Application Note AN-949

Current Rating of Power Semiconductors

Table of Contents

The current rating of an electrical device, whether a circuit breaker or a motor or a transformer, is the current at which the temperature within the electrical device reaches a value that may impair the reliability or functionality of the device itself. The manufacturer knows the temperature limits of the materials used in the device, but he does not know the temperature of the ambient in which the device will be used. So he makes an assumption on this temperature.

Current Rating of Power Semiconductors

1. What Is Current Rating?

The current rating of an electrical device, whether a circuit breaker or a motor or a transformer, is the current at which the temperature within the electrical device reaches a value that may impair the reliability or functionality of the device itself. The manufacturer knows the temperature limits of the materials used in the device, but he does not know the temperature of the ambient in which the device will be used. So he makes an assumption on this temperature. This has two important consequences:

1) **A current rating is meaningless without the rated temperature.**
2) **The temperature at which the rating applies may, or may not be related to actual operating conditions.**
   If it is, the current rating can be used as **an indication** of the current capability of that device in a real application. If the device is rated at a temperature that is not encountered in a typical operating environment, e.g. 25ºC, **it cannot be trusted to provide an indication of actual device capability in an application.** It can only be used to compare the ratings of similar devices rated at the same temperature.

The rating of electrical devices like motors and circuit breakers are dictated by various agreements and regulations. The ratings of many other devices, like transformers, resistors and semiconductors are specified in their data sheets. As a result, the user must, *minimally* verify that the device is capable of operating:

a. at the maximum current
b. at the maximum ambient temperature
c. without exceeding its maximum temperature

2. Current Ratings for Power Semiconductors

Like any other electrical device, power semiconductors must be operated within their maximum temperature. Since the vast majority of power semiconductors operate at large power densities, they require a heatsink. It is the task of the designer to identify the heatsink, or other cooling method that fulfills the requirements of a, b and c noted in the previous section. This task is normally referred to as “thermal design.”

Power semiconductors have, however, some additional limitations normally associated with their capability of handling high voltages and high currents at the same time under static or dynamic conditions. These limitations are peculiar to the specific type of semiconductor, e.g. SOA for transistors, dv/dt, di/dt and t_q for thyristors, t_{rr} for diodes. Information on these limitations is normally contained in publications that are specific to the particular device.

Bipolar transistors have one additional limitation that is not common to other power semiconductors: gain. To operate a bipolar transistor at its headlined “rated” continuous current would require an inconveniently large amount of drive current, and the saturation voltage and switching times would be hard to live with in a practical design.

Other power semiconductors are not limited by gain. IGBTs for motor drive applications are, by design, limited in gain to current levels much beyond normal operating conditions to reduce the current under short-circuit conditions.

3. CONTINUOUS CURRENT RATINGS

The continuous rating of a power semiconductor is based on *heat removal when conducting a fixed amount of current.* This is determined by the fundamental equation for temperature rise (see AN-936), with no switching losses present. Rated I_D, for a MOSFET is therefore:

$$I_D = \sqrt{\frac{T_{J\,max} - T_C}{R_{DS(on)} R_{th(JC)}}}$$

where $R_{DS(on)}$ is the limiting value of the on-resistance *at rated* $T_{J\,max}$ *, at the appropriate value of* I_D. Rth_{JC} is the maximum value of internal junction-to-case thermal resistance, and Tc is the case temperature.

Similarly, the continuous current rating of a diode, or a thyristor, or an IGBT is calculated from the basic equation of temperature rise. The power dissipation is calculated from voltage drop and continuous current.

Except for water-cooled sinks, it is very difficult to keep the case temperature of a power semiconductor at less than 90ºC. Thus, the usable *continuous direct* current of a power device for most practical is whatever is applicable to a case temperature of 90 to 110° C. This allows a sufficient differential between case and ambient temperature for the heat dissipater to handle the heat transfer.

The “headlined” continuous current rating shown on the datasheets of most power transistors is usually larger than the practically usable level of *continuous* drain current. This is because the case temperature adopted by the industry, to which the “headlined” continuous I_D rating applies, is 25°C.

Figure 1 shows typical heatsinks for TO-3 and TO-220 packaged HEXFET® Power MOSFETs that allow them to operate in a 40°C ambient at a *continuous direct* drain current that is 60 to 70% of the rated continuous drain current at Tc = 25°C; the corresponding *steady* case temperature is about 100°C.

