2 High-Power Semiconductor Devices

2.1 A VIEW OF THE POWER SEMICONDUCTOR MARKET

Power semiconductor components are at the core of any power electronic converter. They have a history of more than 50 yrs, a well-developed reach in technology, and have been a market success. As the technology behind these devices is not new, the differences in technology between the newly released components and the role of these changes are not always easy to understand. For this reason, a brief market survey is presented with the goal of outlining the efforts taken to set certain performance parameters for power semiconductor devices.

It is also important to understand the specifics of the semiconductor industry. As production is based on large capital equipment, the technology is developed in a cyclical manner. Different reports show that the global sales of semiconductors in 2003 increased by 14–16% to \$160–163 billion. The year 2004 showed an even higher growth of 16–19%, though we expect a slightly decelerated growth of 12.6% in 2005. Most data researchers estimate that average growth rate for the power semiconductor segment will outperform the growth rates of the overall semiconductor market.

Power semiconductor devices are at the heart of many modern industrial and consumer end-use applications and come in different size and ratings. The application objectives range from low power supplies of tens of watts to 4 MW locomotives or 10 MW steel rollers. The total market for power semiconductor devices is estimated at around \$18 billion per year in direct sales, with an estimated \$570 billion of sales of other products that power electronics directly enables [1]. This global market is estimated to increase to over \$36 billion before 2010 and it includes devices that are not the subject of this book due to their application nature or power level. It is worthwhile to note the estimated \$7 billion power discrete market and the \$5.2 billion analog power management market. All these numbers are estimates only and may differ from source to source. However, it is important to understand their correlation with the development of technology.

The power semiconductor devices most related to our book topic are MOSFETs, Insular Gate Bipolar Transistors (IGBTs), and diverse modern variations of thyristors such as silicon-controlled rectifiers (SCRs). Since 20 years ago, there has been spectacular improvement in technology and performance. The technological S-curves related to the power device capacity are shown in Figure 2.1 [2].

The power MOSFET device was introduced in the early 1980s with starting parameters of 3–5 A for the drain current, up to 400 V breakdown voltage, and

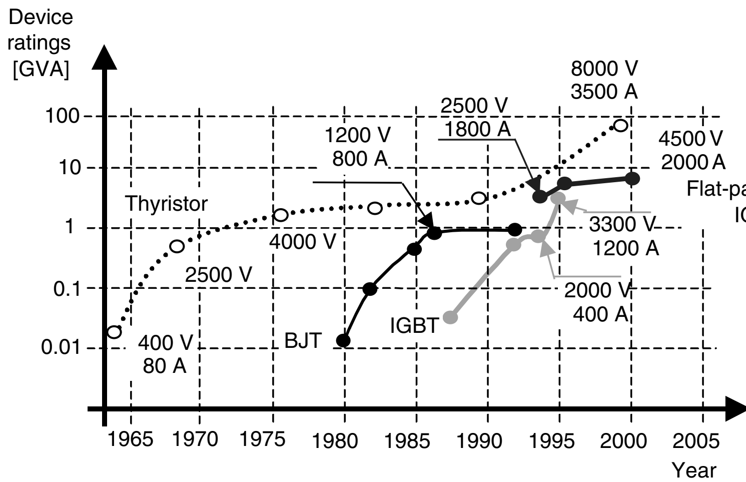

FIGURE 2.1 Technology S-curves with maximum device ratings as parameters.

turn-off time in the range of 1.2 msec. Technology development allowed improvement of ratings to different sets of 9 A/600 V or 100 A/50 V and decrease of the turn-off time to 600 nsec. The most recent technology advances include the CoolMOS devices, which are able to switch 20 A/600 V with a turn-off time of around 100 nsec. The overall market for power MOSFET devices was around $4.0 billion in 2004, with spectacular increases of up to 40% per year, during the last 10 years. Among all sorts of MOSFET devices, the largest market increase is now seen in the high-current applications in which new devices are released continuously.

IGBT devices combine the advantages of bipolar and MOSFET transistors into a device dedicated to power-switching converters operated under high current and high voltage. These devices are the most useful for the class of converters presented in this book and we will dedicate more space to the presentation of IGBT parameters.

