industrial and consumer applications such as temperature, light and speed control; process and remote controls; warning systems; capacitive discharge circuits and MPU interface.

- Center Gate Geometry for Uniform Current Density
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Low Trigger Currents, 200 μA Maximum for Direct Driving from Integrated Circuits





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage, Note 1 (T_J = −40 to 110°C) (1/2 Sine Wave, R_{GK} = 1 kΩ)	V_{DRM} or V_{RRM}		Volts
MCR72 -2		50	
MCR72 -3		100	
MCR72 -4		200	
MCR72 -6		400	
MCR72 -8		600	
MCR72 -10		800	
On-State RMS Current (T_C = 83°C)	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (1/2 Cycle, 60 Hz, T_J = −40 to 110°C)	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Voltage (t ≤ 10 μs)	V_{GM}	±5	Volts
Peak Gate Current (t ≤ 10 μs)	I_{GM}	1	Amp
Peak Gate Power (t ≤ 10 μs)	P_{GM}	5	Watts
Average Gate Power	$P_{G(AV)}$	0.75	Watts
Operating Junction Temperature Range	T_J	−40 to +110	°C

Note 1. Ratings apply for negative gate voltage or R_{GK} = 1 kΩ. Devices shall not have a positive gate voltage concurrently with a negative voltage on the anode. Devices should not be tested with a constant current source for forward and reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

(cont.)

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque	—	8	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, R_{GK} = 1 kΩ unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current, Note 1 (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 110°C	I_{DRM}, I_{RRM}	 — —	 — —	 10 500	 μA μA
On-State Voltage (I_{TM} = 16 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	1.7	2	Volts
Gate Trigger Current (Continuous dc), Note 2 (V_D = 12 V, R_L = 100 Ω)	I_{GT}	—	30	200	μA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ω) (V_D = Rated V_{DRM}, R_L = 10 kΩ, T_J = 110°C)	V_{GT}	 — 0.1	 0.5 —	 1.5 —	Volts
Holding Current (V_D = 12 V, I_{TM} = 100 mA)	I_H	—	—	6	mA
Critical Rate of Rise of Forward Blocking Voltage (V_D = Rated V_{DRM}, T_J = 110°C, Exponential Waveform)	dv/dt	—	10	—	V/μs
Gate Controlled Turn On Time (V_D = Rated V_{DRM}, I_{TM} = 16 A, I_G = 2 mA)	t_{gt}	—	1	—	μs

3

Notes: 1. Ratings apply for negative gate voltage or R_{GK} = 1 kΩ. Devices shall not have a positive gate voltage concurrently with a negative voltage on the anode. Devices should not be tested with a constant current source for forward and reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. Does not include R_{GK} current.

FIGURE 1 – AVERAGE CURRENT DERATING

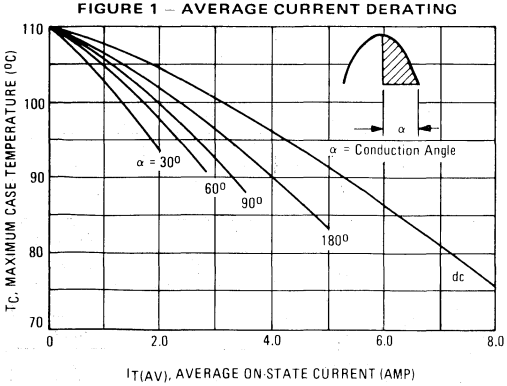


FIGURE 2 – ON-STATE POWER DISSIPATION

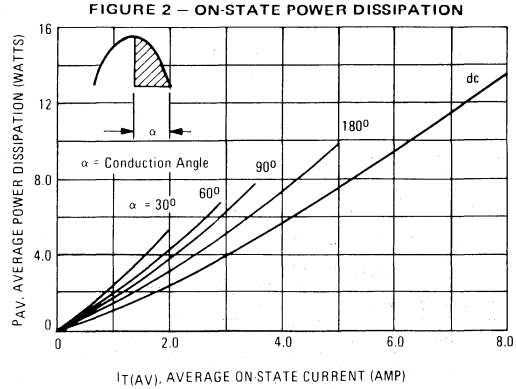


FIGURE 3 – NORMALIZED GATE CURRENT

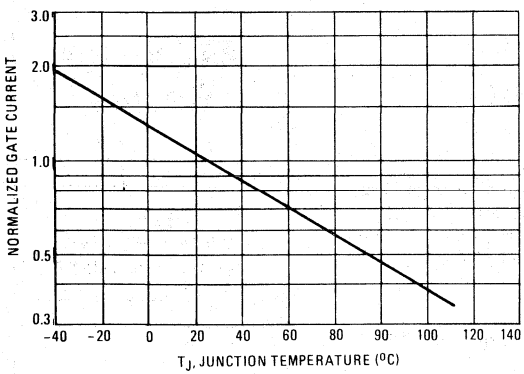
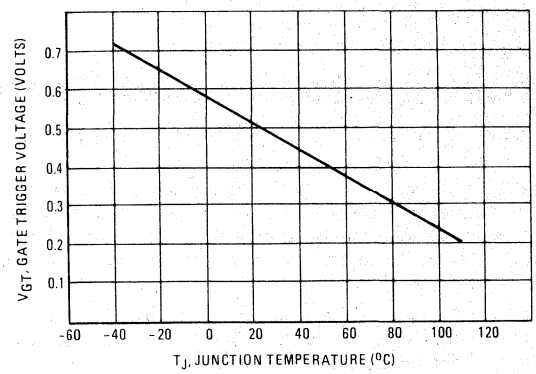


FIGURE 4 – GATE VOLTAGE



3

Silicon Controlled Rectifiers

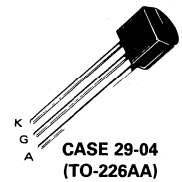
Reverse Blocking Triode Thyristors

MCR100 Series

SCRs
0.8 AMPERE RMS
100 thru 600 VOLTS

PNPN devices designed for high volume, line-powered consumer applications such as relay and lamp drivers, small motor controls, gate drivers for larger thyristors, and sensing and detection circuits. Supplied in an inexpensive plastic TO-226AA package which is readily adaptable for use in automatic insertion equipment.

- Sensitive Gate Trigger Current — 200 μA Maximum
- Low Reverse and Forward Blocking Current — 100 μA Maximum, T_C = 125°C
- Low Holding Current — 5 mA Maximum
- Glass-Passivated Surface for Reliability and Uniformity
- Also Available with TO-5 or TO-226AA (TO-18) Leadforms



CASE 29-04
(TO-226AA)
STYLE 10

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage MCR100-3 MCR100-4 MCR100-6 MCR100-8	V_{RRM}	 100 200 400 600	Volts
Forward Current RMS (See Figures 1 & 2) (All Conduction Angles)	$I_{T(RMS)}$	0.8	Amps
Peak Forward Surge Current, T_A = 25°C (1/2 cycle, Sine Wave, 60 Hz)	I_{TSM}	10	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	0.415	A^2s
Peak Gate Power — Forward, T_A = 25°C	P_{GM}	0.1	Watts
Average Gate Power — Forward, T_A = 25°C	$P_{GF(AV)}$	0.01	Watt
Peak Gate Current — Forward, T_A = 25°C (300 μs, 120 PPS)	I_{GFM}	1	Amp
Peak Gate Voltage — Reverse	V_{GRM}	5	Volts
Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−65 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Lead Solder Temperature (< 1/16″ from case, 10 s max)	—	+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS (R_{GK} = 1000 Ohms)

Characteristic		Symbol	Min	Max	Unit
Peak Forward Blocking Voltage (T_C = 125°C)		V_{DRM}			Volts
	MCR100-3		100	—	
	MCR100-4		200	—	
	MCR100-6		400	—	
	MCR100-8		600	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM})		I_{DRM}, I_{RRM}			
	T_C = 25°C		—	10	μA
	T_C = 125°C		—	100	μA
Forward "On" Voltage, Note 1 (I_{TM} = 1 A peak @ T_A = 25°C)		V_{TM}	—	1.7	Volts
Gate Trigger Current (Continuous dc), Note 2 (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	200	μA
Gate Trigger Voltage (Continuous dc)	T_C = 25°C	V_{GT}	—	0.8	Volts
(Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = −40°C		—	1.2	
(Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms)	T_C = 125°C		0.1	—	
Holding Current	T_C = 25°C	I_H	—	5	mA
(Anode Voltage = 7 Vdc, initiating current = 20 mA)	T_C = −40°C		—	10	

Notes: 1. Forward current applied for 1 ms maximum duration, duty cycle ≤ 1%.
2. R_{GK} current is not included in measurement.

3

FIGURE 1 – MCR100-7, MCR100-8 CURRENT DERATING (REFERENCE: CASE TEMPERATURE)

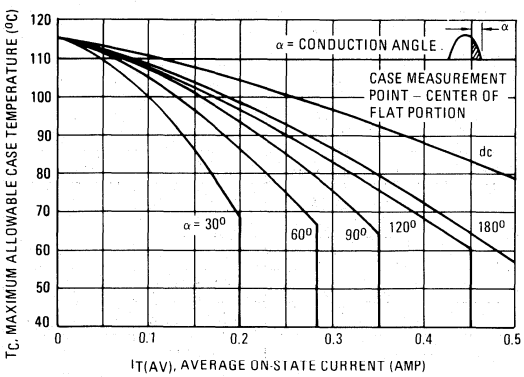
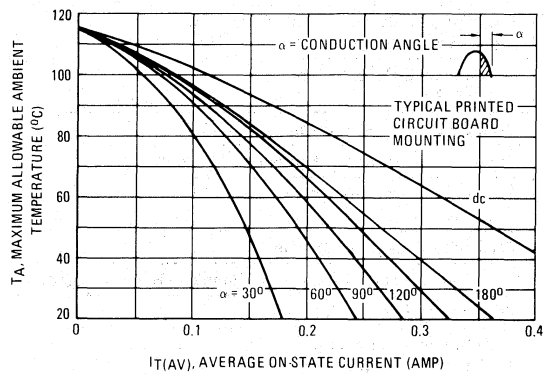


FIGURE 2 – MCR100-7, MCR100-8 CURRENT DERATING (REFERENCE: AMBIENT TEMPERATURE)



Silicon Controlled Rectifiers

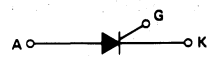
Reverse Blocking Triode Thyristors

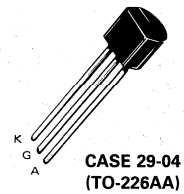
MCR102
MCR103

SCRs
0.8 AMPERES RMS
30 and 60 VOLTS

Annular PNPN devices designed for low cost, high volume consumer applications such as relay and lamp drivers, small motor controls, gate drivers for larger thyristors, and sensing and detection circuits. Supplied in an inexpensive plastic TO-226AA package which is readily adaptable for use in automatic insertion equipment.

- Sensitive Gate Trigger Current — 200 μA Maximum
- Low Reverse and Forward Blocking Current — 100 μA Maximum, T_C = 85°C
- Low Holding Current — 5 mA Maximum
- Passivated Surface for Reliability and Uniformity
- Also Available with TO-5 or TO-206AA (TO-18) Leadforms





CASE 29-04
(TO-226AA)
STYLE 10

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage, Note 2 (R_{GK} = 1000 ohms, T_C = +85°C) MCR102 / MCR103	V_{RRM}	30 / 60	Volts
Forward Current RMS (See Figures 1 & 2) (All Conduction Angles)	$I_{T(RMS)}$	0.8	Amps
Peak Forward Surge Current, T_A = 25°C (1/2 cycle, Sine Wave, 60 Hz)	I_{TSM}	10	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	0.415	A^2s
Peak Gate Power — Forward, T_A = 25°C	P_{GM}	0.1	Watt
Average Gate Power — Forward, T_A = 25°C	$P_{GF(AV)}$	0.01	Watt
Peak Gate Current — Forward, T_A = 25°C (300 μs, 120 PPS)	I_{GFM}	1	Amp
Peak Gate Voltage — Reverse	V_{GRM}	4	Volts
Operating Junction Temperature Range @ Rated V_{RRM} and V_{DRM}	T_J	−40 to +85	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Lead Solder Temperature (< 1/16″ from case, 10 s max)	—	+230	°C

Notes: 1. Temperature reference point for all case temperature is center of flat portion of package. (T_C = +85°C unless otherwise noted.)

2. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage but positive gate voltage shall not be applied concurrently with a negative potential on the anode. When checking forward or reverse blocking capability, thyristor devices should not be tested with a constant current source in a manner that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	75	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	200	°C/W

ELECTRICAL CHARACTERISTICS (R_{GK} = 1000 Ohms)

Characteristic		Symbol	Min	Max	Unit
Peak Forward Blocking Voltage (T_C = 85°C)	MCR102 MCR103	V_{DRM}	30 60	— —	Volts
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM})	T_C = 25°C T_C = 85°C	I_{DRM}, I_{RRM}	— —	10 100	μA μA
Forward "On" Voltage, Note 1 (I_{TM} = 1 A peak @ T_A = 25°C)		V_{TM}	—	1.7	Volts
Gate Trigger Current (Continuous dc), Note 2 (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C	I_{GT}	—	200	μA
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 7 Vdc, R_L = 100 Ohms)	T_C = 25°C T_C = −65°C T_C = 85°C	V_{GT} V_{GD}	— — 0.1	0.8 1.2 —	Volts
Holding Current (Anode Voltage = 7 Vdc, initiating current = 20 mA)	T_C = 25°C T_C = −65°C	I_H	— —	5 10	mA

Notes: 1. Forward current applied for 1 ms maximum duration, duty cycle ≤ 1%.
2. R_{GK} current is not included in measurement.

3

FIGURE 1 – CURRENT DERATING
(REFERENCE: CASE TEMPERATURE)

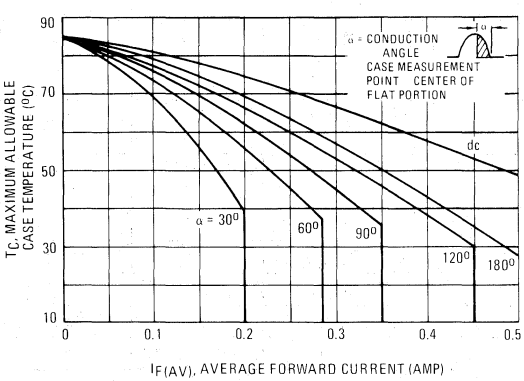
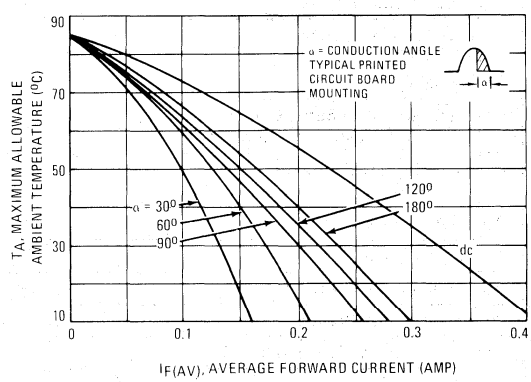


FIGURE 2 – CURRENT DERATING
(REFERENCE: AMBIENT TEMPERATURE)



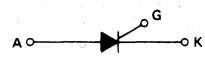
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

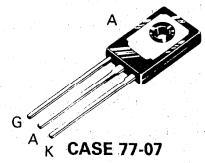
MCR106 Series

SCRs
4 AMPERES RMS
60 thru 600 VOLTS

PNPN devides designed for high volume consumer applications such as temperature, light, and speed control; process and remote control, and warning systems where reliability of operation is important.

- Glass-Passivated Surface for Reliability and Uniformity
- Power Rated at Economical Prices
- Practical Level Triggering and Holding Characteristics
- Flat, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability





CASE 77-07
(TO-225AA)
STYLE 2

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage, Note 1	V_{RRM}		Volts
MCR106-2		60	
-3		100	
-4		200	
-6		400	
-8		600	
RMS Foward Current (All Conduction Angles)	$I_{T(RMS)}$	4	Amps
Average Forward Current T_C = 93°C or T_A = 30°C	$I_{T(AV)}$	2.55	Amps
Peak Non-Repetitive Surge Current (1/2 cycle, 60 Hz, T_J = −40 to +110°C)	I_{TSM}	25	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	2.6	A^2s
Peak Gate Power	P_{GM}	0.5	Watt
Average Gate Power	$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current	I_{GM}	0.2	Amp
Peak Reverse Gate Voltage	V_{RGM}	6	Volts
Operating Junction Temperature Range	T_J	−40 to +110	°C

Note 1. Ratings apply for zero or negative gate voltage but positive gate voltage shall not be applied concurrently with a negative potential on the anode. When checking forward or reverse blocking capability, thyristor devices should not be tested with a constant current source in a manner that the voltage applied exceeds the rated blocking voltage. (cont.)

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque, Note 1	—	6	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	75	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and R_{GK} = 1000 ohms unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (T_J = 110°C, Note 1) MCR106-2	V_{DRM}	60	—	—	Volts
-3		100	—	—	
-4		200	—	—	
-6		400	—	—	
-8		600	—	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	200	μA
Forward "On" Voltage (I_{TM} = 4 A Peak)	V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc), Note 2 (V_{AK} = 7 Vdc, R_L = 100 ohms)	I_{GT}	—	—	200	μA
(V_{AK} = 7 Vdc, R_L = 100 ohms, T_C = −40°C)		—	—	500	
Gate Trigger Voltage (Continuous dc) (V_{AK} = 7 Vdc, R_L = 100 ohms, T_C = 25°C)	V_{GT}	—	—	1	Volts
Gate Non-Trigger Voltage (V_{AK} = Rated V_{DRM}, R_L = 100 ohms, T_J = 110°C)	V_{GD}	0.2	—	—	Volts
Holding Current (V_{AK} = 7 Vdc, T_C = 25°C)	I_H	—	—	5	mA
Forward Voltage Application Rate (T_J = 110°C)	dv/dt	—	10	—	V/μs

Notes: 1. Torque rating applies with use of compression washer (B52200-F006 or equivalent). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Anode lead and heatsink contact pad are common. (See AN209B).

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +200°C. For optimum results, an activated flux (oxide removing) is recommended.

2. R_{GK} current is not included in measurement.

3

CURRENT DERATING

FIGURE 1 – MAXIMUM CASE TEMPERATURE

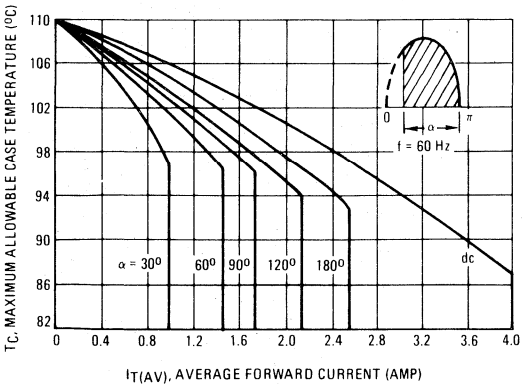
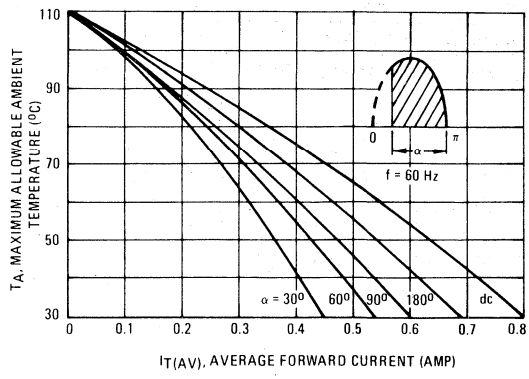


FIGURE 2 – MAXIMUM AMBIENT TEMPERATURE



3

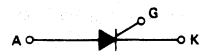
Thyristors
Silicon-Controlled Rectifiers

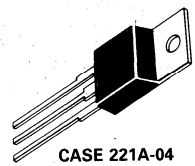
MCR218 Series

SCRs
8 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass-Passivated Junctions
- Blocking Voltage to 800 Volts
- TO-220 Construction — Low Thermal Resistance, High Heat Dissipation and Durability





CASE 221A-04
(TO-220AB)
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 Repetitive Peak Reverse Voltage MCR218-2 -3 -4 -6 -8 -10	V_{RRM} V_{DRM}	 50 100 200 400 600 800	Volts
Forward Current RMS (All Conduction Angles)	$I_{T(RMS)}$	8	Amps
Peak Forward Surge Current (1/2 Cycle, Sine Wave, 60 Hz)	I_{TSM}	80	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	26	A^2s
Forward Peak Gate Power	P_{GM}	5	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt
Forward Peak Gate Current	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
Peak On-State Voltage, Note 1 (I_{TM} = 16 A Peak)	V_{TM}	—	1.5	1.8	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 V, R_L = 100 Ohms)	I_{GT}	—	10	25	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 V, R_L = 100 Ohms) (Rated V_{DRM}, R_L = 1000 Ohms, T_J = 125°C)	V_{GT}	— 0.2	— —	1.5 —	Volts
Holding Current (Anode Voltage = 24 Vdc, Peak Initiating On-State Current = 0.5 A, 0.1 to 10 ms Pulse, Gate Trigger Source = 7 V, 20 Ohms)	I_H	—	16	30	mA
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Exponential Waveform, Gate Open, T_J = 125°C)	dv/dt	—	100	—	V/μs

Note 1. Pulse Test: Pulse Width = 1 ms, Duty Cycle ≤ 2%.

FIGURE 1 — CURRENT DERATING

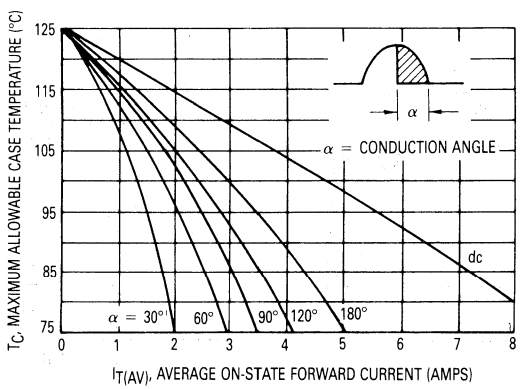


FIGURE 2 — ON-STATE POWER DISSIPATION

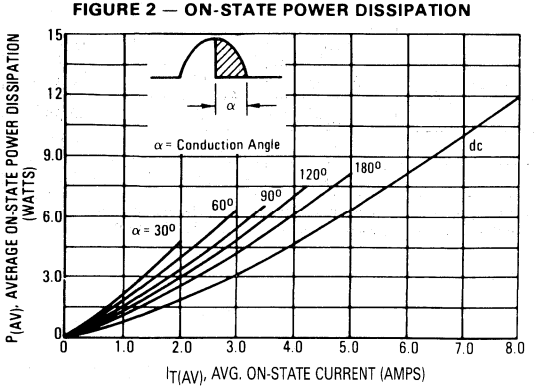


FIGURE 3 — NORMALIZED GATE TRIGGER CURRENT

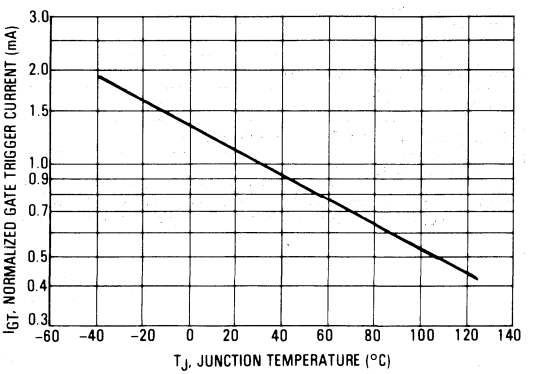


FIGURE 4 — NORMALIZED GATE TRIGGER VOLTAGE

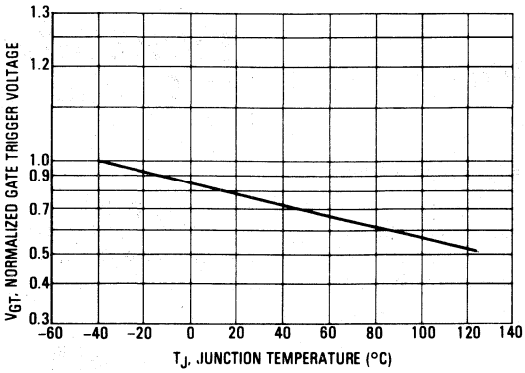
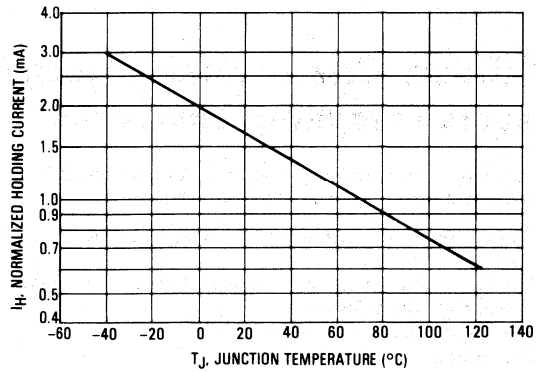


FIGURE 5 — NORMALIZED HOLDING CURRENT



3

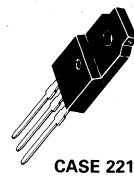
Silicon Controlled Rectifiers
Reverse Blocking Thyristors

MCR218FP Series

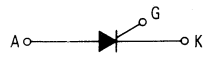
ISOLATED SCRs
8 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supply crowbar circuits.

- Glass Passivated Junctions with Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Constructed for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts
- 80 A Surge Current Capability
- Insulated Package Simplifies Mounting



CASE 221C-02
STYLE 2



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) Repetitive Peak Reverse Voltage	V_{DRM} V_{RRM}		Volts
MCR218-2FP		50	
MCR218-4FP		200	
MCR218-6FP		400	
MCR218-8FP		600	
MCR218-10FP		800	
On-State RMS Current (T_C = +70°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	8	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +70°C) Preceded and followed by rated current	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Power (T_C = +70°C, Pulse Width = 10 μs)	P_{GM}	5	Watts
Average Gate Power (T_C = +70°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current (T_C = +70°C, Pulse Width = 10 μs)	I_{GM}	2	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

Thermowatt is a trademark of Motorola Inc.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Current T_J = 25°C (Rated V_{DRM} @ T_J = 125°C)	I_{DRM}	— —	— —	10 2	μA mA
Peak Reverse Blocking Current (Rated V_{RRM} @ T_J = 125°C)	I_{RRM}	—	—	2	mA
Forward "On" Voltage, Note 1 (I_{TM} = 16 A Peak)	V_{TM}	—	1	1.8	Volts
Gate Trigger Current (Continuous dc) T_C = 25°C (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	I_{GT}	—	10	25	mA
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	V_{GT}	—	—	1.5	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)	V_{GD}	0.2	—	—	Volts
Holding Current (Anode Voltage = 12 Vdc)	I_H	—	16	30	mA
Turn-On Time (I_{TM} = 8 A, I_{GT} = 40 mAdc)	t_{gt}	—	1.5	—	μs
Turn-Off Time (V_{DRM} = Rated Voltage) (I_{TM} = 8 A, I_R = 8 A) (I_{TM} = 8 A, I_R = 8 A, T_J = 125°C)	t_q	 — —	 15 35	 — —	μs
Critical Rate of Rise of Off-State Voltage (Gate Open, Rated V_{DRM}, Exponential Waveform)	dv/dt	—	100	—	V/μs

Note 1. Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.

