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Determining MOSFET Driver Needs for Motor Drive Applications

Author: Jamie Dunn
Microchip Technology Inc.

INTRODUCTION

Electronic motor control for various types of motors represents one of the main applications for MOSFET drivers today. This application note discusses some of the fundamental concepts needed to obtain the proper MOSFET driver for your application.

The bridging element between the motor and MOSFET driver is normally in the form of a power transistor. This can be a bipolar transistor, MOSFET or an Insulated Gate Bipolar Transistor (IGBT). In some small Brushless DC motor or stepper motor applications, the MOSFET driver can be used to directly drive the motor. For this application note, though, we are going to assume that a little more voltage and power capability is needed than what the MOSFET drivers can handle.

The purpose of motor speed control is to control the speed, direction of rotation or position of the motor shaft. This requires that the voltage applied to the motor is modulated in some manner. This is where the power-switching element (bipolar transistor, MOSFET, IGBT) is used. By turning the power-switching elements on and off in a controlled manner, the voltage applied to the motor can be varied in order to vary the speed or position of the motor shaft. Figures 1 through 5 show diagrams of some typical drive configurations for DC Brush, DC Brushless, Stepper, Switch Reluctance and AC Induction motors.

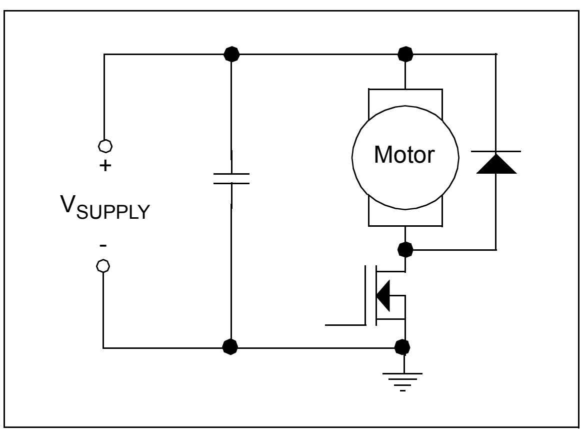

FIGURE 1: Drive Configuration for a DC Brush Motor.

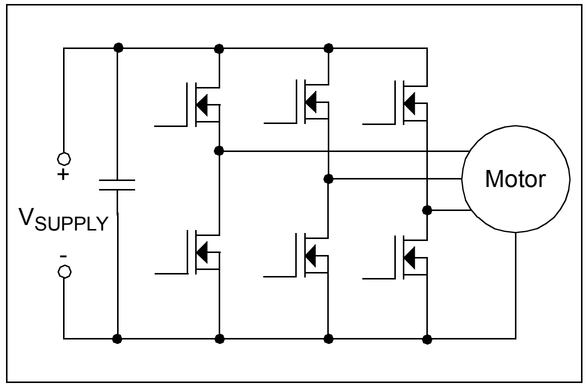

FIGURE 2: Drive Configuration for a DC Brushless Motor.

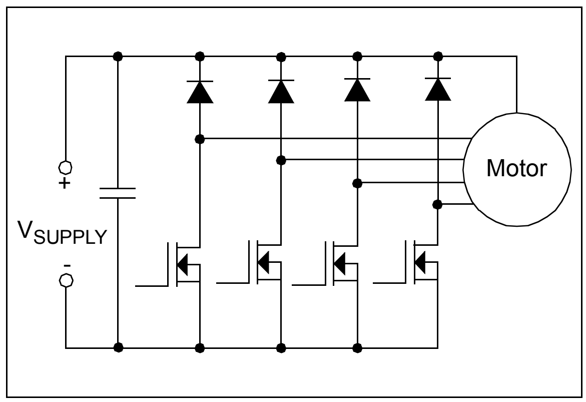

FIGURE 3: Drive Configuration for a 4-Wire Stepper Motor.

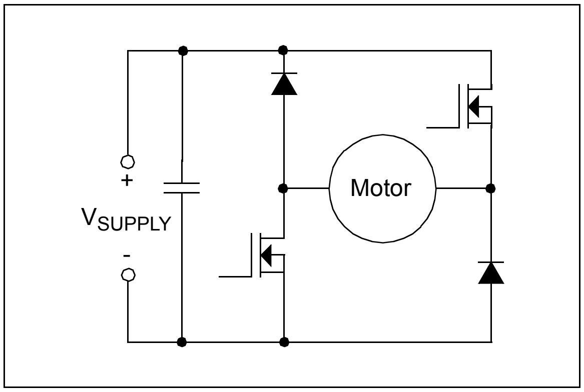

FIGURE 4: Drive Configuration for Each Winding of a Switch Reluctance Motor.

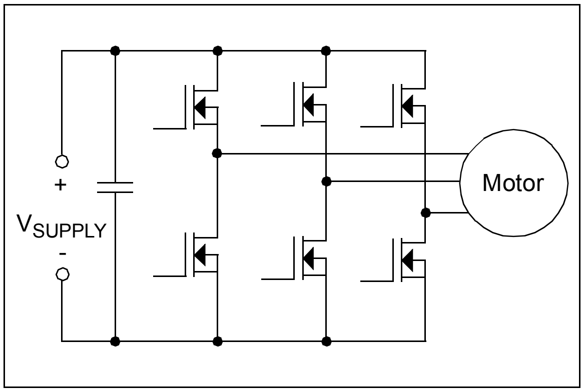

FIGURE 5: Drive Configuration for an AC Induction Motor.

As seen in Figures 1 through 5, even though the motor type changes, the purpose of the drive circuitry is to provide voltage and current to the windings of the motor. The voltage and current level will vary depending on what type and size of motor is being used, but the fundamentals of selecting the power-switching element and the MOSFET driver are the same.

SELECTING THE POWER-SWITCHING ELEMENT

The first stage in selecting the correct power-switching element for your motor drive application is understanding the motor being driven. Understanding the ratings of the motor is an important step in the process as it is often the corner points of operation that will determine the choice of the power switching element. A sample of motor ratings for the motor types listed earlier is shown in Table 1. When dealing with motors, it is often useful to remember that 1 Horse Power (HP) is equal to 746 Watts.

From the ratings in Table 1, the voltage, current and power ratings vary significantly with the different types of motors. Motor ratings can also vary significantly within the same motor type. A key point to note in Table 1 is the value of the start-up current (sometimes given as stall current or locked-rotor current). The startup current value can be up to three times the value of the steady-state operating current. As mentioned previously, it is these corner points of operation that will determine the necessary ratings of the drive element. Because of the various voltage and current ratings for the various motor types, the selected drive device ratings will have to vary as well, depending on the application and design goals.

TABLE 1: MOTOR RATINGS

Motor Type	Horse Power Rating (HP)	Voltage Rating	Current Rating (A)		Efficiency (%)	Power Factor	Slip Factor	Torque lb*ft	Full Load RPM
			Full Load	Locked Rotor					
DC Brushless	0.54	48 VDC	8.7	120	87	NA	NA	0.53	5000
DC Brush	0.40	60 VDC	6.0	106	84	NA	NA	0.70	4000
Stepper	0.01	24 VDC	0.3	1.0	65	NA	NA	0.1	300 to 600
Switch Reluctance	1.20	24 VDC	37.5	NG	94	NA	NA	1.8	15,000
AC Induction	2.00	230 VAC	10.0	65.0	79	0.81	1.15	3.0	3450

MOSFET OR IGBT, WHAT'S BEST FOR YOUR APPLICATION?

The two main choices for power-switching elements for motor drives are the MOSFET and IGBT. The bipolar transistor used to be the device of choice for motor control due to it's ability to handle high currents and high voltages. This is no longer the case. The MOSFET and IGBT have taken over the majority of the applications. Both the MOSFET and IGBT devices are voltage controlled devices, as opposed to the bipolar transistor, which is a current-controlled device. This means that the turn-on and turn-off of the device is controlled by supplying a voltage to the gate of the device, instead of a current. This makes control of the devices much easier.