The history of IGBTs also starts in the early 1980s but the real technological advent was in late 1980s and early 1990s when several generations of IGBT devices were developed by a number of companies [2]. Snapshots of performance evolution are as follows:

- 1986: starting parameters 50 A/600 V/3 ms
- 1990: commonly from 50 A to 400 A/1000 V/1.8 ms
- 1995: commonly from 50 A to 400 A/1200 V/1.3 ms
- 1996: 800 A/1600 V/1.6 ms
- 1997: 1200 A/3.3 kV/2.2 ms
- 2000: 50–400 A/1200 V/0.4–0.8 ms
- 2000: 1000 A/3.3 kV are available in smaller series
- 2004: commonly 1000 A/1700 V/1.2 ms, in small series up to 1200 A/6 kV

Given their application to high-power converters, the focus was on the improvement of parameters that relate to power conversion. During the last 20 yrs we have seen technology evolution with effects in:

- Current-handling capability, which increased four times since 1982
- Voltage-handling capabilities, which increased four times
- Turn-off time dropped 20 times, to around 100 nsec today
- Switching frequencies from 2 kHz in the early 1980s to 150 kHz in 1999 and 200 kHz nowadays

The evolution of the IGBT market has also been impressive over the last 10 yrs. The 1995 world market for IGBT was estimated at $200 million with the European market taking the largest share (approximately 45%). The global market increased to $800 million in 2003, and it is estimated to top $1 billion in 2005. This has caused a reduction in price for the final customer at an average rate of 10% per year, compared to the price levels in the last five years.

The success of power semiconductor devices in existing applications and the appearance of new applications encouraged the development of new concepts. Today, emerging high-frequency power semiconductor devices (e.g., 1 to 10 kW switched at 100 s kHz) are a very hot R&D topic. The early adopter segments accounted for about $298 million in power semiconductor sales in 1999 and were projected to grow to at least $765 million in five year, an annual growth rate of 21%. This can be compared with traditional devices that account for a $4.0 billion worldwide MOSFET market.

A special market segment refers to integrated circuits dedicated to power management and motor control. This sector is very dynamic, having seen large investments over the last years. The global market for motor control integrated circuits was approximately $910 million in 2000. This market was expected to grow at an average annual rate of 9% through 2005.

2.2 POWER MOSFETs

2.2.1 Operation

Power MOSFET devices are faster than bipolar transistors, as they do not have excess minority carrier that should be moved during turn-on and turn-off. A positive voltage is applied at turn-on on the gate circuit. The equivalent gate capacitance is charged through an external gate resistor. When this gate voltage rises above the $V_{GS(th)}$, a current starts circulating in the drain circuit with a (di/dt) determined by both the internal semiconductor structure and the external circuit. During this time interval, charge is stored within both C_{ds} (drain-source) and C_{gs} (gate-source). This state ends when drain current reaches the level of the current determined by the external circuit (the current is clamped at the load current). As no variation of the current is possible, the voltage across the gate-source circuit remains constant at a level depending on the load-circuit current. This level is called the Miller plateau. During this state, the gate-source capacitance has a constant voltage and all the

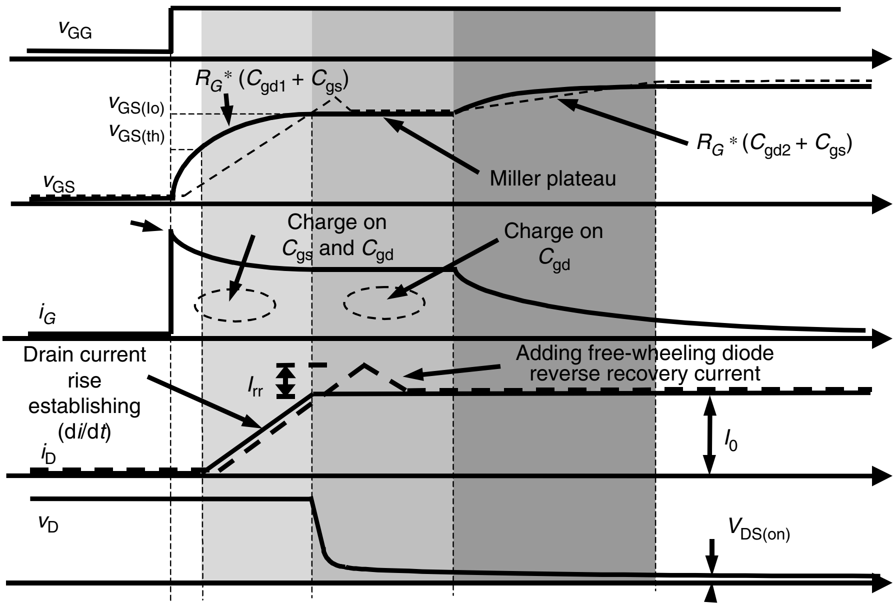

FIGURE 2.2 Model for the transient analysis in cut-off and active regions.

gate current charges the gate-drain capacitance. This determines the trip of the drain-source voltage towards the ground. When this voltage reaches a low level, the gate-source voltage increases to the level of the control voltage.

During the first two states of the turn-on transient, electrical charges are moved through the stray capacitances or depletion-layer capacitances, and the equivalent circuit model for transient analysis in cut-off and active regions is shown in Figure 2.2.