3

TYPICAL CHARACTERISTICS

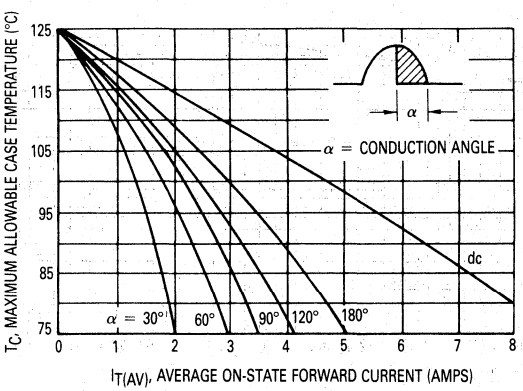


Figure 1. Current Derating

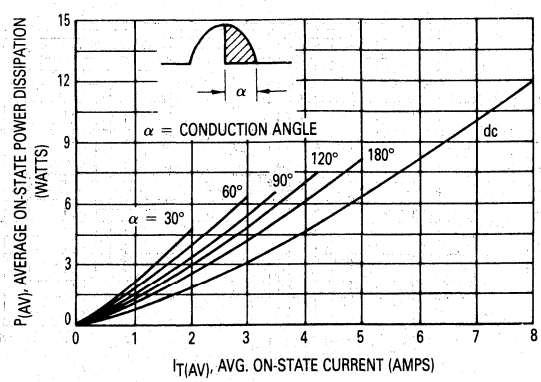


Figure 2. On-State Power Dissipation

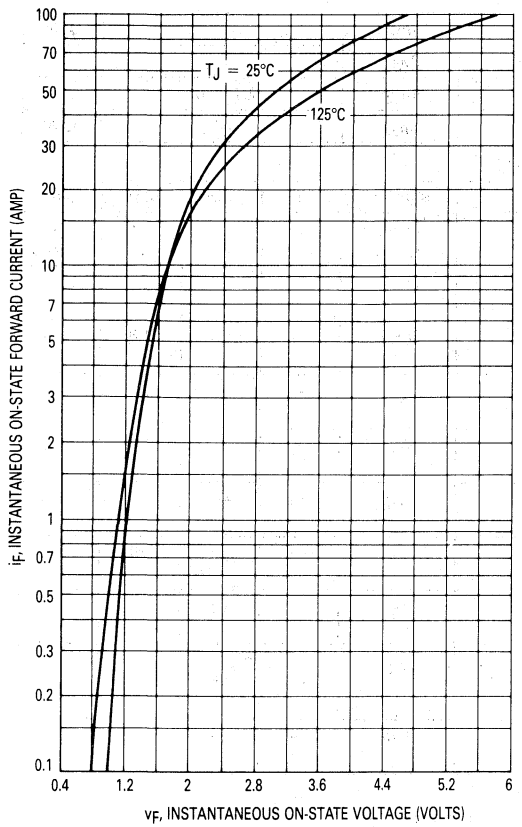


Figure 3. On-State Characteristics

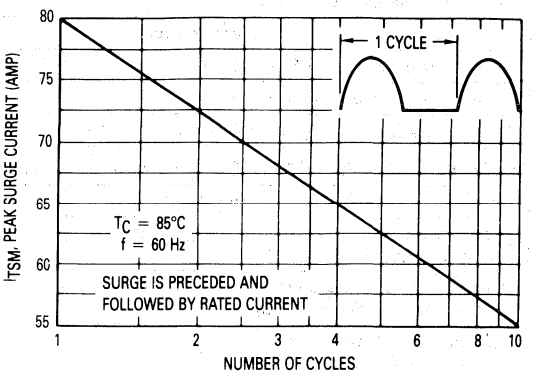


Figure 4. Maximum Non-Repetitive Surge Current

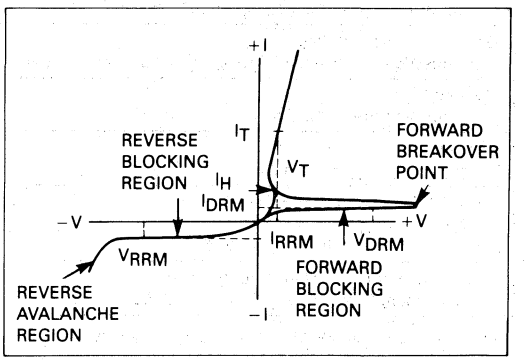


Figure 5. Characteristics and Symbols

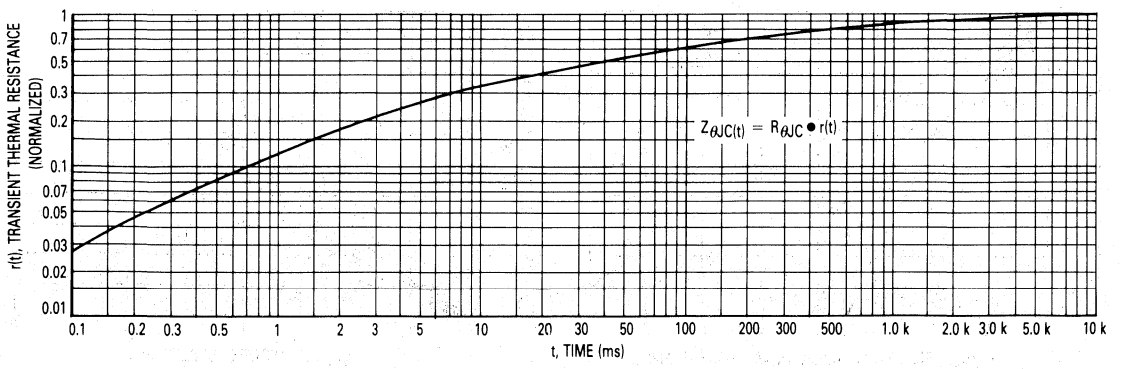


Figure 6. Thermal Response

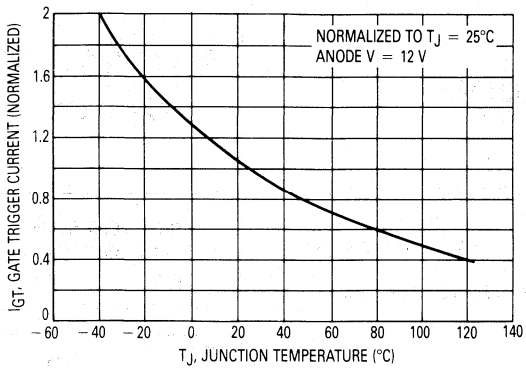


Figure 7. Gate Trigger Current versus Temperature

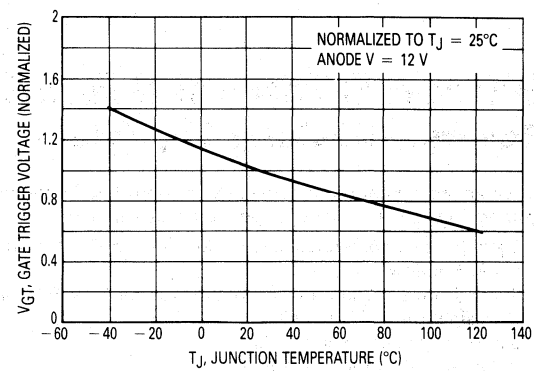


Figure 8. Gate Trigger Voltage versus Temperature

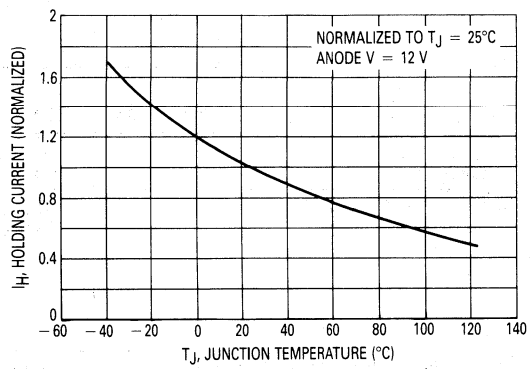


Figure 9. Holding Current versus Temperature

3

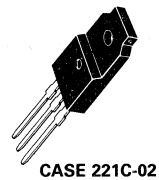
Silicon Controlled Rectifiers
Reverse Blocking Thyristors

MCR225FP Series

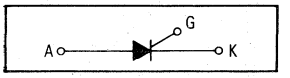
. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supply crowbar circuits.

- Glass Passivated Junctions with Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Constructed for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts
- 300 A Surge Current Capability
- Insulated Package Simplifies Mounting

ISOLATED SCRs
25 AMPERES RMS
50 thru 800 VOLTS



CASE 221C-02
STYLE 2



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +125°C) Repetitive Peak Reverse Voltage MCR225-2FP MCR225-4FP MCR225-6FP MCR225-8FP MCR225-10FP	V_{DRM} V_{RRM}	 50 200 400 600 800	Volts
On-State RMS Current (T_C = +70°C) Full Cycle Sine Wave 50 to 60 Hz, Note 2	$I_{T(RMS)}$	25	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +70°C) Preceded and followed by rated current	I_{TSM}	300	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	375	A^2s
Peak Gate Power (T_C = +70°C, Pulse Width = 10 μs)	P_{GM}	20	Watts
Average Gate Power (T_C = +70°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current (T_C = +70°C, Pulse Width = 10 μs)	I_{GM}	2	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	$V_{(ISO)}$	1500	Volts
Operating Junction Temperature	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

Thermowatt is a trademark of Motorola Inc.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.5	°C/W
Thermal Resistance, Case to Sink	$R_{\theta CS}$	2.2 (typ)	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Current T_J = 25°C (Rated V_{DRM} @ T_J = 125°C)	I_{DRM}	— —	— —	10 2	μA mA
Peak Reverse Blocking Current (Rated V_{RRM} @ T_J = 125°C)	I_{RRM}	—	—	2	mA
Forward "On" Voltage, Note 1 (I_{TM} = 50 A)	V_{TM}	—	—	1.8	Volts
Gate Trigger Current (Continuous dc) T_C = 25°C (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	I_{GT}	—	—	40	mA
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	V_{GT}	—	0.8	1.5	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)	V_{GD}	0.2	—	—	Volts
Holding Current (Anode Voltage = 12 Vdc)	I_H	—	20	40	mA
Turn-On Time (I_{TM} = 25 A, I_{GT} = 40 mAdc)	t_{gt}	—	1.5	—	μs
Turn-Off Time (V_{DRM} = Rated Voltage) (I_{TM} = 25 A, I_R = 25 A) (I_{TM} = 25 A, I_R = 25 A, T_J = 125°C)	t_q	— —	15 35	— —	μs
Critical Rate of Rise of Off-State Voltage (Gate Open, Rated V_{DRM}, Exponential Waveform)	dv/dt	—	100	—	V/μs

Note 1. Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.

3

TYPICAL CHARACTERISTICS

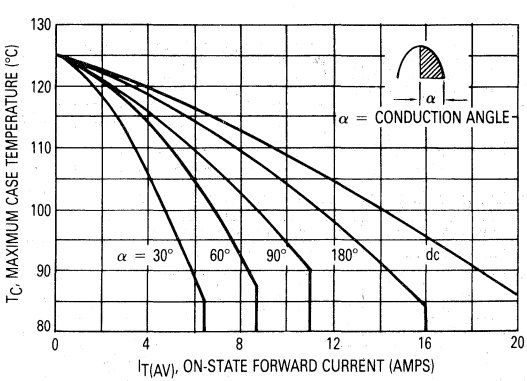


Figure 1. Average Current Derating

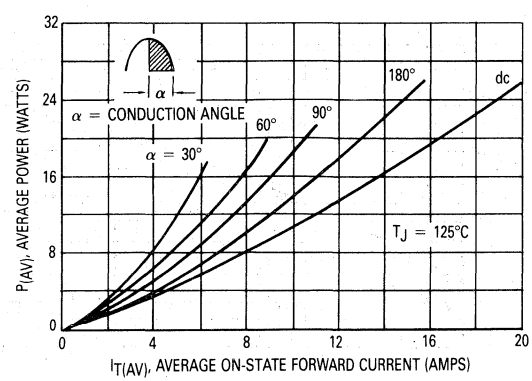


Figure 2. Maximum On-State Power Dissipation

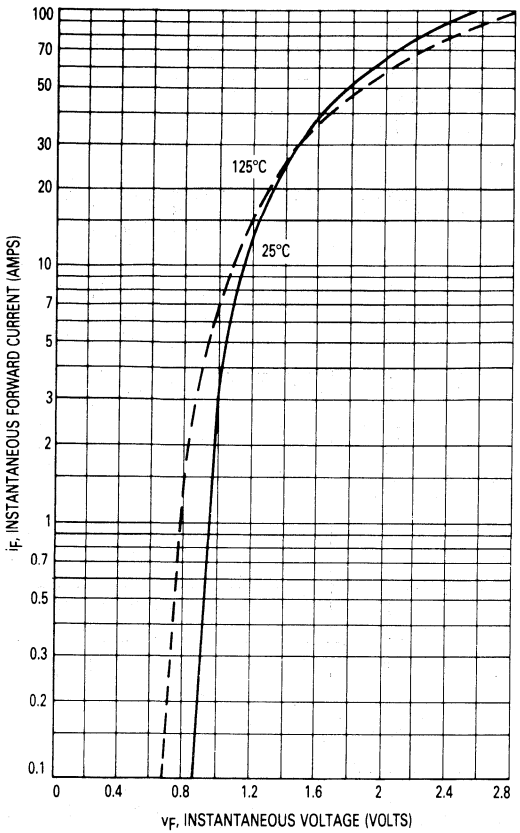


Figure 3. Maximum Forward Voltage

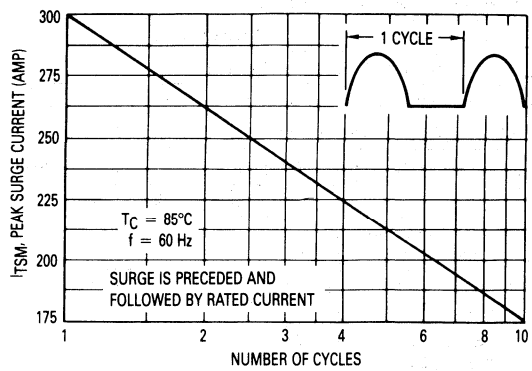


Figure 4. Maximum Non-Repetitive Surge Current

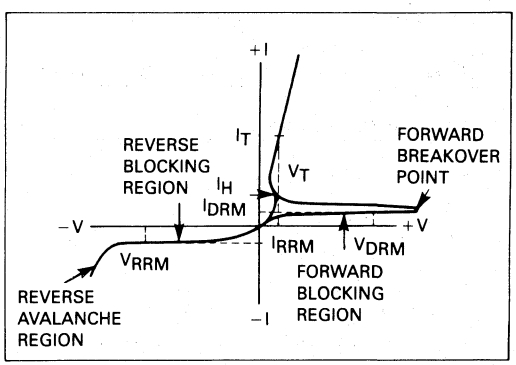


Figure 5. Characteristics and Symbols

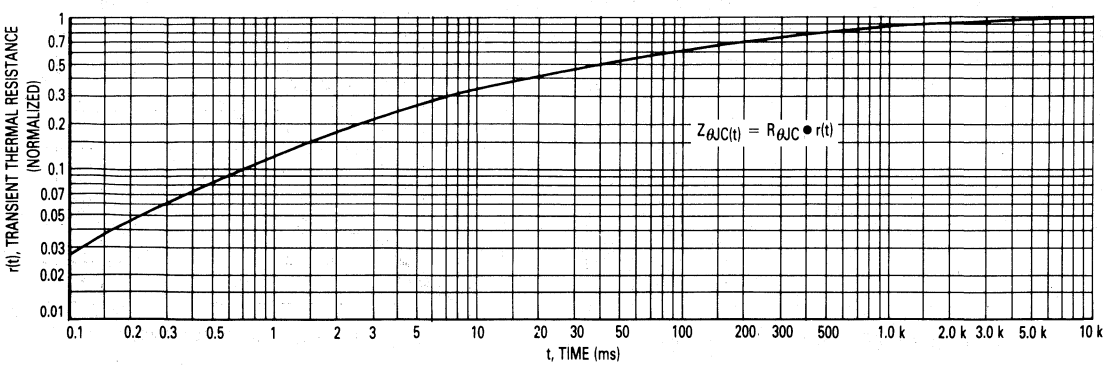


Figure 6. Thermal Response

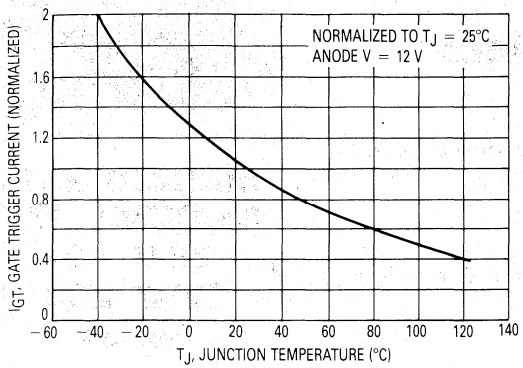


Figure 7. Gate Trigger Current versus Temperature

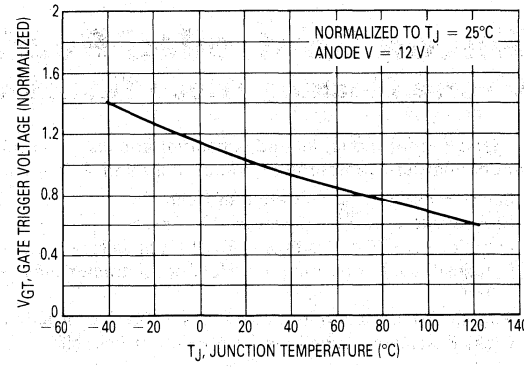


Figure 8. Gate Trigger Voltage versus Temperature

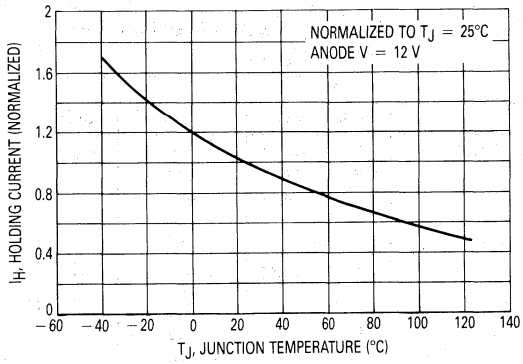


Figure 9. Holding Current versus Temperature

3

Silicon Controlled Rectifiers

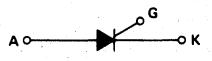
Reverse Blocking Triode Thyristors

MCR229 Series
(See C228)

MCR230, 231
MCR230( )3,
231( )3
C232, 233
Series

. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 300 Amps
- Low Forward "On" Voltage – 1.2 V (Typ) @ I_{TM} = 25 Amps
- Practical Level Triggering and Holding Characteristics — 10 mA (Typ) @ T_C = 25°C
- Rugged Construction in Either Pressfit, Stud, or Isolated Stud
- Glass Passivated Junctions for Maximum Reliability

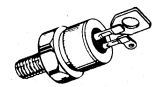


SCRs
25 AMPERES RMS
50 thru 600 VOLTS

MAXIMUM RATINGS

Rating	Suffix	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_C = –40 to +100°C) All Types	F A B D M	V_{DRM} and V_{RRM}	50 100 200 400 600	Volts
Non-Repetitive Reverse Voltage (T_C = –40 to 100°C) All Types	F A B D M	V_{RSM}	75 150 300 500 720	Volts
Forward Current RMS		$I_{T(RMS)}$	25	Amps
Peak Surge Current (One Cycle, 60 Hz, T_C = –40 to 100°C)		I_{TSM}	250	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	260	A^2s
Peak Gate Power		P_{GM}	5	Watts
Average Gate Power		$P_{G(AV)}$	0.5	Watt
Peak Forward Gate Current		I_{GM}	2	Amps
Operating Junction Temperature Range		T_J	–40 to +100	°C
Storage Temperature Range		T_{stg}	–40 to +125	°C
Stud Torque		—	30	in. lb.

CASE 174-04
(TO-203AA)
STYLE 1
C232 and C233 Series



CASE 263-04
STYLE 1
MCR230 and 231 Series

CASE 311-02
STYLE 1
MCR230( )3 and
MCR231( )3 Series

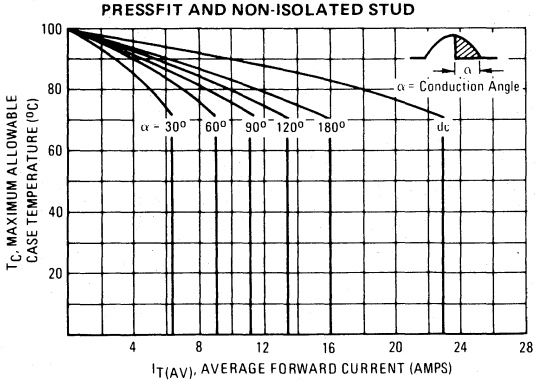
3

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	1	mA
Forward "On" Voltage (I_{TM} = 100 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.9	Volts
Gate Trigger Current, MCR230, MCR230( )3, C232 series (V_D = 12 Vdc, R_L = 120 Ohms)	I_{GT}	—	—	25	mA
(V_D = 12 Vdc, R_L = 60 Ohms) T_C = −40°C		—	—	40	
Gate Trigger Current, MCR231, MCR231( )3, C233 (Continuous dc) (V_D = 12 Vdc, R_L = 120 Ohms)	I_{GT}	—	—	9	mA
(V_D = 12 Vdc, R_L = 60 Ohms) T_C = −40°C		—	—	20	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 120 Ohms)	V_{GT}	—	—	1.5	Volts
(V_D = 12 Vdc, R_L = 60 Ohms) T_C = −40°C		—	—	2	
(V_D = Rated V_{DRM}, R_L = 1000 Ohms) T_C = +100°C		0.2	—	—	
Holding Current (V_D = 24 V, gate open, I_T = 0.5 A)	I_H	—	—	50	mA
T_C = −40°C		—	—	100	
Turn-On Time (t_d + t_r) (I_{TM} = 25 Adc, I_{GT} = 40 mAdc, V_D = Rated V_{DRM})	t_{gt}	—	1	—	μs
Turn-Off Time (I_{TM} = 10 A, I_R = 10 A, Pulse Width = 50 μs, dv/dt = 20 V/μs, V_D = Rated V_{DRM})	t_q	—	25	—	μs
T_C = 100°C		—	35	—	
Forward Voltage Application Rate (V_D = Rated V_{DRM}) T_C = 100°C	dv/dt	—	100	—	V/μs

3

FIGURE 1 – CURRENT DERATING FOR PRESSFIT AND NON-ISOLATED STUD



NOTE: Derating is for Pressfit and Stud Devices. Isolated stud devices must be derated an additional 15%. For example, the max T_C @ 16 A (180° conduction angle) is 70°C, a derating of 30°C. Isolated stud devices must be derated 34.5°C; therefore, the maximum T_C is 65.5°C.

FIGURE 2 – ON-STATE POWER DISSIPATION versus ON-STATE CURRENT

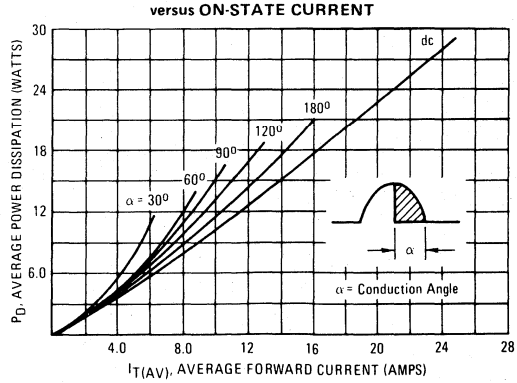


FIGURE 3 – GATE CURRENT VARIATION WITH TEMPERATURE

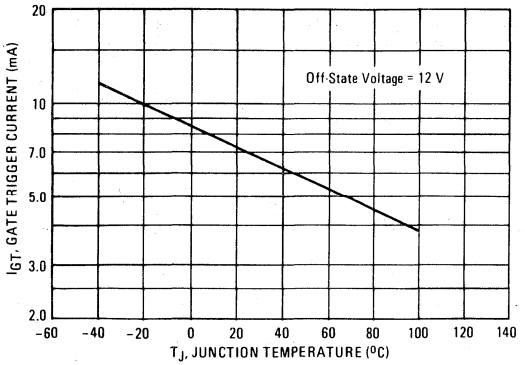
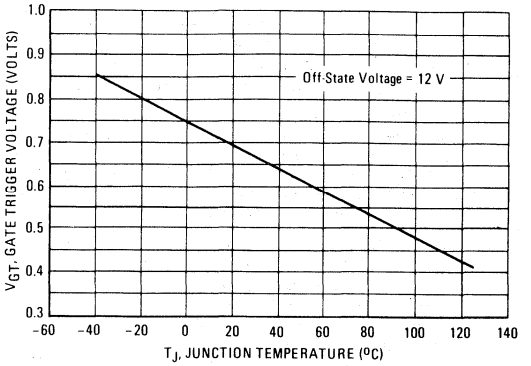


FIGURE 4 – GATE VOLTAGE VARIATION WITH TEMPERATURE



3

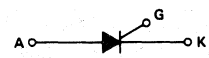
Thyristors
Silicon Controlled Rectifiers

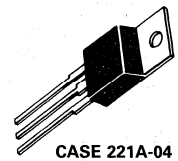
MCR264-2 thru MCR264-10

SCRs
40 AMPERES RMS
50 thru 800 VOLTS

. . . designed for back-to-back SCR output devices for solid state relays or applications requiring high surge operation.

- Photo Glass Passivated Blocking Junctions for High Temperature Stability, Center Gate for Uniform Parameters
- 400 Amperes Surge Capability
- Blocking Voltage to 800 Volts





CASE 221A-04
(TO-220AB)
STYLE 3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage, Note 1	V_{RRM}		Volts
MCR264-2		50	
MCR264-3		100	
MCR264-4		200	
MCR264-6		400	
MCR264-8		600	
MCR264-10		800	
Forward Current (T_C = 80°C) (All Conduction Angles)	$I_{T(RMS)}$ $I_{T(AV)}$	40* 25*	Amps
Peak Nonrepetitive Surge Current — 8.3 ms (1/2 Cycle, Sine Wave) 1.5 ms	I_{TSM}	400 450	Amps
Forward Peak Gate Power	P_{GM}	20	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt
Forward Peak Gate Current (300 μs, 120 PPS)	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

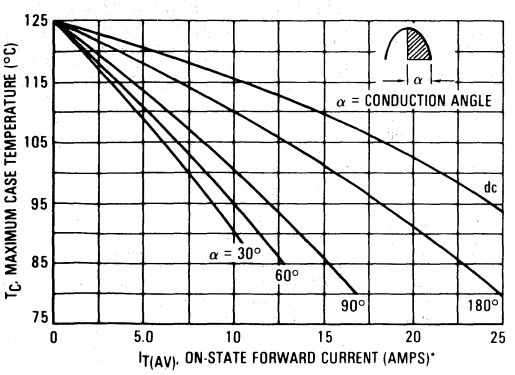
ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (T_J = 125°C)	V_{DRM}				Volts
MCR264-2		50	—	—	
MCR264-3		100	—	—	
MCR264-4		200	—	—	
MCR264-6		400	—	—	
MCR264-8		600	—	—	
MCR264-10		800	—	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Forward "On" Voltage, Note 1 (I_{TM} = 80 A)	V_{TM}	—	1.6	2	Volts
Gate Trigger Current (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms, T_C = −40°C)	I_{GT}	—	15	50	mA
		—	30	90	
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	V_{GT}	—	1	1.5	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)	V_{GD}	0.2	—	—	Volts
Holding Current (Anode Voltage = 12 Vdc)	I_H	—	30	60	mA
Turn-On Time (I_{TM} = 40 A, I_{GT} = 60 mAdc)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (Gate Open, Rated V_{DRM}, Exponential Waveform)	dv/dt	—	50	—	V/μs

Note 1. Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.

3

FIGURE 1 — AVERAGE CURRENT DERATING



*This device is rated for use in applications subject to high surge conditions. Care must be taken to insure proper heat sinking when the device is to be used at high sustained currents.