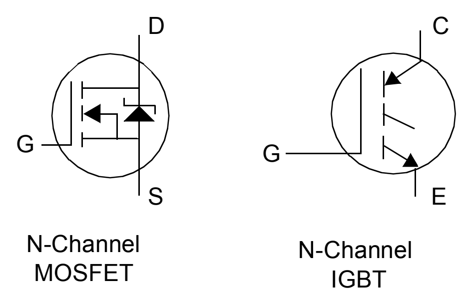

FIGURE 6: MOSFET and IGBT Symbols.

The similarities between the MOSFET and the IGBT end with the turn-on and turn-off of the devices being controlled by a voltage on the gate. The rest of the operation of these devices is very different. The main difference being that the MOSFET is a resistive channel from drain-to-source, whereas the IGBT is a PN junction from collector-to-emitter. This results in a difference in the way the on-state power dissipations are calculated for the devices. The conduction losses for these devices are defined as follows:

MOSFET

$P_{LOSS} = I_{rms}^2 * R_{DS\text{-}ON}$

where:

$R_{DS\text{-}ON}$ = drain-to-source on-state resistance

I_{rms} = drain-to-source rms current

IGBT

$P_{LOSS} = I_{ave} * V_{CE\text{-}SAT}$

$V_{CE\text{-}SAT}$ = collector-to-emitter saturation voltage

I_{ave} = collector-to-emitter average current

The key difference seen in these two equations for power loss is the squared term for current in the MOSFET equation. This requires the $R_{DS\text{-}ON}$ of the MOSFET to be lower, as the current increases, in order to keep the power dissipation equal to that of the IGBT. In low voltage applications, this is achievable as the $R_{DS\text{-}ON}$ of MOSFETs can be in the 10's of milli-ohms. At higher voltages (250V and above), the $R_{DS\text{-}ON}$ of MOSFETs do not get into the 10's of milli-ohms. Another key point when evaluating on-state losses is the temperature dependence of the $R_{DS\text{-}ON}$ of the MOSFET versus the $V_{CE\text{-}SAT}$ of an IGBT. As temperature increases, so does the $R_{DS\text{-}ON}$ of the MOSFET, while the $V_{CE\text{-}SAT}$ of the IGBT tends to decrease (except at high current). This means an increase in power dissipation for the MOSFET and a decrease in power dissipation for the IGBT.

Taking all of this into account, it would seem that the IGBT would quickly take over the applications of the MOSFET at higher voltages, but there is another element of power loss that needs to be considered. That is the losses due to switching. Switching losses occur as the device is turned on and off with current ramping up or down in the device with voltage from drain-to-source (MOSFET) or collector-to-emitter (IGBT). Switching losses occur in any hard-switched application and can often dominate the power losses of the switching element.

The IGBT is a slower switching device than the MOSFET and, therefore, the switching losses will be higher. An important point to note at this juncture is that as IGBT technology has progressed over the past 10 years, various changes have been made to improve the devices with different applications. This is also true of MOSFETs, but even more so for IGBTs. Various companies have multiple lines of IGBTs. Some are optimized for slow-speed applications that have lower $V_{CE\text{-}SAT}$ voltages, while others are optimized for higher-speed applications (60 kHz to 150 kHz) that have lower switching losses, but have higher $V_{CE\text{-}SAT}$ voltages. The same is true for MOSFETs. Over the past 5 years, a number of advances have been made in MOSFET technology which have increased the speed of the devices and lowered the $R_{DS\text{-}ON}$. The net result of this is that, when doing a comparison between IGBTs and MOSFETs for an application, make sure the devices being compared are best suited to the application. This assumes, of course, that the devices also fit within your budget.

Although the IGBT is slower than the MOSFET at both turn-on and turn-off, it is mainly the turn-off edge that is slower. This is due to the fact that the IGBT is a minority carrier recombination device in which the gate of the device has very little effect in driving the device off (will vary depending on the version of the IGBT, fast, ultra-fast, etc.). This can be seen in the equivalent circuit for the IGBT shown in Figure 7. When the gate is turned on (driven high), the N-channel MOSFET pulls low on the base of the PNP transistor, effectively driving the device on. During turn-off, however, when the gate of the device is pulled low, only the minority carrier recombination of the device is effecting the turn-off speed. By varying some of the parameters of the device (such as oxide thickness and doping) the speed of the device can be changed. This is the essence of the various families of IGBTs that are available from multiple suppliers. Increases in speed often result in higher $V_{CE\text{-}SAT}$ voltages and reduced current ratings for a given die size.

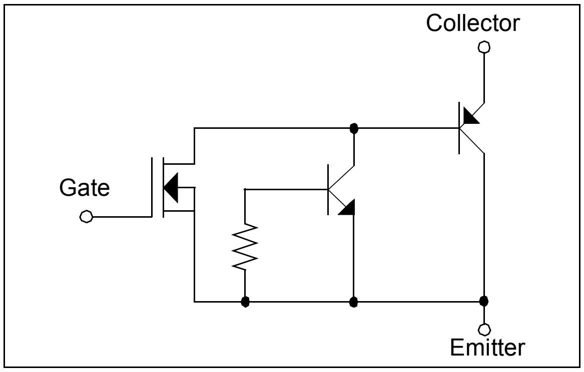

FIGURE 7: *Equivalent Circuit for an IGBT.*

Calculating switching losses for IGBTs is not as straightforward as it is for the MOSFET. For this reason, switching losses for IGBTs are typically characterized in the device data sheet. Switching losses are typically given in units of Joules. This allows the user to multiply the value by frequency in order to get power loss.

Switching losses is the biggest limiting factor that keep IGBTs out of many high-voltage, high switching-frequency applications. Because of the relatively low modulation/switching frequencies of motor control applications (typically less than 50 kHz), the switching losses are kept in check and the IGBT is as good or better than the MOSFET.

Since this application note does not cover all the pros and cons of MOSFETs versus IGBTs, listed below are other application notes written about this topic.

- "IGBTs vs. HEXFET Power MOSFETs For Variable Frequency Motor Drives", AN980, International Rectifier.
- "Application Characterization of IGBTs" (this one will help you apply the IGBT and understand the device), AN990, International Rectifier.
- "IGBT Characteristics". This one goes into the fundamentals of the IGBT and compares it with the MOSFET, AN983, International Rectifier.
- "IGBT or MOSFET: Choose Wisely". This one discusses the crossover region of applications based on voltage rating of the device and operating frequency, White Paper, International Rectifier.
- "IGBT Basic II". This application note covers IGBT basics and discusses IGBT gate drive design and protection circuits, AN9020, Fairchild Semiconductor.
- Application Manual from Fuji Semiconductor for their 3rd-Generation IGBT modules. This covers many topics from IGBT basics to current sharing.

To summarize some of the discussion so far, some of the generally accepted boundaries of operation when comparing the IGBT and MOSFET are:

- For application voltages < 250V, MOSFETs are the device of choice. In searching many IGBT suppliers, you will find that the selection of IGBTs with rated voltages below 600V is very small.
- For application voltages > 1000V, IGBTs are the device of choice. As the voltage rating of the MOSFET increases, so does the $R_{DS\text{-}ON}$ and size of the device. Above 1000V, the $R_{DS\text{-}ON}$ of the MOSFET can no longer compete with the saturated junction of the IGBT.
- Between the 250V and 1000V levels described above, it becomes an application-specific choice that revolves around power dissipation, switching frequency and cost of the device.

When evaluating the MOSFET versus the IGBT for an application, be sure to look at the performance of the device over the entire range. As discussed previously, the resistive losses of the MOSFET increase with temperature, as do the switching losses for the IGBT.