The last state shown in Figure 2.3 corresponds to a drain-source voltage $v_{DS} < v_{GS} - v_{GSth}$, when the MOSFET device enters the ohmic region. In power-switching converters, $v_{GS} \gg v_{GSth}$ (typically, 15 V > 4 V) and the boundary for the ohmic region is sometimes approximated with $v_{DS} < v_{GS}$, the equivalent circuit model

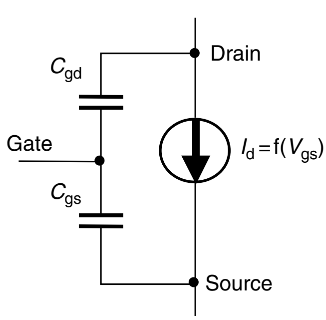

FIGURE 2.3 Generic turn-on waveforms for an IGBT/MOSFET power device.

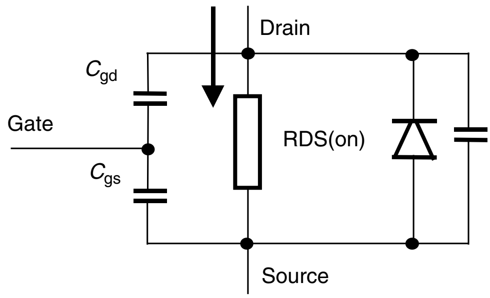

FIGURE 2.4 Model for the analysis of the ohmic region of a MOSFET device.

for the ohmic region shown in Figure 2.4. The drain-source resistance corresponds to the conduction loss, mostly arising from the drain-drift region. This is the most important performance index for MOSFET devices. Modern MOSFETs go as low as 5 mΩ $R_{DS(on)}$.

The capacitances C_{gd} and C_{gs} are not constant during the transient. A better model can be defined with values varying with the voltage across them. The capacitance C_{gd} shows a substantial change that can be approximated with a two-step variation (Figure 2.5).

The gate-source capacitance is constant on the first interval, increases with voltage on the second interval because of the gate oxide capacitance of drain overlap, and it is constant during and after the third interval. The final value is three to four times higher than the initial value (both values are in the range of few nF).

MOSFET datasheets provide values of C_{ISS}, C_{RSS}, and C_{OSS}. The following relationships help relate these parameters to inter-junction parasitic capacitances $C_{gd} = C_{RSS}$, $C_{gs} = C_{ISS} - C_{RSS}$, $C_{ds} = C_{OSS} - C_{RSS}$.

The switching speed is not only determined within the input capacitance and gate resistor circuit, but the Miller threshold level and the device transconductance

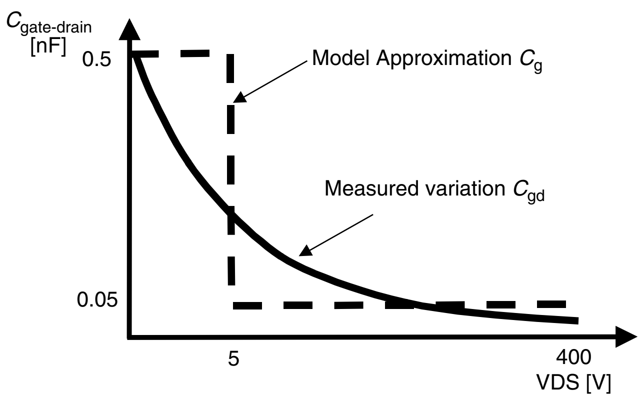

FIGURE 2.5 Variation of the gate-drain capacitance.

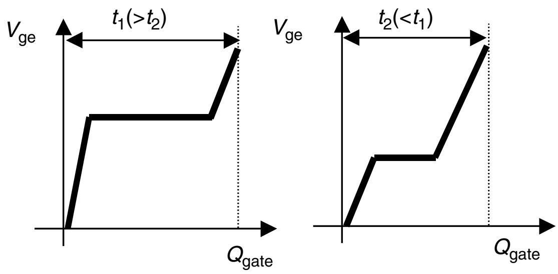

FIGURE 2.6 Gate-switching characteristics for two devices.

also matter. This is illustrated in Figure 2.6. The first slope (from zero to the Miller threshold) is determined by the input capacitance, which is higher for the second device. However, the second device has a higher transconductance and, therefore, requires less voltage at its gate for a given amount of collector current. The device with the smaller input capacitance is not always faster.