FIGURE 2 — MAXIMUM ON-STATE POWER DISSIPATION

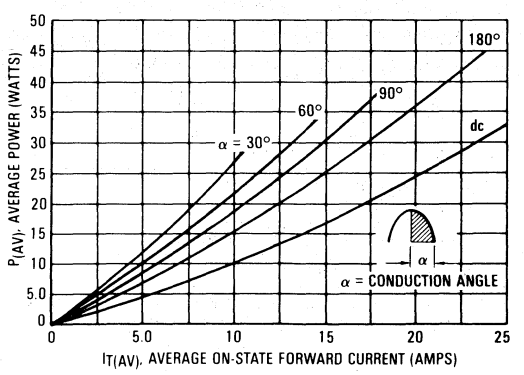


FIGURE 3 — GATE TRIGGER CURRENT

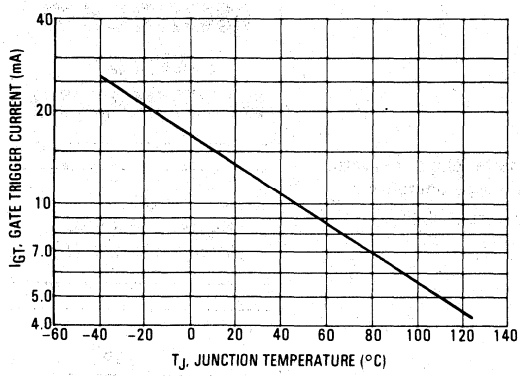


FIGURE 4 — NEW GATE TRIGGER VOLTAGE

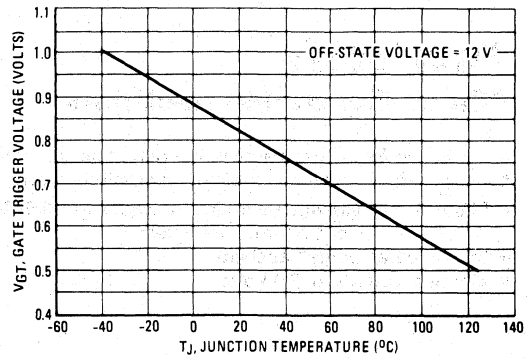


FIGURE 5 — HOLDING CURRENT

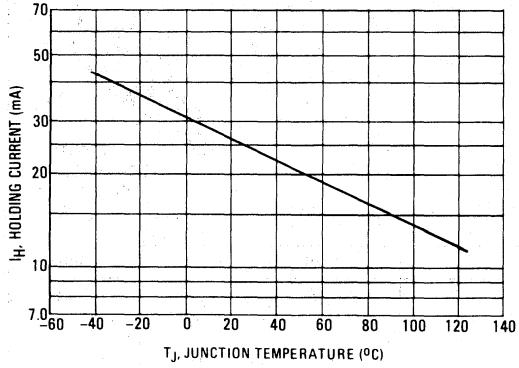


FIGURE 6 — TYPICAL FORWARD VOLTAGE

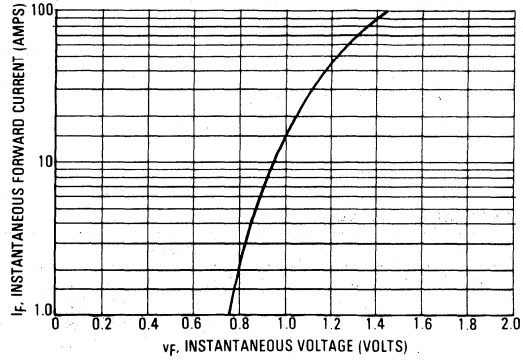
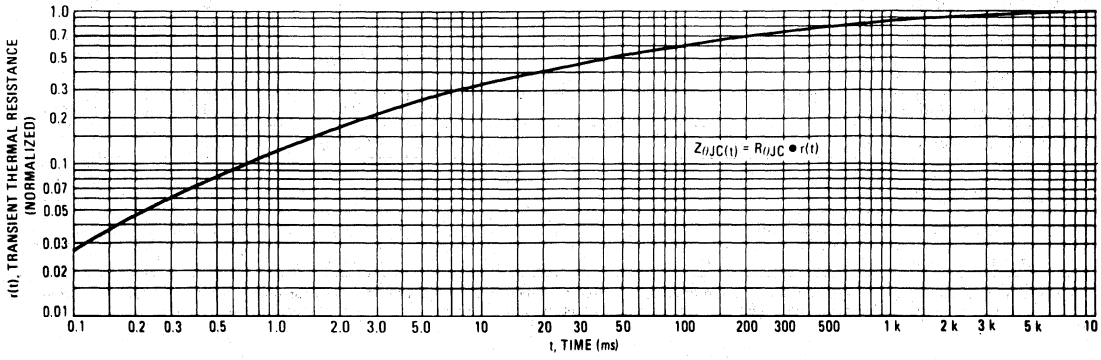


FIGURE 7 — THERMAL RESPONSE



3

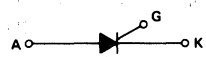
Thyristors
Silicon Controlled Rectifiers

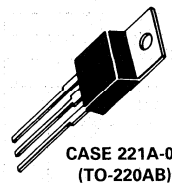
MCR265-2 thru MCR265-10

SCRs
55 AMPERES RMS
50 thru 800 VOLTS

. . . designed for inverse parallel SCR output devices for solid state relays, welders, battery chargers, motor controls or applications requiring high surge operation.

- Photo Glass Passivated Blocking Junctions for High Temperature Stability, Center Gate for Uniform Parameters
- 550 Amperes Surge Capability
- Blocking Voltage to 800 Volts





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage, Note 1	V_{RRM}		Volts
MCR265-2		50	
MCR265-3		100	
MCR265-4		200	
MCR265-6		400	
MCR265-8		600	
MCR265-10		800	
Forward Current (T_C = 70°C) (All Conduction Angles)	$I_{T(RMS)}$ $I_{T(AV)}$	55 35	Amps
Peak Nonrepetitive Surge Current — 8.3 ms (1/2 Cycle, Sine Wave)	I_{TSM}	550	Amps
Forward Peak Gate Power	P_{GM}	20	Watts
Forward Average Gate Power	$P_{G(AV)}$	0.5	Watt
Forward Peak Gate Current (300 μs, 120 PPS)	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	0.9	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (T_J = 125°C)	V_{DRM}				Volts
MCR265-2		50	—	—	
MCR265-3		100	—	—	
MCR265-4		200	—	—	
MCR265-6		400	—	—	
MCR265-8		600	—	—	
MCR265-10		800	—	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 125°C		—	—	2	mA
Forward "On" Voltage, Note 1 (I_{TM} = 110 A)	V_{TM}	—	1.5	1.9	Volts
Gate Trigger Current (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	I_{GT}	—	20	50	mA
(T_C = −40°C)		—	40	90	
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, R_L = 100 Ohms)	V_{GT}	—	1	1.5	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM}, R_L = 100 Ohms, T_J = 125°C)	V_{GD}	0.2	—	—	Volts
Holding Current (Anode Voltage = 12 Vdc)	I_H	—	30	75	mA
Turn-On Time (I_{TM} = 55 A, I_{GT} = 200 mAdc)	t_{gt}	—	1.5	—	μs
Critical Rate of Rise of Off-State Voltage (Gate Open, Rated V_{DRM}, Exponential Waveform)	dv/dt	—	50	—	V/μs

Note 1. Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.

3

FIGURE 1 — AVERAGE CURRENT DERATING

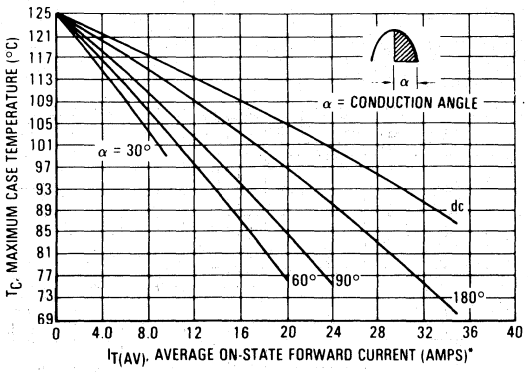
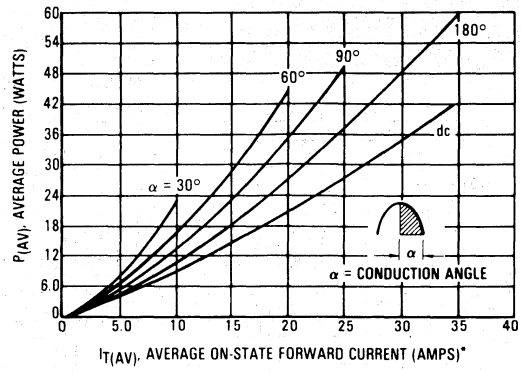


FIGURE 2 — MAXIMUM ON-STATE POWER DISSIPATION



* This device is rated for use in applications subject to high surge conditions. Care must be taken to insure proper heat sinking when the device is to be used at high sustained currents.

FIGURE 3 — GATE TRIGGER CURRENT

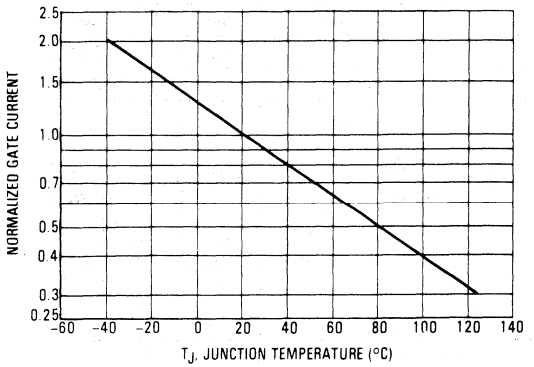


FIGURE 4 — GATE TRIGGER VOLTAGE

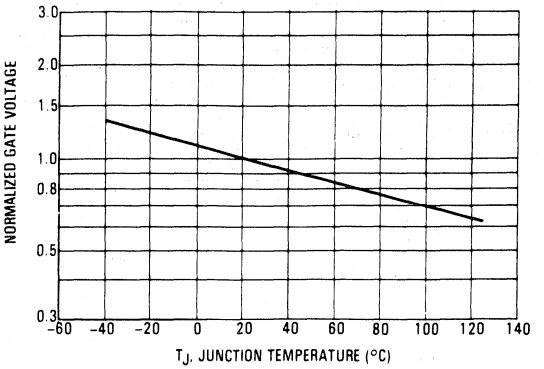


FIGURE 5 — HOLDING CURRENT

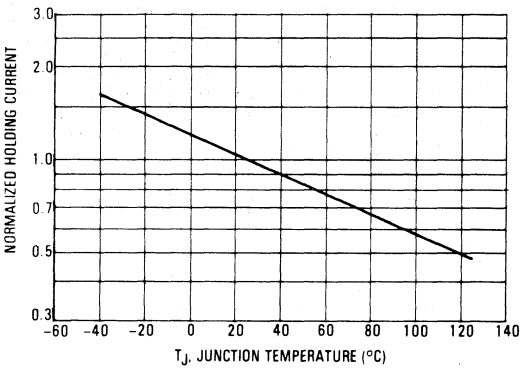


FIGURE 6 — ON-STATE CHARACTERISTICS

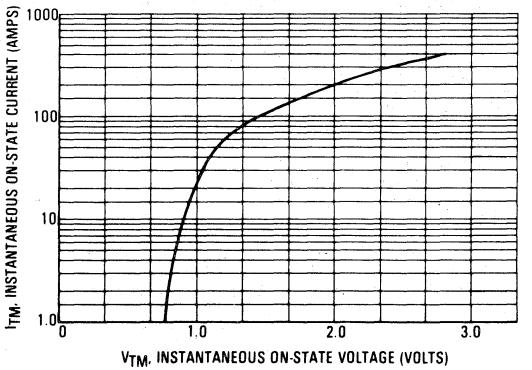
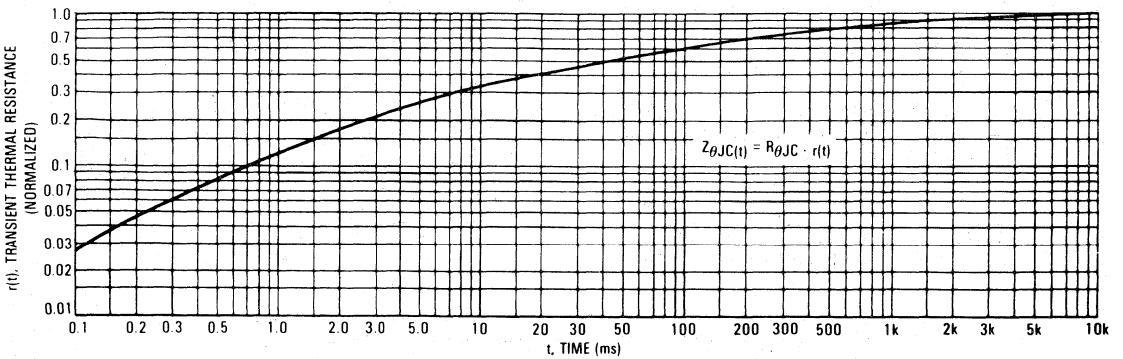


FIGURE 7 — THERMAL RESPONSE



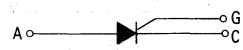
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

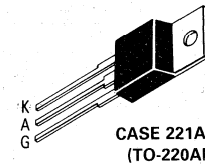
MCR310 Series

SCRs
10 AMPERES RMS
50 thru 800 VOLTS

. . . designed for industrial and consumer applications such as temperature, light and speed control; process and remote controls; warning systems; capacitive discharge circuits and MPU interface.

- Center Gate Geometry for Uniform Current Density
- All Diffused and Glass-Passivated Junctions for Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Low Trigger Currents, 200 μA Maximum for Direct Driving from Integrated Circuits





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage (Note 1) (T_J = −40 to 110°C) (1/2 Sine Wave, R_{GK} = 1 kΩ)	V_{DRM} or V_{RRM}		Volts
MCR310-2		50	
-3		100	
-4		200	
-6		400	
-8		600	
-10		800	
On-State RMS Current (T_C = 75°C)	$I_{T(RMS)}$	10	Amps
Peak Non-Repetitive Surge Current (1/2 Cycle, 60 Hz, T_J = −40 to 110°C)	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Voltage (t ≤ 10 μs)	V_{GM}	±5	Volts
Peak Gate Current (t ≤ 10 μs)	I_{GM}	1	Amp
Peak Gate Power (t ≤ 10 μs)	P_{GM}	5	Watts
Average Gate Power	$P_{G(AV)}$	0.75	Watt
Operating Junction Temperature Range	T_J	−40 to +110	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Mounting Torque	—	8	in-lb

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

NOTE 1: Ratings apply for negative gate voltage or R_{GK} = 1 kΩ. Devices shall not have a positive gate voltage concurrently with a negative voltage on the anode. Devices should not be tested with a constant current source for forward and reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C, R_{GK} = 1 kΩ unless otherwise noted)

Characteristic		Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Current (Note 1) (T_J = 110°C, V_D = Rated V_{DRM})	T_C = 110°C T_C = 25°C	I_{DRM}	— —	— —	500 10	μA μA
Peak Reverse Blocking Current (Note 1) (T_J = 110°C, V_R = Rated V_{RRM})	T_C = 110°C T_C = 25°C	I_{RRM}	— —	— —	500 10	μA μA
On-State Voltage (I_{TM} = 20 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)		V_{TM}	—	1.7	2.2	Volts
Gate Trigger Current, Continuous dc (Note 2) (V_D = 12 V, R_L = 100 Ω)		I_{GT}	—	30	200	μA
Gate Trigger Voltage, Continuous dc (V_D = 12 V, R_L = 100 Ω) (V_D = Rated V_{DRM}, R_L = 10 kΩ, T_J = 110°C)		V_{GT}	— 0.1	0.5 —	1.5 —	Volts
Holding Current (V_D = 12 V, I_{TM} = 100 mA)		I_H	—	—	6	mA
Critical Rate of Rise of Forward Blocking Voltage (V_D = Rated V_{DRM}, T_J = 110°C, Exponential Waveform)		dv/dt	—	10	—	V/μs
Gate Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 20 A, I_G = 2 mA)		t_{gt}	—	1	—	μs

NOTES:
1. Ratings apply for negative gate voltage or R_{GK} = 1 kΩ. Devices shall not have a positive gate voltage concurrently with a negative voltage on the anode. Devices should not be tested with a constant current source for forward and reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. Does not include R_{GK} current.

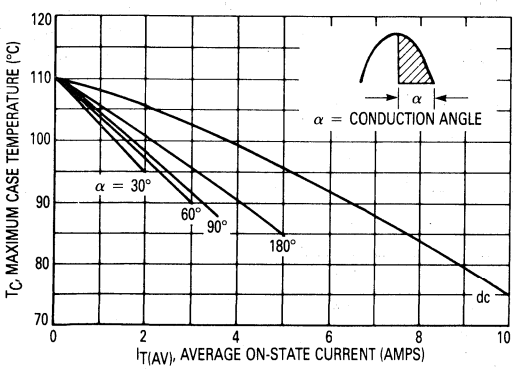


Figure 1. Average Current Derating

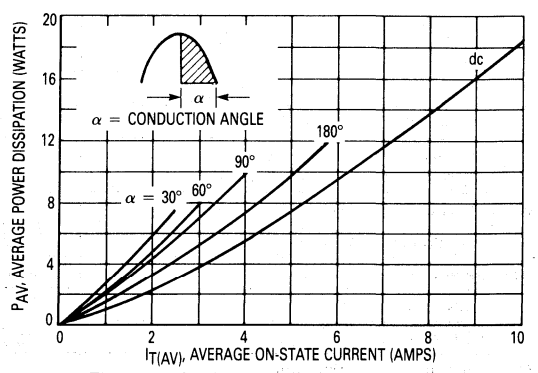


Figure 2. On-State Power Dissipation

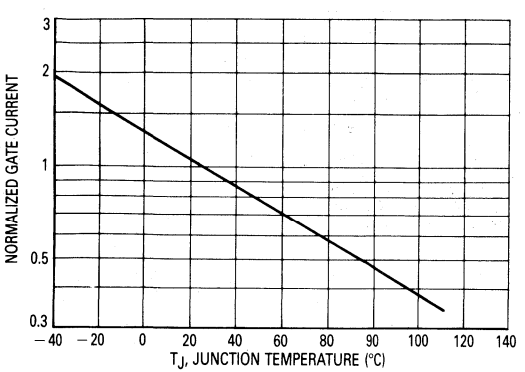


Figure 3. Normalized Gate Current

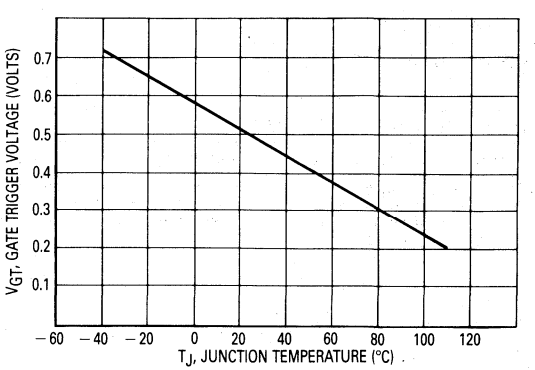


Figure 4. Gate Voltage

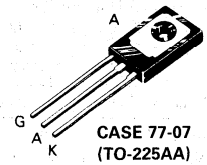
Silicon Controlled Rectifiers

MCR506 Series

SCRs
6 AMPERES RMS
50 thru 600 VOLTS

. . . PNPN devices designed for high volume consumer applications such as temperature, light, and speed control; process and remote control, and warning systems where reliability of operation is important.

- Passivated Surface for Reliability and Uniformity
- Power Rated at Economical Prices
- Practical Level Triggering and Holding Characteristics
- Flat, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability.



CASE 77-07
(TO-225AA)
STYLE 2

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Reverse Blocking Voltage, Note 1	MCR506-2 -3 -4 -6 -8	V_{RRM}	50 100 200 400 600	Volts
RMS Forward Current (All Conduction Angles)		$I_{T(RMS)}$	6	Amp
Average Forward Current (T_C = 93°C)		$I_{T(AV)}$	3.82	Amp
Peak Non-Repetitive Surge Current (1/2 cycle, 60 Hz, T_J = −40 to 110°C)		I_{TSM}	40	Amp
Circuit Fusing Considerations (t = 8.3 ms)		I^2t	2.6	A^2s
Peak Gate Power		P_{GM}	0.5	Watt
Average Gate Power		$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current		I_{GM}	0.2	Amp
Peak Reverse Gate Voltage		V_{RGM}	6	Volts
Operating Junction Temperature Range		T_J	−40 to 110	°C
Storage Temperature Range		T_{stg}	−40 to 150	°C
Mounting Torque (Note 2)		—	6	in. lb.

Notes: 1. Ratings apply for zero or negative gate voltage but positive gate voltage shall not be applied concurrently with a negative potential on the anode. When checking forward or reverse blocking capability, thyristor devices should not be tested with a constant current source in a manner that the voltage applied exceeds the rated blocking voltage.

2. Torque rating applies with use of torque washer (Shakeproof WD19523 or equivalent). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Anode lead and heat sink contact pad are common. (See AN290 B)

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +225°C. For optimum results, an activated flux (oxide removing) is recommended.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	75	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted, R_{GK} = 1000 ohms.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (T_J = 110°C) MCR506-2	V_{DRM}	60	—	—	Volts
-3		100	—	—	
-4		200	—	—	
-6		400	—	—	
-8		600	—	—	
Peak Forward Blocking Current (Rated V_{DRM}, T_J = 110°C)	I_{DRM}	—	—	200	μA
Peak Reverse Blocking Current (Rated V_{RRM}, T_J = 110°C)	I_{RRM}	—	—	200	μA
Forward "On" Voltage (I_{TM} = 12 A Peak)	V_{TM}	—	—	1.9	Volts
Gate Trigger Current (Continuous dc) (V_{AK} = 7 Vdc, R_L = 100 ohms)	I_{GT}	—	—	200	μA
(V_{AK} = 7 Vdc, R_L = 100 ohms, T_C = −40°C)		—	—	500	
Gate Trigger Voltage (Continuous dc) (V_{AK} = 7 Vdc, R_L = 100 ohms, T_C = 25°C)	V_{GT}	—	—	1	Volts
Gate Non-Trigger Voltage (V_{AK} = Rated V_{DRM}, R_L = 100 ohms, T_J = 110°C)	V_{GD}	0.2	—	—	Volts
Holding Current (V_{AK} = 7 Vdc, T_C = 25°C)	I_H	—	—	5	mA
Forward Voltage Application Rate (T_J = 110°C)	dv/dt	—	10	—	V/μs

3

FIGURE 1 — CURRENT DERATING

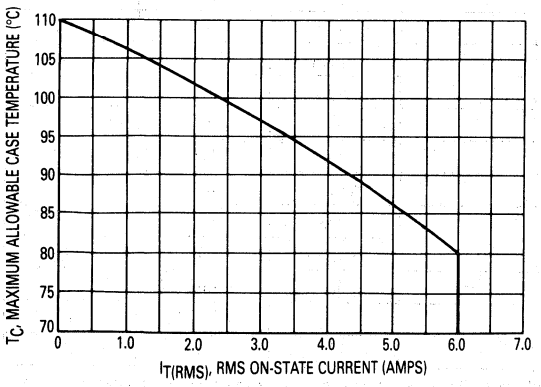
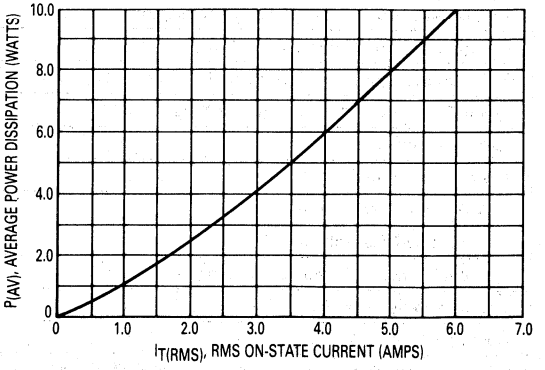


FIGURE 2 — POWER DISSIPATION



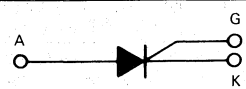
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

MCR703A thru MCR708A

SCRs
4.0 AMPERES RMS
100 thru 600 VOLTS

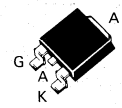
. . . PNPN devices designed for high volume, low cost consumer applications such as temperature, light and speed control; process and remote control; and warning systems where reliability of operation is critical.

- Small Size
- Passivated Die Surface for Reliability and Uniformity
- Low Level Triggering and Holding Characteristics
- Recommend Electrical Replacement for C106
- Available in Two Package Styles:
  Surface Mount Leadforms — Case 369A
  Miniature Plastic Package — Straight Leads — Case 369

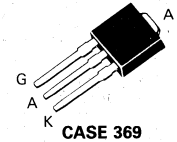


ORDERING INFORMATION

- To Obtain "DPAK" in Surface Mount Leadform (Case 369A):
  Shipped in Sleeves — No Suffix, i.e., MCR706A
  Shipped in 16 mm Tape and Reel — Add "RL" Suffix to Device Number, i.e., MCR706ARL
- To Obtain "DPAK" in Straight Lead Version:
  Shipped in Sleeves — Add '1' Suffix to Device Number, i.e., MCR706A1



CASE 369A
STYLE 5



CASE 369
STYLE 5

MAXIMUM RATINGS (T_C = 110°C unless otherwise noted.)

Characteristic	Symbol	Value	Unit
Repetitive Peak Forward and Reverse Blocking Voltage, Note 1 (1/2 Sine Wave) (R_{GK} = 1000 ohms, T_C = −40 to +110°C) MCR703A1, MCR703A	V_{DRM} or V_{RRM}	100	Volts
MCR704A1, MCR704A		200	
MCR706A1, MCR706A		400	
MCR708A1, MCR708A		600	
Non-Repetitive Peak Reverse Blocking Voltage (1/2 Sine Wave, R_{GK} = 1000 ohms, T_C = −40 to +110°C) MCR703A1, MCR703A	V_{RSM}	150	Volts
MCR704A1, MCR704A		250	
MCR706A1, MCR706A		450	
MCR708A1, MCR708A		650	
Average On-State Current (T_C = −40 to +90°C) (T_C = +100°C)	$I_{T(AV)}$	2.6 1.6	Amps
Surge On-State Current (1/2 Sine Wave, 60 Hz, T_C = +90°C) (1/2 Sine Wave, 1.5 ms, T_C = +90°C)	I_{TSM}	25 35	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	2.6	A^2s
Peak Gate Power (Pulse Width = 10 μs, T_C = 90°C)	P_{GM}	0.5	Watt
Average Gate Power (t = 8.3 ms, T_C = 90°C)	$P_{G(AV)}$	0.1	Watt
Peak Forward Gate Current	I_{GM}	0.2	Amp
Peak Reverse Gate Voltage	V_{RGM}	6	Volts
Operating Junction Temperature Range	T_J	−40 to +110	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode. Devices should not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

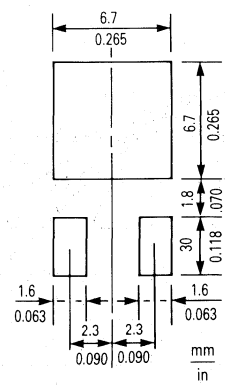


Figure 1. Minimum Pad Sizes for Surface Mounting

3

THERMAL CHARACTERISTICS

Characteristic	Symbol	Min	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	—	8.33	°C/W
Thermal Resistance Junction to Ambient (Case 369A-04) (Note 1)	$R_{\theta JA}$	—	80	°C/W
Thermal Resistance Junction to Ambient (Case 369-03) (Note 2)	$R_{\theta JA}$	—	85	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and R_{GK} = 1000 ohms unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C T_C = 110°C	I_{DRM}, I_{RRM}	 — —	 — —	 10 200	μA
Peak Forward "On" Voltage (I_{TM} = 8.2 A Peak, Pulse Width = 1.0 to 2.0 ms, 2% Duty Cycle)	V_{TM}	—	—	2.2	Volts
Gate Trigger Current (Continuous dc) (Note 3) (V_{AK} = 12 Vdc, R_L = 24 Ohms) (V_{AK} = 12 Vdc, R_L = 24 Ohms, T_C = −40°C)	I_{GT}	 — —	 25 —	 75 300	μA
Gate Trigger Voltage (Continuous dc) (Source Voltage = 12 V, R_S = 50 Ohms) (V_{AK} = 12 Vdc, R_L = 24 Ohms, T_C = −40°C)	V_{GT}	—	—	1.0	Volts
Gate Non-Trigger Voltage (V_{AK} = Rated V_{DRM}, R_L = 100 Ohms, T_C = 110°C)	V_{GD}	0.2	—	—	Volts
Holding Current (V_{AK} = 12 Vdc, I_{GT} = 2.0 mA) T_C = 25°C (Initiating On-State Current = 200 mA) T_C = −40°C	I_H	 — —	 — —	 5.0 10	mA
Total Turn-On Time (Source Voltage = 12 V, R_S = 6.0 k Ohms) (I_{TM} = 8.2 A, I_{GT} = 2.0 mA, Rated V_{DRM}) (Rise Time = 20 ns, Pulse Width = 10 μs)	t_{gt}	—	2.0	—	μs
Forward Voltage Application Rate (V_D = Rated V_{DRM}, T_C = 110°C)	dv/dt	—	10	—	V/μs

Notes: 1. Case 369A-04 when surface mounted on minimum pad sizes recommended.
2. Case 369-03 standing in free air.
3. R_{GK} current not included in measurement.