Other hints for design and derating are:

- Voltage rating of the device is derated to 80% of its value. This would make a 500V MOSFET usable to 400V. Any ringing in the drain-to-source voltage in an application should also be taken into account.
- The maximum junction temperature of the device should not exceed 120ºC at maximum load and maximum ambient. This will prevent any thermal runaway. Some sort of overtemperature protection should also be incorporated.
- Care should be taken in the layout of the printed circuit board to minimize trace inductance going to the leads of the motor from the drive circuitry. Board trace inductance and lead inductance can cause ringing in the voltage that is applied to the motors terminals. The higher voltages can often lead to breakdown in the motor insulation between windings.
- The current rating for the switching element must also be able to withstand short circuit and start-up conditions. The start-up current rating of a motor can be three to six times higher than the steady state operating current.

GATE DRIVE SCHEMES

The type of motor, power-switching topology and the power-switching element will generally dictate the necessary gate drive scheme. The two fundamental categories for gate drive are high-side and low-side. High-side means that the source (MOSFET) or emitter (IGBT) of the power element can float between ground and the high-voltage power rail. Low-side means the source or emitter is always connected to ground. An example of both of these types can be seen in a half-bridge topology, shown in Figure 8. In this configuration, Q_1 and Q_2 are always in opposite states. When Q_1 is on, Q_2 is off and vice-versa. When Q_1 goes from being off to on, the voltage at the source of the MOSFET goes from ground up to the high-voltage rail. This means that the voltage applied to the gate must float up as well. This requires some form of isolated, or floating, gate drive circuitry. Q_2, however, always has its source or emitter connected to ground so the gate drive voltage can also be referenced to ground. This makes the gate drive much more simple.

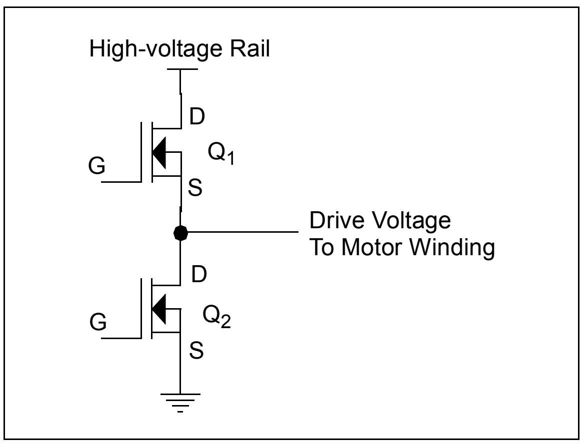

FIGURE 8: *Example of a High Side (Q_1) and Low Side (Q_2) Gate Drive Requirement.*

Various schemes exist for high-side gate drive applications. These include single-ended or double-ended gate drive transformers, high-voltage bootstrap driver ICs, floating bias voltages and opto-isolator drive. Examples of these drive schemes are shown in Figures 9 through 12.

The Microchip MOSFET drivers that are shown in Table 2 on page 15 fit a wide variety of applications using the gate drive schemes shown in Figures 9, 10 and 12. The single output drivers, which have ratings of 0.5A up to 9.0A, work well for the single-ended gate drive needs for the circuits in Figures 9 and 12. The dual output drivers provide an excellent solution for the gate drive solution shown in Figure 10. The selection process for the MOSFET drivers is discussed later in this application note.

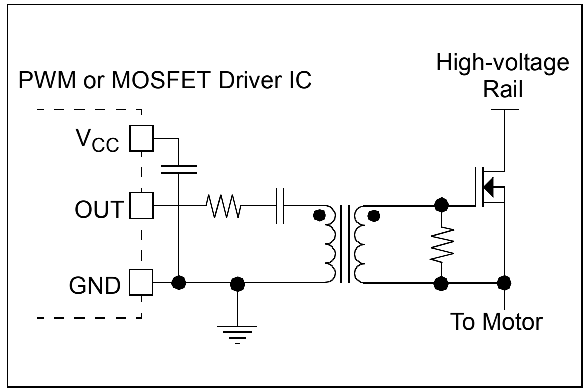

FIGURE 9: *Single-Ended Gate Drive Transformer.*

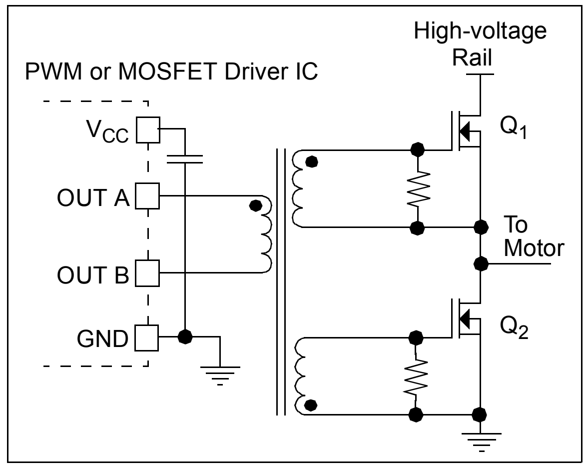

FIGURE 10: *Double-Ended Gate Drive Transformer.*

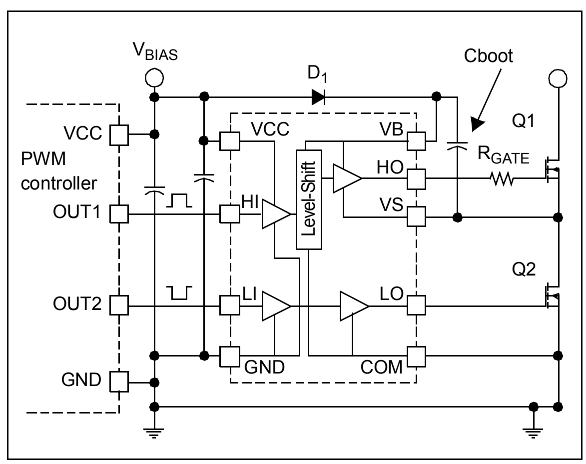

FIGURE 11: *High Voltage Bootstrap Driver IC.*

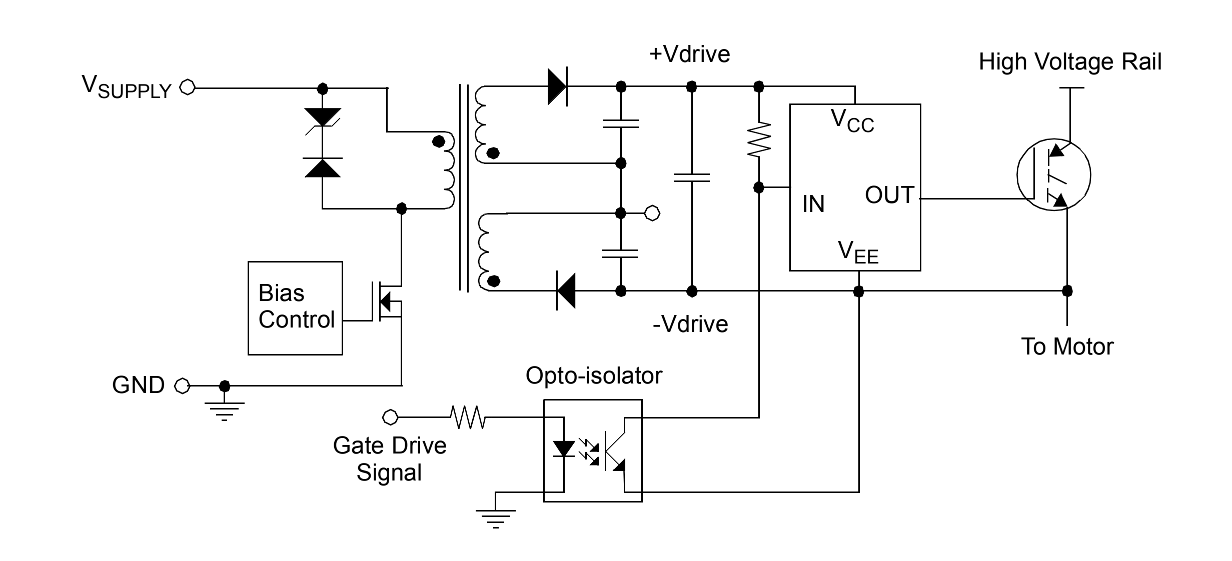

FIGURE 12: *Floating Bias Gate Drive Circuit.*

The gate drive transformer solutions shown in Figures 9 and 10 provide a number of good features. The first feature is that they solve the high-side drive problem. The drive winding(s) that drive the gate of the power MOSFET/IGBT can float at any potential (only limitations to this are the insulation ratings of the wire).

