Current Ratings, Safe Operating Area, and High Frequency Switching Performance of Power HEXFETs®

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Summary

This application note discusses the current handling capability, safe operating area, and power dissipation of a HEXFET power MOSFET. It is shown that the HEXFET's ability to carry current is essentially limited only by junction heating, both for the "switched" and "linear" modes of operation — unlike the bipolar transistor, which is limited by gain and second breakdown. For this reason, peak current ratings of HEXFETs are phenomenally high by comparison with those of bipolar transistors.

Examples are given which show how the HEXFET's current carrying ability can be utilized, and how the power dissipation of a HEXFET compares with that of a fast switching bipolar transistor, as a function of operating frequency.

Introduction

International Rectifier HEXFETs are well established in a variety of applications which previously have been served by bipolar transistors, and are continuing to find many new applications. Designers who are familiar with the practical derating factors that need to be applied when designing with bipolar transistors frequently do not realize that the criteria for determining HEXFET ratings are quite different, and as a result often select a HEXFET which is oversized for the job. This can have a significant bearing on the cost effectiveness of the design.

The purpose of this application note is to explain the basis of the current ratings and Safe Operating Area (SOA) of power HEXFETs, and thus enable the user to make a properly informed choice of HEXFET for his particular application.

A practical comparison of the power losses of a HEXFET and a bipolar transistor is also given. Whereas the conduction losses of a bipolar are generally lower than those of the HEXFET, the switching losses are significantly higher. Base drive power for the bipolar also reduces efficiency.

Test results are presented which illustrate the difference in losses of the HEXFET and the bipolar transistor as a function of frequency. The HEXFET is shown to be generally more efficient above frequencies in the 20 to 40 kHz range.

Bipolar Transistor Current Ratings

It will help to set the stage by first considering the basis of the current ratings of a bipolar transistor. Whereas the continuous and peak current ratings of a bipolar that are "headlined" in the data sheet are theoretically valid, they are hardly ever usable in practice. A basis for specifying the current ratings of bipolar transistors has been adopted in the industry which unfortunately is not representative of usable current levels; it simply provides a yardstick for making comparisons between different products on a reasonably common basis.

The Achilles' heel of the bipolar's current carrying capability is the critical question of the attendant gain, saturation voltage and switching time at *elevated operating temperature.* These supporting parameters are usually specified at "rated" current at a junction temperature of 25° C (where they give the appearance of acceptability), but the data sheet usually does not specify their values at higher "operating" junction temperature (where they are usually not so acceptable).

In reality the bipolar is not intended to be used at its headlined "rated" continuous current. To do so 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, in which the normal junction operating temperature would, of course, be well in excess of 25° C.

A good maximum design operating level for a bipolar transistor is typically 60 to 70% of the headlined "continuous" collector current rating; experienced users know this and design to it. Device manufacturers know it, too; this is why the data sheet specifies the minimum gain, maximum saturation voltage and maximum switching times at elevated junction temperature (usually 100° C), at a collector current which is 60 to 70% of the headlined "rated" value, but not at the "rated" current itself.

An example will illustrate this. The industry-standard 2N6542/3 bipolar transistor has a "headlined" continuous collector rating of 5A. The maximum value of $V_{CE(SAT)}$, the corresponding forced gain, and the maximum switching times at elevated temperature (T_C = 100° C) are, however, specified at a collector current of only 3A. If the designer really wants to use this device at its headlined "rated" current, he will have to refer to the manufacturer to determine the critical "worst case" supporting data needed to design the circuit; this information will not be found on the data sheet.

The rated *peak* collector current of a bipolar is even more tenuous than the rated continuous value. This is usually specified without reference to the required base drive current. Consider the 2N6542/3. The peak collector current rating headlined on the data sheet is 10A. Not specified is the base current needed to produce this collector current.

The DC gain curve, reproduced from the data sheet in Figure 1, terminates at the "continuous" collector current rating of 5A. Bearing in mind that this is anyway a typical curve, it is a matter of conjecture what the minimum gain will be at a collector current of 10A, at elevated operating temperature — and hence what base current will be needed to support the 10A peak collector current rating.

In reality the gain will likely be less than the unity. The 2N6542/3 device would therefore have to be driven with a *base* current of at least 10A in order to utilize its peak collector current rating of 10A — an untenable situation for most practical designs.

Current Ratings of MOSFETs

Continuous Ratings

The MOSFET is quite a different device to a bipolar, and its continuous I_D rating is based upon quite different considerations. Whereas the usable current of a bipolar is basically limited by gain, this is not the case with a power MOSFET. Figure 2 shows a typical relationship between transconductance of a HEXFET and a drain current. Transconductance increases with increasing drain current — just the opposite situation than with a bipolar transistor. Obviously the HEXFET — unlike the bipolar — is not going to "run out of gain" as the drain current increases.

Switching speed is generally much faster than that of a bipolar. With proper drive circuit design, switching speed of a HEXFET varies relatively slightly as the current increases, and is not a factor in determining the rated current. This can be deduced from Figure 3, which shows a typical relationship between gate charge, gate voltage, and drain current for a HEXFET. For a given gate charging current, switching speed is directly proportional to gate charge. The gate charge required for switching, and hence switching speed itself, is not influenced greatly by the amplitude of the drain current, and not at all by junction operating temperature.

The major criterion on which the continuous rating of a HEXFET is based is *heat removal.* The HEXFET will carry as much current as the cooling system will permit, while keeping peak junction temperature within the rated maximum value. The more efficient the heat dissipator to which the HEXFET is attached, the lower the case temperature will be, the greater the permitted case-to-junction temperature rise, the greater the permitted internal power dissipation, and the greater permissible current. These considerations are, of course, exactly the same as those which apply to other non-gain-limited power semiconductor devices, such as rectifiers and thyristors.

Usable current, I_D, for a HEXFET is therefore:

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

where $R_{DS(on)}$ is the limiting value of the on-resistance *at rated* $T_{(Jmax)}$, *at the appropriate value of* I_D, R_{thJC} is the maximum value of internal junction-to-case thermal resistance, and T_C is the case temperature.

Figure 4 shows the continuous current rating of the IRF330 HEXFET as a function of case temperature. Note that below a case temperature of 25° C, the continuous I_D rating is limited by the current carrying capacity of the internal source bonding wire. But this is not a practical limitation.