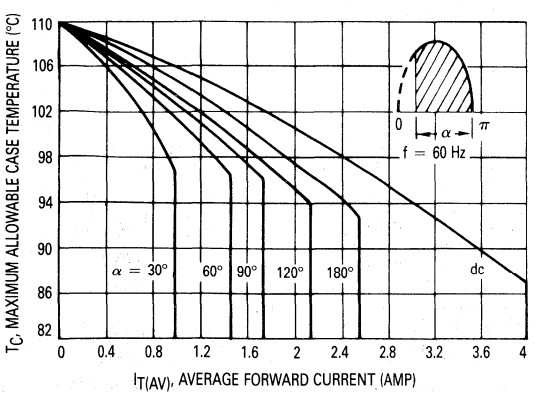


Figure 2. Maximum Case Temperature

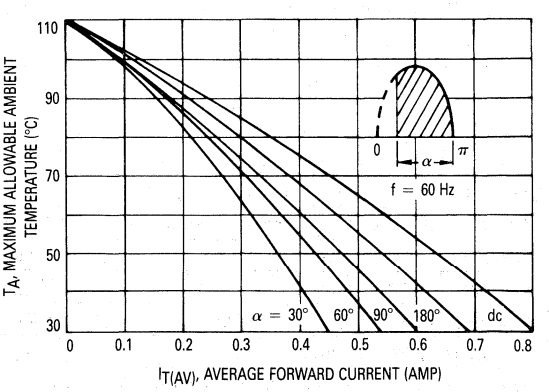


Figure 3. Maximum Ambient Temperature

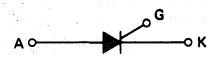
Silicon Controlled Rectifiers
Reverse Blocking Triode Thyristors

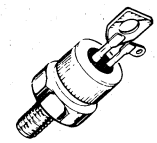
MCR1718-5 thru MCR1718-8

SCRs
25 AMPERES RMS
300 thru 600 VOLTS

. . . fast switching, high-voltage thyristors especially designed for pulse modulator applications.

- High-Voltage Capability from 300 to 600 Volts
- Repetitive Pulse Current to 1000 Amps
- Pulse Repetition as High as 4000 pps
- Current Application Rate as High as 1000 A/μs





CASE 263-04
STYLE 1

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Forward or Reverse Blocking Voltage, Note 1	V_{DRM} V_{RRM}		Volts
MCR1718-5		300	
-6		400	
-7		500	
-8		600	
Peak Reverse Blocking Voltage (Transient) (Non-Recurrent 5 ms (max))	V_{RSM}		Volts
MCR1718-5		400	
-6		500	
-7		600	
-8		700	
Forward Current RMS	$I_{T(RMS)}$	25	Amps
Peak Forward Surge Current (1–10 μs Pulse Width)	I_{TSM}	1000	Amps
Current Application Rate (up to 1000 A peak)	di/dt	1000	A/μs
Circuit Fusing (t = 8.3 ms)	I^2t	250	A^2s
Dynamic Average Power (T_C = 65°C)	$P_{F(AV)}$	30	Watts
Peak Gate Power — Forward	P_{GM}	20	Watts
Average Gate Power — Forward	$P_{G(AV)}$	1	Watt
Peak Gate Current — Forward	I_{GM}	5	Amps

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage.

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Peak Gate Voltage	V_{GM}	10	Volts
Operating Junction Temperature Range	T_J	−65 to +125	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C T_J = 125°C	I_{DRM}, I_{RRM}	— —	— —	10 8	μA mA
Forward "On" Voltage (I_{TM} = 25 Adc)	V_{TM}	—	1.1	1.3	Volts
Dynamic Forward On Voltage (I_{GT} = 500 mA, I_{pulse} = 500 Amps) (1 μs after start (10% pt.) of I_{pulse}) (5 μs after start (10% pt.) of I_{pulse})	v_{TM}	— —	30 5	— —	
Gate Trigger Current (Continuous dc) (V_D = 7 Vdc, R_L = 50 Ohms)	I_{GT}	—	10	50	mA
Gate Trigger Voltage (Continuous dc) (V_D = 7 Vdc, R_L = 50 Ohms) (V_D = Rated V_{DRM}, R_L = 50 Ohms, T_J = 125°C)	V_{GT} V_{GD}	— 0.25	0.8 —	1.5 —	Volts
Holding Current (V_D = 7 Vdc, Gate Open) (V_D = 7 Vdc, Gate Open, T_J = 125°C)	I_H	5 —	15 6	— —	mA
Circuit Commutated Turn-Off Time (I_F = 500 A, I_R = 10 A, dv/dt = 20 V/μs V_D = Rated V_{DRM}, V_R = Rated V_{RRM}) (Conductive Charging Circuit — Circuit dependent)	t_q	—	20	—	μs
Critical Exponential Rate of Rise (Gate Open, T_J = 125°C, V_D = Rated V_{DRM})	dv/dt	—	100	—	V/μs

3

MCR1906 Series

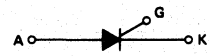
Silicon Controlled Rectifiers

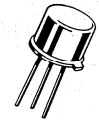
Reverse Blocking Triode Thyristors

These devices are glassivated planar construction designed for applications in control systems and sensing circuits where low-level gating and holding characteristics are necessary.

- Low-Level Gate Characteristics — I_{GT} = 1 mA (Max) @ T_C = 25°C
- Low Holding Current — I_H = 5 mA (Max) @ T_C = 25°C
- Glass-to-Metal Bond for Maximum Hermetic Seal

SCRs
1.6 AMPERES RMS
50 thru 600 VOLTS





CASE 79-04
(TO-205AD)
STYLE 3

MAXIMUM RATINGS (T_J = 100°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Repetitive Peak Reverse Blocking Voltage MCR1906-2 MCR1906-3 MCR1906-4 MCR1906-6 MCR1906-8	V_{RRM}	 50 100 200 400 600	Volts
RMS On-State Current (All Conduction Angles)	$I_{T(RMS)}$	1.6	Amp
Peak Non-Repetitive Surge Current (One Cycle, 60 Hz, T_J = −40 to +110°C) Preceded and followed by rated current and voltage	I_{TSM}	15	Amps
Peak Gate Power	P_{GM}	0.1	Watt
Average Gate Power	$P_{GF(AV)}$	0.01	Watt
Peak Gate Current	I_{GM}	0.1	Amp
Peak Gate Voltage	V_{GM}	6	Volts
Operating Junction Temperature Range	T_J	−65 to +110	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Lead Solder Temperature (>1/16″ From Case, 10 s max)	—	+230	°C

3

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage (R_{GK} = 1000 Ohms)	V_{DRM}				Volts
MCR1906-2		50	—	—	
MCR1906-3		100	—	—	
MCR1906-4		200	—	—	
MCR1906-6		400	—	—	
MCR1906-8		600	—	—	
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, R_{GK} = 1000 Ohms) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 110°C		—	—	500	μA
Peak On-State Voltage (Pulsed, 1 ms max, Duty Cycle ≤ 1%) (I_F = 1 Adc peak)	V_{TM}	—	—	1.75	Volt
Gate Trigger Current (Continuous dc) (V_{AK} = 7 V, R_L = 100 ohms)	I_{GT}	—	—	1	mAdc
Gate Trigger Voltage (Continuous dc)	V_{GT}				Volt
(V_{AK} = 7 V, R_L = 100 ohms)		—	—	1	
(V_{AK} = Rated V_{DRM}, R_L = 100 ohms, R_{GK} = 1000 Ohms, T_J = 110°C)		0.1	—	—	
Holding Current (V_{AK} = 7 V, R_{GK} = 1000 ohms)	I_H	—	—	5	mA
Turn-On Time	t_{gt}				μs
(I_{GT} = 10 mA, I_F = 1 A)		—	0.8	—	
(I_{GT} = 20 mA, I_F = 1 A)		—	0.6	—	
Turn-Off Time (I_F = 1 A, I_R = 1 A, dv/dt = 20 V/μs, T_J = 110°C)	t_q	—	10	—	μs

3

CURRENT DERATING

FIGURE 1 – CASE TEMPERATURE REFERENCE

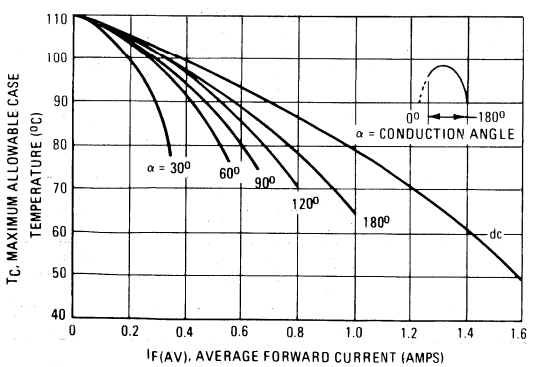
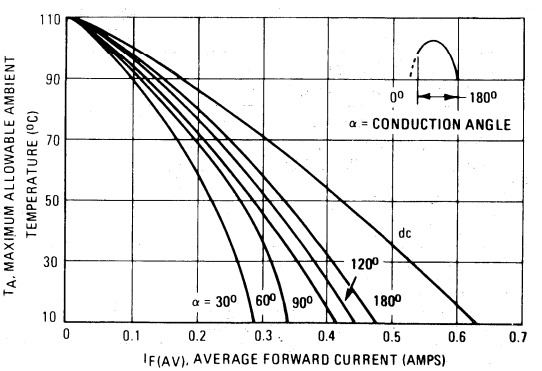


FIGURE 2 – AMBIENT TEMPERATURE REFERENCE



Silicon Controlled Rectifiers

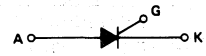
Reverse Blocking Triode Thyristors

MCR3818 Series
MCR3918-( )A Series

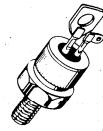
. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Supplied in Either Pressfit or Stud Package
- High Surge Current Rating — I_{TSM} = 240 Amps
- Low On-State Voltage — 1.2 V (Typ) @ I_{TM} = 20 Amps
- Practical Level Triggering and Holding Characteristics — 40 mA (Max) and 50 mA (Max) @ T_C = 25°C

SCRs
25 AMPERES RMS
50 thru 800 VOLTS



CASE 174-04
(TO-203AA)
STYLE 1
MCR3818 Series



CASE 263-04
STYLE 1
MCR3918-( )A Series

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Voltage, Note 1	V_{DRM} or V_{RRM}		Volts
MCR3818- 2		50	
MCR3918-( )A 3		100	
4		200	
6		400	
8		600	
10		800	
Non-Repetitive Reverse Blocking Voltage	V_{RSM}		Volts
MCR3818- 2		75	
MCR3918-( )A 3		150	
4		300	
6		500	
8		700	
10		900	
On-State Current RMS	$I_{T(RMS)}$	20	Amps
Average On-State Current (T_C = 67°C)	$I_{T(AV)}$	13	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	235	A^2s
Peak Non-Repetitive Surge Current (One Cycle, 60 Hz, T_J = −40 to +100°C)	I_{TSM}	240	Amps
Peak Gate Power (Maximum Pulse Width = 10 μs)	P_{GM}	5	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Peak Forward Gate Current (Maximum Pulse Width = 10 μs)	I_{GM}	2	Amps

Note 1. V_{DRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. These devices should not be tested with a constant current source for forward or reverse blocking capability such that the voltage applied exceeds the rated blocking voltage.

3

MAXIMUM RATINGS — continued

Rating	Symbol	Value	Unit
Peak Gate Voltage	V_{GM}	10	Volts
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C
Stud Torque		30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Typ	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$			°C/W
Pressfit Package		1	1.5	
Stud Package		1.1	1.6	

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Max	Unit
Peak Forward or Reverse Blocking Current	I_{DRM}, I_{RRM}			
(Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C		—	10	μA
T_J = 100°C		—	5	mA
Gate Trigger Current (Continuous dc)	I_{GT}			mA
(V_D = 7 Vdc, R_L = 100 Ω)		—	40	
(V_D = 7 Vdc, R_L = 100 Ω, T_C = −40°C)		—	75	
Gate Trigger Voltage (Continuous dc)	V_{GT}			Volts
(V_D = 7 Vdc, gate open)		—	1.5	
(V_D = 7 Vdc, R_L = 100 Ω, T_C = −40°C)		—	2.5	
(V_D = Rated V_{DRM}, R_L = 100 Ω, T_J = 100°C)		0.2	—	
Peak On-State Voltage (Pulse Width = 1 ms max, duty cycle ≤ 1%)	V_{TM}			Volts
(I_{TM} = 20 A)		—	1.5	
(I_{TM} = 41 A)		—	1.7	
Holding Current	I_H			mA
(V_D = 7 Vdc, gate open)		—	50	
(V_D = 7 Vdc, gate open, T_C = −40°C)		—	90	
		Typical		
Gate Controlled Turn-On Time (t_d + t_r) (I_{TM} = 20 A, I_{GT} = 40 mAdc, V_D = Rated V_{DRM})	t_{gt}	1		μs
Circuit Commutated Turn-Off Time	t_q			μs
(I_{TM} = 10 A, I_R = 10 A)		20		
(I_{TM} = 10 A, I_R = 10 A, T_J = 100°C)		30		
(V_D = V_{DRM} = rated voltage) (dv/dt = 30 V/μs)				
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Wave Form, Gate open, T_J = 100°C)	dv/dt	50		V/μs

3

EFFECT OF TEMPERATURE UPON TYPICAL TRIGGER CHARACTERISTICS

FIGURE 1 – GATE TRIGGER CURRENT

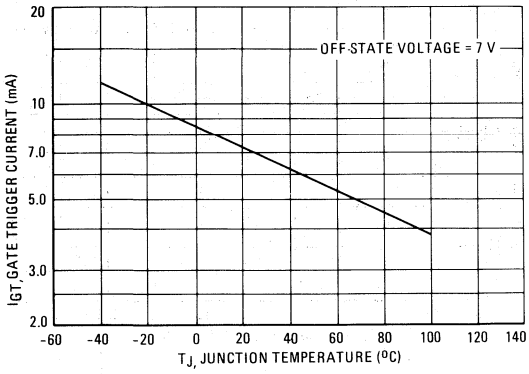
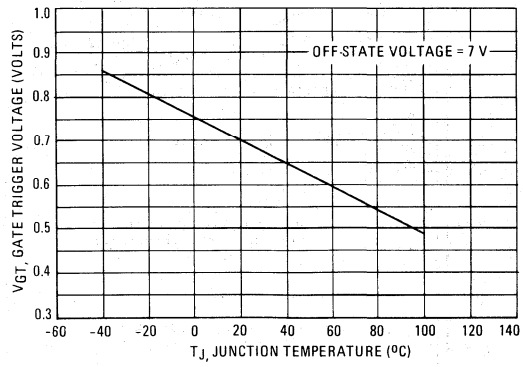


FIGURE 2 – GATE TRIGGER VOLTAGE



MAXIMUM ALLOWABLE NON-REPETITIVE SURGE CURRENT

FIGURE 3 – 60 Hz SURGES

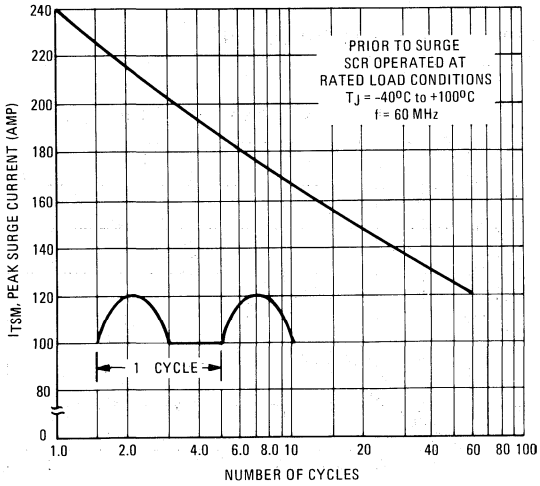


FIGURE 4 – SUB-CYCLE SURGES

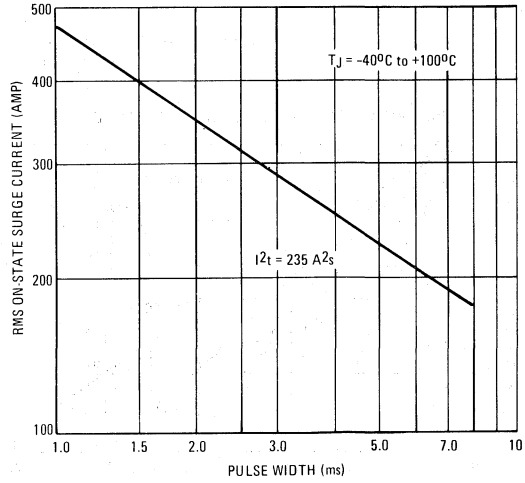


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FIGURE 5 – GATE TRIGGER CHARACTERISTICS

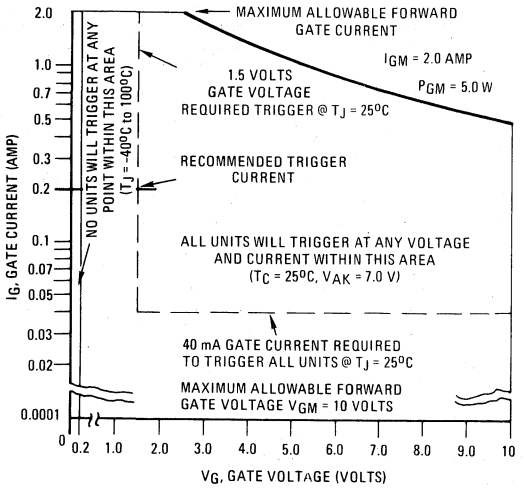
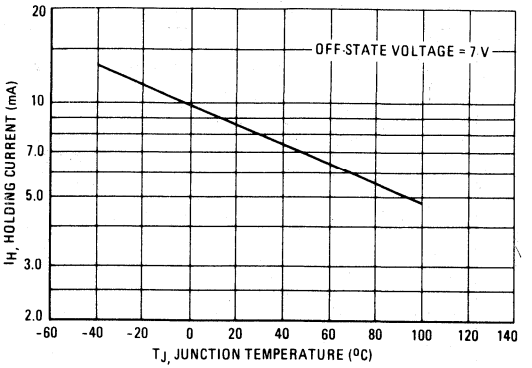


FIGURE 6 – EFFECT OF TEMPERATURE ON TYPICAL HOLDING CURRENT



DERATING AND DISSIPATION FOR RESISTIVE AND INDUCTIVE LOADS (f = 60 to 400 Hz, SINE WAVE)

FIGURE 7 – AVERAGE CURRENT DERATING

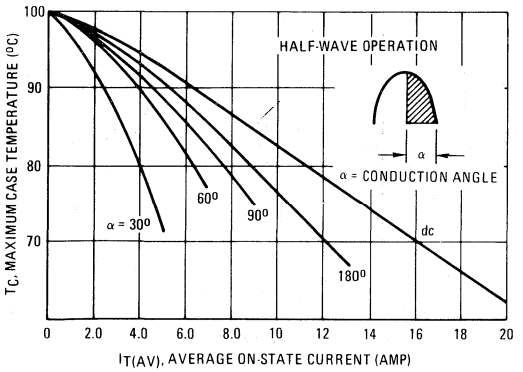


FIGURE 8 – ON-STATE POWER DISSIPATION

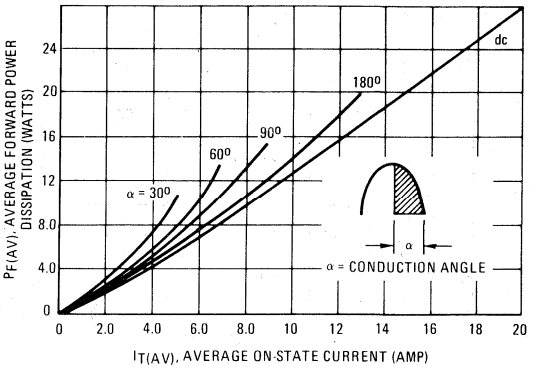


FIGURE 9 – ON-STATE CHARACTERISTICS

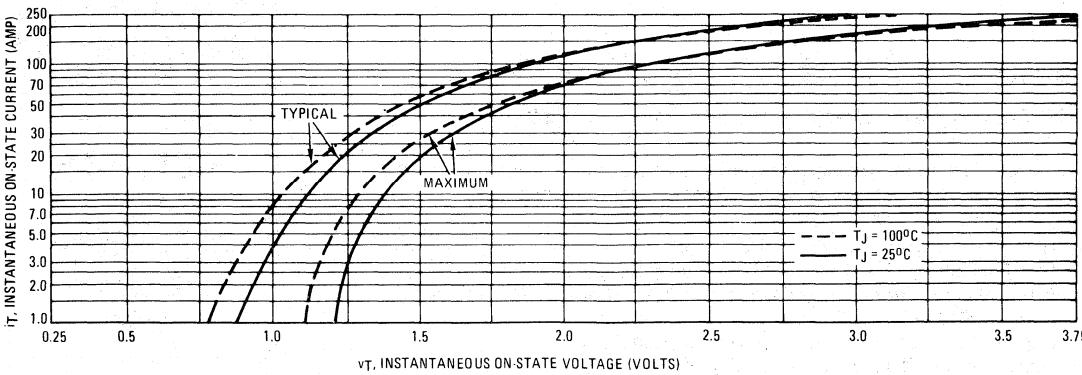


FIGURE 10 – TYPICAL THERMAL RESISTANCE OF PLATES

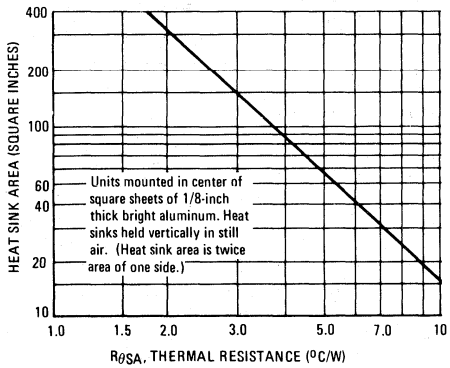
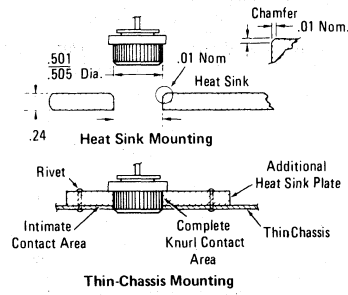


FIGURE 11 – MOUNTING DETAILS FOR PRESSFIT THYRISTORS



The hole edge must be chamfered as shown to prevent shearing off the knurled edge of the rectifier during press-in. The pressing force should be applied evenly on the shoulder ring to avoid tilting or canting of the rectifier case in the hole during the pressing operation. Also, the use of a thermal joint compound will be of considerable aid. The pressing force will vary from 250 to 1000 pounds, depending upon the heat sink material. Recommended hardnesses are: copper – less than 50 on the Rockwell F scale; aluminum – less than 65 on the Brinell scale. A heat sink as thin as 1/8" may be used, but the interface thermal resistance will increase in proportion to the reduction of contact area. A thin chassis requires the addition of a back-up plate.

3

MCR3896–99
(See 2N3870)

Silicon Controlled Rectifiers

Reverse Blocking Triode Thyristors

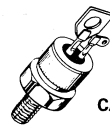
MCR3835 Series
MCR3935-( )A Series

SCRs
35 AMPERES RMS
50 thru 800 VOLTS

. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 350 Amps
- Low Forward "On" Voltage — 1.2 V (Typ) @ I_{TM} = 35 Amps
- Practical Level Triggering and Holding Characteristics — 10 mA (Typ) @ T_C = 25°C
- Rugged Construction in Either Pressfit or Stud Package
- Glass Passivated Junctions for Maximum Reliability

CASE 174-04
(TO-203AA)
STYLE 1
MCR3835 Series



CASE 263-04
STYLE 1
MCR3935-( )A Series

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Forward and Reverse Blocking Voltage Note 1	V_{DRM} V_{RRM}		Volts
MCR3835-2		50	
-8		600	
-10		800	
MCR3935-( )A2		50	
3		100	
4		200	
6		400	
8		600	
10		800	
Peak Non-Repetitive Reverse Blocking Voltage (t ≤ 5 ms)	V_{RSM}		Volts
MCR3835-2		35	
-8		700	
-10		900	
MCR3935-( )A2		75	
3		150	
4		300	
6		500	
8		700	
10		900	
Forward Current RMS	$I_{T(RMS)}$	35	Amps
Peak Surge Current (One Cycle, 60 Hz, T_J = −40 to +125°C)	I_{TSM}	350	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	510	A^2s
Peak Gate Power	P_{GFM}	5	Watts
Average Gate Power	$P_{GF(AV)}$	0.5	Watt
Peak Forward Gate Current	I_{GFM}	2	Amps
Peak Gate Voltage — Forward	V_{GFM}	10	Volt
Reverse	V_{GRM}	10	
Operating Junction Temperature Range	T_J	−40 to +125	°C

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode.