At turn-off, the gate voltage goes to zero and the gate's equivalent capacitance starts to discharge through the gate resistance (Figure 2.7). Both C_{gd} and C_{gs} are discharged at the first interval. When the gate voltage reaches the Miller plateau, it is clamped until the drain voltage increases to the bus voltage. During this interval, charge is changed with the C_{gd} capacitance only. Finally, the current decreases to zero at the last interval, whereas the drain-source voltage remains at the bus voltage level. The device can be considered turned-off when the gate voltage goes below the threshold voltage. Figure 2.7 also presents the turn-off characteristic of

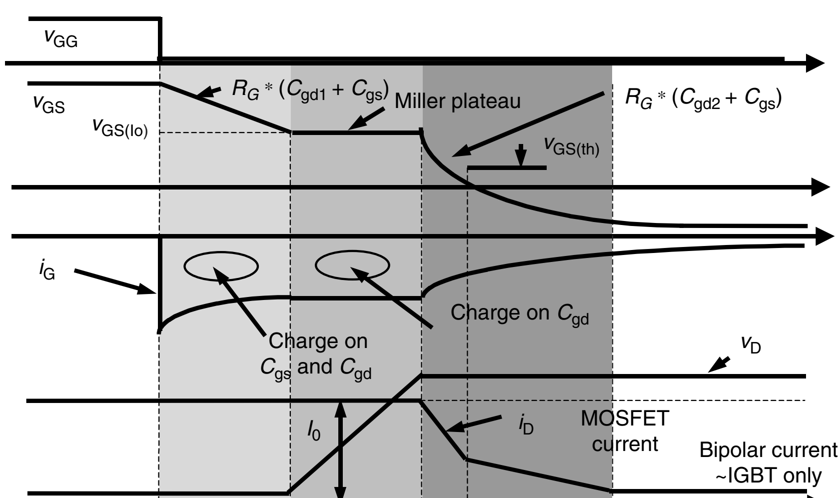

FIGURE 2.7 Generic turn-off waveforms for an IGBT/MOSFET power device.

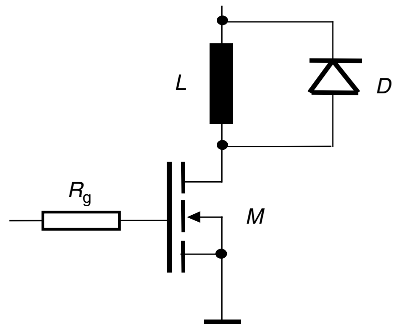

FIGURE 2.8 Using free-wheeling diodes for inductive switching.

an IGBT device. IGBT devices will be described in the next section. Notice the difference between the MOSFET turn-off with the current tail due to bipolar effect and the use of a bipolar voltage for the gate control.

The MOSFET's switching characteristics also depend on the external circuit. Switching currents on inductive loads impose special precautions, including a free-wheeling diode for the reactive current (Figure 2.8). As this diode is not an ideal switch, its reverse recovery current has an important influence on the switching characteristics. The dotted lines in both Figure 2.2 and Figure 2.7 illustrate how the reverse recovery current of the free-wheeling diode influences the switching of the power MOSFET.

There is also another reason for the use of the antiparallel free-wheeling diode. The internal structure of the MOSFET device features a p–n junction between the source and the drain. Under certain conditions, a negative current may free-wheel through this parasitic diode, which may happen especially within an inverter leg when a MOSFET turns-off and the other one turns-on. The conduction of the parasitic diode becomes a problem because of its slow turn-off (or long reverse recovery time) when the opposing MOSFET tries to turn-on. If the body diode of one MOSFET conducts when the opposing device is switched on, then a short circuit occurs similar to the shoot-through condition.

The historical solution to this problem consists in using two additional diodes for each MOSFET. A fast diode (can be a Schottky diode) is connected in series with the MOSFET source preventing the body diode from turning-on. A second fast diode is used in parallel with the MOSFET to allow a path for the free-wheeling current. Schottky diodes are nowadays available up to 200 V, whereas other fast recovery diodes are available at higher voltages. Moreover, MOSFET devices are mainly sold with the fast diode integrated within the same package for ease of use.

Numerous modern MOSFET devices eliminate this problem by creating a fast body diode. For instance, International Rectifier has introduced a 500 V HEXFET in the power MOSFET family, with fast body-diode characteristics that eliminate the need for additional Schottky and high-voltage diodes, reducing component

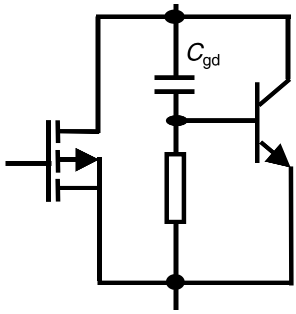

FIGURE 2.9 Parasitic bipolar transistor.

count, cost, and layout space. The maximum reverse recovery time for the body diodes in the L-Series HEXFET devices is less than 250 nsec, and even shorter for lower-current devices.

Note that the MOSFET semiconductor structure has a parasitic bipolar transistor formed with the body region of the MOSFET as the base, the source as the bipolar emitter, and the drain as the bipolar collector. The base of such transistors should be kept at a low voltage, which can otherwise cause negative effects.