Figure 4 also shows the relationship between HEXFET internal power dissipation and drain current. Power is proportional to the square of the current, so rises quite rapidly as cur-

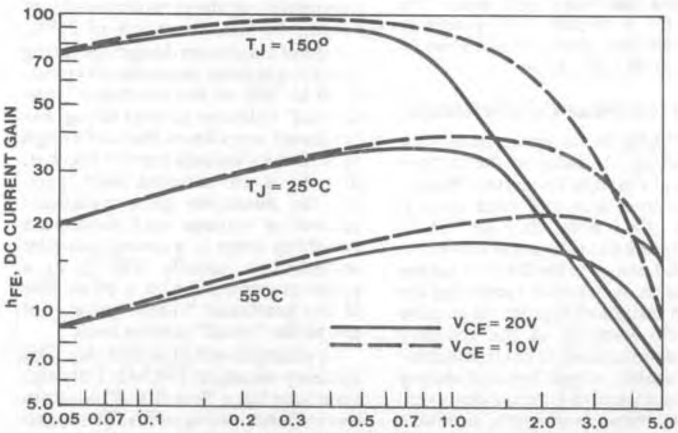


Figure 1. Typical DC Current Gain, 2N6542/3 Bipolar Transistor

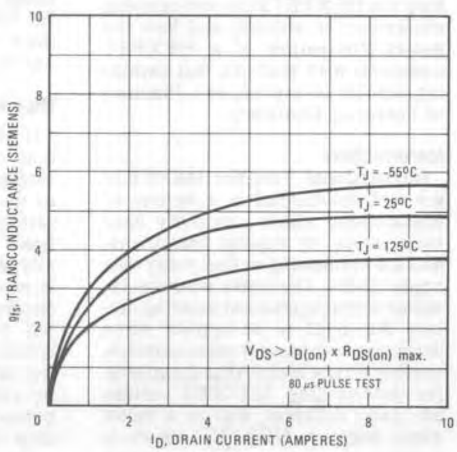


Figure 2. Transconductance Vs. Drain Current of the IRF330.

rent increases. The required heatsink DC thermal resistance decreases quite rapidly with increasing continuous drain current, for two reasons. First, permissible case-to-ambient temperature decreases; and second, power dissipation increases.

For this reason the usable *continuous direct* current of a power MOSFET for most practical purposes relates to a case temperature around 90 to 100° C. This allows a sufficient differential between case and ambient temperature for the heat dissipator to handle the heat transfer, maintaining the case temperature at or below the permitted maximum.

The "headlined" continuous current rating shown on the data sheets of most power MOSFETs is usually *greater* than the above 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 only 25° C.

Figure 5 shows typical heatsinks for TO-3 and TO-220 packaged HEXFETs 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 T_C = 25° C; the corresponding *steady* case temperature is about 100° C.

Actually, the continuous current rating of a MOSFET is often of little direct use to the designer, other than as a benchmark. This is because in many switching applications the MOSFET operates at a switching duty cycle considerably less than 100%, and what is really of interest is the current-carrying capability of the device under the *actual* "switched" operating conditions. This is discussed in the next section.

Switching "Duty Cycle" Ratings

As has been seen, the basic criterion that determines the current-carrying capability of a HEXFET is *junction heating*. For most practical purposes, the HEXFET can carry any waveform of current under any "duty cycle", just so long as the peak junction temperature is kept within the rated $T_{(Jmax)}$ (150°C). (The RMS content of the current wave must not exceed the continuous I_D rating, in order not to exceed the RMS current carrying capability of the source bonding wire. Compliance with this will generally be a natural result of compliance with the condition above.)

Peak junction temperature for any "duty cycle" application can be calculated directly from the transient thermal impedance characteristics for the device, as given in the data sheet. Transient thermal impedance curves for the IRF330 HEXFET are shown in Figure 6. Each of these curves is normalized to the steady DC junction-to-case thermal resistance (1.67 deg. C/Watts for the IRF330).

The curve labelled "single pulse" shows the rise of junction temperature per watt of power dissipation as a function of pulse duration. As expected, junction temperature rise increases as pulse duration increases — leveling off to a steady value for pulse durations above 1 second or so.

The "single pulse" curve is useful for determining transient junction temperature rise for single or very low duty cycle pulses of power; it is not *directly* usable for repetitive power pulses, such as are usually encountered in switching applications. The remaining curves in Figure 6 show effective thermal impedance for repetitive operation at different duty cycles, and allow peak junction temperature rise for repetitive operation to be calculated directly. These curves are approximately related to the single pulse curve, by the following relationship:

Effective normalized thermal impedance.
= D+(1 - D) x (normalized transient thermal impedance for single pulse of duration t).

The effective 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.

As seen from Figure 6, the effective thermal impedance for any duty cycle D increases as pulse duration increases, showing that the peak junc-

**Thermal Impedance = Normalized Value x DC Thermal Resistance*

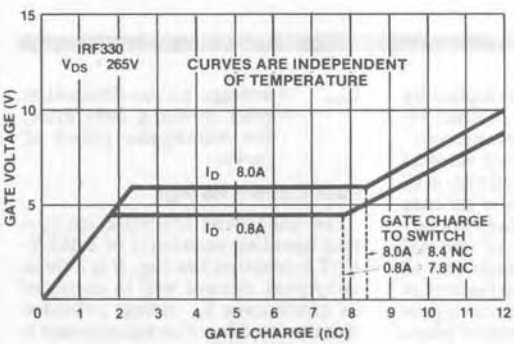


Figure 3. Typical Relationships for IRF330 HEXFET Between Gate Charge, Gate Voltage and Amplitude of Drain Current Being Switched.

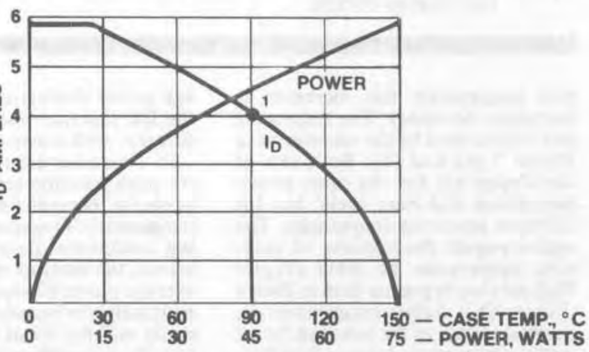
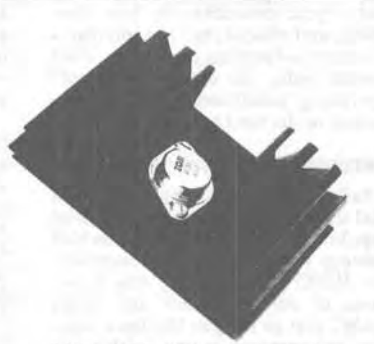


Figure 4. Case Temperature and Power as Function of I_D for IRF330 HEXFET.

(a) Type 621-A heatsink gives 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 B-4 heatsink gives 1A continuous rating for IRF710 with natural convection cooling in 40° C ambient.