The second feature is that it provides both a positive and negative gate drive voltage. As with any transformer, there must be volt-time balancing. With the solution shown in Figure 9, the capacitor, in series with the winding, is charged during the on time of the drive signal and then provides the negative bias/drive voltage to the transformer during the off time. This acts as the reset mechanism for the transformer and also the mechanism to provide the negative gate drive voltage to the power-switching element, which is often very useful and needed, if an IGBT is being used. If a MOSFET is being used as the switching element and the negative drive is not desired (negative drive often increases delay times), a few additional components can be added to the circuit to fix this issue, as shown in Figure 13. With the addition of the diode and N-channel FET (low voltage, small signal type FET), the main N-channel MOSFET still sees the same positive level drive signal as before (minus a diode drop), but is clamped to zero volts during the off time. The diode blocks the negative bias that now turns on the small signal FET that clamps the gate-to-source voltage to zero.

The second gate drive transformer drive configuration shown in Figure 10 is a double-ended type drive, meaning that the transformer is driven in both directions. This type of drive is often used for half-bridge and full-bridge topologies. The bidirectional drive, coupled with the dot polarity of the transformer, drives Q_1 on and Q_2 off and vice versa. If the duty cycles of the MOSFETs are modulated differently, additional gate drive circuitry may be required to balance the volt-time of the transformer. The same negative bias-blocking circuitry shown in Figure 13 can also be used in the double-ended drive scheme.

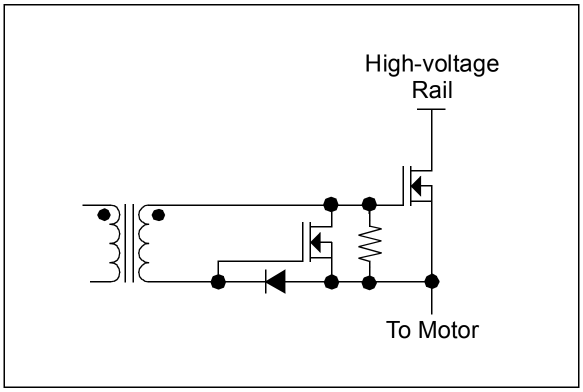

FIGURE 13: *Removal of Negative Drive Voltage.*

The other feature of the gate drive transformer is that it can be driven from the secondary side with ground referenced circuitry. This means that it can provide a high voltage isolation boundary and allow the drive circuitry (PWM and MOSFET driver) to be ground-referenced and near the control circuitry, which is typically on the secondary side. This makes interfacing between the small signal-sensing circuitry (temperature-sensing, feedback loops, shutdown circuits) and the PWM very easy. With the drive circuitry now ground-referenced, low-side MOSFET drivers can be used. This expands the selection of available devices and will reduce the cost of the driver.

The high-voltage half-bridge driver IC, shown in Figure 11, provides a solution to the high-side drive issue and does not require the user to have any knowledge of transformers. These types of ICs utilize high-voltage, level-shifting circuitry, in conjunction with a “bootstrap” capacitor, to provide the high-side gate drive. During the on-time of FET/IGBT Q_2, the source/emitter of Q_1 is at ground potential. This allows capacitor Cboot to be charged through diode D_1 from the bias supply V_{BIAS}. When Q_2 is turned off and Q_1 is turned on, the voltage at the source of Q_1 begins to rise. The Cboot capacitor now acts as the bias source for the high-side drive portion of the driver and provides the current-to-charge the gate of Q_1. The level shifting circuitry of the driver allows the high-side drive stage to float up with the source voltage of Q_1. These types of drivers are often rated to handle up to 600V (with respect to ground) on the high-side drive portion of the circuitry. One of the draw-backs to many of these types of drivers is the long propagation delay times between the input signal and the high-side drive turning on/off. This is a result of the level shifting circuitry. These delays can be between 500 nsec. and 1 µsec. This can cause problems for some higher-frequency applications as the delay times take up too much of the overall period. Though, for most motor drive applications that are operating below 50 kHz, it is not an issue.

The circuit shown in Figure 12 is often used in very high power applications where IGBT/MOSFET modules are being used. In these applications, the IGBT modules are often located a slight distance from all of the control circuitry. This makes it difficult to bus the gate drive signal to the module as the inductance in the wires will cause ringing at the gate of the module. For this reason, the isolated bias circuit is often built on a separate PC card and mounted directly to the IGBT/MOSFET module. With this scheme, the only signal that needs to be brought to the module is the small signal line that drives the opto-isolator. This is more easily accomplished since there is less current flowing in this line.

The negative bias is often required for these applications in order to keep the IGBT in the off state. This will be described more in the following sections that discuss the gate properties of the IGBT. Though this scheme does require much more circuitry, it does provide a very robust solution for driving large gate capacitances in high-power applications. The V_{supply} voltage that feeds the flyback topology can be a low voltage or high voltage supply. A low voltage supply of 10V or less will make the flyback design easier, as biasing of the control circuitry can be done directly off of this voltage. High voltage flyback ICs that incorporate the high voltage MOSFET and biasing circuitry are available, which make low-power flybacks like this one easy to design.

MOSFET AND IGBT GATE PROPERTIES

As stated earlier, the MOSFET and IGBT are voltage-controlled devices. Both devices are characterized in the same manner, with data sheets supplying values for Gate Threshold Voltages (voltage at which the drain to source/collector to emitter channels begin to conduct) and Total Gate Charge.

Figures and show the Electrical Characteristics section of data sheets for a MOSFET and an IGBT device that are rated for 500V and 20A, and 600V and 20A, respectively.

Some key differences in the Electrical Characteristics table when comparing MOSFETs and IGBTs are:

- The gate threshold voltage for the IGBT is slightly higher than that of the MOSFET. For the two devices being compared, the IGBT is specified for 3.0V to 6.0V (min. to max) where the MOSFET is specified for 2.0V to 4.0V. For most power devices, these thresholds are fairly standard. A key difference between the two devices is the temperature dependency of the gate-to-emitter threshold for the IGBT. This is shown as the “Temperature Coefficient of Threshold Voltage” in the IGBT data sheet. For this particular device, it is 13 mV/ºC. So as the junction temperature of this device heats up to 125°C (100°C rise above the specification temperature for the 3.0V to 6.0V range), the new range for the gate threshold voltage becomes 1.7V to 4.7V. This will make the device more susceptible to transient conditions which try to turn the gate on when it is supposed to be off. This is often the reason why negative gate drive voltages are used with IGBTs.
- In the “Conditions” column, note that for the IGBT many of the conditions are for a V_{GE} of 15V where the MOSFET is for a V_{GS} of 10V. This is for good reason. Even though both devices are rated for ±20V from gate-to-source/emitter, the MOSFETs operation does not really improve with gate voltages above 10V (R_{DS-ON} of the device no longer decreases with an increase in gate voltage). This can be seen by looking at the MOSFET typical characteristic curves for Drain-to-Source Current versus Drain-to-Source Voltage. There is very little difference between the curves once V_{GS} is 10V and above. For the IGBT, the curve for Collector-to-Emitter Voltage versus Gate-to-Emitter Voltage show that the device’s capability to handle more current continues to increase as the gate voltage is raised above 10V. This is important to remember when doing a comparison between the two devices. Many of the gate drive devices available today have an upper operating limit of 18V. Running 15V on V_{CC} leaves very little room for adding a negative bias for IGBT turn-off.

