50 - 100v DC INVERTERS FOR VEHICULAR ACTUATOR SYSTEMS;- THE USE OF MOSFETs AND IGBTs COMPARED.

Fenton L. Rees
Consultant
Everett, WA

John M. Miller
FORD Motor Co
Scientific Research Lab.
Dearborn, MI

ABSTRACT.

The use of either MOSFETs or IGBTs for a generic 3kw inverter/motor/actuator system operating from either a 100v or 50v DC bus was investigated. The aim was to determine which semiconductor type resulted in the least costly design, when the semiconductors were used in such a way that the total inverter losses and EMI/RFI impact were comparable. When 200v MOSFETs and 250v IGBTs were compared for operation from the 100v DC bus, it was found that the cost of the IGBT based design was approximately half that when MOSFETs were used. For operation from the 50v DC bus, 100v MOSFETs were compared to the same 250v IGBTs, (lowest voltage rating available). In this case the advantage that the IGBT had at 100v was not as evident, and depended on the particular load condition and PWM frequency used.

INTRODUCTION.

The actuators required by such functions as power steering or 4 wheel steering have historically been powered by engine driven hydraulics. Because this results in a load on the engine even when the actuator is at rest, there is a significant impact on overall fuel economy. Similarly, air-conditioning has been typically accomplished by an engine driven compressor, the size of which is mainly governed by the air-conditioning capacity required at engine idle. The system is therefore typically oversized for non idle speeds, resulting in less than optimal efficiency, and a greater impact on fuel economy than desired. Both of the above functions can be implemented by using a solid state variable frequency inverter controlling either an induction motor or a PM DC motor. In the case of the actuator, it only consumes significant power while actually moving, resulting in an average engine load lower than that obtainable with the hydraulic system.

Likewise an electric air-conditioner allows the maximum capacity to be set independent of engine speed, and therefore optimally sized. Future actuator functions will also benefit from the use of the variable speed electric motor topology.

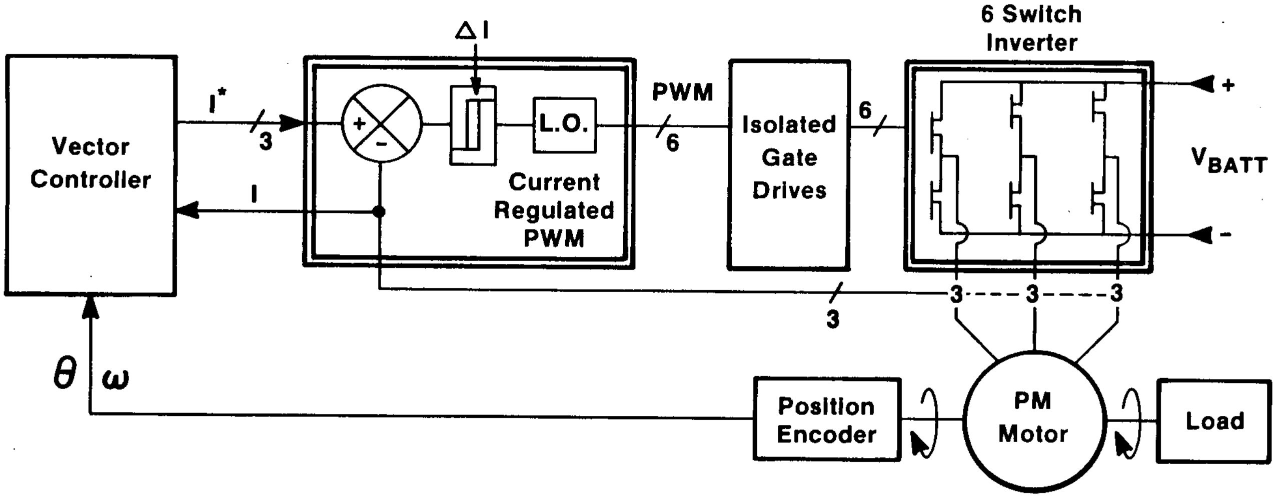

Figure 1: Simplified Block Diagram of the Inverter and Motor of the Generic Actuator System

The last few years have seen the convergence of a number of factors that increasingly make the use of electrically driven actuators a realistic alternative. These include:

- The better performance and lower cost of the MOSFET and IGBT semiconductor switches.
- Microprocessor and ASIC technology which allows the necessary controls to be implemented relatively inexpensively.
- The development of less costly rare earth magnets for the PM motors.
- And the continuing requirement for ever better fuel economy.

To better understand the performance benefits to be gained from using electric actuation, and the potential obstacles to its cost effective implementation, a generic inverter/motor/actuator system was built. The initial design called for operation from a 100v DC bus, a peak power output of 3kW, and accurate control of torque. MOSFETs were initially chosen as the power semiconductor switch, partly as a result of the DC bus voltage, and partly because of the expected maximum PWM frequency of 30kHz. As the project progressed and the capabilities of IGBTs continued to improve, it became obvious that IGBT's could perhaps be a better choice of power semiconductor. This paper discusses the relative advantages of using either MOSFETs or IGBTs for a DC bus voltage of 100v, and then looks at the effects of reducing the DC bus voltage to 50v.