Figure 5. Typical Heatsinks for HEXFETS. (Heatsinks by Wakefield)

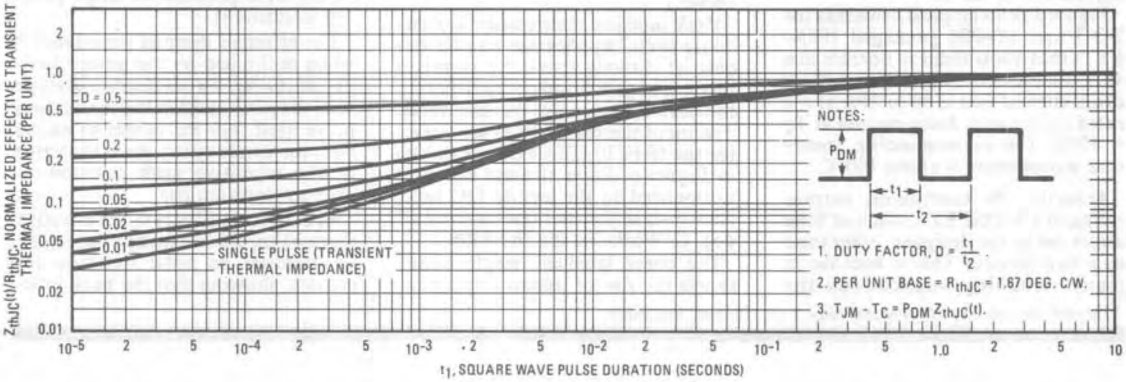


Figure 6. Normalized Transient Thermal Impedance Curves for IRF330 HEXFET. Curves are normalized to DC Thermal Resistance (1.67° C/W for IRF330).

tion temperature rise increases as frequency decreases. The reason for this is illustrated by the waveforms in Figure 7 (a) and (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 7[a]) are clearly greater than at 200Hz (Figure 7[b]). 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 usually 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.

To determine the absolute value of the peak junction temperature, it is, of course, necessary to know the case temperature T_C under steady 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:

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

where:

T_A = ambient temperature
$R_{thC\text{-}S}$ = case-to-sink thermal resistance
$R_{thS\text{-}A}$ = sink-to-ambient thermal resistance
P_{AV} = average power dissipation = peak power x duty cycle, for rectangular pulses of power

Peak Current Ratings

The underlying limitation on current handling capability of a HEXFET is junction heating. It is able to carry peak current well in excess of its continuous I_D rating, provided that the rated junction temperature is not exceeded. There is, however, an upper limit on the permissible current, defined by the rated I_{DM}. Most HEXFETs have an I_{DM} rating that is about 4X the continuous I_D rating at $T_C = 25°$ C. This is a very substantial peak current carrying capability by comparison with the I_{CM} rating of a bipolar — especially when it is rec-

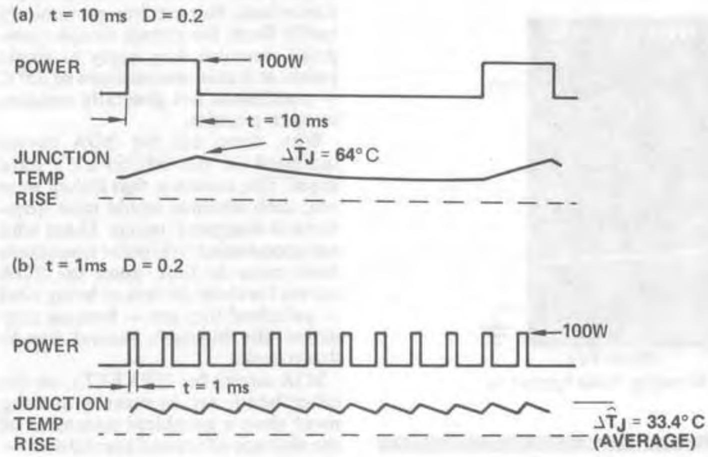


Figure 7. Waveforms of Power and Junction Temperature for Repetitive Operation, showing that Peak Junction Temperature is Function of Operating Frequency. IRF330.

ognized that the I_{DM} rating of a HEXFET is *usable*, whereas the I_{CM} rating of a bipolar generally is not.

The I_{DM} limit of a HEXFET is determined by the fact that it is, after all, fundamentally a "linear" device. As drain current increases, the point eventually is reached at which the HEXFET goes into "linear" operation and starts 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. I_{DM} ratings of all HEXFETs are achievable with an applied gate voltage that is equal to the maximum permissible gate-to-source voltage of 20V.

Designers often do not know how to interpret the I_{DM} rating. Data sheets typically give little or no supporting information, and no direct indication of whether this is a non-repetitive or repetitive rating. The fact is that the I_{DM} rating of all HEXFETs can be used *both* for *repetitive* and *non-repetitive* operation, so 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 (shown in Figure 6). The I_{DM} rating is simply a "ceiling"; below this ceiling, the designer is free to move, provided the T_{Jmax} rating is not violated.

Use of the HEXFET's peak current ratings is illustrated by the oscillograms in Figures 8 through 10. Figure 8 shows operation of the 500V rated IRF450 at a repetitive peak current of 48A. The conduction time of the rectangular current pulse is 7 μs, and the operating frequency is 1kHz. The rated continuous I_D (at T_C = 25°C) of this device is 13A, and its rated I_{DM} is 52A.

Figure 9 illustrates the use of the I_{DM} rating of the 100V IRF150 HEXFET for a "single shot" low duty cycle application, such as capacitor charging or motor starting. The peak current is 150A, decaying to 50A in approximately 10 milliseconds. Figure 10 illustrates similar duty, but in this case, the initial peak current is 100A, decreasing to 30A in approximately 400 milliseconds. The rated continuous I_D (at T_C = 25°C) of the IRF150 is 40A, and its rated I_{DM} is 160A.

It should be pointed out that the on-resistance of any MOSFET does increase as current increases. As shown in Figure 11, the on-resistance of a 100V rated HEXFET 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 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.

Safe Operating Area of MOSFET

It has been tacitly assumed so far that the HEXFET is operated as a "closed switch" in the "fully en-

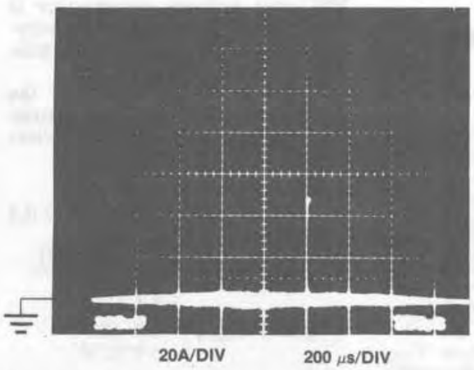


Figure 8. 48A, 7 μs Pulses Being Carried by the IRF450 at Frequency of 1kHz.