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Storage Temperature Range	T_{stg}	−40 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case MCR3835 MCR3935	$R_{\theta JC}$	 1.2 1.3	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C T_J = 100°C	I_{DRM}, I_{RRM}	 — —	 — 1	 10 5	 μA mA
Forward "On" Voltage (I_{TM} = 35 A Peak)	V_{TM}	—	1.2	1.5	Volts
Gate Trigger Current (Continuous dc) (V_D = 7 V, R_L = 100 Ω)	I_{GT}	—	10	40	mA
Gate Trigger Voltage (Continuous dc) (V_D = 7 V, R_L = 100 Ω) (V_D = Rated V_{DRM}, R_L = 100 Ω, T_J = 100°C)	V_{GT} V_{GD}	 — 0.2	 0.7 —	 1.5 —	Volts
Holding Current (V_D = 7 V, gate open)	I_H	—	10	50	mA
Turn-On Time (t_d + t_r) (I_{TM} = 35 Adc, I_{GT} = 40 mAdc)	t_{on}	—	1	—	μs
Turn-Off Time (I_{TM} = 10 A, I_R = 10 A) (I_{TM} = 10 A, I_R = 10 A, T_J = 100°C)	t_q	 — —	 20 30	 — —	μs
Forward Voltage Application Rate (V_D = Rated V_{DRM}, T_J = 100°C)	dv/dt	—	50	—	V/μs

3

FIGURE 1 – CURRENT DERATING

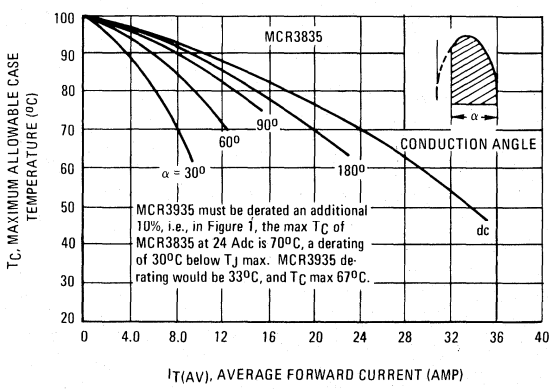


FIGURE 2 – TYPICAL POWER DISSIPATION

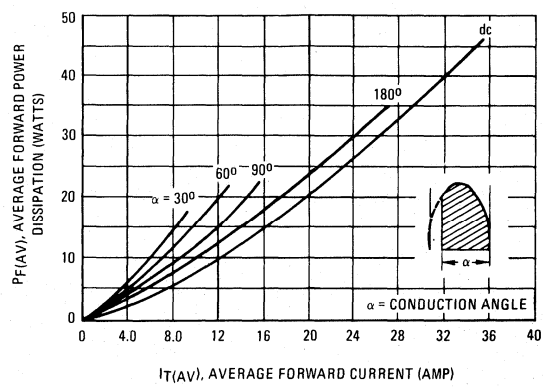


FIGURE 3 – TYPICAL GATE TRIGGER CURRENT

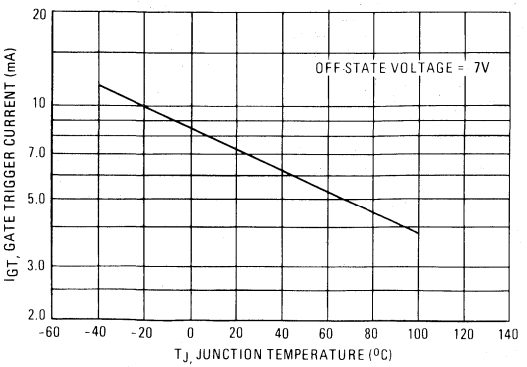
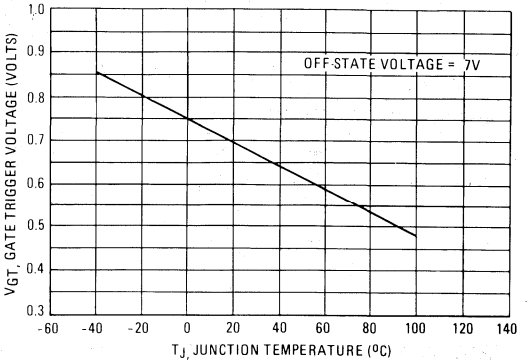


FIGURE 4 – TYPICAL GATE TRIGGER VOLTAGE



3

Thyristors
Silicon Controlled Rectifiers

. . . designed for industrial and consumer applications such as power supplies, battery chargers, temperature, motor, light and welder controls.

- Economical for a Wide Range of Uses
- High Surge Current — I_{TSM} = 200 Amps
- Low Forward "On" Voltage — 1.2 V (Typ) @ I_{TM} = 20 Amps
- Practical Level Triggering and Holding Characteristics — 10 mA (Typ) @ T_C = 25°C
- Rugged Construction in Either Pressfit, Stud or Isolated Stud Package
- Glass Passivated Junctions for Maximum Reliability

MCR5164–67
(See 2N5168)

MCR6200 Series

SCRs
20 AMPERES RMS
100 thru 600 VOLTS

CASE 174-04
(TO-203AA)
STYLE 1
MCR6200 SERIES

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 Repetitive Peak Reverse Voltage, Note 1 MCR6200 A MCR6200 B MCR6200 D MCR6200 M	V_{DROM} V_{RROM}	 100 200 400 600	Volts
Non-Repetitive Peak Off-State Voltage, Note 1 Non-Repetitive Peak Reverse Voltage, Note 1 MCR6200 A MCR6200 B MCR6200 D MCR6200 M	V_{DSOM} V_{DROM}	 150 250 500 700	Volts
RMS On-State Current (T_C = 75°C)	$I_{T(RMS)}$	20	Amps
Peak Non-Repetitive Surge Current (One Full Cycle of surge current at 60 Hz, preceded and followed by rated current, T_C = 75°C)	I_{TSM}	200	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	170	A^2s
Peak Gate Power (10 μs Max)	P_{GM}	40	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Operating Junction Temperature Range	T_J	−65 to +100	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case MCR6200	$R_{\theta JC}$	1.2 1.4	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
Intantaneous Forward Breakover Voltage (Gate Open, T_C = 100°C)		$V_{(BO)O}$				Volts
MCR6200	A		100	—	—	
MCR6200	B		200	—	—	
MCR6200	D		400	—	—	
MCR6200	M		600	—	—	
Peak Blocking Current (Rated V_{DROM} @ T_C = 100°C)	T_C = 25°C	I_{DOM}, I_{RROM}	—	—	10	μA
	T_C = 100°C		—	—	2	mA
Peak On-State Voltage (I_T = 100 A Peak)		V_T	—	—	2.4	Volts
Gate Trigger Current (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 30 Ohms)		I_{GT}	—	—	15	mA
Gate Trigger Voltage (Continuous dc) (Main Terminal Voltage = 12 Vdc, R_L = 30 Ohms)		V_{GT}	—	—	2	Volts
Holding Current (Either Direction) (Main Terminal Voltage = 12 Vdc, Gate Open)		I_{HO}	—	—	20	mA
Gate Controlled Turn-On Time (V_D = $V_{(BO)O}$, I_T = 30 A Peak, I_{GT} = 200 mA, Rise Time = 0.1 μs)		t_{gt}	—	2	—	μs
Critical Rate-of-Rise of Off-State Voltage (V_D = $V_{(BO)O}$, Exponential Voltage Rise, Gate Open, T_C = 100°C)		dv/dt				V/μs
MCR6200	A,D		10	100	—	
MCR6200	B		10	150	—	
MCR6200	M		10	75	—	

3

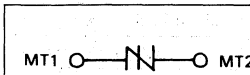
Sidac High Voltage Bilateral Triggers

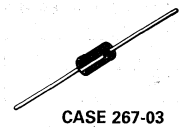
MK1V115
MK1V125
MK1V135

SIDACs
1 AMPERE RMS
104 thru 135 VOLTS

. . . designed for direct interface with the ac power line. Upon reaching the breakover voltage in each direction, the device switches from a blocking state to a low voltage on-state. Conduction will continue like an SCR until the main terminal current drops below the holding current. The plastic axial lead package provides high pulse current capability at low cost. Glass passivation insures reliable operation. Applications are:

- High Pressure Sodium Vapor Lighting
- Strobes and Flashers
- Ignitors
- High Voltage Regulators
- Line Transient Clippers
- Pulse Generators





CASE 267-03
SURMETIC 50
PLASTIC AXIAL
STYLE 1

3

MAXIMUM RATINGS

Rating	Symbol	Min	Max	Unit
Repetitive Breakover Voltage MK1V115 MK1V125 MK1V135	$V_{(BO)}$	 104 110 120	 115 125 135	Volts
Off-State Repetitive Voltage	V_{DRM}	—	±90	Volts
On-State RMS Current (All Conduction Angles)	$I_{T(RMS)}$	—	1	Amp
On-State Surge Current (Nonrepetitive) (60 Hz One Cycle Sine Wave, Peak Value)	I_{TSM}	—	20	Amps
Operating Junction Temperature Range	T_J	−40	+125	°C
Storage Temperature Range	T_{stg}	−40	+150	°C
Lead Solder Temperature (Lead Length ≥ 1/16″ from case, 10 s Max)	—	—	+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Min	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	—	15	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	—	45	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted; both directions.)

Characteristic	Symbol	Min	Typ	Max	Unit
Breakover Current (60 Hz Sine Wave)	$I_{(BO)}$	—	—	200	μA
Repetitive Peak Off-State Current (60 Hz Sine Wave, V = V_{DRM})	I_{DRM}	—	—	10	μA
Repetitive Peak On-State Current (T_C = 25°C, Pulse Width = 10 μs, Repetition Frequency, f = 1 kHz)	I_{TRM}	—	20	—	Amps
Forward "On" Voltage (I_{TM} = 1 A peak)	V_{TM}	—	1.1	1.5	Volts
Dynamic Holding Current (60 Hz Sine Wave, R_S = 0.1 kΩ)	I_H	—	—	100	mA
Switching Resistance	R_S	0.1	—	—	kΩ
Maximum Rate of Change of On-State Current	di/dt	—	50	—	A/μs

CURRENT DERATING

FIGURE 1 — MAXIMUM CASE TEMPERATURE

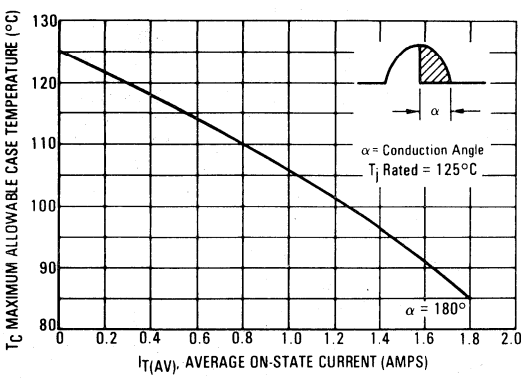


FIGURE 2 — MAXIMUM AMBIENT TEMPERATURE

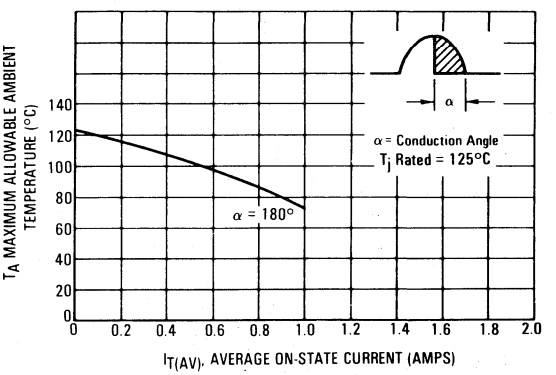


FIGURE 3 — TYPICAL FORWARD VOLTAGE

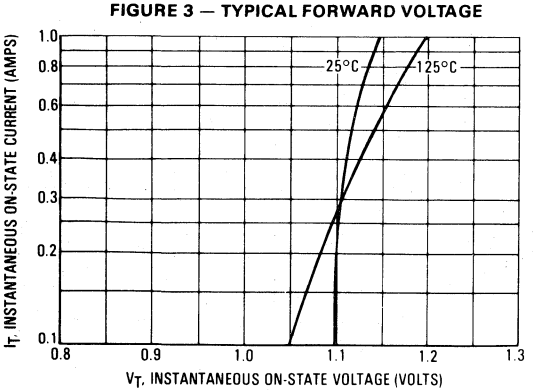
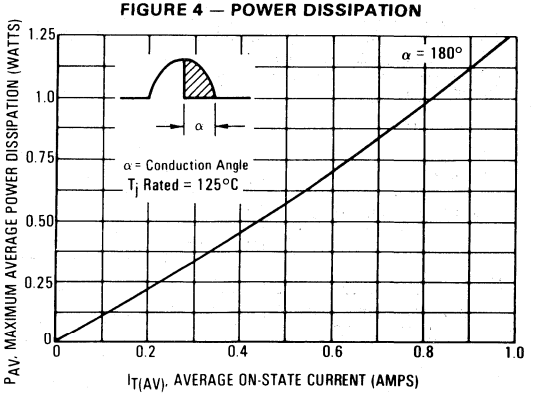
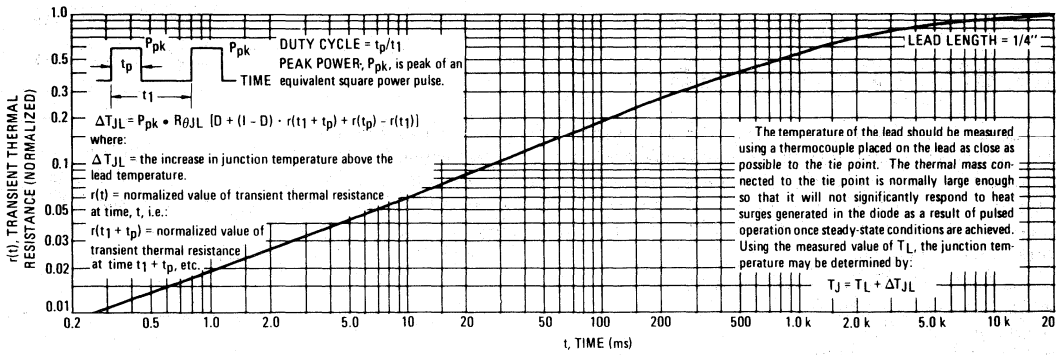


FIGURE 4 — POWER DISSIPATION



THERMAL CHARACTERISTICS

FIGURE 5 — THERMAL RESPONSE



TYPICAL CHARACTERISTICS

FIGURE 6 — BREAKOVER CURRENT

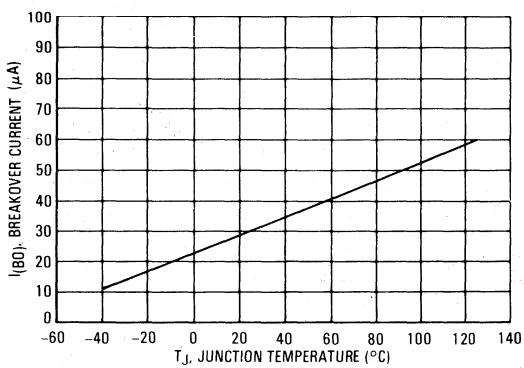


FIGURE 7 — HOLDING CURRENT

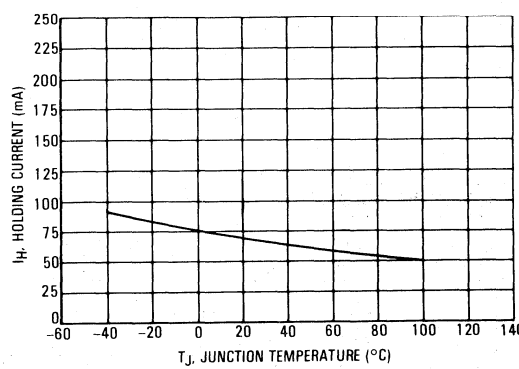
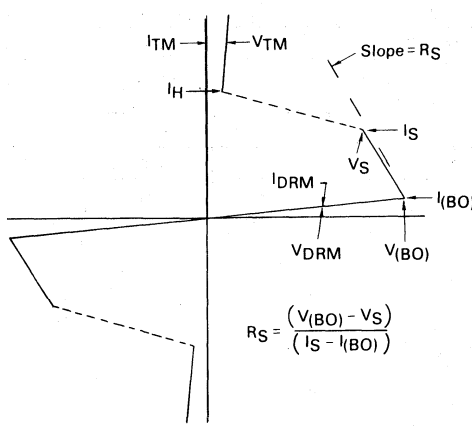


FIGURE 8 — V-1 CHARACTERISTICS



3

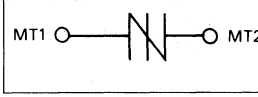
Sidac High Voltage Bilateral Triggers

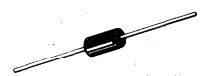
MK1V240
MK1V260
MK1V270

SIDACs
1 AMPERE RMS
220 thru 280 VOLTS

. . . designed for direct interface with the ac power line. Upon reaching the breakover voltage in each direction, the device switches from a blocking state to a low voltage on-state. Conduction will continue like an SCR until the main terminal current drops below the holding current. The plastic axial lead package provides high pulse current capability at low cost. Glass passivation insures reliable operation. Applications are:

- High Pressure Sodium Vapor Lighting
- Strobes and Flashers
- Ignitors
- High Voltage Regulators
- Pulse Generators





CASE 267-03
SURMETIC 50
PLASTIC AXIAL
STYLE 1

3

MAXIMUM RATINGS

Rating	Symbol	Min	Max	Unit
Off-State Repetitive Voltage	V_{DRM}	—	±180	Volts
On-State Current RMS (T_L = 100°C, LL = 3/8", 60 Hz Sine Wave Conduction Angle = 180°)	$I_{T(RMS)}$	—	1	Amp
On-State Surge Current (Nonrepetitive) (60 Hz One Cycle Sine Wave, Peak Value)	I_{TSM}	—	20	Amps
Operating Junction Temperature Range	T_J	−40	+125	°C
Storage Temperature Range	T_{stg}	−40	+150	°C
Lead Solder Temperature (Lead Length ≥ 1/16" from Case, 10 s Max)	—	—	+230	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Min	Max	Unit
Thermal Resistance, Junction to Lead (LL = 3/8")	$R_{\theta JL}$	—	15	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted; both directions.)

Characteristic	Symbol	Min	Typ	Max	Unit
Breakover Voltage	$V_{(BO)}$				Volts
MK1V240		220	—	250	
MK1V260		240	—	270	
MK1V270		250	—	280	
Breakover Current (60 Hz Sine Wave)	$I_{(BO)}$	—	—	200	μA
Repetitive Peak Off-State Current (60 Hz Sine Wave, V = V_{DRM})	I_{DRM}	—	—	10	μA
Forward "On" Voltage (I_{TM} = 1 A peak)	V_{TM}	—	1.1	1.5	Volts
Dynamic Holding Current (60 Hz Sine Wave, R_S = 0.1 kΩ)	I_H	—	—	100	mA
Switching Resistance	R_S	0.1	—	—	kΩ
Maximum Rate of Change of On-State Current	di/dt	—	50	—	A/μs

CURRENT DERATING

FIGURE 1 — MAXIMUM LEAD TEMPERATURE

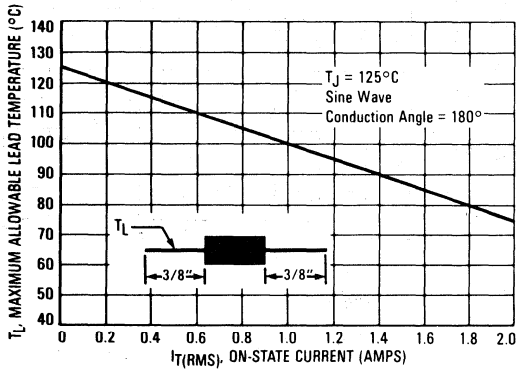


FIGURE 2 — MAXIMUM AMBIENT TEMPERATURE

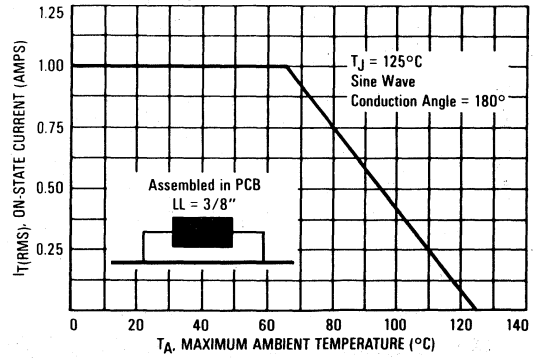


FIGURE 3 — TYPICAL ON-STATE VOLTAGE

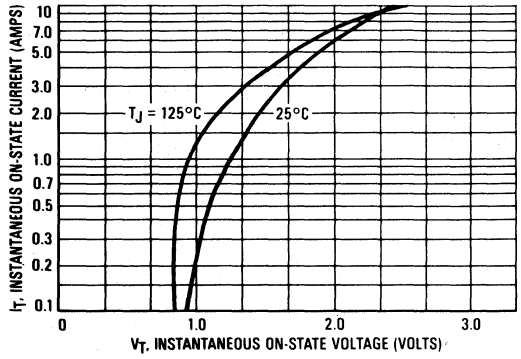
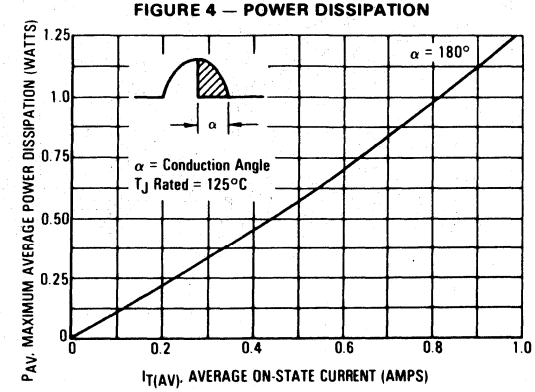


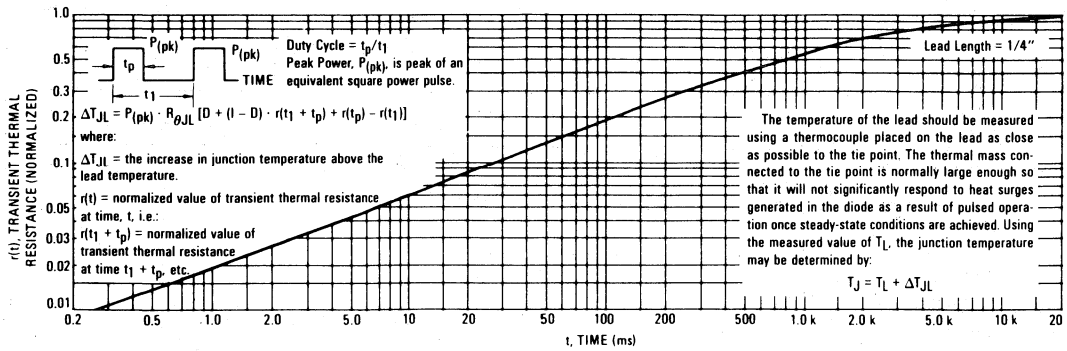
FIGURE 4 — POWER DISSIPATION



3

THERMAL CHARACTERISTICS

FIGURE 5 — THERMAL RESPONSE



TYPICAL CHARACTERISTICS

FIGURE 6 — BREAKOVER VOLTAGE

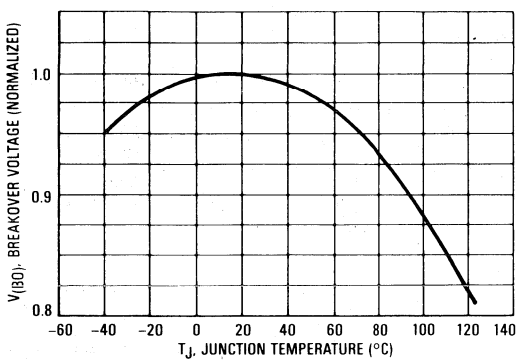


FIGURE 7 — HOLDING CURRENT

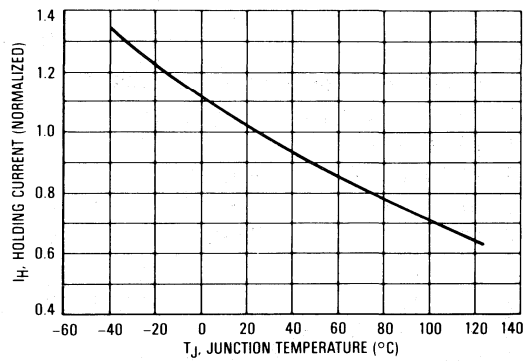


FIGURE 8 — PULSE RATING CURVE

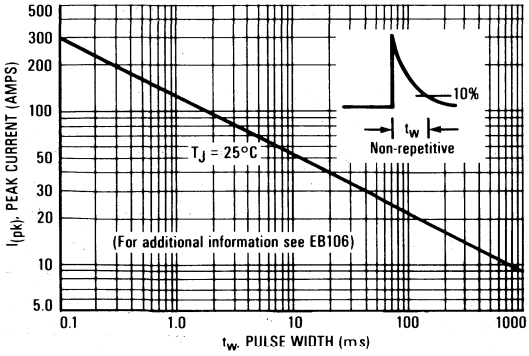
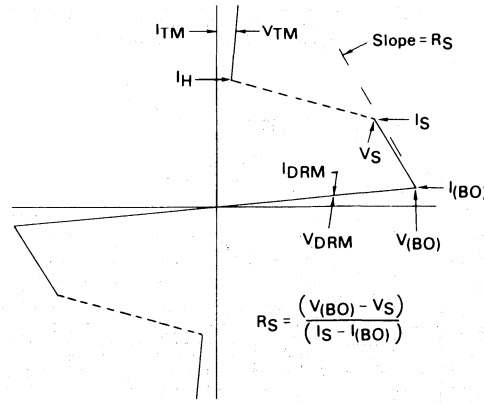


FIGURE 9 — V-1 CHARACTERISTICS



Sidac High Voltage Bilateral Triggers

MKP9V120
MKP9V130
MKP9V240
MKP9V260
MKP9V270

SIDACs
0.9 AMPERES RMS
110 thru 280 VOLTS

. . . designed for direct interface with the ac power line. Upon reaching the breakover voltage in each direction, the device switches from a blocking state to a low voltage on-state. Conduction will continue like an SCR until the main terminal current drops below the holding current. The plastic axial lead package provides high pulse current capability at low cost. Glass passivation insures reliable operation. Applications are:

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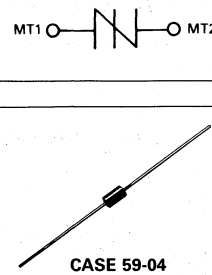
- High Voltage Regulators
- Pulse Generators



CASE 59-04
(DO-41)
Polarity denoted by cathode band

3

MAXIMUM RATINGS

Rating	Symbol	MKP9V120 MKP9V130	MKP9V240 MKP9V260 MKP9V270	Unit
Off-State Repetitive Voltage	V_{DRM}	±90	±180	Volts
On-State Current RMS (T_L = 80°C, LL = 3/8", conduction angle = 180°, 60 Hz Sine Wave)	$I_{T(RMS)}$	0.9		Amp
On-State Surge Current (Nonrepetitive) (60 Hz One Cycle Sine Wave, Peak Value)	I_{TSM}	4		Amps
Operating Junction Temperature Range	T_J	−40 to +125		°C
Storage Temperature Range	T_{stg}	−40 to +150		°C
Lead Solder Temperature (Lead Length ≥ 1/16" from case, 10 s max)	—	230		°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Unit	
Thermal Resistance, Junction to Lead LL = 3/8"	$R_{\theta JL}$	40	°C/W

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted; both directions.)