Initial Inverter Design Using 200V MOSFETs

Figure 1 is a block diagram of the inverter, motor and the associated controls. The key elements of the design included:

- Operation from a nominally 100v DC bus, where the bus voltage could rise to a maximum of 140v during regeneration. MOSFETs with a voltage rating of at least 200v were therefore required.
- Accurate control of torque and full four quadrant operation. Vector control of the motor was therefore necessary, and the hysteresis method of PWM was chosen so that motor current could be accurately controlled under all dynamic conditions.
- A Permanent magnet motor with Neodymium-Boron-Iron (NdBFe) magnets in order to achieve the required power to weight ratio at reasonable cost.
- A load vs. time profile such that for the vast majority of the time the actuator was operating at approx. 10% torque, with the maximum torque required for only short periods. Inverter efficiency at light load was therefore important, both to minimize the heatsinking requirements and to effect overall fuel economy as little as possible.
- Forced air cooling of the inverter, with a maximum ambient air temperature of 65°C.

At low values of torque, accurate control of the torque and therefore the motor current was deemed to be necessary, and so a relatively small hysteresis current band of ±2A was chosen. At maximum motor torque, where the peak motor current was 130A, it was thought that a larger current band of ±8A would still result in acceptable torque accuracy. Because PWM frequency and the size of the current band are inversely related for hysteresis current band PWM, the larger current band at 130A would allow the switching losses to be reduced. For the experimental inverter it was decided that the International Rectifier IRFC260 size 6 die would be a convenient building block. Rated at 200v and 0.06Ω (at 25°C), many die in parallel per inverter switch were obviously required. With a $R_{DS(on)}$ of approx. 0.102Ω at 100°C, a $R_{\theta(J\text{-}C)}$ of 0.7°C/w (assuming an alumina substrate), and with a maximum expected heatsink temperature of 80°C at the highest ambient, it was decided to use SIX die in parallel in order to conduct the maximum RMS motor current of $130/\sqrt{2} = 92$A and still have T_J substantially less than 125°. (ie. When the larger ±8A current band is used at a motor current of 130A, the switching frequency is low enough [3kHz] that the switching losses can be almost be ignored compared to the conduction losses.)

Some power circuits, such as DC to AC inverters, require reverse current flow through the active device. When bipolar transistors or IGBTs are used, it is necessary to add a fast recovery diode in "anti-parallel" across the device. This diode must exhibit good reverse recovery characteristics with a small reverse recovered charge to keep the power dissipation low, and a reverse recovery waveform that does not contain an abrupt change in current.(ie. snap-off). When power MOSFETs are used as the semiconductor switch in a DC - AC inverter, the possibility of using the reverse conducting diode inherent in the structure becomes attractive because it eliminates the complexity and cost of adding an external diode. However early generations of MOSFETs had a tendency to fail if the body diode was used to conduct freewheel current and then subject to significant dV/dT at the time of diode recovery [1]. Since the advent in 1987 of "3rd generation" devices with specified avalanche ratings, re-applied dV/dT's up to about 5V/ns have been permissible [2]. If it was found that the MOSFET's body diode could not be used, the addition of an external fast recovery diode would also require the addition of another diode in series with the MOSFET in order to prevent the body diode conducting in parallel with the added diode; Figure 2. Typically the diode in series with the MOSFET is a Schottky, because it's peak reverse voltage is limited by the added anti-parallel diode.

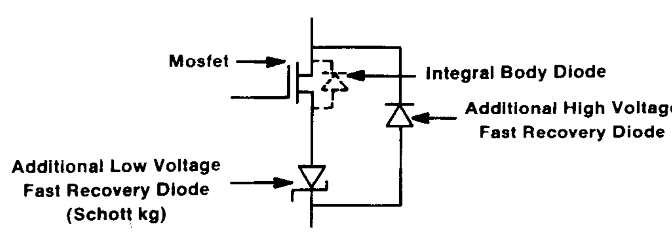

Figure 2: Additional Circuitry Required If the Mosfet's Integral Body Diode has to be Prevented from Conducting Freewheel Current

For the above described experimental inverter, the possibility of using the body diode for freewheeling was doubly attractive;- not only because it eliminated the need for an added diode, but also because the added components shown in Fig 2 result in much higher conduction losses at low load currents, when $I * R_{DS(on)}$ is less than the V_F of a diode. It has already been mentioned that the losses at low load currents were considered to be important because they set both the heatsinking requirements and the average efficiency.

It was found that the standard IRFC260 die exhibited relatively poor reverse recovery characteristics, with a t_{rr} of about 350ns (25°C, 60A/us) and a somewhat "snappy" recovery. To improve upon this performance, I-R supplied die that had been electron irradiated (12 to 18 Megarads), in accordance with established procedure [3]. The t_{rr} was approximately halved and the snappiness somewhat reduced.