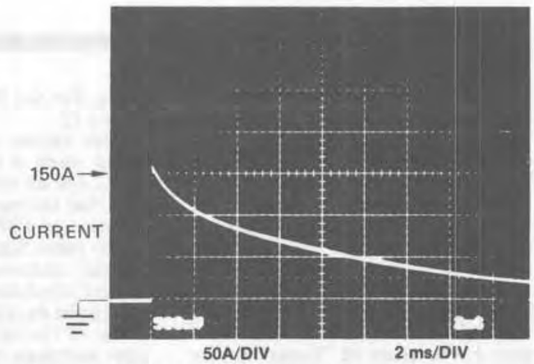


Figure 9. Single Shot 150A Peak Exponentially Decaying Pulse Applied to the IRF150, Time Constant ≐ 8msec

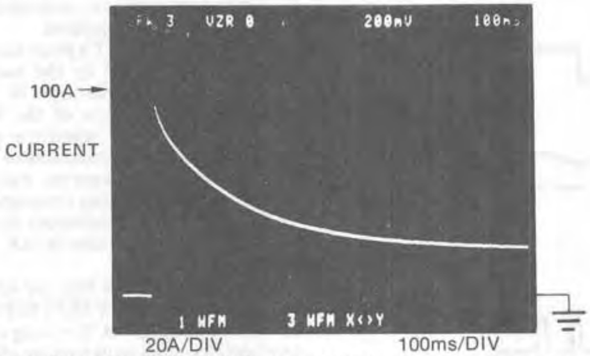


Figure 10. Single Shot 100A peak Exponentially Decaying Pulse Applied to IRF150. Time Constant ≑ 300msec

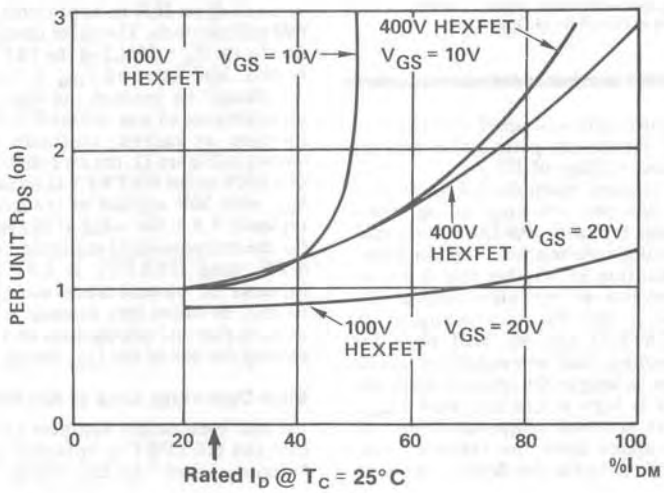


Figure 11. Typical Variation of On-Resistance with Drain Current. 100% I_{DM} ≑ 4 x I_D @ T_C = 25 °C

hanced" mode; the amount of current that the switch can handle has been shown to be calculatable for any specific design situation from a knowledge of the conduction losses, the effective transient thermal impedance, and the heatsink thermal resistance.

MOSFET data sheets generally show a graph of Safe Operating Area, for single pulses of power of varying duration, which for the most part cover areas of "linear" rather than "fully enhanced" operation. These curves embrace drain current and voltage values up to rated I_{DM} and V_{DS}, respectively. A typical SOA curve, for the IRF330, is shown in Figure 12.

SOA curves for HEXFETs are based upon a case temperature of 25°C, and an internal power dissipation that increases the junction temperature to 150°C at the end of the power pulse. Since HEXFETs, unlike bipolar transistors, do not exhibit second breakdown, SOA curves for each pulse duration invariably follow a line of constant power at all voltages less than rated maximum V_{DS} and more than the "fully enhanced" $V_{DS(on)} = I_D \times R_{DS(on)}$.

The SOA curves for HEXFETs in reality are redundant, because they can be calculated directly from the single pulse transient thermal impedance data. Nor are they particularly useful from the circuit design viewpoint, because they apply to single pulses at a case temperature of 25°C — conditions not generally encountered in practice.

Why, then, are the SOA curves included in the MOSFET's data sheet? The reason is that if they were not, their absence would raise questions in designers' minds. Users who are accustomed to bipolar transistors have come to look upon the SOA curves for these devices as being vital — as indeed they are — because they define the bipolar's second breakdown limits.

SOA curves for HEXFETs, on the other hand, are in essence nothing more than a graphical statement of the absence of second breakdown — vital information, to be sure, but information which in reality need not be conveyed through a set of somewhat arbitrary curves.

The oscillograms in Figure 13 (a) and (b) are a verification of the HEXFET's SOA data. Figure 13 (a) shows a 10 microsecond 150A pulse of current being applied to the 100V IRF150 HEXFET, with an applied drain-to-source voltage of *80V*. Figure 13 (b) shows a 10 microsecond 50A pulse of current being applied to the 500V rated IRF450 with an applied drain-to-source voltage of *400V*.

Design Examples

The following examples illustrate typical design procedures:

Repetitive Operation — 30% Duty Cycle

A 400V rated HEXFET and a corresponding heatsink are required for continuous operation with a rectangular current waveform. Amplitude of the current is 3.5A, duty cycle is 30%, and ambient temperature is 45°C. Switching losses and cycle-by-cycle fluctuations of junction temperature can be ignored.

Candidate devices would be the IRF332 and IRF320. Key ratings and characteristics for these devices are shown in Table 1.

Conduction losses for IRF332:
= 3.5 x 11.55 x 0.3
= 12.1W

$$\text{Required } R_{thJ\text{-}A} = \frac{(150 - 45)}{12.1}$$
= 8.7°C/W

Required $R_{thS\text{-}A}$ = 8.7 - 1.8
6.9°C/W

Conduction losses for IRF320:
= 3.5 x 13.9 x 0.3
= 14.6W

Required $R_{thJ\text{-}A} = \frac{(150 - 45)}{14.6}$
$= 7.2°C/W$

Required $R_{thS\text{-}A} = 7.2 - 3.2$
$= 4°C/W$

These calculations show that either of the candidate HEXFETs could serve the application. The smaller IRF320 (almost half the chip size of the IRF332) would require a relatively larger (though quite practical) heatsink, and would dissipate 14.6 instead of 1.21W giving about a 1% reduction in overall system efficiency.

The final choice of device will depend upon trade-offs between economics, size, and performance. The main purpose of this example has been to demonstrate that there *is* a choice, and that either of two HEXFET types are viable candidates.

Repetitive Operation at High Peak Current, Low Duty Cycle

It is required to find the thermal resistance of the heatsink needed to operate the 400V, 5.5A (continuous) rated IRF330 HEXFET with a repetitive rectangular current waveform of amplitude 18A. On-time is 10 microseconds, and duty cycle is 1%. Ambient temperature is 40°C.