Characteristic		Symbol	Min	Typ	Max	Unit
Breakover Voltage		V_{BO}				Volts
	MKP9V120		110	—	125	
	MKP9V130		120	—	135	
	MKP9V240		220	—	250	
	MKP9V260		240	—	270	
	MKP9V270		250	—	280	
Repetitive Peak Off-State Current (60 Hz Sine Wave, V = V_{DRM})		I_{DRM}	—	—	5	μA
	T_J = 125°C		—	—	50	
Forward "On" Voltage (I_T = 1 A)		V_{TH}	—	1.3	1.5	Volts
Dynamic Holding Current (60 Hz Sine Wave)		I_H	—	—	100	mA
Switching Resistance		R_S	0.1	—	—	kΩ
Breakover Current (60 Hz Sine Wave)		I_{BO}			200	μA
Maximum Rate of Change of On-State Current	MKP9V120, 130	di/dt	—	90	—	A/μs
	MKP9V240, 260, 270		—	50	—	

FIGURE 1 — MAXIMUM LEAD TEMPERATURE

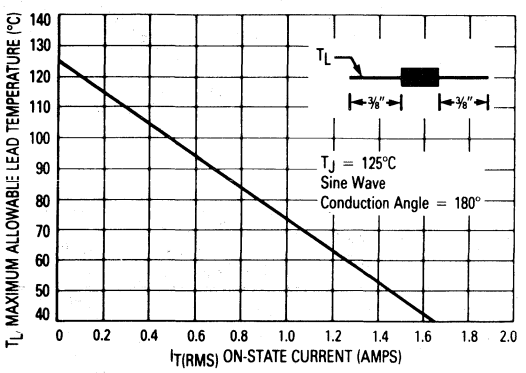


FIGURE 2 — MAXIMUM AMBIENT TEMPERATURE

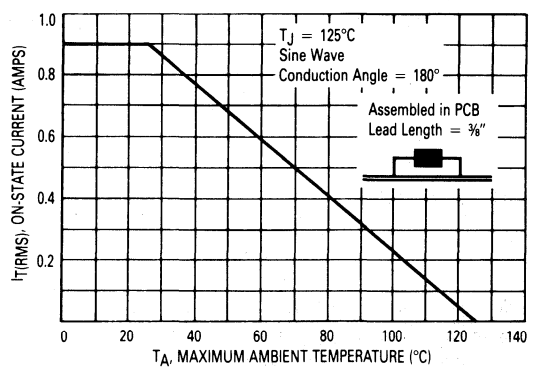


FIGURE 3 — TYPICAL ON-STATE VOLTAGE

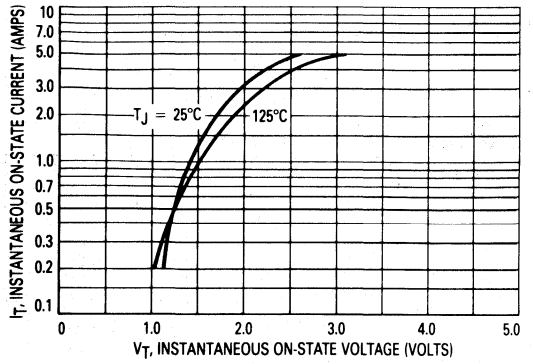
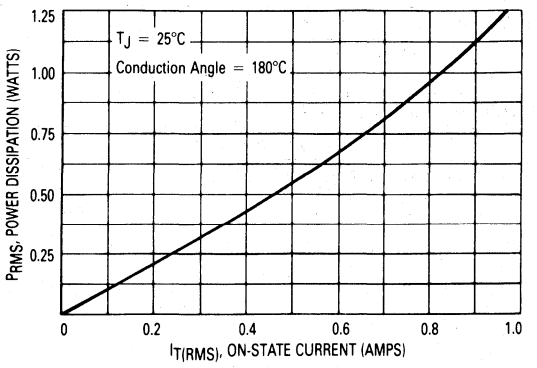
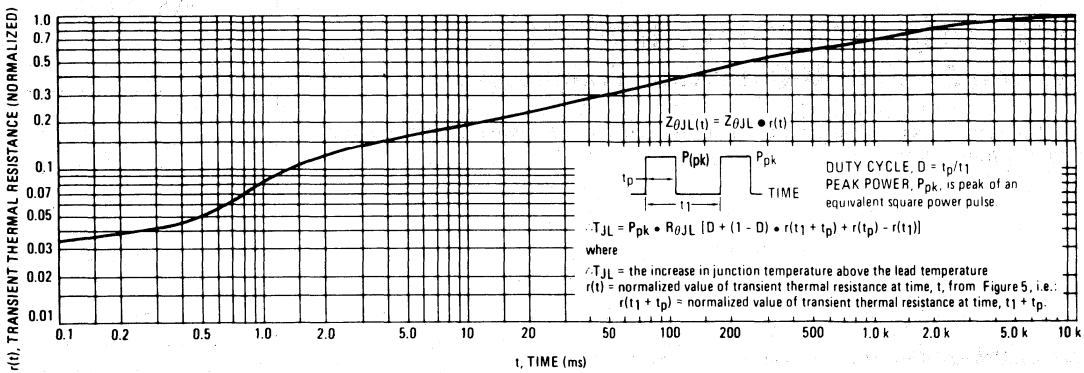


FIGURE 4 — POWER DISSIPATION



THERMAL CHARACTERISTICS

FIGURE 5 — THERMAL RESPONSE



TYPICAL CHARACTERISTICS

FIGURE 6 — BREAKOVER VOLTAGE

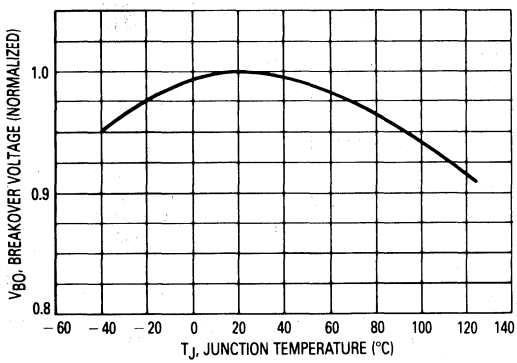


FIGURE 7 — HOLDING CURRENT

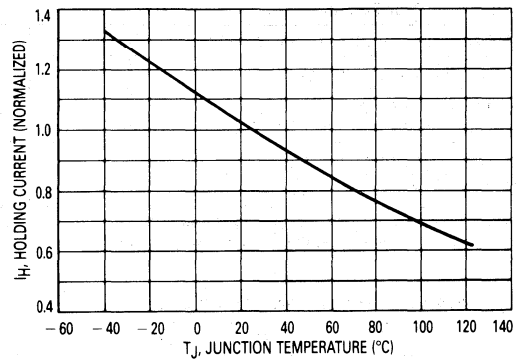


FIGURE 8 — PULSE RATING CURVE

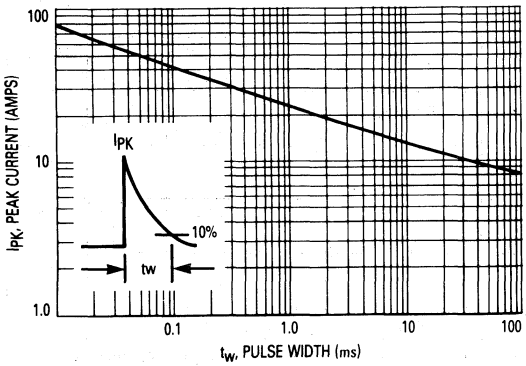
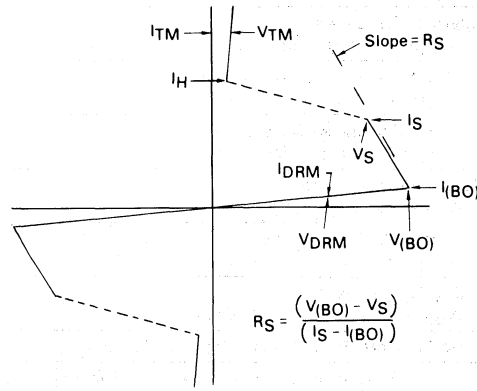


FIGURE 9 — V-I CHARACTERISTICS



3

NOT RECOMMENDED FOR NEW DESIGNS

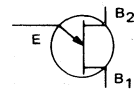
PN Unijunction Transistors

Silicon Annular Unijunction Transistors

. . . designed for economical, general purpose use in pulse, timing, oscillator and thyristor trigger circuits.

MU10

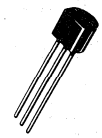
PN UJTs



MAXIMUM RATINGS (T_A = 25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
RMS Power Dissipation, Note 1	P_D	300	mW
RMS Emitter Current	I_e	50	mA
Peak-Pulse Emitter Current, Note 2	i_e	1	Amp
Emitter Reverse Voltage	V_{EB2}	30	Volts
Interbase Voltage Based upon Power Dissipation at T_A = 25°C	V_{B2B1}	35	Volts
Operating Junction Temperature Range	T_J	−65 to +125	°C
Storage Temperature Range	T_{stg}	−65 to +150	°C

Notes: 1. Derate 3 mW/°C increase in ambient temperature.
2. Duty Cycle ≤ 1%, PRR = 10 PPS (See Figure 5).



CASE 29-04
STYLE 9
(TO-226AA)
MU10

3

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Intrinsic Standoff Ratio (Test Circuit Figure 4), Note 1 (V_{B2B1} = 10 V)	η	0.50	—	0.85	—
Interbase Resistance (V_{B2B1} = 3 V, I_E = 0)	r_{BB}	4	—	10	kΩ
Emitter Saturation Voltage, Note 2 (V_{B2B1} = 10 V, I_E = 50 mA)	$V_{EB1(sat)}$	—	2	—	Volts
Modulated Interbase Current (V_{B2B1} = 10 V, I_E = 50 mA)	$I_{B2(mod)}$	—	—	50	mA
Emitter Reverse Current (V_{EB2} = 30 V, I_{B1} = 0)	I_{EB20}	—	—	1	μA
Peak-Point Emitter Current (V_{B2B1} = 25 V)	I_P	—	—	5	μA
Valley-Point Current** (V_{B2B1} = 20 V, R_{B2} = 100 Ohms)	I_V	1	—	—	mA
Base-One Peak Pulse Voltage (Test Circuit Figure 3)	V_{OB1}	3	—	—	Volts

Notes: 1. η, intrinsic standoff ratio, is defined in terms of the peak-point voltage, V_P, by means of the equation: $V_P = \eta V_{B2B1} + V_F$, where V_F is about 0.45 volt at 25°C @ I_F = 10 μA and decreases with temperature at about 2.5 mV/°C. The test circuit is shown in Figure 4. Components R_1, C_1, and the UJT form a relaxation oscillator; the remaining circuitry serves as a peak-voltage detector. The forward drop of Diode D_1 compensates for V_F. To use, the "cal" button is pushed, and R_3 is adjusted to make the current meter, M_1, read full scale. When the "cal" button is released, the value of η is read directly from the meter, if full scale on the meter reads 1.

2. Pulse Test: Pulse Width ≈ 300 μs, Duty Cycle ≤ 2% to avoid internal heating, which may result in erroneous readings.

FIGURE 1 – UNIJUNCTION TRANSISTOR SYMBOL AND NOMENCLATURE

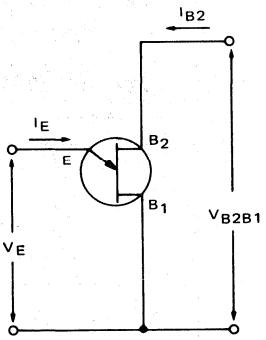


FIGURE 2 – STATIC EMITTER CHARACTERISTICS CURVES

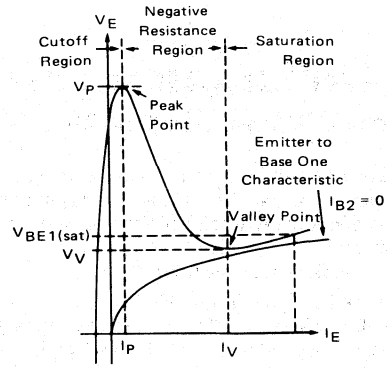


FIGURE 3 – V_{OB1} TEST CIRCUIT

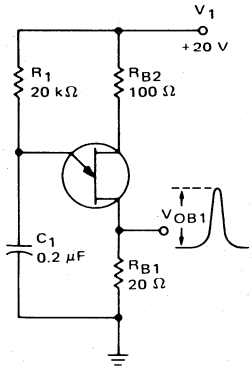
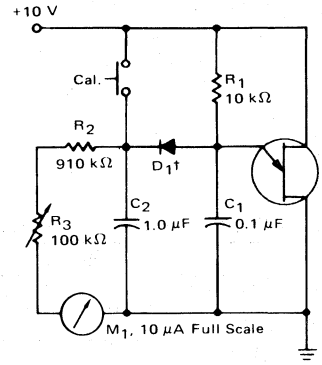
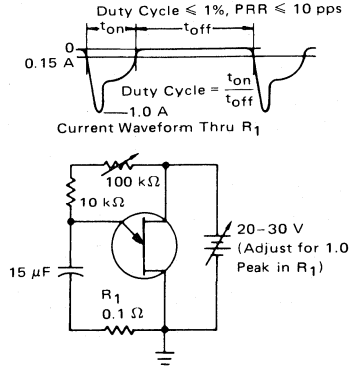


FIGURE 4 – η TEST CIRCUIT



† D_1 diode with the following characteristics:
$V_F = 0.45$ V @ $I_F = 10$ μA
$I_R \leq 2.0$ μA @ $V_R = 20$ V

FIGURE 5 – PRR TEST CIRCUIT AND WAVEFORM



NOT RECOMMENDED FOR NEW DESIGNS

MU4891 thru MU4894

PN Unijunction Transistors

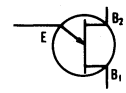
Silicon Plastic Unijunction Transistors

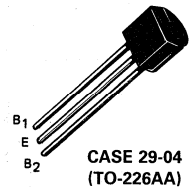
PN UJTs

. . . designed for military and industrial use in pulse, timing, triggering, sensing, and oscillator circuits. The annular process provides low leakage current, fast switching and low peak-point currents as well as outstanding reliability and uniformity.

Recommended usage includes:

- Long-time Delay Circuits — MU4894
- Silicon Controlled Rectifier Triggering Circuits — MU4893
- High-frequency Relaxation-Oscillator Circuits — MU4892
- General-Purpose Unijunction Applications — MU4891





CASE 29-04
(TO-226AA)
STYLE 9

3

MAXIMUM RATINGS (T_A = 25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
RMS Power Dissipation, Note 1	P_D	300	mW
RMS Emitter Current	I_e	50	mA
Peak Pulse Emitter Current, Note 2	i_e	1	Amp
Emitter Reverse Voltage	V_{B2E}	30	Volts
Storage Temperature Range	T_{stg}	−65 to +150	°C

Notes: 1. Derate 3 mW/°C increase in ambient temperature. Total power dissipation (available power to Emitter and Base-Two) must be limited by external circuitry. Interbase voltage (V_{B2B1}) limited by power dissipation,

$V_{B2B1} = \sqrt{R_{BB} \cdot P_D}$

2. Capacitance discharge current must fall to 0.37 Amp within 3 ms and PRR ≤ 10 PPS.

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
Intrinsic Standoff Ratio (V_{B2B1} = 10 V), Note 1	MU4892	η	0.51	—	0.69	—
	MU4891, MU4893		0.55	—	0.82	
	MU4894		0.74	—	0.86	
Interbase Resistance (V_{B2B1} = 3 V, I_E = 0)	MU4891, MU4892	R_{BB}	4	7	9.1	k ohms
	MU4893, MU4894		4	7	12	
Interbase Resistance Temperature Coefficient (V_{B2B1} = 3 V, I_E = 0, T_A = −65°C to +100°C)		αR_{BB}	0.1	—	0.9	%/°C
Emitter Saturation Voltage (V_{B2B1} = 10 V, I_E = 50 mA), Note 2		$V_{EB1(sat)}$	—	2.5	4	Volts
Modulated Interbase Current (V_{B2B1} = 10 V, I_E = 50 mA)		$I_{B2(mod)}$	10	15	—	mA
Emitter Reverse Current (V_{B2E} = 30 V, I_{B1} = 0)		I_{EB2O}	—	5	10	nA
Peak Point Emitter Current (V_{B2B1} = 25 V)	MU4891	I_P	—	0.6	5	μA
	MU4892, MU4893		—	0.6	2	
	MU4894		—	0.6	1	
Valley Point Current (V_{B2B1} = 20 V, R_{B2} = 100 Ohms), Note 2	MU4891, MU4893, MU4894	I_V	2	4	—	mA
	MU4892		2	3	—	
Base-One Peak Pulse Voltage (Note 3, Figure 3)	MU4891, MU4892, MU4894	V_{OB1}	3	5	—	Volts
	MU4893		6	8	—	

Notes:

1. Intrinsic standoff ratio, η, is defined by equation:

$$\eta = \frac{V_P - V_{(EB1)}}{V_{B2B1}}$$

Where V_P = Peak Point Emitter Voltage
V_{B2B1} = Interbase Voltage
$V_{(EB1)}$ = Emitter to Base-One Junction Diode Drop (≈0.5 V @ 10 μA)

2. Use pulse techniques: PW ≈ 300 μs, duty cycle ≤ 2% to avoid internal heating due to interbase modulation which may result in erroneous readings.
3. Base-One Peak Pulse Voltage is measured in circuit of Figure 3. This specification is used to ensure minimum pulse amplitude for applications in SCR firing circuits and other types of pulse circuits.

3

FIGURE 1 — UNIJUNCTION TRANSISTOR SYMBOL AND NOMENCLATURE

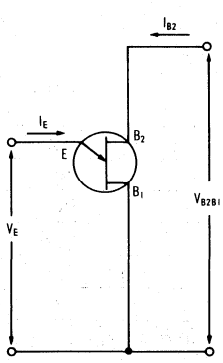


FIGURE 2 — STATIC EMITTER CHARACTERISTICS CURVES

(Exaggerated to Show Details)

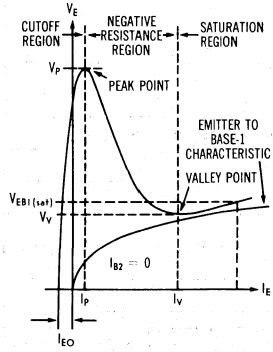
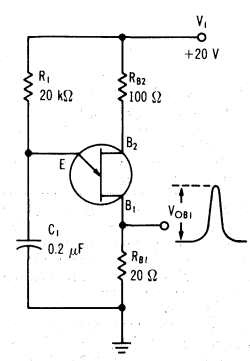


FIGURE 3 — V_{OB1} TEST CIRCUIT

(Typical Relaxation Oscillator)



Silicon Controlled Rectifiers

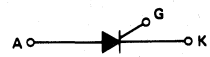
Reverse Blocking Triode Thyristors

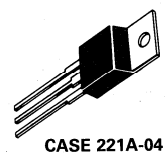
S2800 Series

SCRs
10 AMPERES RMS
50 thru 800 VOLTS

. . . designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies; or wherever half-wave silicon gate-controlled, solid-state devices are needed.

- Glass Passivated Junctions and Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts





CASE 221A-04
(TO-220AB)
STYLE 3

3

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Peak Repetitive Reverse Voltage, Note 1 Peak Repetitive Off-State Voltage		V_{RRM} V_{DRM}		Volts
	S2800 F		50	
	S2800 A		100	
	S2800 B		200	
	S2800 D		400	
	S2800 M		600	
	S2800 N		800	
Non-Repetitive Peak Reverse Voltage Non-Repetitive Off-State Voltage		V_{RSM} V_{DSM}		Volts
	S2800 F		75	
	S2800 A		125	
	S2800 B		250	
	S2800 D		500	
	S2800 M		700	
	S2800 N		900	
RMS Forward Current (All Conduction Angles)	T_C = 75°C	$I_{T(RMS)}$	10	Amps
Peak Forward Surge Current (1 Cycle, Sine Wave, 60 Hz, T_C = 80°C)		I_{TSM}	100	Amps
Circuit Fusing Considerations (t = 8.3 ms)		I^2t	40	A^2s
Forward Peak Gate Power (t ≤ 10 μs)		P_{GM}	16	Watts
Forward Average Gate Power		$P_{G(AV)}$	0.5	Watt
Operating Junction Temperature Range		T_J	−40 to +100	°C
Storage Temperature Range		T_{stg}	−40 to +150	°C

Note 1. V_{DRM} and V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices shall not have a positive bias applied to the gate concurrently with a negative potential on the anode.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}) T_C = 25°C T_C = 100°C	I_{DRM}, I_{RRM}	— —	— —	10 2	μA mA
Instantaneous On-State Voltage (I_{TM} = 30 A Peak, Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)	V_T	—	1.7	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 30 Ohms)	I_{GT}	—	8	15	mA
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 30 Ohms)	V_{GT}	—	0.9	1.5	Volts
Holding Current (Gate Open, V_D = 12 Vdc, I_T = 150 mA)	I_H	—	10	20	mA
Gate Controlled Turn-on Time (V_D = Rated V_{DRM}, I_{TM} = 2 A, I_{GR} = 80 mA)	t_{gt}	—	1.6	—	μs
Circuit Commutated Turn-Off Time (V_D = V_{DRM}, I_{TM} = 2 A, Pulse Width = 50 μs, dv/dt = 200 V/μs, di/dt = 10 A/μs, T_C = 75°C)	t_q	—	25	—	μs
Critical Rate-of-Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Rise, T_C = 100°C)	dv/dt	—	100	—	V/μs

3

FIGURE 1 – CURRENT DERATING

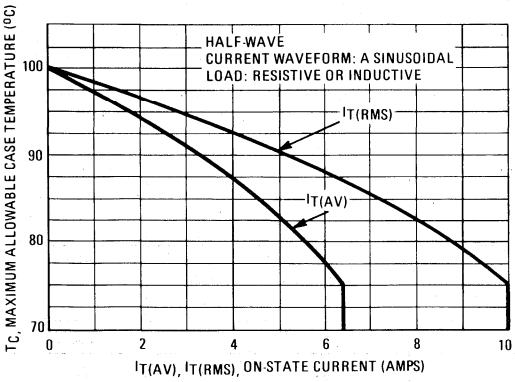
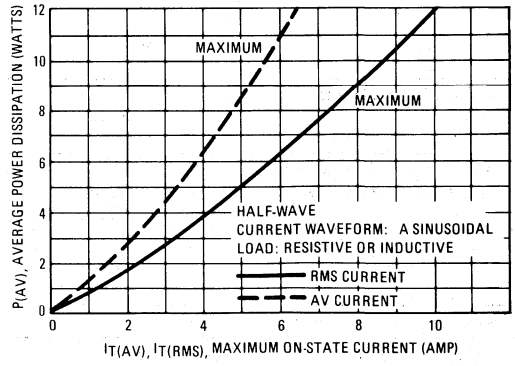


FIGURE 2 – POWER DISSIPATION



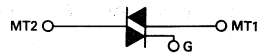
Triacs
Bidirectional Triode Thyristors

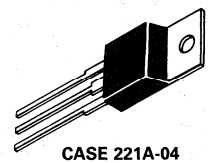
SC141
SC146

TRIACs
6 and 10 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Triggering Specified in Three Quadrants
- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Gate Open	V_{DRM}		Volts
B		200	
SC141 D		400	
SC146 M		600	
N		800	
RMS On-State Current (T_C = 80°C) SC141	$I_{T(RMS)}$	6	Amps
SC146		10	
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz) SC141	I_{TSM}	80	Amps
SC146		120	
Circuit Fusing Considerations (t = 8.3 ms) SC141	I^2t	26.5	A^2s
SC146		60	
Peak Gate Power (Pulse Width = 10 μs)	P_{GM}	10	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current (Pulse Width = 10 μs)	I_{GM}	3.5	Amps
Peak Gate Voltage	V_{GM}	10	Volts
Operating Junction Temperature Range	T_J	−40 to +125	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$		°C/W
SC141		2.2	
SC146		1.5	

ELECTRICAL CHARACTERISTICS (T_C = +25°C, Either Polarity of MT2 to MT1 Voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current	I_{DRM}, I_{RRM}				
(Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C		—	—	10	μA
T_C = +100°C		—	—	0.5	mA
Peak On-State Voltage	V_{TM}				Volts
(Pulse Width ≤ 1 ms, Duty Cycle ≤ 2%)					
SC141 I_{TM} = 8.5 A Peak		—	—	1.83	
SC146 I_{TM} = 14 A Peak		—	—	1.65	
Critical Rate of Rise of Off-State Voltage	dv/dt	—	50	—	V/μs
(V_D = Rated V_{DRM}, Gate Open-Circuited, Exponential Waveform T_C = +100°C					
Critical Rate-of-Rise of Commutating Off-State Voltage (1)	dv/dt(c)				V/μs
($I_{T(RMS)}$ = Rated $I_{T(RMS)}$, V_D = Rated V_{DRM}, Gate Open-Circuited T_C = +80°C					
SC141 Commutating di/dt = 3.2 A/ms		4	—	—	
SC146 Commutating di/dt = 5.4 A/ms		4	—	—	
DC Gate Trigger Current (Continuous dc)	I_{GT}				mAdc
(V_D = 12 Vdc, Trigger Mode)					
MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms		—	—	50	
MT2(+), G(−); R_L = 50 Ohms		—	—	50	
MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms T_C = −40°C		—	—	80	
MT2(+), G(−); R_L = 25 Ohms T_C = −40°C		—	—	80	
DC Gate Trigger Voltage (Continuous dc)	V_{GT}				Vdc
(V_D = 12 Vdc, Trigger Mode)					
MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms		—	—	2.5	
MT2(+), G(−); R_L = 50 Ohms		—	—	2.5	
MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms T_C = −40°C		—	—	3.5	
MT2(+), G(−); R_L = 25 Ohms T_C = −40°C		—	—	3.5	
(V_D = Rated V_{DRM}; R_L = 1000 Ohms) All Polarities					
T_C = +100°C		0.2	—	—	
Holding Current	I_H				mAdc
(V_D = 24 Vdc, I_T = 0.5 A)					
(Pulse Width = 1 ms, Duty Cycle ≤ 2%)					
(Gate Trigger Source = 7 V, 20 Ohms) T_C = +25°C		—	—	50	
T_C = −40°C		—	—	100	
Latching Current	I_L				mAdc
(V_D = 24 Vdc)					
(Gate Trigger Source = 15 V, 100 Ohms, Trigger Mode)					
MT2(+), G(+); MT2(−), G(−)		—	—	100	
MT2(+), G(−)		—	—	200	
MT2(+), G(+); MT2(−), G(−) T_C = −40°C		—	—	200	
MT2(+), G(−) T_C = −40°C		—	—	400	

3

FIGURE 1 – RMS CURRENT DERATING

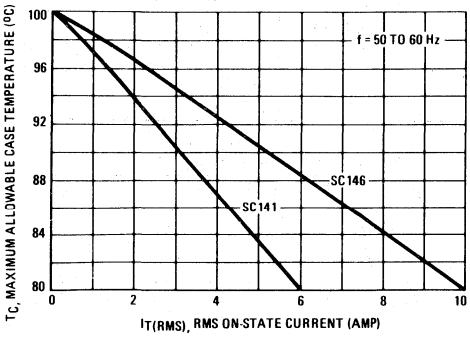
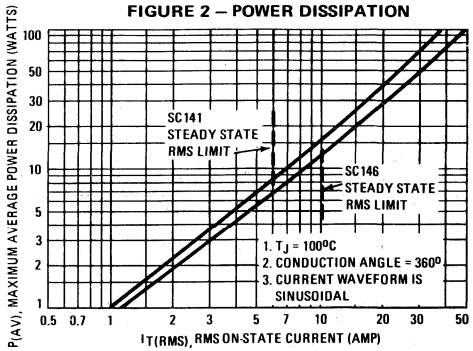


FIGURE 2 – POWER DISSIPATION



3

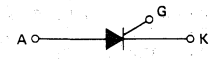
Triacs
Silicon Bidirectional Thyristors

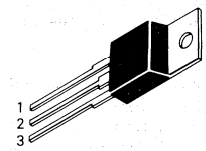
SC143 Series
SC149 Series

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Blocking Voltage to 600 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability

TRIACs
8 and 12 AMPERES RMS
200 thru 600 VOLTS





CASE 221A-04
TO-220AB

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Gate Open B SC143 D SC149 E M	V_{DRM}	 200 400 500 600	Volts
Peak Gate Voltage	V_{GM}	—	Volts
RMS On-State Current T_C = 80°C SC143 SC149	$I_{T(RMS)}$	 8 12	Amps
Peak Non-Repetitive Surge Current One Full Cycle, 60 Hz SC143 SC149	I_{TSM}	 120 120	Amps
Circuit Fusing Considerations t = 1 ms SC143 SC149	I^2t	 20 25	A^2s
Critical Rate of Rise of On-State Current	di/dt	10	A/μs
Peak Gate Power (Pulse Width = 10 μs)	P_{GM}	10	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.5	Watt
Peak Gate Current (Pulse Width = 10 μs)	I_{GM}	3.5	Amps
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

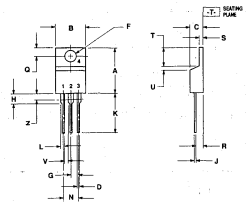
ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Off-State Current (1) Rated V_{DRM} = Peak Off-State Voltage, Gate Open-Circuited T_C = +25°C T_C = +100°C	I_{DRM}	 — —	 — —	 0.1 0.5	mA
Peak On-State Voltage (1) Pulse Width = 1 ms, Duty Cycle < 2%. I_{TM} = 11 A Peak I_{TM} = 17 A Peak	V_{TM}	 — —	 — —	 1.55 1.65	Volts

NOTE 1. Values apply for either polarity of Main Terminal 2 characteristics referenced to Main Terminal 1. (cont.)