- The MOSFET breakdown voltage will be reduced to the collector–emitter voltage of this transistor;
- The bipolar transistor can turn-on accidentally without any possibility of being turned-off by control (this is called MOSFET latch-up);
- A fast turn-off of the MOSFET would produce the turn-on of the parasitic bipolar transistor through the portion of the gate-drain capacitance than would connect the base to the collector. This can be prevented with series diodes on each drain.

Modern technology avoids the presence of this parasitic bipolar transistor (Figure 2.9). For instance, the modern MOSFET devices have $(\mathrm{d}v/\mathrm{d}t)$ larger than 10 000 V/μsec.

2.2.2 Control

Design of gate drivers depends on the switching characteristics. The switching times given in datasheets as electrical characteristics are for resistive load switching. The performance curves are for half-bridge inductive load, as they are the most prevalent application of IGBTs.

Circuits used to control power MOSFET devices are called gate drivers. A MOSFET gate driver has the simple task of providing a voltage for the gate control, and it does not require a large amount of current. The gate current is large at the beginning and limited by the resistance at the gate circuit. Depending on the level of the gate threshold voltage, gate control is usually performed with voltages at logic level (5 V) or from complementery metal oxide semiconductor buffers (15–20 V). Many integrated gate-driver circuits are available for both situations, along with protection circuits for fast shutdown (for e.g., TPS2812).

Given the small amount of capacitance to be charged, MOSFET gate drivers should ensure a fast variation of the control voltage with slopes below 20 nsec.

The gate-source voltage should not be higher than a datasheet parameter, $V_{GS(max)}$. This is determined by the requirement that the gate oxide not be broken down by a large electric field. Another consideration is the paralleling of MOSFETs, which is presented in Chapter 12.

2.3 INSULATED GATE BIPOLAR TRANSISTORS

2.3.1 Operation

IGBTs combines the advantages of bipolar transistors, such as low conduction losses, with the merits of MOSFETs, such as shorter switching times. For this reason, the switching behavior of IGBTs can be analyzed based on MOSFET models described earlier. The conduction interval is better modeled with the characteristics of a saturated bipolar transistor. Because of the smaller voltage drop at conduction, IGBT devices are used at higher voltages than the MOSFET devices (Figure 2.10).

Without entering into the details of the semiconductor structure, let us focus on the IGBT model presented in Figure 2.11. It considers the IGBT formed as a Darlington combination of a main MOSFET device and a pnp transistor. Unlike the conventional Darlington, the MOSFET device carries most of the current. The parasitic npn transistor has the same origin and effect as the parasitic transistor from the MOSFET structure.

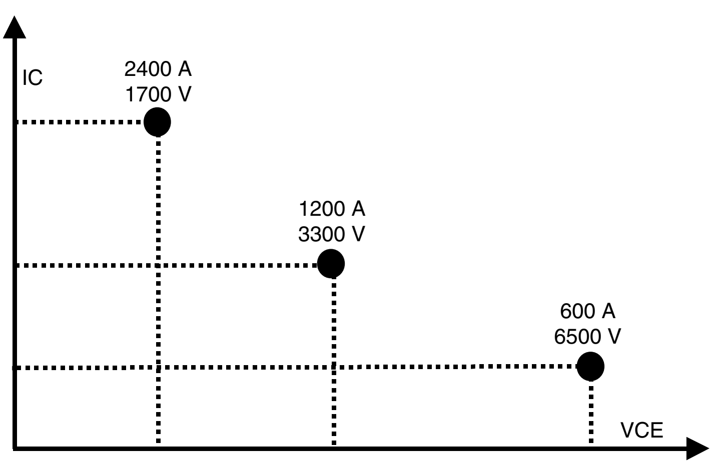

FIGURE 2.10 Present limits of the IGBT technology (e.g., from EUPEC product line).

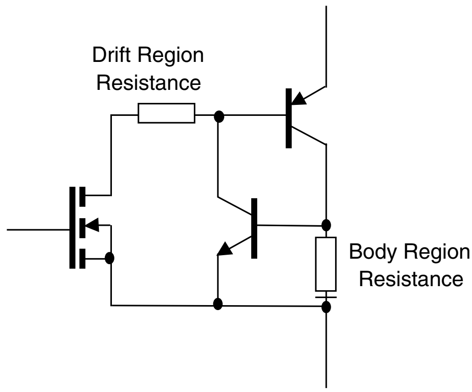

FIGURE 2.11 Equivalent model of an IGBT device.

Switching characteristics of IGBT devices are highly similar to those of the MOSFET devices. The major difference consists in the bipolar effect at turn-off, when a tail current still persists for a certain amount of time. Because of this tail current, the IGBT devices are not very suitable for use within zero-voltage switching (ZVS) applications and generally introduce additional switching loss.