As a first step towards integrating the entire inverter, the 12 MOSFET die for one phase and the associated gate drivers were packaged into a power module by Sanyo, the simplified schematic of which is shown in Figure 3. The key features of this FORD/Sanyo module are:

- The use of the I-R IR2110 gate driver IC configured to provide a +10v/-5v gate signal.
- Minimal added impedance at turn-off in order to minimize turn-off losses.
- A relatively large gate resistance at turn-on in order to control the re-applied dV/dT at reverse recovery.
- In a module with baseplate dimensions of 4.5" X 2.6" and 0.7" high.

To further improve the waveforms at turn-on and to help control the peak V_{DS} voltage at turn-off, it was found necessary to added both a 30uF MLC decoupling capacitor directly across the DC bus at the module, and a 6Ω / 0.03uF RC snubber across each switch. The resulting turn-off, reverse recovery and turn-on waveforms achieved are shown in Fig 3, where it can be seen that the waveforms are reasonably free from excessive ringing, the dV/dT's are quite acceptable, but the turn-on energy is moderately high. Operation of the entire inverter/motor/actuator system with three of the above describes inverter phase modules has to date been reasonably successful. Measurement of the switching losses at lower current levels allows the various components to be expressed as:

$$E_{OFF} \approx 0.023\, I^{0.8} \quad E_{ON} \approx 0.05\, I^{1.25} \quad E_{SNUBBER} \approx 0.3$$

all in mJ/cycle at T_J = 100°C for a DC bus voltage of 100v.

TESTING OF A 250V IGBT.

To help FORD evaluate the possibility of using IGBTs in the above described inverter/motor/actuator system, I-R supplied some experimental "fast" 250v "size 5" IGBT die packaged in TO3s. To be comparable to the 200v MOSFETs, an IGBT with a voltage rating of at least 250v was required in order to compensate for the IGBT's lack of an avalanche rating. 250v also happened to be the lowest voltage rating for the IGBTs available from I-R.

In order to conduct the peak motor current of 130A with an acceptable T_J, at least 2 or 3 die in parallel are obviously required. Consequently the $V_{CE(sat)}$ was not measured beyond a current level of 45A. Figure 4 shows the results of measuring $V_{CE(sat)}$ as a function of current at both 25 and 100°C, where it can be seen that the effect of temperature on $V_{CE(sat)}$ is surprisingly minimal. $V_{CE(sat)}$ at T_J = 100°C can then be represented by:

$$V_{CE(sat)} = 0.58 + 0.136\, I^{0.645}$$

Using the circuit of Figure 5a, the switching energies of the 250V size 5 IGBT were evaluated. The test circuit used:

- Half of an I-R 25JPF30 dual ultra fast recovery diode module, where each diode is rated at 300v, 25A, with t_{rr} (@25°C) of 60ns.
- A FUJI EXB841 hybrid IGBT gate driver IC, which gave a gate drive voltage of +15v/-5v.
- A 0.02Ω non-inductive current shunt to measure the IGBT current.
- A 30uF 300v MLC decoupling capacitor as close as possible to the IGBT and freewheel diode.

Much care was taken in ensuring the leakage inductance around the loop comprised of the IGBT, the 25JPF30 diode, the shunt and the MLC capacitor was as small as possible. The loop containing the above 4 components in the final configuration had a diameter of approximately 1½".

In order to fully utilize the speed capabilities of the IGBT, so that the turn-on and turn-off energies were minimized, I-R initially recommended using a gate resistance of 3Ω. Unfortunately however it was found that such a small value resulted in a totally unacceptable amount of ringing, in both the voltage and current waveforms. Figs 5b and 5c show switching waveforms obtained when R_G = 100Ω. It can be seen that:
- A small amount of ringing (at 25MHz) is still present on both of the current waveforms.
- Snubbers across each switch were NOT used.
- At TURN-ON:
    - ■ The initial drop in V_{CE} of about 10v when the dI/dT is 160A/us allows a leakage inductance of 63nH to be calculated.
    - ■ The relatively high value of R_G results in a relatively long "dynamic saturation" region for V_{CE}.
    - ■ The peak dV/dT of about 2V/ns (which approximates that which would be seen across the fast recovery diode as it recovers) is fairly reasonable. The corresponding value for the MOSFET module was 0.8V/ns (Fig 3)
    - ■ The measured E_{ON} of 1.0mJ is much less than that for the MOSFET module, once the lower current use has been allowed for.