Static @ TJ = 25°C (unless otherwise speicified)						
	Parameter	**Min.**	**Typ.**	**Max.**	**Units**	**Conditions**
$V_{(BR)DSS}$	Drain-to-Source Breakdown Voltage	500	—	—	V	V_{GS} = 0V, I_D = 250 µA
$\Delta V_{(BR)DSS}/\Delta_{TJ}$	Breakdown Voltage Temperature Coefficient	—	0.61	—		V/°C Reference to 25°C, I_D = 1 mA
$R_{DS(on)}$	Static Drain-to-Source On-Resistance	—	—	0.27	Ω	V_{GS} = 10V, I_D = 12A
$V_{GS(th)}$	Gate Threshold Voltage	2.0	—	4.0	V	V_{DS} = V_{GS}. I_D = 250 µA
I_{DSS}	Drain-to-Source Leakage Current	— —	— —	25 250	µA	V_{DS} = 500V, V_{GS} = 0V V_{DS} = 400V, V_{GS} = 0V, T_J = 125°C
I_{GSS}	Gate-to-Source Forward Leakage	—	—	100	nA	V_{GS} = 30V
	Gate-to-Source Reverse Leakage	—	—	-100		V_{GS} = -30V

Dynamic @ TJ = 25°C (unless otherwise specified)						
	Parameter	**Min.**	**Typ.**	**Max.**	**Units**	**Conditions**
g_{FS}	Forward Transconductance	11	—	—	S	V_{DS} = 50V, I_D = 12A
Qg	Total Gate Charge	—	—	105	nC	I_D = 20A V_{DS} = 400V V_{GS} = 10V, **See Figures 6 and 13**.
Q_{gs}	Gate-to-Source Charge	—	—	26		
Q_{gd}	Gate-to-Drain (“Miller”) Charge	—	—	42		
$t_{d(on)}$	Turn-on Delay Time	—	18	—	ns	V_{DD} = 250V I_D = 20A R_G = 4.3W R_D = 13W, **See Figure 10**.
t_r	Rise Time	—	55	—		
$t_{d(off)}$	Turn-off Delay Time	—	45	—		
t_r	Fall Time	—	39	—		
C_{iss}	Input Capacitance	—	3100	—	pF	V_{GS} = 0V V_{DS} = 25V f = 1.0 MHz, **See Figure 5**.
C_{oss}	Ouput Capacitance	—	480	—		
C_{rss}	Reverse Transfer Capacitance	—	18	—		
C_{oss}	Ouput Capacitance	—	4430	—		V_{GS} = 0V, V_{DS} = 1.0V, f = 1.0 MHz
C_{oss}	Ouput Capacitance	—	130	—		V_{GS} = 0V, V_{DS} = 400V, f = 1.0 MHz
C_{oss} eff.	Effective Ouput Capacitance	—	140	—		V_{GS} = 0V, V_{DS} = 0V to 400V

Avalanche Characteristics				
	Parameter	**Typ.**	**Max.**	**Units**
E_{AS}	Single Pulse Avalanche Energy	—	960	mJ
I_{AR}	Avalanche Current	—	20	A
E_{AR}	Repetitive Avalanche Energy	—	28	mJ

FIGURE 14: *500V, 20A MOSFET Electrical Characteristics Table.*

Electrical Characteristics @ T_J = 25°C (unless otherwise specified)						
	Parameter	**Min.**	**Typ.**	**Max.**	**Units**	**Conditions**
$V_{(BR)CES}$	Collector-to-Emitter Breakdown Voltage	600	—	—	V	V_{GE} = 0V, I_C = 250 µA
$V_{(BR)ECS}$	Emitter-to-Collector Breakdown Voltage	18	—	—	V	V_{GE} = 0V, I_C = 1.0A
$\Delta V_{(BR)CES}/\Delta TJ$	Temperature Coefficient of Breakdown Voltage	—	0.44	—	V/°C	V_{GE} = 0V, I_C = 1.0 mA
$V_{CE(ON)}$	Collector-to-Emitter Saturation Voltage	—	2.05	2.5	V	I_C = 20A; V_{GE} = 15V
		—	2.36	2.5		I_C = 40A; **See Figures 2, 5**
		—	1.90	—		I_C = 20A, T_J = 150°C
$V_{GE(th)}$	Gate Threshold Voltage	3.0	—	6.0		V_{CE} = V_{GE}, I_C = 250 µA
$\Delta V_{GE(th)}/\Delta T_J$	Temperature Coefficient of Threshold Voltage	—	13	—	mV/°C	V_{CE} = V_{GE}, I_C = 250 µA
g_{fe}	Forward Transconductance	18	28	—	S	V_{CE} = 100V, I_C = 20A
I_{CES}	Zero Gate Voltage Collector Current	—	—	250	µA	V_{GE} = 0V, V_{CE} = 600V
		—	—	2.0		V_{GE} = 0V, V_{CE} = 10V, T_J = 25°C
		—	—	2500		V_{GE} = 0V, V_{CE} = 600V, T_J = 150°C
I_{GES}	Gate-to-Emitter Leagage Current	—	—	±100	nA	V_{GE} = ±20V

Switching Characteristics @ TJ = 25°C (unless otherwise specified)						
	Parameter	**Min.**	**Typ.**	**Max.**	**Units**	**Conditions**
Q_g	Total Gate Charge (turn-on)	—	98	147	nC	I_C = 20A V_{CC} = 400V **See Figure 8**. V_{GE} = 15V
Q_{ge}	Gate - Emitter Charge (turn-on)	—	12	18		
Q_{gc}	Gate - Collector Charge (turn-on)	—	36	54		
$t_{d(on)}$	Turn-on Delay Time	—	27	—	ns	T_J = 25°C I_C = 20A, V_{CC} = 480V V_{GE} = 15V, R_G = 10Ω Energy losses include "tail" **See Figures 9,10,14**
t_r	Rise Time	—	22	—		
$t_{d(off)}$	Turn-off Delay Time	—	100	150		
t_f	Fall Time	—	74	110		
E_{on}	Turn-on Switching Loss	—	0.11	—	mJ	
E_{off}	Turn-off Switching Loss	—	0.23	—		
E_{ts}	Total Switching Loss	—	0.34	0.45		
$t_{d(on)}$	Turn-on Delay Time	—	25	—	ns	T_J = 150°C I_C = 20A, V_{CC} = 480V V_{GE} = 15V, R_G = 10Ω Energy losses include "tail" **See Figures 10,11,14**
t_r	Rise Time	—	23	—		
$t_{d(off)}$	Turn-off Delay Time	—	170	—		
t_f	Fall Time	—	124	—		
E_{ts}	Total Switching Loss	—	0.85	—	mJ	
L_E	Internal Emitter Inductance	—	13	—	nH	Measured 5 mm from package
C_{ies}	Input Capacitance	—	1900	—	pF	V_{GE} = 0V V_{CC} = 30V, **See Figure 7.** f = 1.0 MHz
C_{oes}	Output Capacitance	—	140	—		
C_{rss}	Reverse Transfer Capacitance	—	35	—		

FIGURE 15: *600V, 20A IGBT Electrical Characteristics Table.*

- When comparing gate charge values, again note the possible difference in gate voltage values used for the measurement. In this particular example, the gate charge for the IGBT is done with 15V, whereas the MOSFET uses 10V which makes the gate charge value lower. Q = C*V. This is important for the application when calculating losses in the gate drive circuitry.
- Turn-on Delay Time, Rise Time, Turn-off Delay Time and Fall Time are not measured the same way for the MOSFET and IGBT. For the MOSFET, the times are relationships between gate voltage and Drain-to-Source voltage. For the IGBT, the times are relationships between gate voltage and collector current. Further explanation of this can be seen in any MOSFET and IGBT data sheet where the switching waveform is explained.