The tail current is also the source of an interesting design trade-off. This current exists due to the charge stored in the drift region. As the MOSFET section is OFF and there is no reverse voltage applied to the device to generate a negative drain current, there is no possibility of removing the stored charge. This can finally be removed by recombination within the IGBT. Here comes the trade-off. The excess carrier lifetime is required to be large for a small voltage drop in the conduction state. This would determine a slow recombination and a long existence of the tail current.

The most-used method to minimize the magnitude of the tail current, or the magnitude of the bipolar current within the IGBT device. The device is designed to have 90% or more drain current passing through the MOSFET structure and only a small amount of current through the bipolar transistor, which can be achieved with a low beta of the pnp transistor.

An alternative technological solution to this problem is the so-called punch-through (PT) technology. The PT-IGBT minimizes current-tailing by shortening the duration of the tailing time, with an $n+$ layer acting like a sink for the excess holes. This buffer layer allows the drift region to be smaller than that of the NPT-IGBTs, resulting, consequently, in a reduced voltage drop in the conduction state.

2.3.2 Control, Gate-Drivers

The major requirements for the IGBT gate driver are highly similar to those of a MOSFET device. However, IGBT devices usually require a negative gate voltage for turning-off, with the exception of a class of IGBT devices designed for operation with unipolar voltage. Negative OFF-state control voltage and appropriate gate

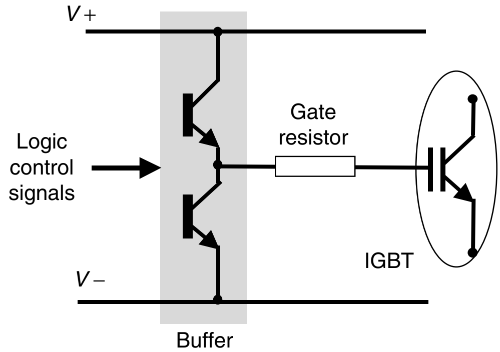

FIGURE 2.12 Gate-driver concept.

resistance can prevent cross-conduction. When the high-side IGBT turns-on within an inverter leg, the voltage across the low-side IGBT increases with a high $(\mathrm{d}v/\mathrm{d}t)$. This induces a current in the gate of the IGBT that may produce turn-on of the low-side device short of the DC bus. A negative gate voltage prevents this by providing a different path for the gate current. The same possible turn-on due to $(\mathrm{d}v/\mathrm{d}t)$ can take place within a MOSFET as well, but the input capacitance is different and the chances of cross-conduction are minimal. This can be understood by looking at the ratio of reverse transfer capacitance to the input capacitance, which is larger for IGBTs $(C_{\mathrm{res}}/C_{\mathrm{ies}})$. This produces an increased Miller effect and a larger noise is coupled from collector to gate. However, certain low-power IGBTs do not need negative gate voltage for turn-off, as their design minimizes C_{res} (reverse transfer). Another reason for the negative gate voltage at IGBTs is of the operation at higher voltages with increased $(\mathrm{d}v/\mathrm{d}t)$ coupling of noise.

Figure 2.12 shows the minimal requirements of the gate-driver circuit: a power supply able to ensure enough gate current, a gate-driving circuit, and a gate resistor. As the IGBT can float with respect to ground at the power stage, both the power supply and the gate circuitry should be isolated from the inverter ground. This gives room to a limited number of gate-driver configurations [4,5].

- Gate drivers with potential separation
  - Gate driver with inductive transfer of power (power supply of up to 1 MHz intermediate frequency) and a direct information transfer
  - Gate driver with inductive transfer of energy (power supply of up to 20 kHz intermediate frequency) and optocoupler transfer of information
- Gate drivers without potential separation
  - Gate driver with bootstrap for power supply of high-side and level shifter for switching control of information

In all these designs, a series resistor is employed at both turn-on and turn-off (Figure 2.12) that is usually implemented with a passive resistor. Advanced gate-driver design requires different resistors for turn-on and turn-off. The value of the

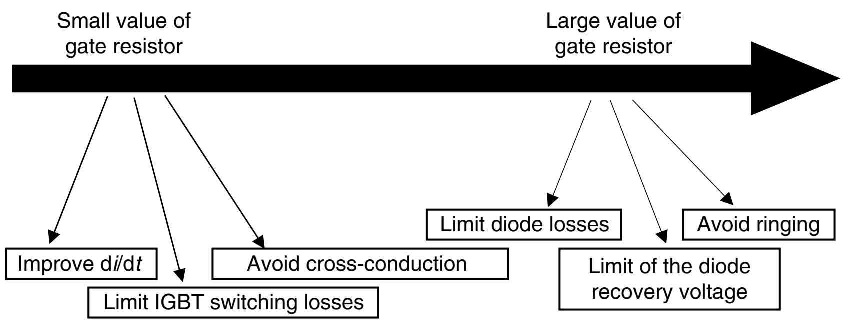

FIGURE 2.13 Effect of the gate resistor.

gate resistor within the range of values suggested by the IGBT/MOSFET manufacturer influences different aspects of the switching process. Figure 2.13 illustrates this graphically [3].