The continuous current rating of power transistors is, however, of little direct use to the designer, other than as a benchmark, for the following three reasons:

1) Power transistors are normally operated in switch mode, with duty cycles considerably less than l00%, and what is really of interest is the current-carrying capability of the device under the actual “switched” operating conditions.
2) When operated in switch mode, power transistors have switching losses, which have to be calculated and added to the conduction losses, as indicated in AN-936.
3) The selection of the power device may be dictated by surge requirements that make the continuous current rating irrelevant.

If this were not enough, advances in the low-voltage MOSFET technology have reduced conduction losses to the point that the package has become the limiting factor in their continuous current rating. This is explained in DT93-4

4. SWITCHING “DUTY CYCLE” RATINGS

The basic thermal equation of AN-936 determines the basic rating of a “thermal system” in a practical application. This equation can be used to determine how much power can be dissipated by a (known) thermal system or the junction temperature for a given set of electrical operating conditions (power dissipation). Since the power entered in that equation is the “average” power, it remains valid as long as the frequency of operation is high with respect to the thermal inertia of the system. When the frequency is very low (tens of Hz), the transient thermal response curve is used, as described below in Section 5.

The power dissipation is normally divided in conduction and switching. Conduction losses in a power MOSFET, being resistive in nature, can be calculated as $(I_{RMS})^2$ x R. The RMS content of waveforms of different shape can be found in the Appendix. Switching losses can be calculated from the switching waveforms, from the gate charge or from analytical methods. Conduction and switching losses for IGBTs are more complex, as explained in AN-990.

5. JUNCTION TEMPERATURE UNDER PULSED CONDITIONS

Under surge conditions the junction temperature rises exponentially, according to its thermal inertia. Rather than using the thermal resistance, that is appropriate for steady state operation, we use the Transient Thermal Impedance (or, more correctly, Thermal Response Curve), as the one shown in Figure 2. For a surge of given duration (x axis), this curve gives a thermal response factor (y axis). The peak junction temperature due to the surge condition can be calculated as indicated in the figure itself. The power dissipation is normally calculated from the voltage and current across the device during the surge.

This curve is also useful for determining the peak junction temperature for power or pulses with a very low repetition rate, when the method described in the previous section is not applicable. The reason for this is illustrated by the waveforms in Figures 3(a) and 3(b). Both sets of waveforms are for the same power dissipation and duty cycle, but for different operating frequencies. The cycle-by-cycle fluctuations of junction temperature at 20Hz (Figure 3a) are clearly greater than at 200Hz (Figure 3b). As frequency increases, thermal inertia of the junction “irons out” instantaneous temperature fluctuations and the junction responds more to average, rather than peak power dissipation. At frequencies above a few kHz and duty cycles above 20% or so, cycle-by-cycle temperature fluctuations become small, and peak junction temperature rise becomes equal to the average power dissipation multiplied by the DC junction-to-case thermal resistance, within one or two percent.

For pulses with low repetition rate the remaining curves in Figure 2 show effective thermal impedance at different duty cycles. These curves are approximately related to the single pulse curve, by the following relationship:

Effective normalized thermal impedance = D + (l - D) x (transient thermal impedance for single pulse of duration t). The thermal impedance, when multiplied by the power dissipation *during the conduction period t* (i.e., the power *within* the conduction pulse itself, *not* the power averaged over the whole cycle), gives the value of the repetitive peak junction-to-case temperature rise.

To determine the absolute value of the peak junction temperature, it is, of course, necessary to know the case temperature T_C under steady-state operating conditions. Because of thermal inertia, the heatsink responds only to average power dissipation (except at extremely low frequencies which generally will not be of practical interest). T_C is therefore given by:

$T_C = T_A + (R_{thC-S} + R_{thS-A})\, P_{AV}$ where:

T_A = ambient temperature

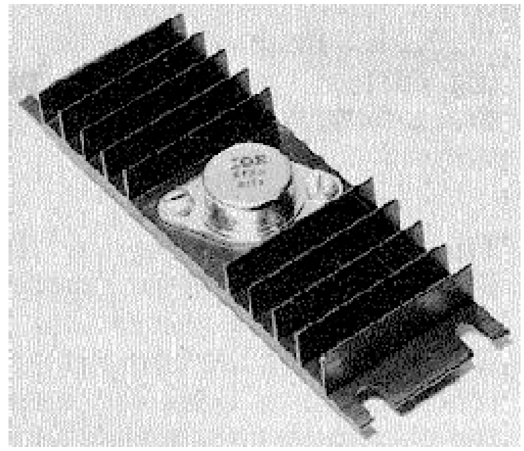
R_{thC-S} = case-to-sink thermal resistance

R_{thS-A} = sink-to-ambient thermal resistance

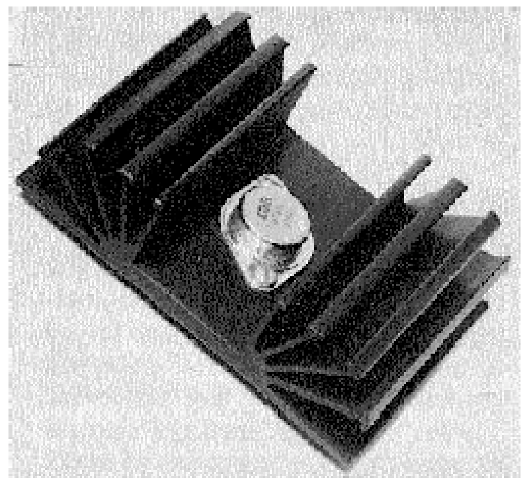
P_{AV} = average power dissipation

also,

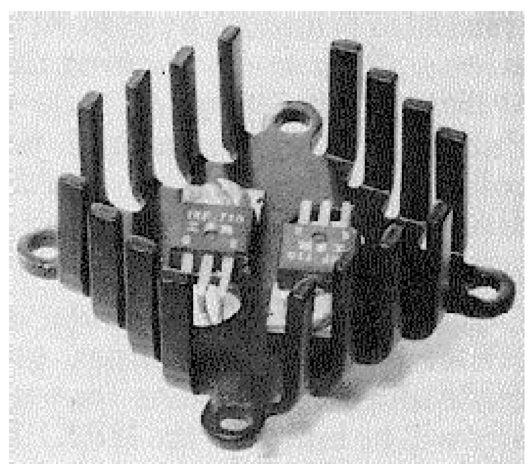
P_{AV} = peak power x duty cycle for rectangular pulses of power



(a) Type 621-A heatsink give 4A continuous rating for IRF331 with 5 CFM airflow in 40°C ambient.



(b) Type 641-A heatsink gives 3.5A continuous rating for IRF331 with natural convection cooling in 40°C ambient.



(c) Type 689-75 e4 heatsink give 1 A continuous rating for IRF710 with natural convection cooling in 40°C ambient.

Figure 1. Typical Heatsinks for HEXFETs (Heatsink by Wakefield)

The transient thermal response curve ***assumes constant case temperature***. This is generally valid for pulses shorter than 10ms. For longer surges the case temperature starts to rise and the results are of questionable accuracy. For operation in free air, case temperature starts to rise within few milliseconds and this curve does not provide any useful information. More sophisticated analytical methods that take the entire thermal system into account are normally used to calculate temperature rise under these conditions.