- As discussed earlier, because of the “tail” in the collector current of the IGBT, it is difficult to predict the switching losses of the IGBT. For this reason, the data sheet often characterizes the switching losses for you. As is seen in Figure , the IGBT data sheet actually characterizes the switching times and switching losses at both room ambient and a junction temperature of 150ºC. The MOSFET data sheet only gives switching times at room ambient and does not give numbers for switching losses. Further characterization of the switching losses of the IGBT is done in the typical characteristic curves of the data sheet. Curves for “Total Switching Losses vs. Gate Resistance”, Total Switching Losses vs. Junction Temperature” with curves for different collector currents, and “Total Switching Losses vs. Collector Current” are often given.
- Another important parameter when it comes to switching times, is the gate resistance that is used for the testing. This is shown in the Conditions column for the various switching times. For a MOSFET, gate resistance will effect both turn-on and turn-off switching times and, therefore, will also effect switching losses. A trade-off is often made between switching losses and the dv/dt of the drain-to-source voltage. The faster the transition means lower switching losses. However, it also means more ringing and induced EMI in the circuit. The turn-on speed of the IGBT is always effected by the gate resistance. The turn-off speed, however, is effected differently depending on the design of the IGBT. For devices designed for faster switching speeds, the turn-off times and losses are effected more by the change in gate resistance. For the IGBT, there is also another aspect that is effected by gate resistance, which is device latch-up. For many IGBT devices, too low of a gate resistance may result in high dv/dt at turn-off, which can lead to dynamic latch-up of the device. For the device represented in Figure , a gate resistance value of 10Ω is used throughout the data sheet. The manufacturer should be consulted about their devices’ susceptibility to dynamic latch-up. If their devices are resistant to latch-up, the gate resistance value can often be decreased in order to obtain lower switching losses. Many times, though, the gate resistance value shown in the data sheet for characterization is the minimum value of gate resistance the manufacturer recommends for stable gate circuit operation and resistance to latch-up. This is an important aspect to understand, as this will set the lower limit of the switching losses in the application.

GATE CHARACTERISTICS OF IGBTS AND MOSFETS

Now that many of the device characteristics of the MOSFET and IGBT have been discussed, we can focus on the requirements for driving the gates of these devices.

When determining the gate drive requirements for the switching device in your application, the key specification to look for is gate charge. Many application notes have been written discussing why gate charge values should be used instead of the gate capacitance values. The main reason for this is the “Miller Effect”. The gate-to-drain capacitance (or Miller capacitance) effect on gate drive for MOSFETs has long been understood and is characterized in the gate charge value. The same effect is true for IGBTs. The gate capacitance model is the same for both devices. These are shown in Figure 16.

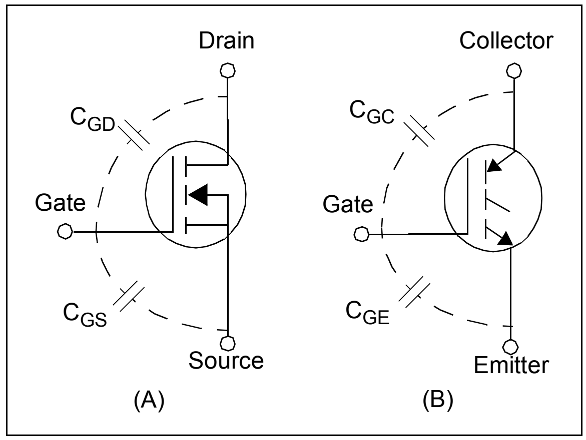

FIGURE 16: *Gate Capacitance Models for the MOSFET (A) and IGBT (B).*

The charging process for the gate of a MOSFET/IGBT can be broken down into three stages. This is shown in Figure 17.

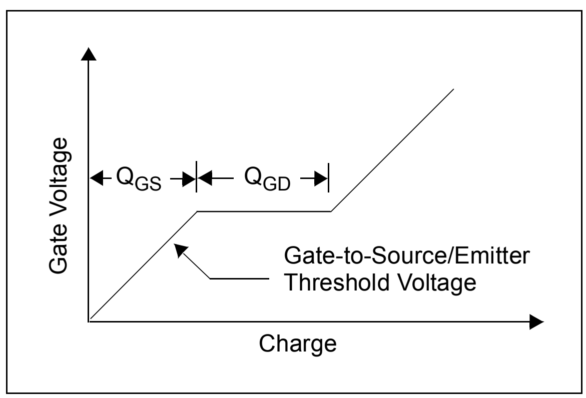

FIGURE 17: *Gate Charge Waveform.*

The first stage of the charging is mainly charging the gate-to-source/emitter capacitance. The gate-to-drain/ collector capacitance is also being charged, but this amount of charge is very low. Once the gate-to-source/ emitter capacitance is charged up to the gate threshold voltage, the device begins to turn on and the current ramps up to the full value of current in the circuit. Once full current is reached, the drain-to-source/collector-to-emitter voltage begins to collapse. It is at this point that the gate voltage flattens out due to the Miller capacitance being charged as the drain/collector voltage falls. Once the drain/collector voltage has fallen to its final level, the gate capacitance (both G-S/E and G-D/C) is charged the rest of the way to the gate drive voltage. Detailed descriptions of the turn-on and turn-off waveforms for IGBTs and MOSFETs can be found in some of the application notes that are listed in the reference section.

The "Total Gate Charge" value in the data sheet brings together all of the pieces of the gate charge puzzle into one easy-to-use number. The conditions for the total gate charge value should be noted as the gate-to-source/emitter voltage and D-S/C-E voltage may be different than your application. For both devices, a typical characteristics curve is normally provided which plots gate charge versus gate voltage.

In order to further understand gate charge better, it can be broken down one more level with the relationship:

$Q_{TOTAL} = C_{GATE} * V_{GATE}$

or

$C_{GATE} = Q_{TOTAL} / V_{GATE}$

Where:

Q_{TOTAL} = Total Gate Charge Value (most of the time given in nano-coulombs)

C_{GATE} = Total Gate Capacitance

V_{GATE} = Gate Drive Voltage

This relationship breaks down the gate charge value into a capacitance value. From here, the charging and discharging of the gate of the MOSFET/IGBT can be viewed as the charging and discharging of a capacitor.

Other important relationships related to gate charge are:

Power Required to Charge Gate Capacitance:

$P_{GATE} = 1/2\ C_{GATE} * V_{GATE}^2 * F$

Where:

F = Switching Frequency

Power Dissipated in the Gate Driver Circuitry:

$P_{DRIVER} = C_{GATE} * V_{GATE}^2 * F$

Where:

F = Switching Frequency

DETERMINING THE MOSFET DRIVER RATING FOR THE GATE DRIVE APPLICATION

Most MOSFET drivers or gate driver circuits are often rated using a peak current. Ratings of 1.5A, 3.0A and 6.0A are commonly used when discussing these devices. What do these ratings mean, and how can they be used to select the appropriate device? These are the questions that will be answered in this section.

At this point, the power-switching element has been chosen and a gate charge value can be found. Depending on the size of the device chosen, the gate charge value can range from tens of nano-coulombs to over 600 nano-coulombs (IGBT and MOSFET modules rated for 100's of amps). In order to meet the conduction and switching losses that led to the selection of the power-switching device, the gate will need to be driven with the proper voltage and at the correct speed.

Most MOSFET drivers are fairly simple devices, from a circuitry point of view. The input stage of the device converts the incoming low-voltage signal (most MOSFET drivers are designed to handle TTL and CMOS level signals) to a full range (GND to V_{DD}) signal that turns on and off a cascaded chain of increasingly stronger drive stages, called the pre-drivers. The final pre-drive stage then drives the gates of the output stage of the driver that are shown as Q_1 and Q_2 in Figure 18. A typical block diagram for a MOSFET driver is shown in Figure 18. MOSFETs Q_1 and Q_2 represent the pull-up and pull-down output drive stage of the MOSFET driver.