The gate current and the appropriate power of the voltage supply depend on the operating frequency, bias control voltages, and total gate charge. The total gate charge is published in IGBT/MOSFET datasheets, depending on gate-control voltage. The gate charge necessary for switching is very important to establish the switching performance of a MOSFET or IGBT. The lower the charge, the lower is the gate-drive current needed for a given switching time [7].

The average gate current can be calculated as $i_S = Q \cdot \text{freq}$. The total gate power can be estimated as $P = i_S \cdot (V_{G+} - V_{G-})$. Therefore, the power requirements for the gate circuit are reduced to a small power level and a high peak gate current. This peak current can be roughly estimated as: $I_{G(Peak)} = (V_{G+} - V_{G-})/R$.

2.3.3 Protection

A power converter equipment includes a set of protection circuits and features. Chapter 5 presents in detail the practical aspects of building a power converter. Before such a design can be accomplished, the datasheet information about the limits of operation should be understood.

The proper operation of an IGBT or MOSFET device is bounded by datasheet limitations. The collector current is limited to avoid latch-up. The maximum gate-emitter voltage is set by the gate oxide breakdown considerations. The maximum current that can flow under short-circuit with a maximum gate-emitter voltage is four to ten times the nominal rated collector current. The maximum collector–emitter voltage of an IGBT device is set with the breakdown voltage of the internal pnp transistor. The maximum junction temperature is 150°C.

A special datasheet information refers to the safe operating area (SOA). Both IGBT and MOSFET devices have square SOAs for short switching times. If the conduction intervals are longer, thermal aspects modify the SOA, as shown in Figure 2.14. Modern IGBT devices can operate at the corner of the SOA for 10 μs. This allows a protection circuit to trigger the gate signal and to protect the

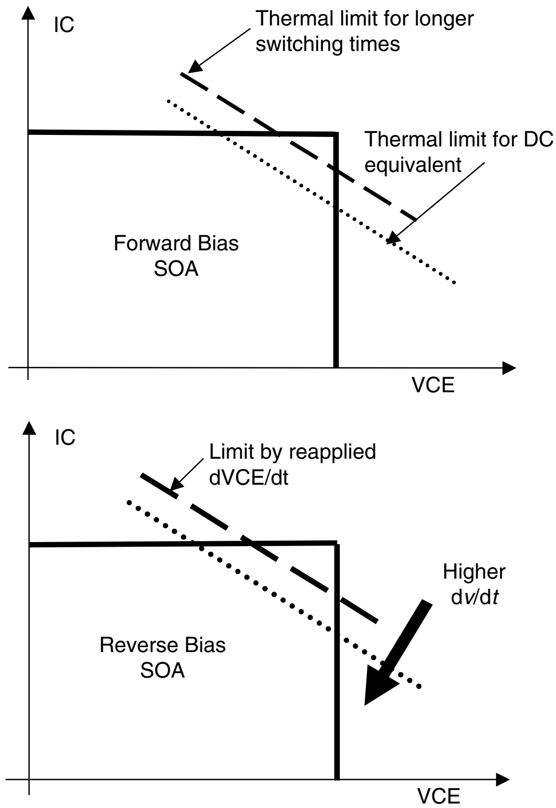

FIGURE 2.14 Ideal SOA and limitations due to special operation conditions.

IGBT at high currents. In other words, the IGBT can operate in short-circuit for up to 10 μs [4].

The parasitic components of the collector–emitter circuit determine a real switching characteristic far from a square one. The designer should make sure that this real characteristic is always inside the SOA (Figure 2.15). These trajectories depend upon stray inductances, parasitic capacitances, and the MOSFET's switching performance as (di/dt), (dv/dt). The IGBT package itself has a stray inductance of about 10–20 nH [4,18].