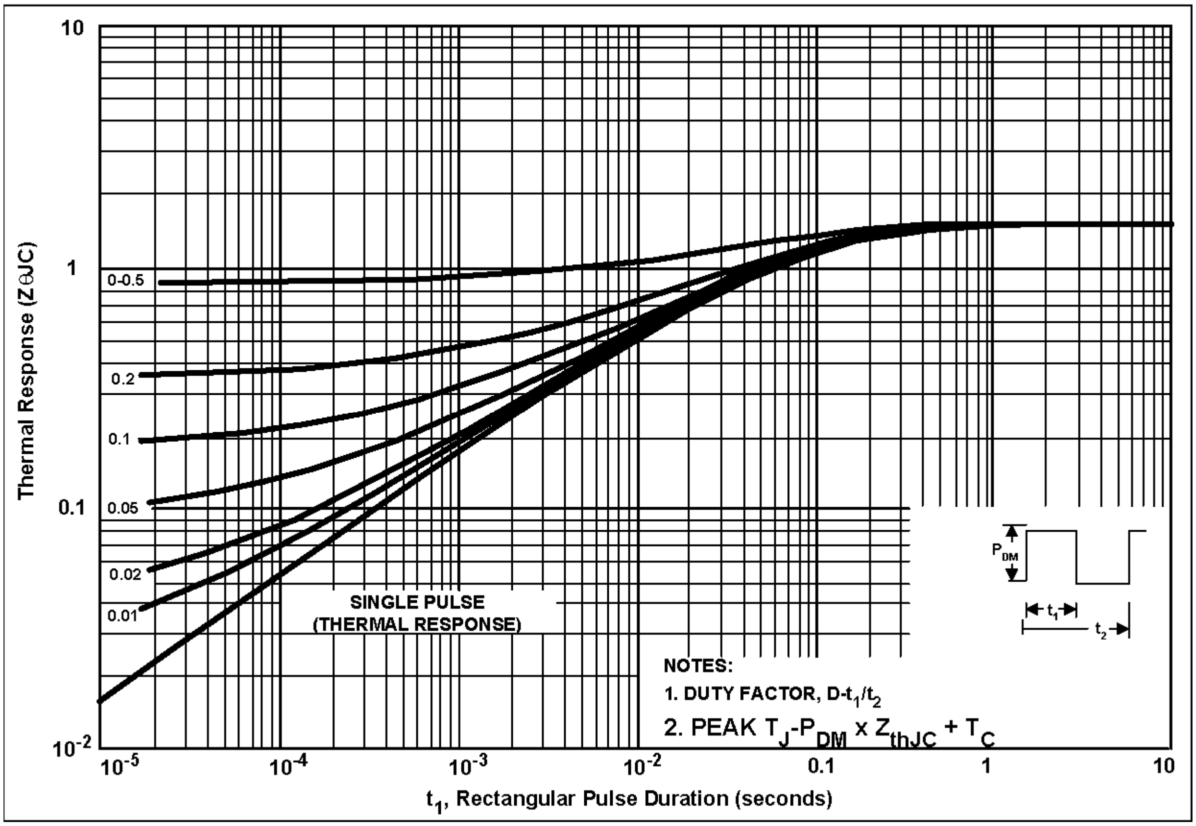


Figure 2. Transient Thermal Impedance Curves for IRF530 HEXFET

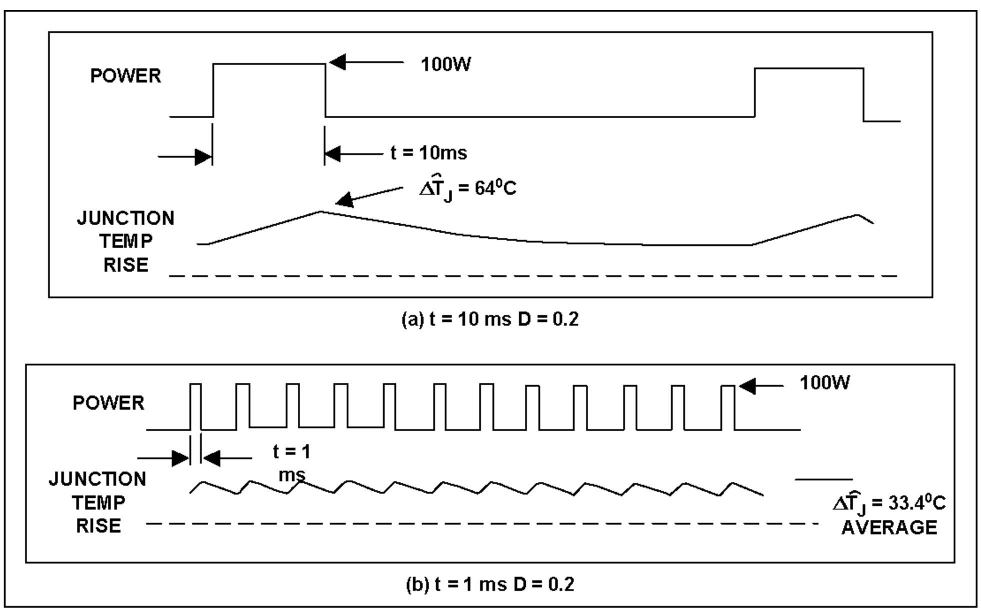


Figure 3. Waveforms of Power and Junction Temperature for Repetitive Operation, showing that Peak Junction Temperature is function of Operating Frequency. IRF330

6. PEAK CURRENT RATINGS

IGBTs and MOSFETs are able to carry peak current well in excess of their continuous current rating, provided that the rated junction temperature is not exceeded. There is, however, an upper limit on the permissible current, defined by the rated peak current. Most devices have a peak rating, that is several times their continuous rating at Tc = 25°C.