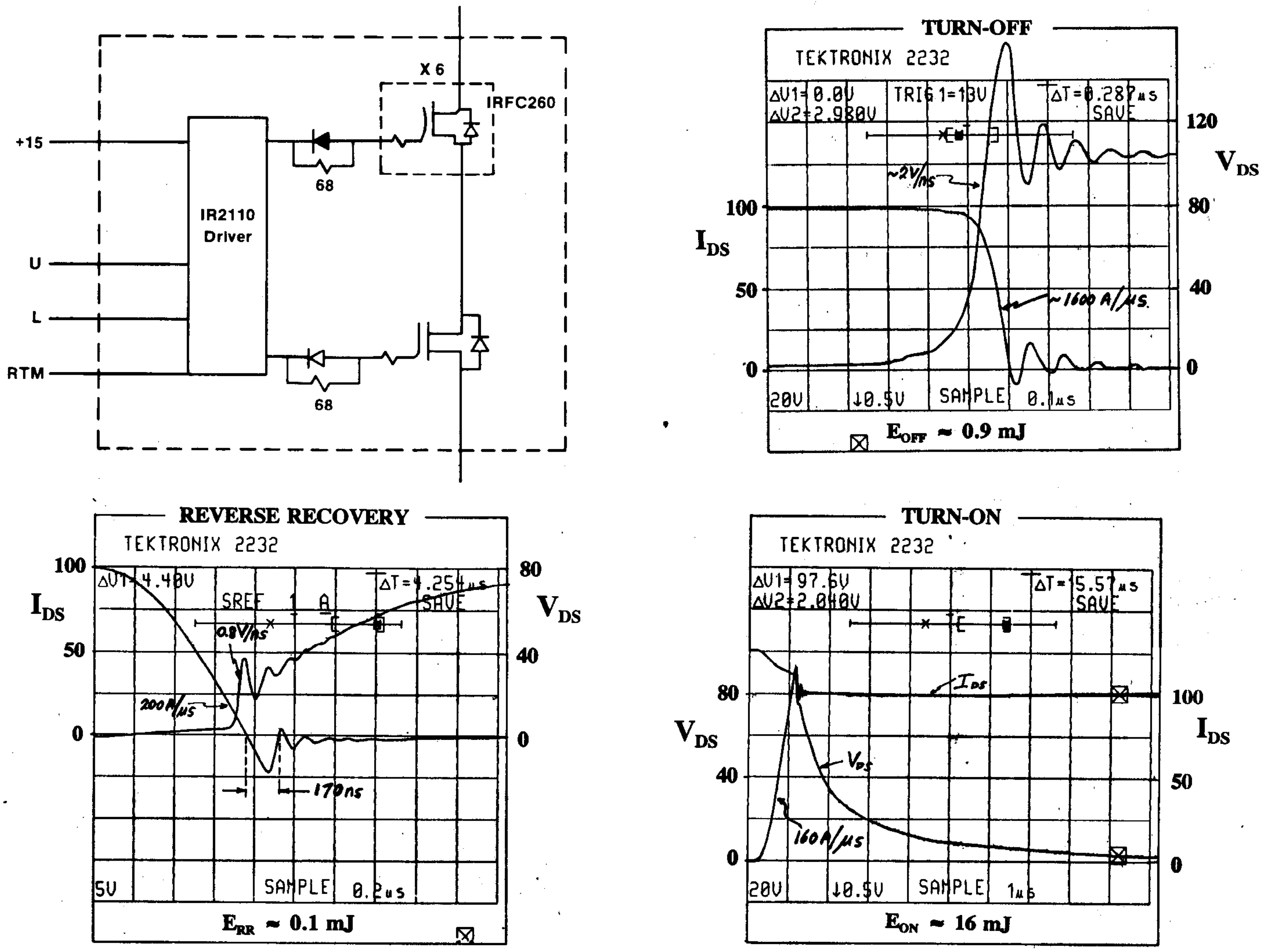

Figure 3. Ford/Sanyo Integrated Power Module for One Phase of Inverter; Simplified Schematic, Switching Waveforms for V_{Bus} = 100 V, I_{Load} = 100 Amp, T_j = 100 C.

- At TURN-OFF:

- The amount of overshoot in V_{CE} (40v above the 100v bus), during the period of maximum dI/dT (≈ 1000A/us), allows a leakage inductance of 40nH to be calculated.
- The time for the IGBT current to fall to zero is a quite respectable 300ns.
- The peak dV_{CE}/dT of 3V/ns compares to the 2V/ns for the MOSFET module.
- The measured E_{OFF} of 0.51mJ is comparable to that of the MOSFET allowing for the lower current used.

It can therefore be concluded that the leakage inductance for the test set-up is in the range 40 - 60nH, probably quite representative of what would be achievable if multiple die were assembled in a power hybrid module like that used for the IRFC260 MOSFET die. Subsequent discussions with I-R lead to the conclusion that the observed oscillations at 25MHz were caused by resonance between the leakage inductance and the output capacitance of the IGBT. Reducing R_G from 100Ω to 25Ω was found to increase the amplitude of the 25MHz ringing, result in only a small reduction in E_{OFF}, but cause E_{ON} to be halved. The switching losses were thought to be small enough that any further reduction at the expense of increased ringing was not warranted. Measurement of the switching losses at lower current levels allows the E_{ON} and E_{OFF} to be expressed as:

$E_{ON} = 0.012\ I^{1.2}$ $\quad E_{OFF} = 0.0042\ I^{1.3}$

mJ/cycle at 100°C

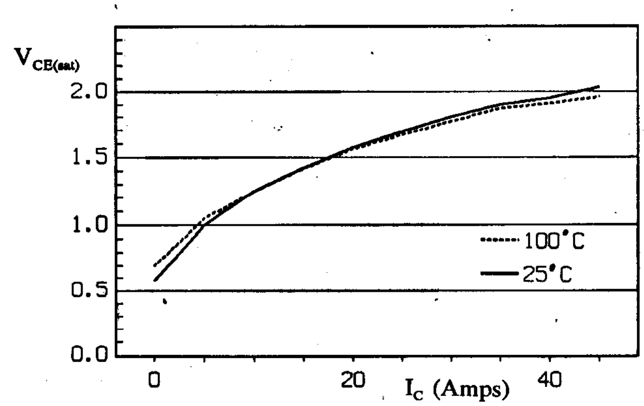

Figure 4. $V_{CE(sat)}$ Characteristics of the 250 V Size 5 IGBT.