The limiting on-resistance of the IRF330 at I_D = 5.5A at 25°C is 1.0 ohm. Knowing that 100% I_{DM} = 22A, the limiting value of $R_{DS(on)}$ at I_D = 18A can be estimated from Figure 11 to be 2.3 ohms at 25°C. From the relationship between $R_{DS(on)}$ and temperature given in the data sheet, $R_{DS(on)}$ at T_J = 150°C and I_D = 18A will be about 5.1 ohms.

Power per pulse $= 18^2 \times 5.1$
$= 1.652 \times 10^3 W$

Junction-to-case transient thermal impedance for 10 μs pulse (from Figure 6):
$= 1.67 \times 0.03$
$= 0.05°C/W$

Table 1. Design details for IRF332 and IRF320 HEXFET's

		IRF332	IRF320
V_{DS}	Volts	400	400
I_D @ T_C = 25°C	Amps	4.5	3.0
$V_{DS(on)}$ @ 5A, 150°C	Volts	11.55	13.9
$R_{thJ\text{-}C}$	°C/W	1.67	3.12
$R_{thC\text{-}S}$	°C/W	0.2	0.2
Approximate die size	mil²	19,250	11,700

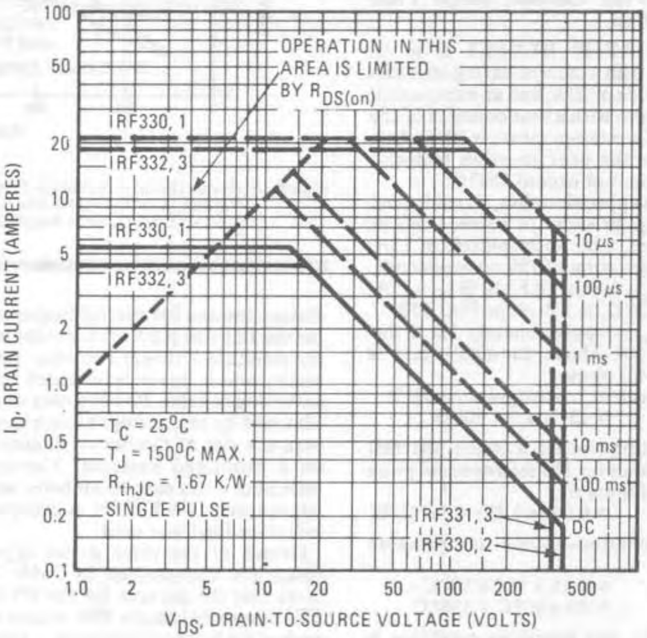
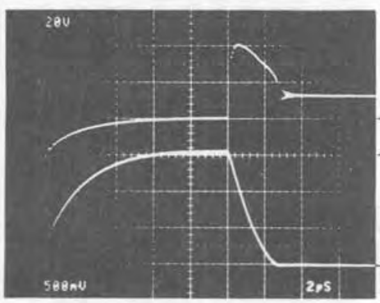
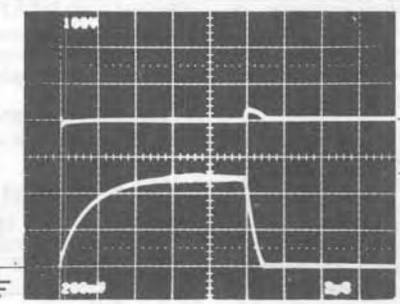


Figure 12. Safe Operating Area, IRF330 HEXFET.



(a) IRF150, 50A/div, 2μs/div



(b) IRF450, 20A/div, 2μs/div

Figure 13. Oscillograms Verifying SOA Curves of HEXFETs, Demonstrating the Absence of Second Breakdown

Junction-to-case temperature rise due to 18A pulse:

$= 1.652 \times 10^3 \times 0.05$
$= 82.6°C$

Maximum permissible case temperature:

$= 150 - 82.6 = 67.4°C$

$T_C - T_A = 67.4 - 40 = 27.4°C$

Average power dissipation:

$= 0.01 \times 1.652 \times 10^3$
$= 16.52W$

$$\therefore R_{thC\text{-}A} = \frac{27.4}{16.52} = 1.66°C/W$$

High Peak Current, Single Pulse Operation

The IRF330 HEXFET is to be pulsed with a current having an initial amplitude of 20A, and an exponential waveform with a time constant of 150 μsec. Case temperature is 30°C. Verify that the peak junction temperature does not exceed 150°C.

As an approximation, an equivalent rectangular pulse of current will be assumed, with an amplitude of 15A, and a duration of 150 microseconds.

$R_{DS(on)}$ for the IRF330 @ $I_D = 15A$, $T_J = 25°C$, is 1.8 ohms.(1) At 150°C, $R_{DS(on)}$ is approximately 2.2 x this value (see above example), and is about 4.0 ohms.

Equivalent "rectangular power":

$= 15^2 \times 4.0 = 900W$

Junction-to-case transient thermal impedance for 150 microsecond pulse (from Figure 6):

$= 0.065 \times 1.67 = 0.11°C/W$

∴ Junction-to-case temperature rise:

$= 0.11 \times 900 = 99°C$

$\therefore T_J = 30 + 99°C = 129°C$

Hence, this operating condition is within the capability of the IRF330.

Comparison of MOSFET and Bipolar Losses

Conduction power in a bipolar transistor is generally lower than in a MOSFET, but switching energy is usually considerably higher. The bipolar, therefore, tends to be more efficient at low frequency, while the MOSFET is more efficient at high frequency.

In an effort to close the gap between bipolar and MOSFET performance in high frequency switching applications, several new types of fast switching bipolar transistors have recently been introduced, with switching times in the order to 100 to 200 nanoseconds. It is pertinent to compare the losses of these new bipolar types with those of comparably rated MOSFETs.

Figure 14 shows measured power dissipation as a function of frequency for the IRF330 HEXFET, the industry-standard 2N6542/3 bipolar transistor, and a newly introduced fast switching bipolar. Power losses were obtained by measuring the case temperature rise of the device mounted on a calibrated heatsink. Thermal resistance from case-to-ambient was approximately 4.5°C/W. A clamped inductive load was used.

Details of the three device types listed are summarized in Table 2. Note that the die area for the HEXFET is approximately 80% of that of each of the bipolar transistors — thus, the comparison is weighted in favor of the larger-die bipolar devices.

Figure 14 shows that the frequency crossover point for the HEXFET and the 2N6542/3 is approximately 25kHz, while it is approximately 35kHz for the HEXFET and the fast-switching bipolar transistor. Operating conditions were: circuit supply voltage = 270V, peak current = 2.5A, duty cycle = 33%.

Note that the "full" curves represent *only* the dissipation *within* the device. Additional power is dissipated in the external base drive circuit of the bipolar. The "dashed" curve for the fast-switching bipolar includes an additional 1.3W of external base drive power. This corresponds to an 8V, 0.5A drive circuit, operating at 33% duty cycle.