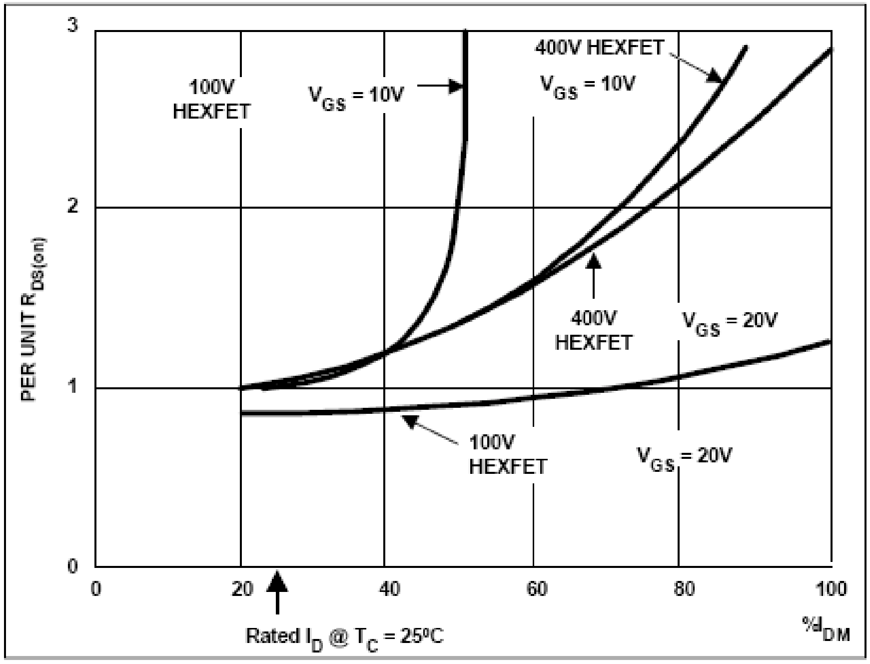


Figure 4. Typical Variation of On-Resistance with Drain Current.

Power transistors are fundamentally “linear” devices, as opposed to “latching” devices. As current increases, the point eventually is reached at which they go into “linear” operation and start to act, in effect, as a current limiter. This point depends upon the drive voltage applied to the gate, the safe limit of which is determined by the thickness of the oxide that insulates the gate from the body of the device. Peak ratings of power devices are normally achievable with an applied gate voltage that is equal to the maximum permissible gate voltage of 20V. They are repetitive ratings, as long as the junction temperature is kept within the rated T_{Jmax}. Peak junction temperature can be calculated from the thermal impedance data for the device, as indicated above.

It should be pointed out that the on-resistance of any MOSFET does increase as current increases. As shown in Figure 4, the on-resistance of a 100V rated HEXFET® Power MOSFET at its rated I_{DM} with 20V applied to the gate is typically 1.4 x the value at the rated I_D; the corresponding multiplier for a 400V rated HEXFET® Power MOSFET is 2.9. This increase of on-resistance must, of course, be taken into account when making thermal calculations and designing for use of the I_{DM} rating.

APPENDIX

Determining the RMS Value of ID Waveforms

To accurately determine the conduction losses in a MOSFET, the RMS value for ID must be known. The current waveforms are rarely simple sinusoids or rectangles, and this can pose some problems in determining the value for I_{RMS}**.** The following equations and procedure can be used to determine IRMS for any waveform that can be broken up into segments for which the RMS value can be calculated individually.

The RMS value of any waveform is defined as $I_{RMS} = \sqrt{\frac{\int_0^t I^2(t)dt}{T}}$ (1)

Figure A-1 shows several simple waveforms and the derivation for I_{RMS} using equation (1).