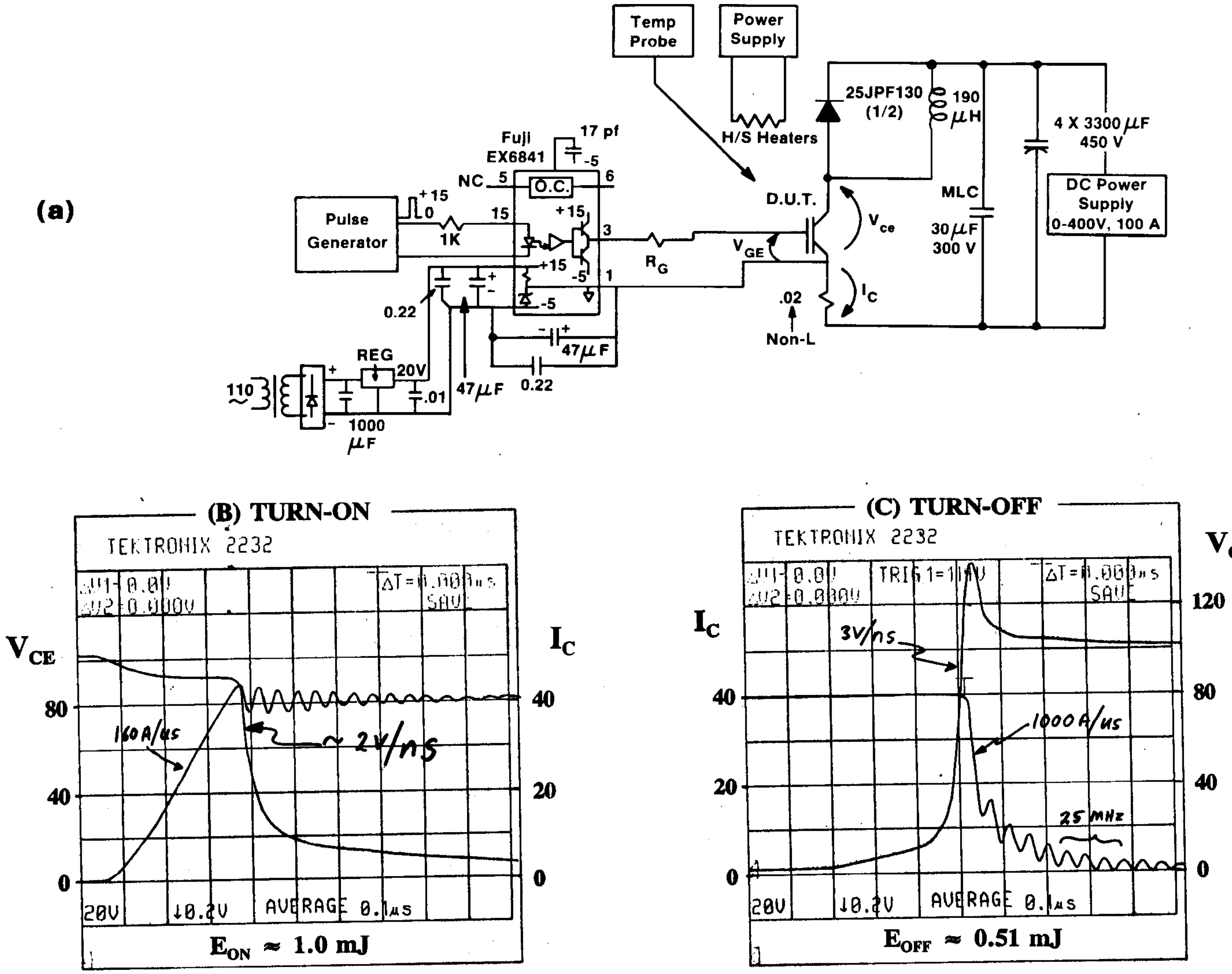

Figure 5. Experimental I-R 250 V Size 5 IGBT; Measurements of Switching Energies. (A) Test Circuit (B) Turn ON Energy for: V_{Bus} = 100 V, I_{Load} = 40 Amp, T_J = 100 C. (C) Turn OFF Energy for: R_G = 100 Ohm, V_{GE} = + 15/-5 V.

SIMULATED USE OF MOSFETS AND IGBTS IN THE ACTUATOR SYSTEM.