A common misconception about MOSFET drivers is that they provide a constant current output. Meaning that if the driver is rated for 1.5A, the output would drive the capacitive load with a constant current of 1.5A until it was fully charged. This is not true. The current rating of a MOSFET driver is a “peak” current rating. The peak current rating of the MOSFET driver is for a given bias voltage (often the maximum V_{DD} voltage) with the output of the driver tied to ground (conditions for the peak pull-up current). Sometimes the peak current rating of the driver is given with the output voltage of the driver at 4V. This gives a representation of the current capability when the gate of the MOSFET would normally be ramping through the region where the Miller capacitance is coming into play, as discussed previously. With either type of driver rating, the bias voltage is a key factor and needs to be considered when selecting the appropriate driver rating.

As is shown in Figure 18, the output stage of the driver consists of a P-channel and a N-channel MOSFET. The P-channel MOSFET provides the pull-up, or charge current for the gate capacitance and the N-channel MOSFET provides the pull-down, or discharge current for the external gate capacitance.

In viewing the output stage of the MOSFET driver as a push-pull pair of MOSFETs, it is now easier to see how the MOSFET driver operates. For a non-inverting driver, when the input signal goes to a high state, the common gate signal of Q_1 and Q_2 is pulled low (referencing Figure 18). The transition of this gate node from a voltage of V_{DD} to GND typically occurs in less than 10 nsec. This fast transition limits the cross conduction time between Q_1 and Q_2 and also gets Q_1 to its fully-enhanced state quickly in order to reach peak current as soon as possible.

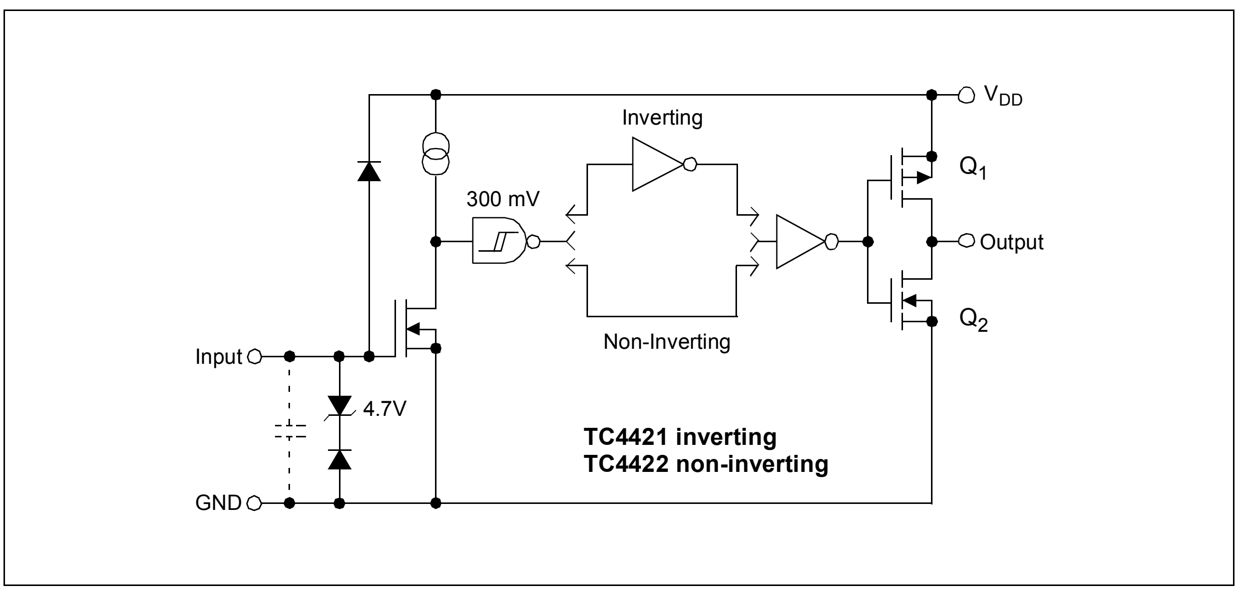

FIGURE 18: *Block Diagram of TC4421/22, 9A MOSFET Driver.*

During the transition time of the gate node from V_{DD} to GND, when Q_1 is turning on, the current flowing through Q_1 is divided between the output of the driver and the lower FET Q_2. The current flowing through FET Q_2 is considered "shoot-through current" or "cross-conduction current", which results in power dissipation within the driver. This is characterized in most MOSFET driver data sheets in a typical characteristic curve as "Cross-over Energy vs. V_{DD}". Once FET Q_2 is off, all of the current flowing through Q_1 goes to charge the gate capacitance of the external FET/IGBT.

Once the P-channel FET (Q_1) is fully enhanced, the system can now be viewed as a resistor charging a capacitor. This would be the R_{DS-ON} of the MOSFET driver charging the gate capacitance of the external FET. A representation of this is shown in Figure 19.

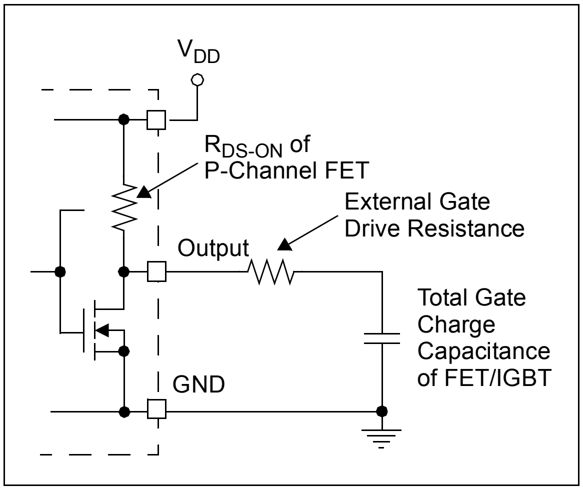

FIGURE 19: *Equivalent circuit of MOSFET Driver Charging an External Gate Capacitance.*

The diagram shown in Figure 19 models the P-channel FET during the charging mode as a resistor. This resistance is the R_{DS-ON} of the FET. This is typically shown in the MOSFET driver data sheet electrical characteristics table. The gate charge of the external MOSFET/IGBT is shown as a lumped capacitance, as was discussed earlier. This lumped capacitance is derived from the total gate charge value given in the data sheet. The external gate drive resistor shown represents any additional resistance that may be needed in the circuit, as was previously discussed.

Other configurations of MOSFET drivers do exist. Some of the earlier FET drivers were bipolar devices in which the output stage consisted of PNP and NPN transistors. These devices tended to draw more bias current during operation than do the newer CMOS devices. They were also slower, having longer propagation delay times.

Another MOSFET driver configuration available now is shown in Figure 20. This configuration has a bipolar drive stage in parallel with a MOSFET drive stage. The MOSFET drive stage gives the faster response time, while the bipolar stage helps in providing the peak currents. The conductivity of the bipolar stage is gated by the size and speed of the pre-drive stage.