If the actual waveform can be approximated satisfactorily by combining the waveforms in Figure A-1, then the RMS value of the waveform can be calculated from:

$$I_{RMS} = \sqrt{I^2RMS(1) + I^2RMS(2) + ... + I^2RMS(N)} \quad (2)$$

This is true to the extent that no two waveforms are different from zero at the same time.
In some applications such as switching regulators, it is possible for the designer to control the wave shape with topology or magnetic design. This can be very beneficial in reducing the value for IRMS in the switch for a given value of average current.

EFFECT OF WAVEFORM SHAPE ON RMS VALUE

In a switch mode converter, the current waveforms through the inductors, transformer windings rectifiers and switches will appear as shown in Figure A-1, ranging from a triangle to a rectangle depending on the value of the averaging inductor and load.

The RMS content of the current waveform changes accordingly and this has a bearing on the MOSFET conduction losses that are proportional to I^2_{RMS}.

A measure of the squareness of the waveform can be obtained from the ratio: K = $\frac{I_a}{I_b}$

It can be shown that: K = $\frac{I_a}{I_b}$ = f (L/L_c) where:

L = inductance of the averaging choke.
L_c = 1 is the critical inductance for a Particular input voltage and load power.

As L is increased, K goes from 0 (triangle) to 1 (rectangle).

From the above expression and $I_{avg} = \frac{I_a + I_b}{2}$

We have: $I_a = \frac{2K}{K+1} I_{avg}$

$I_b = \frac{2}{K+1} I_{avg}$

Substituting into the RMS expression for a trapezoidal waveform, shown in Figure A-1, we have:

$$I_{RMS} = 2\sqrt{D} I_{avg} \sqrt{\frac{1+K+K^2}{3(K+1)^2}}$$

For constant $I_{(avg)}$ and D, the normalized (I_{RMS} = 1 for K = 1) I_{RMS} is as shown in Figure A-3. This curve shows that, for triangular current waveforms, the I^2R losses are 32% higher than for rectangular waveforms. It is also apparent that for $I_a / I_b > 0.6$, the improvement incurred by increasing L is only 2%, so from a practical point-of-view, L needs to only be about twice L_C.