Figure 15 shows collector current and voltage oscillograms for the fast-switching bipolar transistor operating at 100kHz, and Figure 16 shows drain current and voltage oscillograms for the HEXFET at 100kHz. Note the sharper HEXFET waveforms, confirming its faster switching speed.

Oscillograms of base drive current for the bipolar and gate drive current for the HEXFET are shown in Figure 17(a) and (b), respectively. The bipolar requires a significant base drive current both at turn-on — about 1A peak — and at turn-off — 2.5A peak. The HEXFET, by comparison, consumes about 0.3A for a few nanoseconds at turn-on, and about 0.2A for a few nanoseconds at turn-off. This current charges and

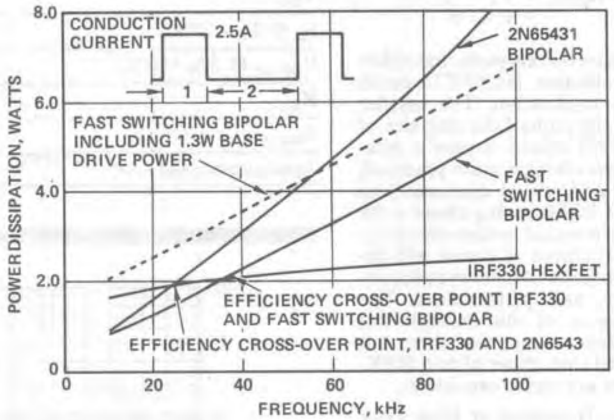
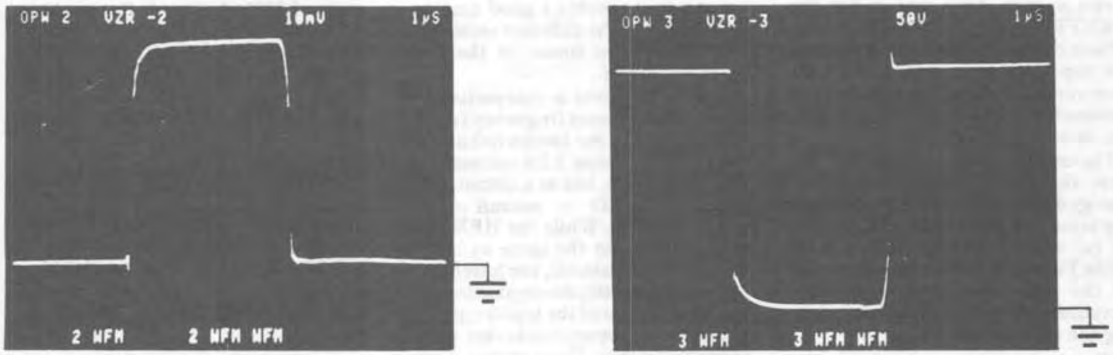


Figure 14. Power Dissipation Versus Frequency for 2N6542/3, Fast Switching Bipolar, and IRF330 HEXFET. Supply Voltage = 270V. Conduction Duty Cycle = 0.33. Current Amplitude = 2.5A.

Table 2. Details of Devices Tested

	IRF330 HEXFET	2N6542/3 Bipolar	Fast-Switching Bipolar	
V_{DS} Volts	400	400	450	$V_{CEO(SUS)}$
$I_{D\ cont}$ A @ $T_C = 25°C$	6	5	5	$I_{C\ cont}$ A @ $T_C = 25°C$
Die Area mil²	19,500	25,000	25,000	Die Area mil²

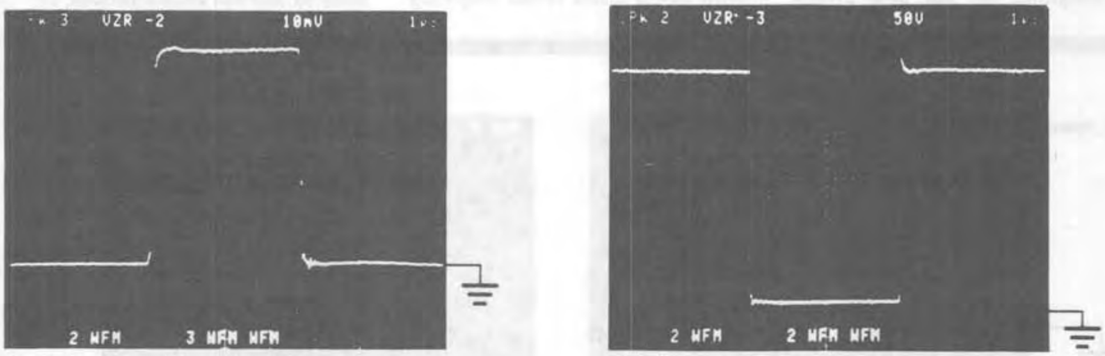
(1)The effect of current on on-resistance can be calculated with the help of Figure 11.



(a) Collector Current. 0.5A/Div, 1μs/Div

(b) Collector Voltage. 50V/Div, 1μs/Div

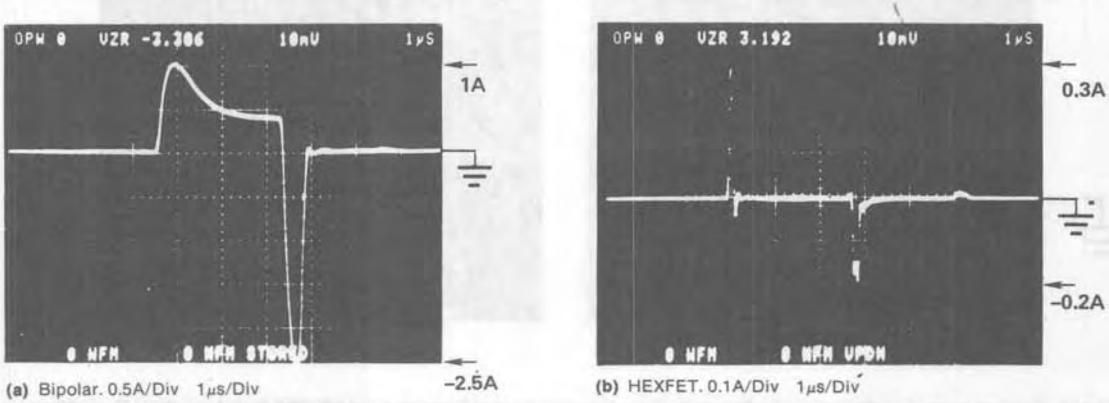
Figure 15. Collector Current and Voltage Waveforms for Fast-Switching Bipolar Operating at 100kHz.



(a) Drain Current. 0.5A/Div, 1μs/Div

(b) Collector Voltage. 50V/Div, 1μs/Div

Figure 16. Drain Current and Voltage Waveforms for IRF330 HEXFET Operating at 100kHz.