Thermowatt is a trademark of Motorola Inc.

OUTLINE DIMENSIONS



STYLE 2:
PIN 1. BASE 3. COLLECTOR
2. EMITTER 4. EMITTER

DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	14.48	15.75	0.570	0.620
B	9.66	10.28	0.380	0.405
C	4.07	4.82	0.160	0.190
D	0.64	0.88	0.025	0.035
F	3.61	3.73	0.142	0.147
G	2.42	2.66	0.095	0.105
H	2.80	3.93	0.110	0.155
J	0.36	0.55	0.014	0.022
K	12.70	14.27	0.500	0.562
L	1.15	1.39	0.045	0.055
N	4.83	5.33	0.190	0.210
Q	2.54	3.04	0.100	0.120
R	2.04	2.79	0.080	0.110
S	1.15	1.39	0.045	0.055
T	5.97	6.47	0.235	0.255
U	0.00	1.27	0.000	0.050
V	1.15	—	0.045	—
Z	—	2.04	—	0.080

CASE 221A-04
TO-220AB

3

ELECTRICAL CHARACTERISTICS — continued (T_C = +25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Critical Rate of Rise of Off-State Voltage (1)	dv/dt				Volts/μs
Rated V_{DRM}, Gate Open-Circuited, SC143 Series		—	150	—	
Exponential Waveform T_C = +100°C SC149 Series		—	200	—	
Critical Rate-of-Rise of Commutating Off-State Voltage (1)	dv/dt(C)				Volts/μs
$I_{T(RMS)}$ = Rated RMS On-State Current					
V_{DRM} = Rated Peak Off-State Voltage,					
Gate Open-Circuited					
SC143 Commutating di/dt = 4.3 A/ms		4	—	—	
SC149 Commutating di/dt = 6.5 A/ms		4	—	—	
T_C = +80°C					
DC Gate Trigger Current (2)	I_{GT}				mAdc
V_D = 12 Vdc					
Trigger Mode					
MT2 (+), Gate (+), R_L = 100 Ohms		—	—	50	
MT2 (−), Gate (−), R_L = 100 Ohms		—	—	50	
MT2 (+), Gate (−), R_L = 50 Ohms		—	—	50	
MT2 (+), Gate (+), R_L = 50 Ohms, T_C = −40°C		—	—	80	
MT2 (−), Gate (−), R_L = 50 Ohms, T_C = −40°C		—	—	80	
MT2 (+), Gate (−), R_L = 25 Ohms, T_C = −40°C		—	—	80	
DC Gate Trigger Voltage (2)	V_{GT}				Vdc
V_D = 12 Vdc					
Trigger Mode					
MT2 (+), Gate (+), R_L = 100 Ohms		—	—	2.5	
MT2 (−), Gate (−), R_L = 100 Ohms		—	—	2.5	
MT2 (+), Gate (−), R_L = 50 Ohms		—	—	2.5	
MT2 (+), Gate (+), R_L = 50 Ohms, T_C = −40°C		—	—	3.5	
MT2 (−), Gate (−), R_L = 50 Ohms, T_C = −40°C		—	—	3.5	
MT2 (+), Gate (−), R_L = 25 Ohms, T_C = −40°C		—	—	3.5	
MT2 (+), Gate (+), R_L = 1000 Ohms, T_C = +100°C (2,3)		0.20	—	—	
MT2 (−), Gate (−), R_L = 1000 Ohms, T_C = +100°C (2,3)		0.20	—	—	
MT2 (+), Gate (−), R_L = 1000 Ohms, T_C = +100°C (2,3)		0.20	—	—	
MT2 (−), Gate (+), R_L = 1000 Ohms, T_C = +100°C (2,3)		0.20	—	—	
Holding Current (1)	I_H				mAdc
Main Terminal Voltage = 24 Vdc,					
Peak Initiating Current = 0.5 A,					
Pulse Width = 1 ms, Duty Cycle ≤ 2%.					
Gate Trigger Source = 7 V, 20 Ohms					
T_C = +25°C		—	—	50	
T_C = −40°C		—	—	100	
Latching Current (2)	I_L				mAdc
Main Terminal Source Voltage = 24 Vdc,					
Gate Trigger Source = 15 V, 100 Ohms,					
Trigger Mode					
MT2 (+), Gate (+)		—	—	100	
MT2 (−), Gate (−)		—	—	100	
MT2 (+), Gate (−)		—	—	200	
MT2 (+), Gate (+), T_C = −40°C		—	—	200	
MT2 (−), Gate (−), T_C = −40°C		—	—	200	
MT2 (+), Gate (−), T_C = −40°C		—	—	400	

NOTES:
1. Values apply for either polarity of Main Terminal 2 characteristics referenced to Main Terminal 1.
2. Main Terminal 1 is the reference terminal.
3. With V_D equal to rated off-state voltage.

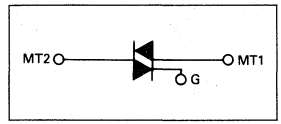
Triacs
Bidirectional Triode Thyristors

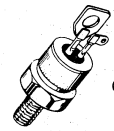
SC260
SC260( )3
MAC261

TRIACs
25 AMPERES RMS
200 thru 600 VOLTS

. . . designed primarily for industrial and military applications for the control of ac loads in applications such as light dimmers, power supplies, heating controls, motor controls, welding equipment and power switching systems; or wherever full-wave, silicon gate controlled solid-state devices are needed.

- All Diffused and Glass Passivated Junctions for Greater Stability
- Pressfit, Stud and Isolated Stud Packages
- Gate Triggering Guaranteed In All 3 Quadrants





CASE 263-04
STYLE 2
SC260

CASE 174-04
(TO-203AA)
STYLE 3
MAC261

CASE 311-02
STYLE 2
SC260( )3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage (T_C = −40°C to +115°C) SC260B, SC260B3, MAC261B SC260D, SC260D3, MAC261D SC260M, SC260M3, MAC261M	V_{DRM}	 200 400 600	Volts
RMS On-State Current	$I_{T(RMS)}$	25	Amps
Peak Non-Repetitive Surge Current (One Cycle, 60 Hz)	I_{TSM}	250	Amps
Circuit Fusing Considerations t = 1 ms t = 8.3 ms	I^2t	 150 260	A^2s
Peak Gate Power (Pulse Width − 10 μs)	P_{GM}	10	Watts
Average Gate Power	$P_{G(AV)}$	0.5	Watt
Peak Gate Power	I_{GM}	2	Amps
Operating Junction Temperature Range	T_J	−40 to +115	°C
Storage Temperature Range	T_{stg}	−40 to +125	°C
Stud Torque	—	30	in. lb.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case SC260, MAC261 SC260( )3	$R_{\theta JC}$	 1.8 1.95	°C/W

3

ELECTRICAL CHARACTERISTICS (T_C = +25°C unless otherwise noted. Values apply for either polarity of Main Terminal 2 Characteristics referenced to Main Terminal 1.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = +115°C		—	—	1	mA
Peak On-State Voltage (I_{TM} = 35 A Peak, Pulse Width = 1 ms, Duty Cycle ≤ 2%)	V_{TM}	—	—	1.58	Volts
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Gate Open-Circuited, Exponential Waveform) T_C = +115°C	dv/dt	50	—	—	V/μs
Critical Rate-of-Rise of Commutating Off-State Voltage ($I_{T(RMS)}$ = Rated RMS On-State Current) (V_{DRM} = Rated Peak Off-State Voltage, Gate Open-Circuited, Commutating di/dt = 13.5 A/ms) T_C = +80°C	dv/dt(c)	5	—	—	V/μs
DC Gate Trigger Current (Continuous dc) (V_D = 12 Vdc) MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms	I_{GT}	—	—	50	mAdc
MT2(+), G(−); R_L = 50 Ohms		—	—	50	
DC Gate Trigger Current (Continuous dc) (V_D = 12 Vdc) T_C = −40°C MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms	I_{GT}	—	—	80	mAdc
MT2(+), G(−); R_L = 25 Ohms		—	—	80	
DC Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc) MT2(+), G(+); MT2(−), G(−); R_L = 100 Ohms	V_{GT}	—	—	2.5	Vdc
MT2(+), G(−); R_L = 50 Ohms		—	—	2.5	
DC Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc) T_C = −40°C MT2(+), G(+); MT2(−), G(−); R_L = 50 Ohms	V_{GT}	—	—	3.5	Vdc
MT2(+), G(−); R_L = 25 Ohms		—	—	3.5	
DC Gate Non-Trigger Voltage (V_D = Rated V_{DRM}, R_L = 1K Ohms, All Trigger Modes) T_C = 115°C	V_{GD}	0.25	—	—	Vdc
Holding Current (V_D = 24 Vdc, Peak Initiating Current = 0.5 A, Pulse Width = 0.1 to 10 ms, Gate Trigger Source = 7 V, 20 Ohms) T_C = +25°C	I_H	—	—	75	mAdc
T_C = −40°C		—	—	100	
Latching Current (V_D = 24 Vdc, Gate Trigger Source = 15 V, 100 Ohms, Pulse Width = 50 μs, 5 μs Maximum Rise and Fall Times) MT2(+), G(+); MT2(−), G(−) T_C = 25°C	I_L	—	—	100	mAdc
MT2(+), G(−) T_C = 25°C		—	—	200	
MT2(+), G(+); MT2(−), G(−) T_C = −40°C		—	—	200	
MT2(+), G(−) T_C = −40°C		—	—	400	

FIGURE 1 — CURRENT DERATING

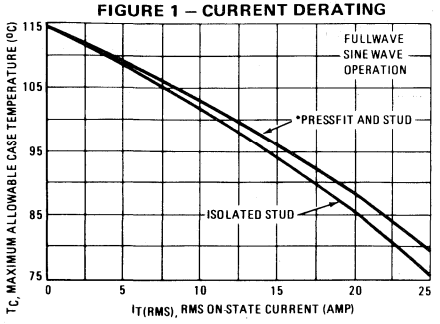
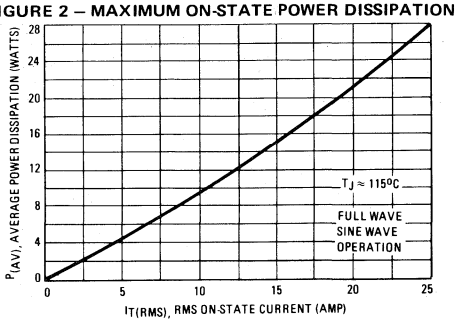


FIGURE 2 — MAXIMUM ON-STATE POWER DISSIPATION



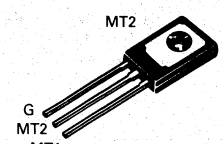
Sensitive Gate Triacs
Silicon Bidirectional Triode Thyristors

T2322
T2323
Series

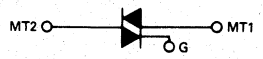
. . . designed primarily for ac power switching. The gate sensitivity of these triacs permits the use of economical transistorized or integrated circuit control circuits, and it enhances their use in low-power phase control and load-switching applications.

- Very High Gate Sensitivity
- Low On-State Voltage at High Current Levels
- Glass-Passivated Chip for Stability
- Small, Rugged Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability

SENSITIVE GATE TRIACs
2.5 AMPERES RMS
200 thru 600 VOLTS



CASE 77-07
(TO-225AA)
STYLE 5



MAXIMUM RATINGS (Apply for T_J = −40 to 100°C unless otherwise noted.)

Rating	Suffix	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 T2322, T2323	B D M	V_{DRM}	200 400 600	Volts
RMS On-State Current (T_C = 70°C) (Full-cycle sine wave 50 to 60 Hz)		$I_{T(RMS)}$	2.5	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)		I_{TSM}	25	Amps
Circuit Fusing (t = 8.3 ms)		I^2t	2.6	A^2s
Peak Gate Power (1 μs)		P_{GM}	10	Watts
Average Gate Power (T_C = 60°C + 38.3 ms)		$P_{G(AV)}$	0.15	Watt
Peak Gate Current (1 μs)		I_{GM}	0.5	Amp
Operating Junction Temperature Range		T_J	−40 to +110	°C
Storage Temperature Range		T_{stg}	−40 to +150	°C
Mounting Torque (6-32 Screw), Note 2		—	8	in. lb.

Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

2. Torque rating applies with use of torque washer (Shakeproof WD19523 or equivalent). Mounting Torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Main terminal 2 and heat-sink contact pad are common.

For soldering purposes (either terminal connection or device mounting), soldering temperatures shall not exceed +200°C, for 10 seconds. Consult factory for lead bending options.

3

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3.5	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	60	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C and either polarity of MT2 to MT1 voltage unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_J = 100°C		—	0.2	0.75	mA
Peak On-State Voltage (I_{TM} = 10 A) T2323 Series	V_{TM}	—	1.7	2.6	Volts
T2322 Series		—	1.7	2.2	
Gate Trigger Current (Continuous dc) (All Modes) (V_D = 12 V, R_L = 30 Ω)	I_{GT}	—	—	10	mA
T2322 Series		—	—	25	
MT2(+), G(+); MT2(−), G(−) T2323 Series		—	—	40	
MT2(+), G(−); MT2(−), G(+) T2323 Series		—	—	—	
Gate Trigger Voltage (Continuous dc) (V_D = 12 Vdc, R_L = 30 Ω, T_C = 25°C)	V_{GT}	—	1	2.2	Volts
(V_D = V_{DROM}, R_L = 125 Ω, T_C = 100°C)		0.15	—	—	
Holding Current (V_D = 12 V, I_{TM} = 150 mA, Gate Open)	I_H	—	15	30	mA
Gate-Controlled Turn-On Time (V_D = Rated V_{DRM}, I_{TM} = 10 A pk, I_G = 60 mA)	t_{gt}	—	1.8	2.5	μs
Critical Rate of Rise of Off-State Voltage (V_D = Rated V_{DRM}, Exponential Waveform, T_C = 100°C)	dv/dt	10	100	—	V/μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, I_{TM} = 3.5 A pk, Commutating di/dt = 1.8 A/ms, Gate Unenergized, T_C = 90°C)	dv/dt(c)	1	4	—	V/μs

3

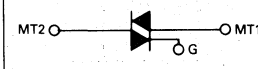
Triacs
Silicon Bidirectional Thyristors

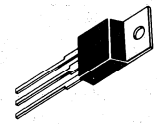
T2500 Series

TRIACs
6 AMPERES RMS
200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +100°C) Gate Open T2500 B D M N	V_{DROM}	 200 400 600 800	Volts
On-State Current RMS (T_C = +80°C) (Full Cycle Sine Wave 50 to 60 Hz)	$I_{T(RMS)}$	6	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_C = +80°C)	I_{TSM}	60	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	15	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 1 μs)	P_{GM}	16	Watts
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.2	Watt
Peak Gate Trigger Current (Pulse Width = 10 μs)	I_{GTM}	4	Amps
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

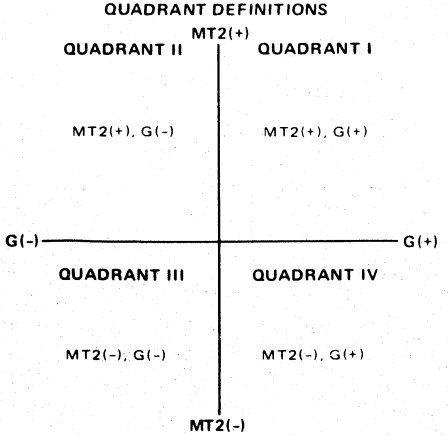
Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.7	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_J = 100°C	I_{DRM}, I_{RRM}	—	—	2	mA
Maximum On-State Voltage (Either Direction) (I_T = 30 A Peak)	V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 12 Ohms)	I_{GT}				mA
$V_{MT2(+)}$, $V_{G(+)}$		—	10	25	
$V_{MT2(+)}$, $V_{G(-)}$		—	20	60	
$V_{MT2(-)}$, $V_{G(-)}$		—	15	25	
$V_{MT2(-)}$, $V_{G(+)}$		—	30	60	
Gate Trigger Voltage (Continuous dc) (All Quadrants) (V_D = 12 Vdc, R_L = 12 Ohms)	V_{GT}	—	1.25	2.5	Volts
(V_D = V_{DROM}, R_L = 125 Ohms, T_C = 100°C)		0.2	—	—	
Holding Current (Either Direction) (Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 150 mA, T_C = 25°C)	I_{HO}	—	15	30	mA
Gate Controlled Turn-On Time (Rated V_{DROM}, I_T = 10 A, I_{GT} = 160 mA, Rise Time = 0.1 μs)	t_{gt}	—	1.6	—	μs
Critical Rate of Rise of Commutation Voltage (Rated V_{DROM}, $I_{T(RMS)}$ = 6 A, Commutating di/dt = 3.2 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	10	—	V/μs
Critical Rate of Rise of Off-State Voltage (Rated V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = 100°C)	dv/dt				V/μs
T2500B		—	100	—	
T2500D,M,N		—	75	—	

3

QUADRANT DEFINITIONS



ELECTRICAL CHARACTERISTICS of RECOMMENDED BIDIRECTIONAL SWITCHES

USAGE	General	
PART NUMBER	MBS4991	MBS4992
V_S	6.0 – 10 V	7.5 – 9.0 V
I_S	350 μA Max	120 μA Max
$V_{S1} - V_{S2}$	0.5 V Max	0.2 V Max
Temperature Coefficient	0.02%/°C Typ	

See AN-526 for Theory and Characteristics of Silicon Bidirectional Switches.

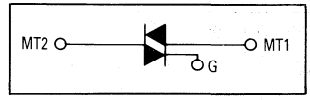
Silicon Bidirectional Triode Thyristors

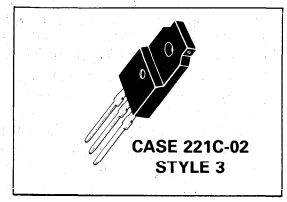
T2500FP Series

ISOLATED TRIACs THYRISTORS 6 AMPERES RMS 200 thru 800 VOLTS

. . . designed primarily for full-wave ac control applications, such as solid-state relays, motor controls, heating controls and power supplies; or wherever full-wave silicon gate controlled solid-state devices are needed. Triac type thyristors switch from a blocking to a conducting state for either polarity of applied anode voltage with positive or negative gate triggering.

- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Isolated Construction for Low Thermal Resistance, High Heat Dissipation and Durability





MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Repetitive Peak Off-State Voltage, Note 1 (T_J = −40 to +100°C) Gate Open	V_{DROM}		Volts
T2500BFP		200	
T2500DFP		400	
T2500MFP		600	
T2500NFP		800	
On-State RMS Current (T_C = +80°C), Note 2 Full Cycle Sine Wave 50 to 60 Hz	$I_{T(RMS)}$	6	Amps
Peak Nonrepetitive Surge Current (One Full Cycle, 60 Hz, T_C = +80°C)	I_{TSM}	100	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 1 μs)	P_{GM}	1	Watt
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.2	Watt
Peak Gate Trigger Current (Pulse Width = 10 μs)	I_{GTM}	4	Amps
RMS Isolation Voltage (T_A = 25°C, Relative Humidity ≤ 20%)	V_{ISO}	1500	Volts
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case, Note 2	$R_{\theta JC}$	2.7	°C/W
Case to Sink	$R_{\theta CS}$	2.2 (typ)	
Junction to Ambient	$R_{\theta JA}$	60	

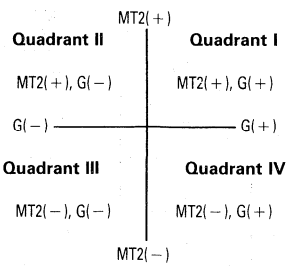
Notes: 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.
2. The case temperature reference point for all T_C measurements is a point on the center lead of the package as close as possible to the plastic body.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Off-State Current (Either Direction) Rated V_{DROM} @ T_J = 100°C, Gate Open	I_{DROM}	—	—	2	mA
Maximum On-State Voltage (Either Direction) I_T = 30 A Peak	V_{TM}	—	—	2	Volts
Gate Trigger Current (Continuous dc) V_D = 12 Vdc, R_L = 12 Ohms $V_{MT2(+)}$, $V_{G(+)}$ $V_{MT2(+)}$, $V_{G(-)}$ $V_{MT2(-)}$, $V_{G(-)}$ $V_{MT2(-)}$, $V_{G(+)}$	I_{GT}	 — — — —	 10 20 15 30	 25 60 25 60	mA
Gate Trigger Voltage (Continuous dc) (All Quadrants) V_D = 12 Vdc, R_L = 12 Ohms V_D = V_{DROM}, R_L = 125 Ohms, T_C = 100°C, All Trigger Modes	V_{GT}	 — 0.2	 1.25 —	 2.5 —	Volts
Holding Current (Either Direction) Main Terminal Voltage = 12 Vdc, Gate Open, Initiating Current = 150 mA, T_C = 25°C	I_{HO}	—	15	30	mA
Gate Controlled Turn-On Time Rated V_{DROM}, I_T = 10 A, I_{GT} = 160 mA, Rise Time = 0.1 μs	t_{gt}	—	1.6	—	μs
Critical Rate of Rise of Commutation Voltage Rated V_{DROM}, $I_{T(RMS)}$ = 6 A, Commutating di/dt = 3.2 A/ms, Gate Unenergized, T_C = 80°C	dv/dt(C)	—	10	—	V/μs
Critical Rate of Rise of Off-State Voltage Rated V_{DROM}, Exponential Voltage Rise, Gate Open, T_C = 100°C	dv/dt	—	100	—	V/μs

3

Quadrant Definitions



Trigger devices are recommended for gating on Triacs. They provide:
1. Consistent predictable turn-on points.
2. Simplified circuitry.
3. Fast turn-on time for cooler, more efficient and reliable operation.

Electrical Characteristics of Recommended Bidirectional Switches

Usage	General	
Part Number	**MBS4991**	**MBS4992**
V_S	6–10 V	7.5–9 V
I_S	350 μA Max	120 μA Max
V_{S1}–V_{S2}	0.5 V Max	0.2 V Max
Temperature Coefficient	0.02%/°C Typ	

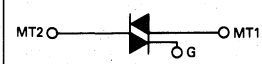
Triacs
Bidirectional Triode Thyristors

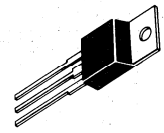
T2800
T2802
Series

TRIACs
8 AMPERES RMS
200 thru 600 VOLTS

. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

- Blocking Voltage to 600 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- T2800 — Four Quadrant Gating
  T2802 — Two Quadrant Gating





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to +100°C) Gate Open	V_{DROM}		Volts
T2800, T2802 B		200	
D		400	
M		600	
RMS On-State Current (T_C = +80°C) (Conduction Angle = 360°)	$I_{T(RMS)}$	8	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz, T_J = +80°C)	I_{TSM}	100	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	40	A^2s
Peak Gate Power (Pulse Width = 1 μs)	P_{GM}	16	Watts
Average Gate Power	$P_{G(AV)}$	0.35	Watt
Peak Gate Trigger Current (Pulse Width = 1 μs)	I_{GTM}	4	Amps
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward or Reverse Blocking Current (Rated V_{DRM} or V_{RRM}, gate open) T_C = 25°C	I_{DRM}, I_{RRM}	—	—	10	μA
T_C = 100°C		—	—	2	mA
Peak On-State Voltage (Either Direction) (I_T = 30 A Peak)	V_{TM}	—	1.7	2	Volts
Gate Trigger Current (Continuous dc) (V_D = 12 Vdc, R_L = 12 Ohms)	I_{GT}				mA
$V_{MT2(+)}$, $V_{G(+)}$ T2800			10	25	
T2802			25	50	
$V_{MT2(+)}$, $V_{G(-)}$ T2800 Only		—	20	60	
$V_{MT2(-)}$, $V_{G(-)}$ T2800		—	15	25	
T2802		—	25	50	
$V_{MT2(-)}$, $V_{G(+)}$ T2800 Only		—	30	60	
Gate Trigger Voltage (Continuous dc) (All Polarities) (V_D = 12 Vdc, R_L = 100 Ohms)	V_{GT}	—	1.25	2.5	Volts
(R_L = 125 Ohms, V_D = V_{DRM}, T_C = 100°C)		0.2	—	—	
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open) T2800	I_{HO}	—	15	30	mA
(I_T = 125 mA) T2802		—	20	60	
Gate Controlled Turn-On Time (Rated V_{DROM}, I_T = 10 A, I_{GT} = 80 mA, Rise Time = 0.1 μs)	t_{gt}	—	1.6	—	μs
Critical Rate of Rise of Commutation Voltage (Rated V_{DRM}, $I_{T(RMS)}$ = 8 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	10	—	V/μs
Critical Rate of Rise of Off-State Voltage (Rated V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 100°C)	dv/dt				V/μs
T2800 B		100	—	—	
T2802 D		75	—	—	
M		60	—	—	

3

FIGURE 1 – CURRENT DERATING

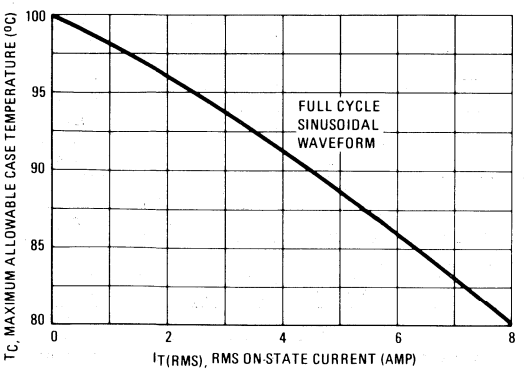
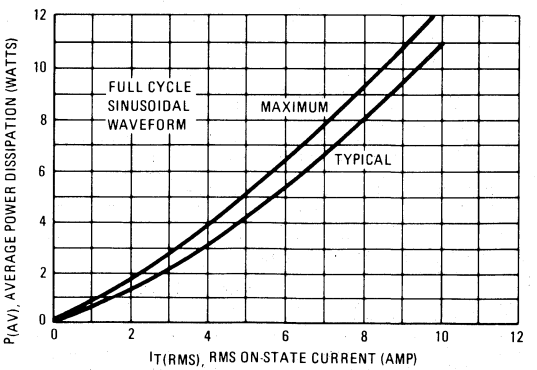


FIGURE 2 – POWER DISSIPATION



Triacs

Bidirectional Triode Thyristors

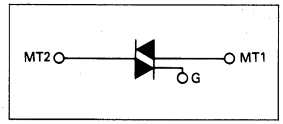
. . . designed primarily for full-wave ac control applications, such as light dimmers, motor controls, heating controls and power supplies.