A computer program was written that modelled the action of the 6 switch inverter with hysteresis band PWM driving the actuator motor. The current waveform over one fundamental cycle so generated then allowed the switching and conduction losses to be computed for either type of semiconductor switch. Two load points were of particular interest:

- Operation at maximum torque, which occurs for only brief periods, => peak motor current of 130A, 600 rpm, larger current band of ±8A.
- Operation at approx 10% max torque, more representative of the average load, => 2000 rpm, 14A peak, smaller current band of ±2A.

For the FORD/Sanyo MOSFET module, complimentary gating of the upper and lower switches was assumed, so that at low load currents synchronous rectification of the freewheel current could take place. The required equations for the calculation of the inverter losses were then:

$$\left.\begin{array}{l} R_{DS(on)} = 0.017\Omega \\ V_F = 0.56 + 0.00173\, I_{DS} \\ E_{ON} = 0.05\, I^{1.25} \text{ mJ/cycle} \\ E_{OFF} = 0.023\, I^{0.8} \text{ mJ/cycle} \\ E_{SNUBBER} = 0.3 \text{ mJ/cycle} \end{array}\right) \begin{array}{l} V_{BUS} = 100\text{v} \\ T_J = 100°C \end{array}$$

Where the derivation of all except V_F have been previously described, and where V_F is the typical data sheet characteristic for 6 IRFC260 die in parallel.

The simulation was designed so that the effect of having various numbers of the size 5 IGBT die in parallel could be investigated. It was assumed that each IRFC260 die had a 25JPF30 fast recovery diode in inverse parallel. Including the typical V_F characteristic for this diode results in the following equations for the IGBT losses:

$V_{CE(sat)} = 0.58 + 0.136\ I^{0.645}$)
$V_F = 0.80 + 0.017\ I_F$) per size 5 IGBT and diode.
$E_{ON} = 0.012\ I^{1.2}$ mJ/cycle) $V_{BUS} = 100v$, $T_J = 100°C$
$E_{OFF} = 0.0042\ I^{1.3}$ mJ/cycle)

It should be noted that for both the FORD/Sanyo MOSFET module and the individual IGBT die, losses associated with the reverse recovery of the diode were ignored, because they were insignificant when compared to the combined turn-on and turn-off losses. The results of the simulation are given in Table 1.

A number of conclusions can be drawn from the results of Table 1:

1/. For the high levels of load current that occur occasionally, between two and three of the size 5 IGBT die in parallel per switch give the same total inverter loss as that for the FORD/Sanyo module with 6 size 6 MOSFET die in parallel. In both cases the conduction losses predominate as the larger current band of ±8A results in the relatively low switching frequency of 3.2kHz.

2/. For low levels of load current, only 2 of the IGBT die in parallel are required to produce total inverter losses that are lower than that for the FORD/Sanyo module. For the FORD/Sanyo module, the switching losses predominate at the 11.7kHz switching frequency; while for the IGBTs the conduction losses are about 3 times as large as those from switching. The much lower conduction losses for the MOSFETs are a result of it's lack of a threshold in the V vs. I characteristic for BOTH the main switch and the freewheel diode. (ie. At low load currents, the complimentary gating assumed allows the MOSFET to synchronously rectify the freewheel current).

3/. Because of the slope of the V_{CE} vs. I characteristic of the IGBT, continuing to parallel more and more die has a relatively weak impact on reducing the conduction losses.

PERFORMANCE vs. COST OF MOSFETs AND IGBTs AT 100V.

If it is assumed that the extra processing required to produce an IGBT die as compared to a MOSFET, results in a cost increment of +40% for the same voltage rating and same die area, then for the above inverter/motor/actuator system, the cost/performance benefits of using IGBTs can be readily shown:

- The switching speeds of the FORD/Sanyo module and the 250v size 5 IGBT die were such that the peak voltages, dV/dT's and dI/dT's were reasonably comparable. Hence either semiconductor switch should result in roughly the same level of radiated and conducted EMI/RFI.

- Table 1 indicates that about 2.5 of the IGBT die in parallel would result in total inverter losses comparable to those obtained with the FORD/Sanyo module for both load conditions.

- The active area of 6 size 6 IFRC260 MOSFET die in parallel is approximately 3 times that of the 2.5 size 5 IGBT die.

Hence the cost of an IGBT based power stage would be about a HALF (ie. 1.4 /3) that of using FORD/Sanyo modules. (ie. where it is assumed that the cost of the freewheel diodes required by the IGBT are relatively insignificant.

TABLE 1. SIMULATED 3 PHASE INVERTER LOSSES vs. SEMICONDUCTOR TYPE.
$V_{BUS} = 100v$, $T_J = 100°C$

LOAD CONDITION	SEMICONDUCTOR TYPE	INVERTER LOSSES (Watts, 3 phase) Conduction MOSFET/IGBT	Conduction Diode	Switching[1]	TOTAL
600 RPM	FORD/Sanyo MOSFET Module	363	34	138	535
130A peak	250v Size 5 IGBTs				
Delta I = ±8A	2 in parallel	463	75	35	573
	3 "	383	64	32	479
F_{SW}=3.2kHz	4 "	337	58	30	425
2000 RPM	FORD/Sanyo MOSFET Module	3	2	34	39
14A peak	250v Size 5 IGBTs				
Delta I = ±2A	2 in parallel	15	11	8	34
	3 "	14	10	7	31
F_{SW}=11.7kHz	4 "	13	10	7	30

1. Including snubber losses.

THE PREFERRED VOLTAGE FOR A VEHICULAR BUS ABOVE 12v.