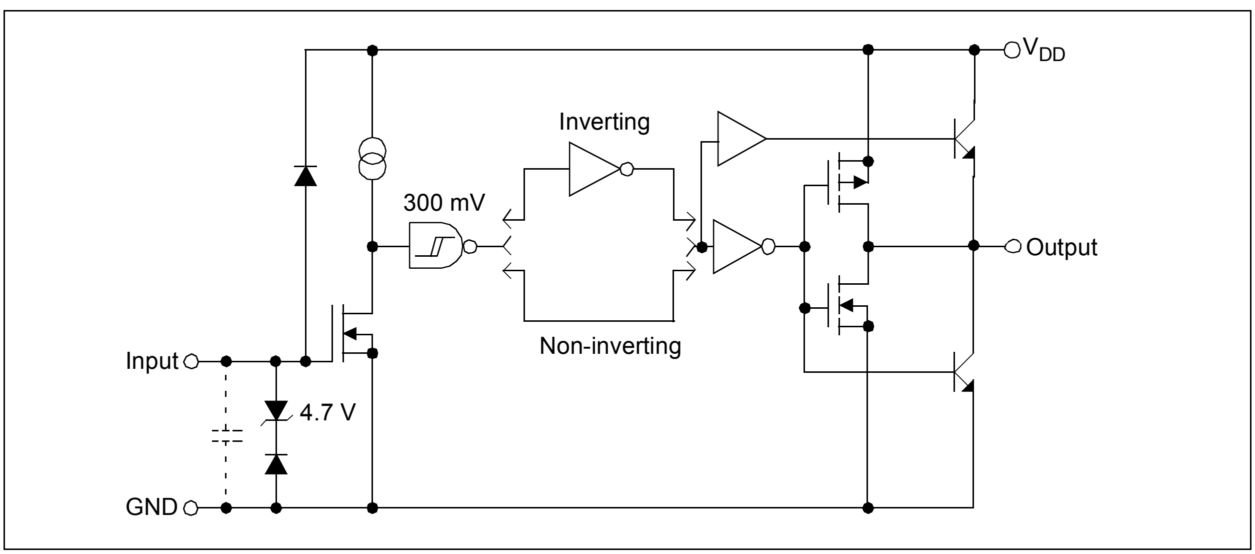

FIGURE 20: *Parallel Bipolar and MOSFET Output Stage MOSFET Driver.*

With the understanding of the MOSFET driver models shown in Figure 18 and Figure 19, as the R_{DS-ON} of the P-channel and N-channel FETs is reduced, the ability to charge and discharge the gate capacitance of the external FET faster increases. Any gate resistance that is external to the MOSFET driver will act to slow down the gate charging and, therefore, the turn-on and turn-off of the MOSFET/IGBT.

The correlation, then, between the MOSFET driver peak current rating and drive stage R_{DS-ON} is that as the driver peak current rating increases, the drive stage R_{DS-ON} decreases and more gate charge can be delivered to the gate of the external FET/IGBT in a shorter amount of time.

As stated at the beginning of this section, at this point, a power-switching element should have been chosen (MOSFET or IGBT), from which the gate charge value can be found. Remembering that the gate charge value must be matched with the gate drive voltage that will be used in the circuit. An assumption for desired turn-on and turn-off times based on desired switching losses is also needed.

When selecting the appropriate driver strength for your application, there are two methods that can be used.

The first method, which is a rough estimate type of method, uses the gate charge value and desired charge time to calculate the needed charging current. The equation below shows an example of this:

Q_{TOTAL} = Icharge * Tcharge

Example:

Q_{TOTAL} = 68 nC

Tcharge = 50 nsec.

Icharge = 68 nC/50 nsec.

Icharge = 1.36A

The charging current calculated using this method is an average/constant current. As described earlier, the current delivered by a MOSFET driver is not a constant current, so the value that is found using this method needs a rule to go with it in order to select the appropriate MOSFET driver rating. A good rule of thumb to use is that the average value found using this method is half the driver peak current rating. So for this case, a driver with a 3A peak current rating would be a good starting point. This method does not take into account any external gate resistance that will lengthen the charging and discharging time. It should also be noted with this method that, if a driver voltage is being used that is much less than the voltage at which the peak current rating of the MOSFET driver is given, additional buffer may be required.

The second method is more of a time constant approach that uses the MOSFET driver resistance (R_{DS-ON} of the P-channel for charging and N-channel for discharging), any external gate resistance and the lumped gate capacitance (from the total gate charge value) to select the appropriate driver. From earlier discussions, most IGBTs require some value of gate resistance to ensure that dynamic latch-up is avoided. In order to obtain the switching losses for the IGBT that are shown in the data sheets, the MOSFET driver can not dominate the gate resistance that is given in the test conditions of the data sheet. The equations for this method are shown below:

Tcharge = ((Rdriver + Rgate) * Ctotal)*TC

Where:

Rdriver = R_{DS-ON} of the output driver stage

Rgate = any external gate resistance between the driver and the gate of the MOSFET or IGBT

Ctotal = the total gate charge value divided by the gate voltage

TC = number of time constants

Example:

Qtotal = 68 nC

Vgate = 10V

Tcharge = 50 nsec.

TC = 3

Rgate = 0 ohms

Rdriver = (Tcharge/TC*Ctotal) - Rgate

Rdriver = (50 nsec./ 3 * 6.8 nF) - 0 ohms

Rdriver = 2.45Ω

Since this equation represents an R-C time constant, using a TC of 3 means that the capacitance will be charged to 95% of the charging voltage after the Tcharge time. Most MOSFETs are fully “on” by the time the gate voltage reaches 6V. Based on this, a TC value of 1 (represents 63% of charging voltage) may be more useful for the application and allow a lower current driver IC to be used.

Using the MOSFET driver information provided in Table 2, the appropriate MOSFET driver can be selected. Given that the gate drive voltage was 10V, the Rout-Hi @10V column should be used. The closest selection for a single output driver would be the TC4420/29, which is a 6.0A peak output current driver. Because the gate drive voltage used was 10V, the R_{DS-ON} of the 3A driver turns out to be too high. If a higher bias voltage was used, the 3A driver would be viable for this application. Or, as discussed in the previous paragraph, if the charge time to 63% of the drive voltage is used (TC = 1), a 3A MOSFET driver could also be used.

TABLE 2: MICROCHIP MOSFET DRIVER RATINGS

Device	Bias Voltage Rating	Peak Current Rating	Rout-Hi @15V	Rout-Lo @15V	Rout-Hi @10V	Rout-Lo @10V
TC1410/N (S)	4.5V - 16V	0.5A	15.0Ω	10.7Ω	18.7Ω	15.0Ω
TC1411/N(S)	4.5V - 16V	1.0A	7.5Ω	4.8Ω	9.8Ω	6.0Ω
TC1412/N (S)	4.5V - 16V	2.0A	3.7Ω	3.1Ω	4.8Ω	4.0Ω
TC1413/N (S)	4.5V - 16V	3.0A	2.6Ω	2.0Ω	3.4Ω	2.7Ω
TC4426/7/8 (D)	4.5V - 18V	1.5A	7.3Ω	7.3Ω	9.1Ω	9.0Ω
TC4426A/7A/8A (D)	4.5V - 18V	1.5A	6.5Ω	5.0Ω	8.0Ω	6.0Ω
TC4423/4/5 (D)	4.5V - 18V	3.0A	2.8Ω	2.8Ω	3.5Ω	3.5Ω
TC4420/9 (S)	4.5V - 18V	6.0A	2.25Ω	1.35Ω	3.15Ω	2.0Ω
TC4421/2 (S)	4.5V - 18V	9.0A	1.5Ω	0.95Ω	2.0Ω	1.25Ω
TC4467/8/9 (Q)	4.5V - 18V	1.2A	10.0Ω	8.5Ω	12.5Ω	10.0Ω

S = Single Output Driver, D = Dual Output Driver, Q = Quad Output Driver

SUMMARY

A large part of motor control design is based in the power portion of the circuit. This application note has touched on a few basic points of contrast and similarity between the MOSFET and IGBT as they apply to motor drive design. As advancements continue to be made with these devices, their use in motor drive applications will become more defined. The constant for either device will be the need for a high peak current drive source to turn the devices on and off. Microchip's line of MOSFET drivers will also continue to evolve to support future motor drive applications.

Further details on motor control circuits, reference designs and application notes can be found on Microchip's web site at www.microchip.com. Reference number 9 is one of the many documents that demonstrate how Microchip's PICmicro® microcontrollers and analog products can be used in a motor control system.

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Milan, Italy
Tel: 39-0331-742611
Fax: 39-0331-466781

Netherlands
P. A. De Biesbosch 14
NL-5152 SC Drunen, Netherlands
Tel: 31-416-690399
Fax: 31-416-690340

United Kingdom
505 Eskdale Road
Winnersh Triangle
Wokingham
Berkshire, England RG41 5TU
Tel: 44-118-921-5869
Fax: 44-118-921-5820

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