(a) Bipolar. 0.5A/Div 1μs/Div

(b) HEXFET. 0.1A/Div 1μs/Div

Figure 17. **(a)** Base Drive Current for Fast Switching Bipolar Operating at 2.5A, 100kHz. V_{CC} = 270V

(b) Gate Drive Current for IRF330 HEXFET Operating at 2.5A, 100kHz. V_{DD} = 270V.

discharges the self-capacitance of the device. Note the change of current scale between Figure 17 (a) and (b). Average gate drive power for the HEXFET is negligible — about one-fiftieth of a watt at 100kHz. Although the bipolar is driven with 1A peak base current, it nonetheless exhibits a noticeable voltage "tailing" at turn-on, as seen in Figure 15 (b).

The oscillograms in Figure 18 compare the instantaneous power and energy dissipation for the fast-switching bipolar and the HEXFET. Figure 18 (a) shows instantaneous power, while Figure 18 (b) shows the integral of the power; in other words, the accumulated energy dissipated during the conduction period.

Clearly the energy expended in the bipolar at turn-on and at turn-off is greater than in the HEXFET, while the energy expended in the HEXFET during the conduction period is greater than in the bipolar. These oscillograms do not give precise quantitative data because of the lack of resolution of the oscilloscope at these fast-switching speeds; they do, nonetheless, provide a good qualitative picture of the different switching and conduction losses in the two types of devices.

Figure 19 shows a comparison of power losses versus frequency for the HEXFET and the fast-switching bipolar, for the same 2.5A current and 33% duty cycle, but at a circuit voltage of only 70V — instead of the previous 270V. While the HEXFET losses are about the same as in the higher voltage circuit, the lower supply voltage greatly de-emphasizes the switching losses of the bipolar, giving a higher frequency crossover point (almost 70kHz). These curves, however, are not representative of a typical operating situation, since the 70V circuit voltage is unrealistically low for 400 to 450V rated devices.

Finally, the curves in Figure 20 show power losses versus frequency for the HEXFET and the fast-switching bipolar, operating at a peak current of 5A in a 270V circuit, at a duty cycle of 33%. Although the conduction losses of the HEXFET are more than 4x greater than with I_D = 2.5A, the switching losses of the bipolar are also significantly greater. Additionally, the bipolar's base drive current has to be increased significantly to maintain acceptable switching performance, as shown by the oscillogram in Figure 21. Interestingly, the frequency "crossover point" is not greatly different from that obtained at 2.5A — about 42kHz versus 35kHz, ignoring external base drive power, and about 20kHz, taking this into account.

Conclusions

The main purpose of this application note has been to show that the current-carrying capability of a power MOSFET is determined essentially by *thermal* considerations, un-

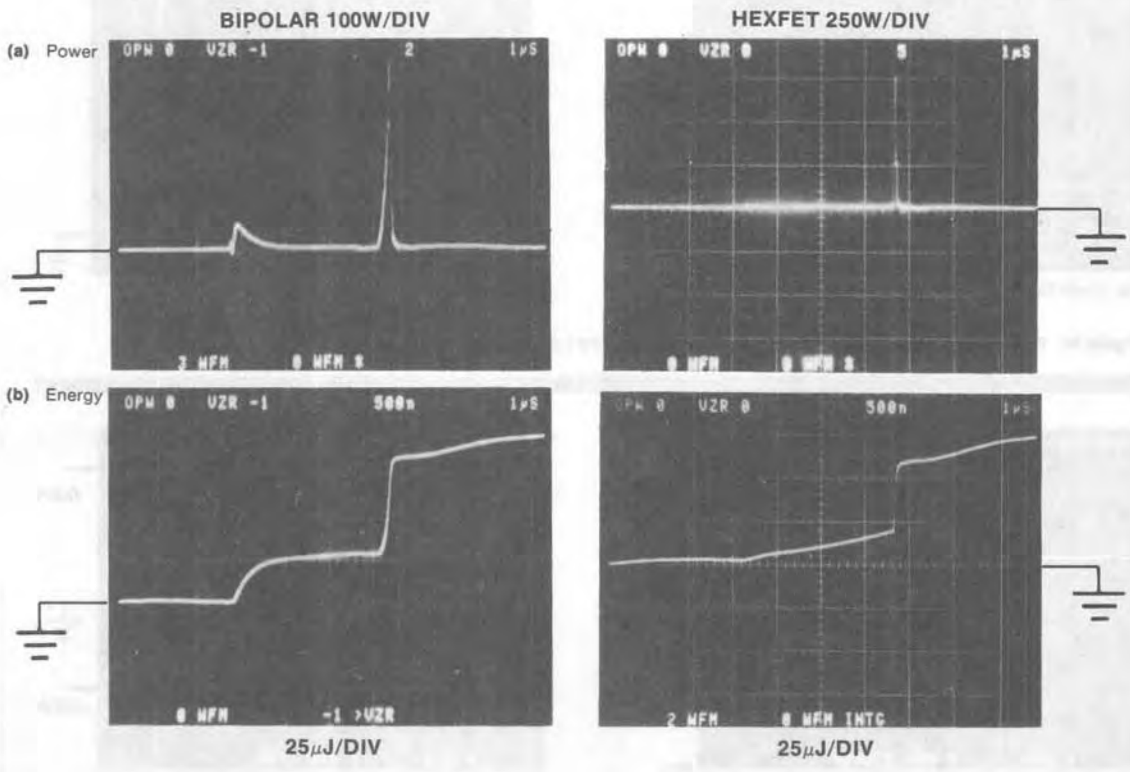


Figure 18. Oscillographs of (a) Power and (b) Energy for Fast-Switching Bipolar Transistor and HEXFET from Turn-On to Turn-Off. Circuit Voltage = 270V. Switched Current = 2.5A. Duty Cycle = 33%. Frequency = 100 kHz.

like that of the bipolar transistor, which is limited by *gain*. With proper thermal design, the HEXFET can be operated at much higher peak current than a comparable bipolar.

A practical comparison of power losses of a HEXFET and a fast-switching bipolar transistor in a switching application shows that the HEXFET is generally more efficient above frequencies in the 20 to 40 kHz range. □

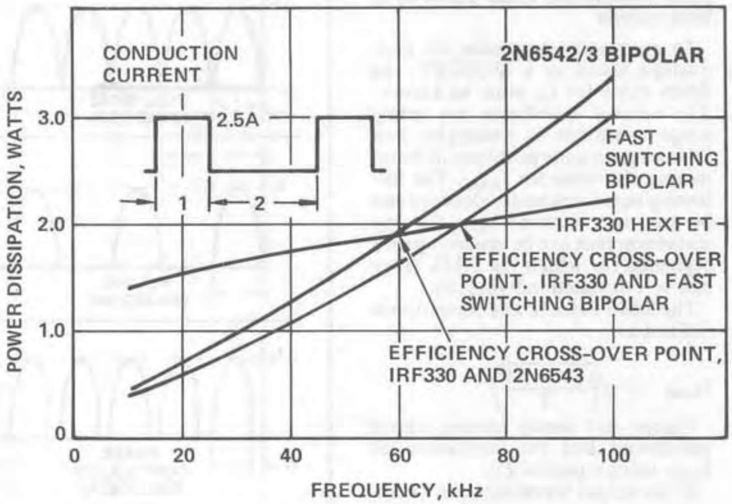


Figure 19. Power Dissipation Versus Frequency for 2N6542/3, Fast-Switching Bipolar and IRF330 HEXFET. Supply Voltage = 70V. Conduction Duty Cycle = 0.33. Current Amplitude = 2.5A.