The SAE's Dual/High Voltage Vehicle Electrical Systems committee has made some initial recommendations as to the preferred voltage of the DC bus to be used for higher power loads [4]. If no special measures are taken to protect against direct contact with the electrical system;- as is the case with the 12v system in vehicles today;- then the maximum nominal battery voltage shall not exceed 50v. Limiting the nominal battery voltage to 50v would then prevent the actual battery voltage from exceeding 65v when charging while cold, the level set as the threshold for safety. The use of battery voltages above 50v was not discouraged per se, however if used, protection against direct contact would be necessary.

THE EFFECT ON THE MOSFET / IGBT COMPARISON OF REDUCING THE BUS VOLTAGE TO 50v.

As a result of the above recommendation from the SAE, it was decided to try and predict the effect of reducing the DC bus voltage from 100v to 50v. Reducing the bus voltage obviously requires the motor be rewound, (ie. twice the current at half the voltage), and at the same time allows semiconductors with lower voltage ratings to be chosen. It was thought that semiconductors with a rating of at least 100v would be adequate.

The parameters of a 100v MOSFET differ from those of a 200v device in the following manner:

$R_{DS(on)}$.

Existing power MOSFETs with voltage ratings between 100 and 200v have a $R_{RD(on)}$ dependence on voltage rating much less than the theoretical 2.5th power.(ie. at 25°C). For the best devices today, the exponent is closer to 1.7. However as the voltage rating of a MOSFET is reduced, a greater proportion of the $R_{DS(on)}$ is due to the resistance of the channel, (the resistance of which is essentially independent of temperature), so that at 100°C, the $R_{DS(on)}$ of the best 100v MOSFET is in fact approximately 1/4 that of a 200v device of the same area. Hence if the FORD/Sanyo module used 6 size 6 100v die, the $R_{DS(on)}$ could be as low as 0.017Ω/4, resulting in the same conduction loss as before even though the current has doubled.

Freewheel Diode.

Published data would tend to imply that to a first approximation, the 100v MOSFET will have a very similar V_F vs I_F curve, and the reverse recovery characteristic after irradiation will be about the same.

Switching Losses.

To a first approximation, if the switching speed is kept the same, then switching the same current at half the bus voltage would result in half the switching loss. However maintaining the switching speeds at their previous value is not at all realistic. For the FORD/Sanyo module with the 200v MOSFET die, turn-on had to be slowed down by using a relatively large gate resistance in order to get acceptable waveforms at reverse recovery. In the absence of improved body diode characteristics, it would seem that the turn-on speed will need to be further reduced when using 100v devices at twice current in order to compensate for the higher dI/dT's. If the turn-on time is therefore doubled and allowance made for the reduction in bus voltage to 50v, then for the same current the turn-on energy is the same.

The turn-off speed for the FORD/Sanyo module was relatively fast, with no gate resistance deliberately added. However with the same die area at V_{BUS} = 100v as at 200v, the leakage inductance around the loop comprised of both inverter switches and the DC bus decoupling capacitor should be the same. Hence turning off twice the current at the same speed would result in the twice the Delta V at turn-off, in a situation where the

TABLE 2. SIMULATED 3 PHASE INVERTER LOSSES vs. SEMICONDUCTOR TYPE.
V_{BUS} = 50v, T_J = 100°C

LOAD CONDITION	SEMICONDUCTOR TYPE	INVERTER LOSSES (watts, 3 phase)			
		Conduction MOSFET/IGBT	Diode	Switching[1]	Total
600 RPM	FORD/Sanyo MOSFET Module	398	38	211	647
260A peak	250v Size 5 IGBTs				
Delta I = ±16A	2 in parallel	1439	128	26	1593
	4 "	1011	89	22	1122
F_{sw} = 2.4kHz	6 "	836	76	20	932
	8 "	737	70	19	826
2000 RPM	FORD/Sanyo MOSFET Module	3	2	80	85
28A peak	250v Size 5 IGBTs				
Delta I = ±4A	2 in parallel	37	23	9	59
	4 "	30	21	8	59
F_{sw} = 11.7kHz	6 "	27	20	7	54

1. Including snubber losses.

"headroom" between DC bus and device rating has decreased from 100v to 50v. Consequently the turn-off speed would have to be reduced by a factor of 4 if it was required to maintain the previous Delta V - headroom ratio. Hence, including the effect of reducing the bus voltage, the turn-off energy will double for the same current.