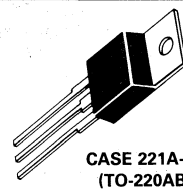
- Blocking Voltage to 800 Volts
- All Diffused and Glass Passivated Junctions for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability

T4100M
(See 2N5571)
T6410, T6420
(See MAC6400)
T6411, T6421
(See MAC6401)

T2801 Series

TRIACs
6 AMPERES RMS
200 thru 800 VOLTS





CASE 221A-04
(TO-220AB)
STYLE 4

3

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Repetitive Off-State Voltage, Note 1 (T_J = −40 to +100°C) Gate Open T2801 B T2801 D T2801 M T2801 N	V_{DRM}	 200 400 600 800	Volts
RMS On-State Current (T_C = +80°C) (Conduction Angle = 360°)	$I_{T(RMS)}$	6	Amps
Peak Non-Repetitive Surge Current (One Full Cycle, 60 Hz)	I_{TSM}	80	Amps
Circuit Fusing (t = 8.3 ms)	I^2t	26	A^2s
Peak Gate Power (T_C = +80°C, Pulse Width = 2 μs)	P_{GM}	0.35	Watt
Average Gate Power (T_C = +80°C, t = 8.3 ms)	$P_{G(AV)}$	0.35	Watt
Peak Gate Trigger Current (Pulse Width = 1 μs)	I_{GTM}	4	Amps
Operating Junction Temperature Range	T_J	−40 to +100	°C
Storage Temperature Range	T_{stg}	−40 to +150	°C

Note 1. Ratings apply for open gate conditions. Thyristor devices shall not be tested with a constant current source for blocking capability such that the voltage applied exceeds the rated blocking voltage.

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.2	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C, Either Polarity of MT2 to MT1 Voltage, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Off-State Current T_J = 25°C (Rated V_{DRM}, Gate Open, T_J = 100°C)	I_{DRM}	— —	— —	10 2	μA mA
Peak On-State Voltage (I_{TM} = 30 A Peak; Pulse Width = 1 to 2 ms, Duty Cycle ≤ 2%)	V_{TM}	—	2	3	Volts
Gate Trigger Current (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 12 Ohms)	I_{GT}	—	25	80	mA
Gate Trigger Voltage (Continuous dc), Note 1 (V_D = 12 Vdc, R_L = 12 Ohms) (V_D = V_{DRM}, R_L = 125 Ohms, T_C = 100°C)	V_{GT}	— 0.2	1.5 —	4 —	Volts
Holding Current (Either Direction) (V_D = 12 Vdc, Gate Open, Initiating Current = 150 mA)	I_H	—	100	—	mA
Turn-On Time (1) (V_D = Rated V_{DRM}, I_T = 10 A, I_{GT} = 80 mA, Rise Time = 0.1 μs)	t_{gt}	—	2.2	—	μs
Critical Rate of Rise of Commutation Voltage (V_D = Rated V_{DRM}, $I_{T(RMS)}$ = 6 A, Commutating di/dt = 4.3 A/ms, Gate Unenergized, T_C = 80°C)	dv/dt(c)	—	10	—	V/μs
Critical Rate of Rise of Off-State Voltage (V_D = V_{DRM}, Exponential Voltage Rise, Gate Open, T_C = 100°C) T2801 B D M N	dv/dt	 50 30 10 10	 — — — —	 — — — —	V/μs

Note 1. Applies for MT2(+), G(+), MT2(−), G(−).

3

FIGURE 1 — CURRENT DERATING

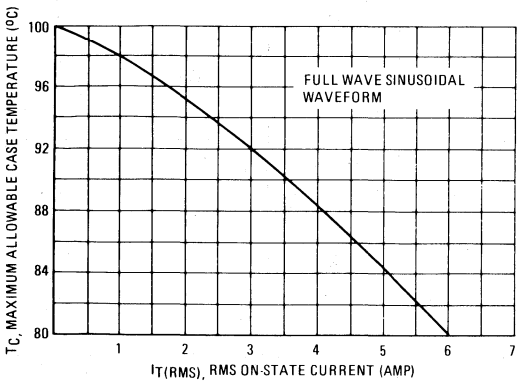
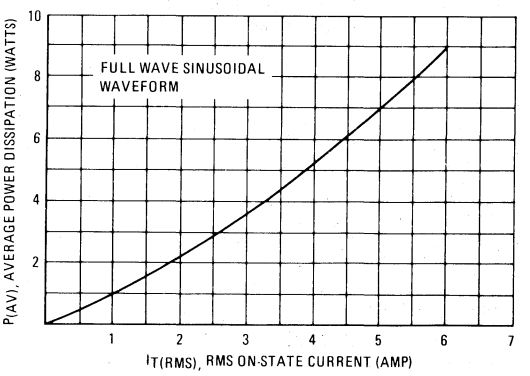
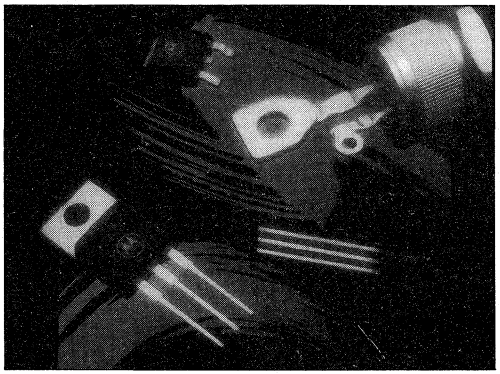


FIGURE 2 — POWER DISSIPATION

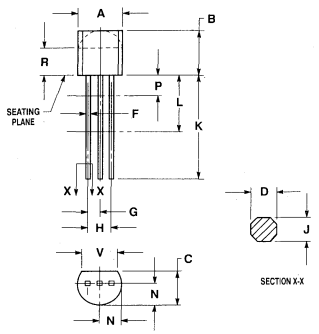




Outline Dimensions and Leadform Options

4

Outline Dimensions



CASE 29-04
TO-226AA (TO-92)
STYLES 3, 9, 10, 12, 16

STYLE 3:
PIN 1. ANODE
2. ANODE
3. CATHODE

STYLE 9:
PIN 1. BASE 1
2. EMITTER
3. BASE 2

STYLE 10:
PIN 1. CATHODE
2. GATE
3. ANODE

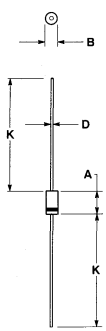
STYLE 12:
PIN 1. MAIN TERMINAL 1
2. GATE
3. MAIN TERMINAL 2

STYLE 16:
PIN 1. ANODE
2. GATE
3. CATHODE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. CONTOUR OF PACKAGE BEYOND DIM R IS UNCONTROLLED.
4. DIM F APPLIES BETWEEN P AND L. DIM D AND J APPLIES BETWEEN L AND K MINIMUM. LEAD DIM IS UNCONTROLLED IN PAD AND BEYOND DIM K MINIMUM.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	4.45	5.20	0.175	0.205
B	4.32	5.33	0.170	0.210
C	3.18	4.19	0.125	0.165
D	0.41	0.55	0.016	0.022
F	0.41	0.48	0.016	0.019
G	1.15	1.39	0.045	0.055
H	2.42	2.66	0.095	0.105
J	0.39	0.50	0.015	0.020
K	12.70	—	0.500	—
L	6.35	—	0.250	—
N	2.04	2.66	0.080	0.105
P	—	2.54	—	0.100
R	2.93	—	0.115	—
V	3.43	—	0.135	—

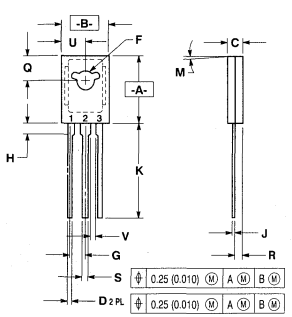
CASE 59-04
DO-41



NOTES:
1. ALL RULES AND NOTES ASSOCIATED WITH JEDEC DO-41 OUTLINE SHALL APPLY.
2. POLARITY DENOTED BY CATHODE BAND.
3. LEAD DIAMETER NOT CONTROLLED WITHIN "F" DIMENSION.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	5.97	6.60	0.235	0.260
B	2.79	3.05	0.110	0.120
D	0.76	0.86	0.030	0.034
K	27.94	—	1.100	—

CASE 77-07
TO-225AA
STYLES 2, 5



STYLE 2:
PIN 1. CATHODE
2. ANODE
3. GATE

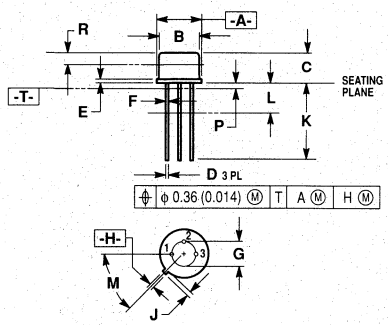
STYLE 5:
PIN 1. MT1
2. MT2
3. GATE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	10.80	11.04	0.425	0.435
B	7.50	7.74	0.295	0.305
C	2.42	2.66	0.095	0.105
D	0.51	0.66	0.020	0.026
F	2.93	3.30	0.115	0.130
G	2.39 BSC		0.094 BSC	
H	1.27	2.41	0.050	0.095
J	0.39	0.63	0.015	0.025
K	14.61	16.63	0.575	0.655
M	3° TYP		3° TYP	
Q	3.76	4.01	0.148	0.158
R	1.15	1.39	0.045	0.055
S	0.64	0.88	0.025	0.035
U	3.69	3.93	0.145	0.155
V	1.02	—	0.040	—

4

CASE 79-04
TO-205AD
STYLE 3

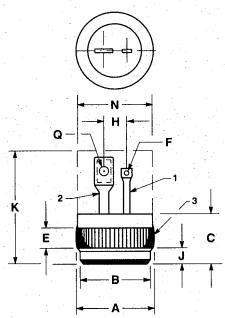


STYLE 3:
PIN 1. CATHODE
2. GATE
3. ANODE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. DIMENSION J MEASURED FROM DIMENSION A MAXIMUM.
4. DIMENSION B SHALL NOT VARY MORE THAN 0.25 (0.010) IN ZONE R. THIS ZONE CONTROLLED FOR AUTOMATIC HANDLING.
5. DIMENSION F APPLIES BETWEEN DIMENSION P AND L. DIMENSION D APPLIES BETWEEN DIMENSION L AND K MINIMUM. LEAD DIAMETER IS UNCONTROLLED IN DIMENSION P AND BEYOND DIMENSION K MINIMUM.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	8.51	9.39	0.335	0.370
B	7.75	8.50	0.305	0.335
C	6.10	6.60	0.240	0.260
D	0.41	0.53	0.016	0.021
E	0.23	1.04	0.009	0.041
F	0.41	0.48	0.016	0.019
G	5.08 BSC		0.200 BSC	
H	0.72	0.86	0.028	0.034
J	0.74	1.14	0.029	0.045
K	12.70	19.05	0.500	0.750
L	6.35	—	0.250	—
M	45° BSC		45° BSC	
P	—	1.27	—	0.050
R	2.54	—	0.100	—

CASE 174-04
STYLES 1, 3

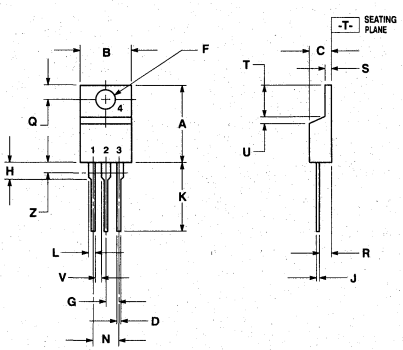


STYLE 1:
TERM. 1. GATE
2. CATHODE
3. ANODE

STYLE 3:
TERM. 1. GATE
2. MAIN TERMINAL 1
3. MAIN TERMINAL 2

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	12.73	12.83	0.501	0.505
B	11.81	12.06	0.465	0.475
C	8.39	9.65	0.330	0.380
E	2.54	—	0.100	—
F	0.89	2.16	0.035	0.085
H	3.75	4.41	0.148	0.174
J	2.04	2.46	0.080	0.097
K	—	20.32	—	0.800
N	—	12.95	—	0.510
Q	1.65	4.06	0.065	0.160

CASE 221A-04
TO-220AB
STYLES 3, 4



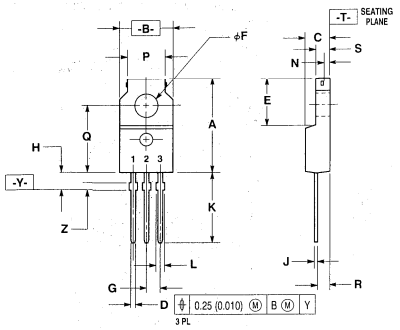
STYLE 3:
PIN 1. CATHODE
2. ANODE
3. GATE
4. ANODE

STYLE 4:
PIN 1. MAIN TERMINAL 1
2. MAIN TERMINAL 2
3. GATE
4. MAIN TERMINAL 2

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. DIM Z DEFINES A ZONE WHERE ALL BODY AND LEAD IRREGULARITIES ARE ALLOWED.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	14.48	15.75	0.570	0.620
B	9.66	10.28	0.380	0.405
C	4.07	4.82	0.160	0.190
D	0.64	0.88	0.025	0.035
F	3.61	3.73	0.142	0.147
G	2.42	2.66	0.095	0.105
H	2.80	3.93	0.110	0.155
J	0.36	0.55	0.014	0.022
K	12.70	14.27	0.500	0.562
L	1.15	1.39	0.045	0.055
N	4.83	5.33	0.190	0.210
Q	2.54	3.04	0.100	0.120
R	2.04	2.79	0.080	0.110
S	1.15	1.39	0.045	0.055
T	5.97	6.47	0.235	0.255
U	0.00	1.27	0.000	0.050
V	1.15	—	0.045	—
Z	—	2.04	—	0.080

CASE 221C-02
STYLES 2, 3



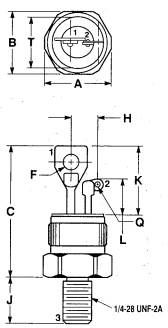
NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. LEAD DIMENSION UNCONTROLLED WITHIN DIMENSION 'Z'.

STYLE 2:
PIN 1. CATHODE
2. NODE
3. GATE

STYLE 3:
PIN 1. MT 1
2. MT 2
3. GATE

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	17.28	17.78	0.680	0.700
B	9.86	10.36	0.388	0.408
C	4.45	4.95	0.175	0.195
D	0.64	1.01	0.025	0.040
E	8.64	9.01	0.340	0.355
F	3.56	3.81	0.140	0.150
G	2.54 BSC		0.100 BSC	
H	2.80	3.93	0.110	0.155
J	0.46	0.71	0.018	0.028
K	12.70	13.97	0.500	0.550
L	1.15	1.77	0.045	0.070
N	1.25	—	0.049	—
P	6.86	7.36	0.270	0.290
Q	12.20	12.70	0.480	0.500
R	2.29	3.04	0.090	0.120
S	2.67	2.92	0.105	0.115
Z	1.78	2.28	0.070	0.090

CASE 263-04
STYLES 1, 2

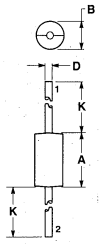


STYLE 1:
PIN 1. CATHODE
2. GATE
3. ANODE

STYLE 2:
PIN 1. MT 1
2. GATE
3. MT 2

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	15.34	15.60	0.604	0.614
B	14.00	14.20	0.551	0.559
C	2.67	30.23	1.050	1.190
F	3.43	4.06	0.135	0.160
H	—	6.73	—	0.265
J	10.67	11.56	0.420	0.455
K	15.75	17.02	0.620	0.670
L	7.62	8.89	0.300	0.350
Q	1.40	2.16	0.055	0.085
T	12.73	12.83	0.501	0.505

CASE 267-03
STYLE 1

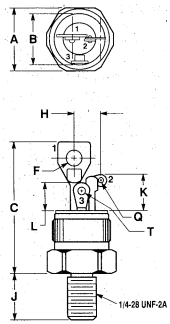


NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

STYLE 1:
PIN 1. CATHODE
2. ANODE

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	9.40	9.65	0.370	0.380
B	4.83	5.33	0.190	0.210
D	1.22	1.32	0.048	0.052
K	25.40	—	1.000	—

CASE 311-02
STYLES 1, 2

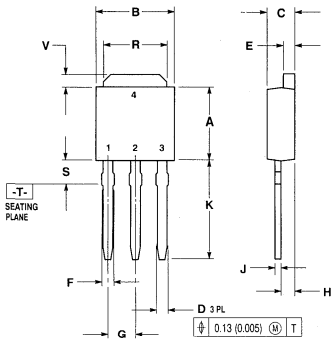


STYLE 1:
PIN 1. CATHODE
2. GATE
3. ANODE

STYLE 2:
PIN 1. MT 1
2. GATE
3. MT 2

DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	14.00	14.20	0.551	0.559
B	12.73	12.83	0.501	0.505
C	—	32.51	—	1.280
F	—	4.06	—	0.160
H	—	6.73	—	0.265
J	10.67	11.56	0.420	0.455
K	7.62	8.89	0.300	0.350
L	6.48	6.99	0.255	0.275
Q	1.40	2.16	0.055	0.085
T	3.43	3.81	0.135	0.150

CASE 369
STYLE 5

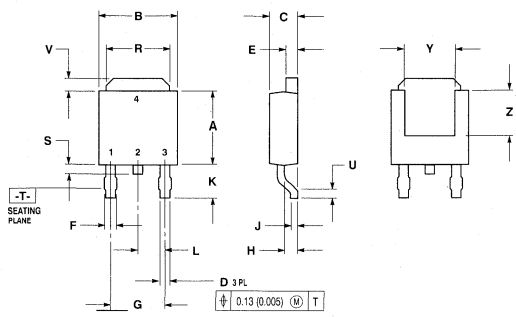


STYLE 5:
PIN 1. GATE
2. ANODE
3. CATHODE
4. ANODE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	5.97	6.22	0.235	0.245
B	6.35	6.73	0.250	0.265
C	2.19	2.38	0.086	0.094
D	0.69	0.88	0.027	0.035
E	0.84	1.01	0.033	0.040
F	0.77	1.14	0.030	0.045
G	2.29 BSC		0.090 BSC	
H	0.87	1.01	0.034	0.040
J	0.46	0.58	0.018	0.023
K	8.89	9.65	0.350	0.380
R	4.45	5.46	0.175	0.215
S	1.91	2.28	0.075	0.090
V	0.77	1.27	0.030	0.050

CASE 369A
STYLE 5



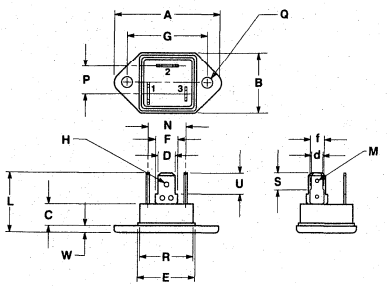
STYLE 5:
PIN 1. GATE
2. ANODE
3. CATHODE
4. ANODE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	5.97	6.22	0.235	0.245
B	6.35	6.73	0.250	0.265
C	2.19	2.38	0.086	0.094
D	0.69	0.88	0.027	0.035
E	0.84	1.01	0.033	0.040
F	0.77	1.14	0.030	0.045
G	4.58 BSC		0.180 BSC	
H	0.87	1.01	0.034	0.040
J	0.46	0.58	0.018	0.023
K	2.60	2.89	0.102	0.114
L	2.29 BSC		0.090 BSCS	
R	4.45	5.46	0.175	0.215
S	0.64	0.88	0.025	0.035
U	0.51	—	0.020	—
V	0.77	1.27	0.030	0.050
Y	4.32	—	0.170	—
Z	3.69	—	0.145	—

OUTLINE DIMENSIONS — continued

CASE 383-01
STYLE 1



STYLE 1:
PIN 1. MT1
2. MT2
3. GATE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: MILLIMETER

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	—	39.20	—	1.543
B	—	23.00	—	0.905
C	—	8.20	—	0.322
D	5.55	7.15	0.219	0.281
E	—	21.60	—	0.850
F	7.45	9.05	0.294	0.356
G	29.90	30.10	1.178	1.185
H	1.65	2.20	0.065	0.086
L	—	22.50	—	0.885
M	1.30	1.57	0.052	0.061
N	13.85	14.15	0.546	0.557
P	10.55	10.85	0.416	0.427
Q	4.10	4.30	0.162	0.169
R	—	20.10	—	0.791
S	6.20	6.50	0.244	0.255
U	7.80	8.10	0.307	0.318
W	2.30	2.90	0.091	0.114
d	4.58	5.08	0.180	0.200
f	5.34	5.58	0.210	0.220

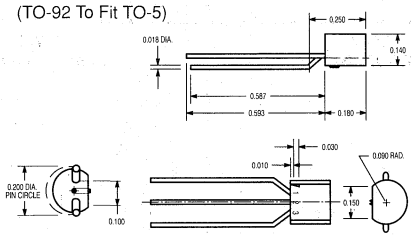
4

Leadform Options

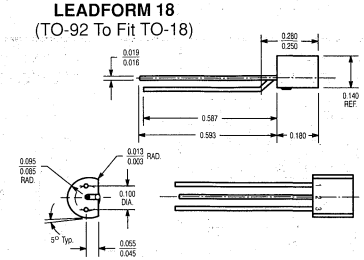
Plastic packaged semiconductors may be leadformed to a variety of configurations for insertion into sockets and boards designed for metal can devices. The following are standard thyristor and trigger leadforms offered by Motorola.

To order leadformed product, determine the form desired, and specify case number and applicable leadform number. A special device title will be assigned by the factory to process your order. Certain standard devices already incorporate a leadform, and may be purchased without consulting the factory.

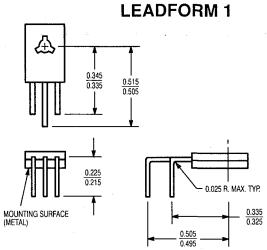
CASE 29
LEADFORM 5
(TO-92 To Fit TO-5)



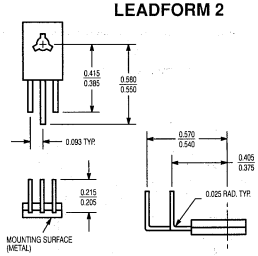
CASE 29
LEADFORM 18
(TO-92 To Fit TO-18)



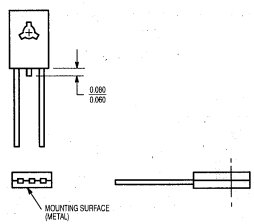
CASE 77
LEADFORM 1



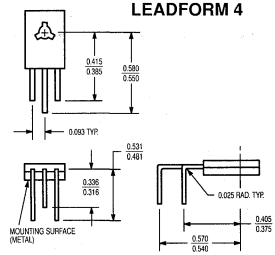
CASE 77
LEADFORM 2



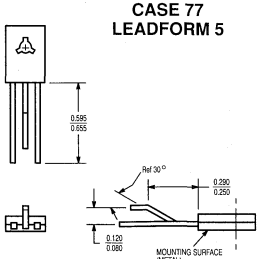
CASE 77
LEADFORM 3



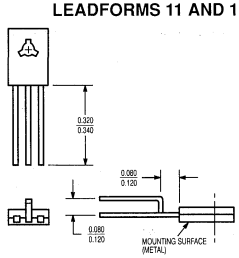
CASE 77
LEADFORM 4



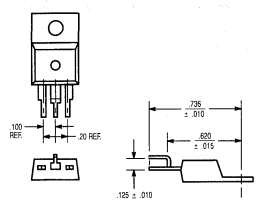
CASE 77
LEADFORM 5



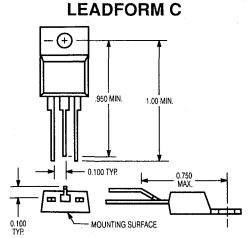
CASE 77
LEADFORMS 11 AND 12



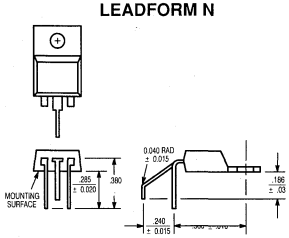
TO-220
LEADFORM S



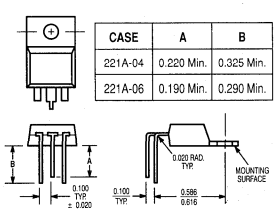
TO-220
LEADFORM C



TO-220
LEADFORM N

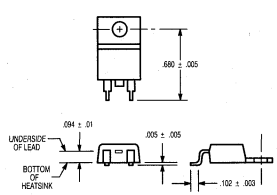


TO-220
LEADFORM A

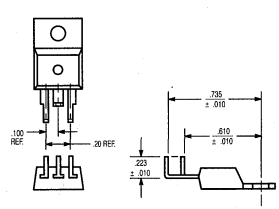


CASE	A	B
221A-04	0.220 Min.	0.325 Min.
221A-06	0.190 Min.	0.290 Min.

TO-220
LEADFORM 16

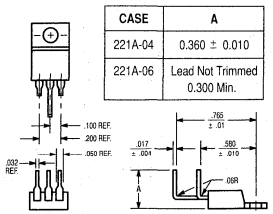


TO-220
LEADFORM T



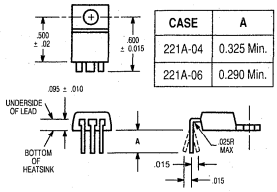
4

TO-220 LEADFORM AJ



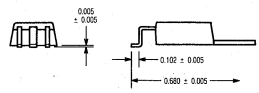
CASE	A
221A-04	0.360 ± 0.010
221A-06	Lead Not Trimmed 0.300 Min.

TO-220 LEADFORM L

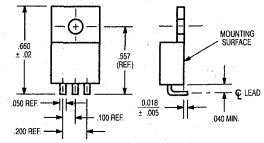


CASE	A
221A-04	0.325 Min.
221A-06	0.290 Min.

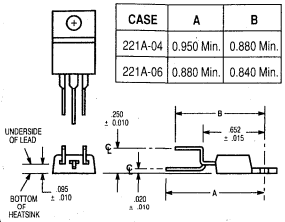
TO-220 LEADFORM A16A



TO-220 LEADFORM AF

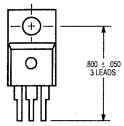


TO-220 LEADFORM D1

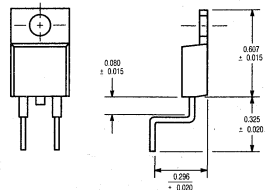


CASE	A	B
221A-04	0.950 Min.	0.880 Min.
221A-06	0.880 Min.	0.840 Min.

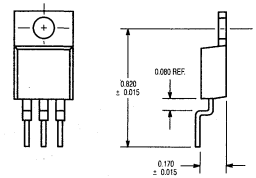
TO-220 LEADFORM W



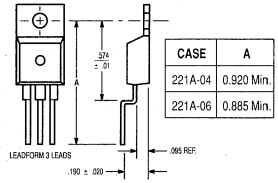
TO-220 LEADFORM U2



TO-220 LEADFORM UA

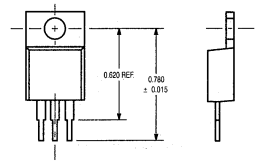


TO-220 LEADFORM U



CASE	A
221A-04	0.920 Min.
221A-06	0.885 Min.

TO-220 LEADFORM WC



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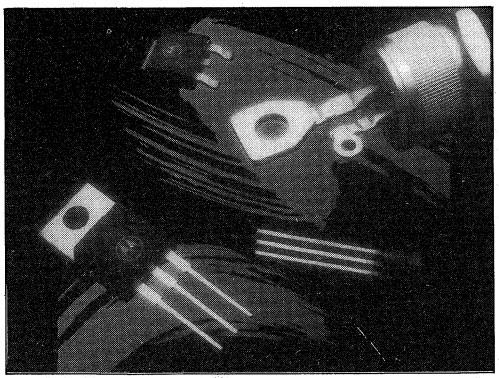

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