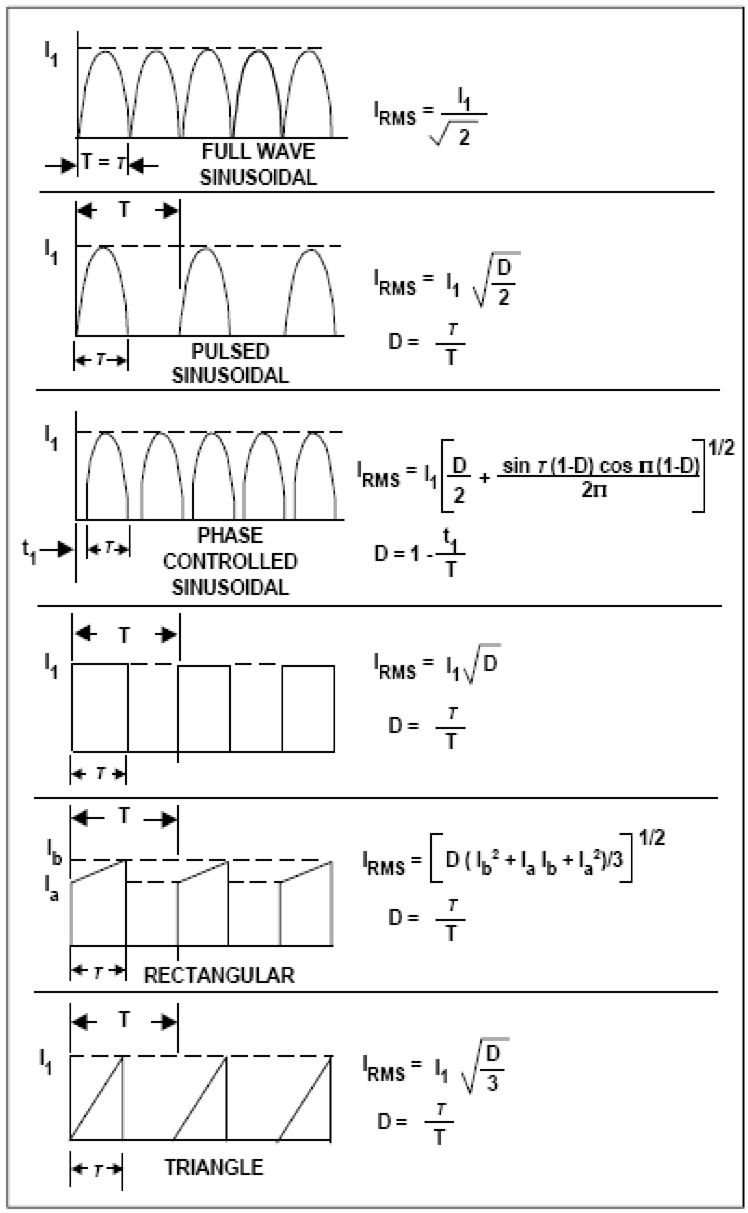


Figure A-1

Increasing the value of I_a / I_b increases the switch turn-on losses but decreases the turn-off losses. Since the turn-off losses tend to be larger than the turn-on losses, increasing I_a / I_b reduces the total switching loss also. For the case of discontinuous inductor current ($L < L_C$), $I_a / I_b = 0$ and is no longer relevant, since the waveforms are now triangles. For a given I_{avg} the RMS current is:

$$I_{RMS} = 2I_{avg}\sqrt{\frac{D}{3}}$$

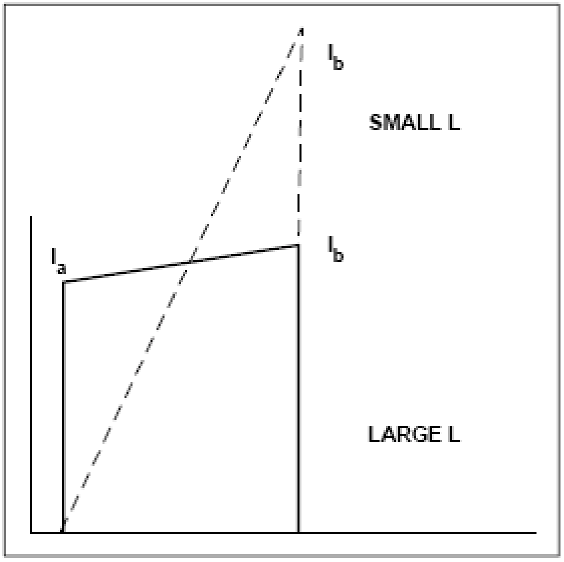


Figure A-2. Current Waveform

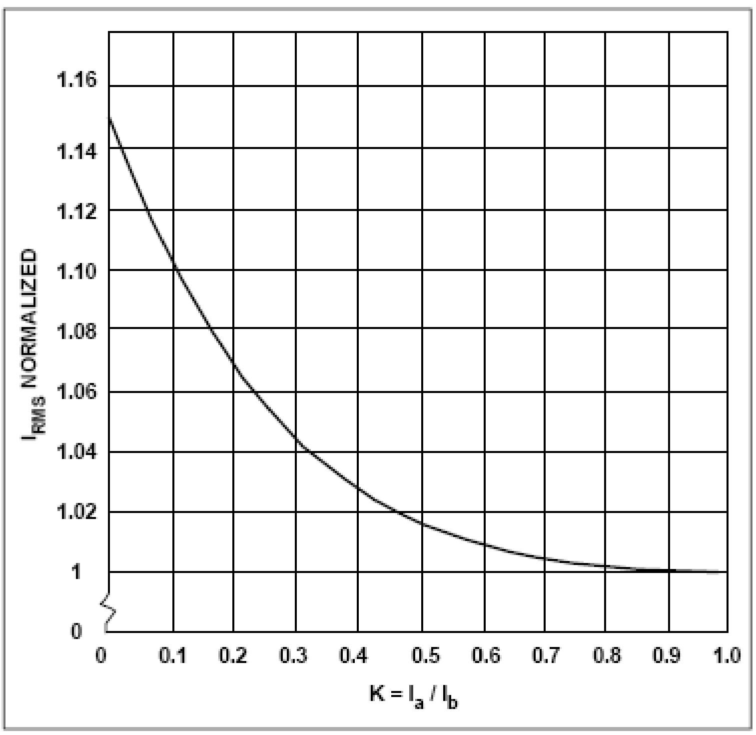


Figure A-3. Variation of IRMS with Squareness Ratio