The 250v rating of the experimental size 5 IGBTs from I-R is the lowest presently available, so these were also used at the reduced bus voltage. Again, for the same switching speed, halving the bus voltage will to a first approximation result in the switching losses being halved at the same current level. If the number of die required to be in parallel did not double, the higher currents per die would result in higher values of dI/dT and peak V_{CE} at turn-off. However with an IGBT rated at 250v operating on a 50v bus, somewhat higher peak voltages would seem perfectly acceptable, provided the bus capacitors can handle them. Consequently it was decided that the previous switching speeds would be acceptable. Hence for the same current per die, the reduction in bus voltage results in half the turn-on and turn-off energies.

It was assumed that the same freewheel diode was again used in anti-parallel with the IGBT, so per die the previously used equations for V_{CE} and V_F were still valid.

Once the model of the 3 phase PM motor was modified to reflect the change in bus voltage, it was then possible to re-run the previous simulations with the modified MOSFET and IGBT parameters included. Obviously for the same torque, the required motor currents had to be doubled, so it was also assumed that the size of the hysteresis band could also be doubled. Table 2 gives the results:

A number on conclusions can be drawn from Table 2:

1/. For high load currents, the dramatic reduction in the MOSFET's $R_{DS(on)}$ that resulted from the change in voltage rating to 50v makes the MOSFET's conduction losses about the same as at $V_{BUS} = 100$v. It's switching losses have almost doubled (as a result of the reduction in switching speed), but the total inverter loss has only increased by only 20%. In contrast, the relative in-effectiveness of paralleling IGBT die to reduce conduction losses results in the IGBT's conduction losses being more than double that of the MOSFET's even when 8 die in parallel are used. The fact that the IGBT's switching losses are still quite low does not help in this situation where the switching frequency is relatively low because of the larger current band used. Obviously at a switching frequencies above about 5kHz, the IGBT would begin to have lower total losses.

2/. For low levels of load current, the relatively high switching losses of the MOSFET completely swamp the low conduction losses, so that only 2 IGBT die in parallel result in lower total inverter losses.

3/. Eight of the 250v size 5 die have approximately the same die area as the 6 size 6 IRFC260 MOSFET die in the FORD/Sanyo module, and so would have the same delta T_J for the same losses. For the high current case, if 8 IGBT die were used, the relative cost of the IGBTs would then be about 1.4 times that of the MOSFETs. If however 6 IGBT die were deemed acceptable, then the relative costs would be the same.

4/. The use of a lower voltage freewheel diode across the IGBT die, with a lower V_F characteristic, (possibly even a Schottky), does not materially effect the above comparison, because the main problem for the IGBT is it's $V_{CE(sat)}$.

CONCLUSIONS.

The use of 200v MOSFETs and 250v IGBTs for operation in a 6 switch inverter for a 3kw PWM inverter/motor/actuator system was compared. It was found that:

- The relatively poor reverse recovery characteristics of the MOSFET's integral body diode required the speed at which it switched to be deliberately slowed down, especially at turn-on.

- The use of a good fast recovery diode in anti-parallel with the IGBT allowed it to be switched relatively quickly.

- At low currents the conduction loss in the MOSFET weree quite low, both because there wss no threshold voltage to be exceeded as for the IGBT, and because synchronous rectification was possible as complimentary gating was used.

Simulation of the inverter operating at two load conditions with either MOSFET or IGBT switches, showed that:

At $V_{BUS} = 100$v,

- For both load conditions, about 2.5 of the size 5 IGBT die in parallel gave the same total inverter losses as 6 size 6 MOSFET die, so that the cost of the IGBT based inverter was approximately HALF that of the MOSFET version. The switching speeds were such that the peak voltages, dV/dT's and dI/dT's were comparable.

At $V_{BUS} = 50$v,

- The same 250v IGBT was used, but a 100v MOSFET was selected and the effect of the reduction in voltage rating on conduction and switching losses estimated.
- At light loads and with the PWM frequency at about 12kHz, only 2 IGBT die in parallel were needed to produce total inverter losses lower than that with MOSFETs.
- At higher loads and with a relatively low PWM frequency of 2.5kHz, even 8 IGBT die in parallel could not produce total inverter losses lower than for the MOSFET version. In this particular case, the use of IGBTs was the more costly option. Once the PWM frequency

exceeded 5kHz however, fewer than 6 IGBT die in parallel were needed to give the same losses, with a cost lower than that for MOSFETs.

ACKNOWLEDGEMENT

The authors wish to acknowledge the assistance of Dr. Steve Clemente of International Rectifier for providing us data and suggestions on the 250 Volt IGBT and to International Rectifier for providing sample devices. The participation of Sanyo Semiconductor in packaging the power MOSFET devices is gratefully acknowledged.

REFERENCES.

[1.] B. Pelly, "The Do's and Dont's of Using Power HEXFET's", I-R Application Note #936.

[2.] D. Grant, "HEXFET III: A New Generation of Power MOSFET", I-R Application Note #966, 1987.

[3.] B. Baliga, "Modern Power Devices", Wiley, NY, 1987, pg 322.

[4.] SAE Information Report # J2232, "Vehicle System Voltage - Initial Recommendations", July 1992.