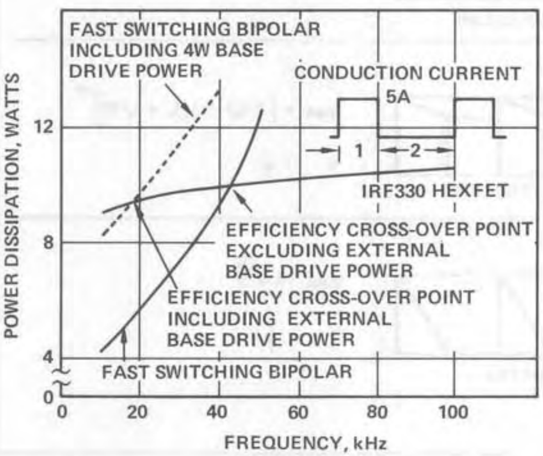


Figure 20. Power Dissipation Versus Frequency for Fast-Switching Bipolar and IRF330. Supply Voltage = 270V. Conduction Duty Cycle = 0.33. Current Amplitude = 5A.

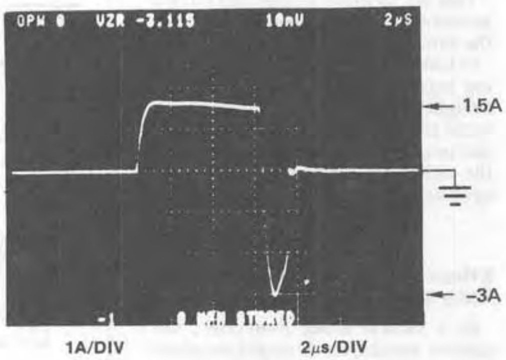


Figure 21. Base Drive Current for Fast-Switching Bipolar. Circuit Voltage = 270V. Collector Current = 5A. Operating Frequency = 50kHz.

APPENDIX

Determining the RMS Value of I_D Waveforms

To accurately determine the conduction losses in a MOSFET, the RMS value for I_D 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 I_{RMS} 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:

$$I_{RMS} = \sqrt{\left(\frac{\int_o^t I^2(t)\,dt}{T}\right)}$$

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:

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

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 to some extent. This can be very beneficial in reducing the value for I_{RMS} in the switch for a given value of average current (I_{AVG}).

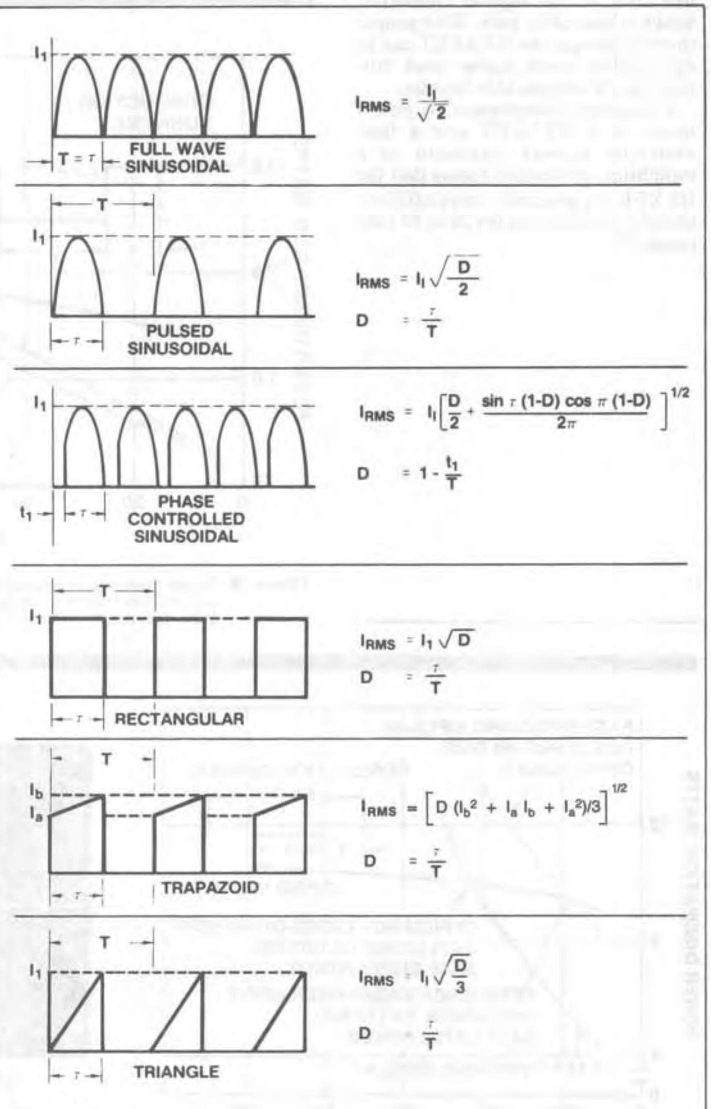


Figure A-1.

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 the 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:

$$K = \frac{I_a}{I_b} = f\,(L/L_c) \quad (3)$$

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:

$$I_{RMS} = 2\sqrt{D}\,I_{avg}\sqrt{\frac{1 + K + K^2}{3\,(K+1)^2}} \quad (4)$$

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 need only be about twice L_c.

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 dominate, 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:

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

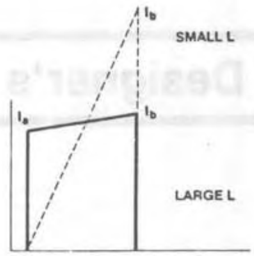


Figure A-2. Current Waveform

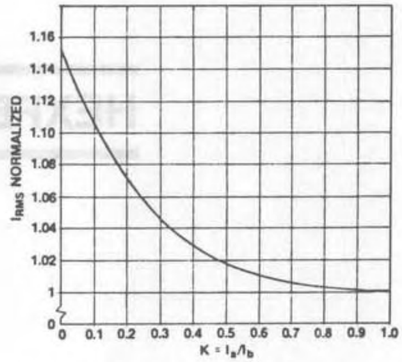


Figure A-3. Variation of I_{RMS} with Squareness Ratio