2.3.4 Power Loss Estimation

As the switching characteristics are the results of nonlinear phenomena, there are different methods to estimate the power loss. Power loss and efficiency of the power stage is very important, given the use of MOSFET and IGBT devices in power-conversion circuits. Observing the collector current and voltage waveforms, switching loss can be derived by calculating the areas of VI regions that correspond

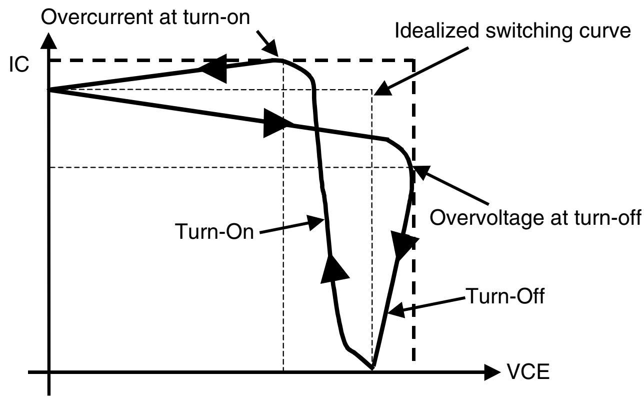

FIGURE 2.15 Switching characteristics.

to the switchings. The switching-loss energy at IGBT turn-on (Figure 2.2) is [8,9]:

$$E_{\mathrm{Ton}} = 0.5\left[V_{\mathrm{DC}} - L_{\mathrm{st}}\left(\frac{\mathrm{d}i}{\mathrm{d}t}\right)_{\mathrm{on}}\right]\frac{(I_{\mathrm{L}} + I_{\mathrm{RM}})^2}{(\mathrm{d}i/\mathrm{d}t)} + 0.5 I_{\mathrm{L}}\frac{V_{\mathrm{DC}}^2}{(\mathrm{d}v/\mathrm{d}t)_{\mathrm{on}}}$$

where we assume L_{st} as the stray inductance, V_{DC} the bus voltage, I_{RM} the peak recovery current of the adjacent diode, and $(\mathrm{d}i/\mathrm{d}t)$ and $(\mathrm{d}v/\mathrm{d}t)$ as the datasheet information about the selected semiconductor device.

The turn-off energy is calculated with (Figure 2.4):

$$E_{\mathrm{Toff}} = 0.5\frac{V_{\mathrm{DC}}^2}{(\mathrm{d}v/\mathrm{d}t)_{\mathrm{off}}} I_{\mathrm{L}} - 0.5\left[V_{\mathrm{DC}} - 2L_{\mathrm{st}}\left(\frac{\mathrm{d}i}{\mathrm{d}t}\right)_{\mathrm{off}}\right] \times \frac{(I_{\mathrm{L}})^2}{(\mathrm{d}i/\mathrm{d}t)_{\mathrm{off}}} + 0.5 k_t V_{\mathrm{DC}} I_{\mathrm{L}} t_{\mathrm{tail}}$$

The last term corresponds to the tail current at the IGBT turn-off and should miss at the same calculation performed for a MOSFET device. It can be seen that the gate drive circuit (especially the gate resistance) influences the switching losses by $\mathrm{d}i/\mathrm{d}t$, $\mathrm{d}v/\mathrm{d}t$, I_{RM}, and overvoltage.

Finally, the diode turn-off within an inverter leg is characterized by loss expressed by:

$$E_{\mathrm{Doff}} = 0.5\left[V_{\mathrm{DC}} + 2L_{\mathrm{st}}\left(\frac{\mathrm{d}i}{\mathrm{d}t}\right)_{\mathrm{diode}}\right]\frac{(I_{\mathrm{L}})^2}{(\mathrm{d}i/\mathrm{d}t)_{\mathrm{diode}}}$$

The conduction loss is calculated as:

$$P_{\mathrm{COND}} = \frac{1}{T}\int_0^T v_{\mathrm{on}}(t)i_{\mathrm{L}}(t)\mathrm{d}t$$

Integral across the fundamental period T can be reduced to integrals across all conduction intervals during a period.

Total losses can be calculated by adding up the switching and conduction losses, by taking into account the inverter topology, the modulation function for each device, and the operation mode or load power factor.

2.3.5 Active Gate-Drivers

The role of a passive gate-driver resistor has already been explained. It has also been shown that the value of the gate resistor influences different characteristics of the switching circuit (Figure 2.13). These performance aspects are not based on simultaneous phenomena and introduce the possibility of an idealistic control through a variable gate resistor. A variable value of the gate resistor can be achieved with an active gate control [6,10–16,19,20].

Historically, the use of active gate control was first mentioned in series connection of IGBT devices. This is required in medium-voltage applications, when the DC bus has values in thousands of voltage range. Series connection of IGBTs raises the problem of unequal voltage-sharing across these devices. The unequal voltage-sharing across the IGBTs is due to:

- Different delay times in gate driver and power semiconductor device
- Small parameter deviation among different devices
- Different reverse recovery behavior of the free-wheeling diodes
- Increased $(\mathrm{d}v/\mathrm{d}t)$ with the number of series-connected devices

Voltage-balancing between series-connected devices is traditionally achieved with individual snubbering of each IGBT device. To reduce the passive component count and volume, modern active-snubbering methods have been reported to limit $(\mathrm{d}v/\mathrm{d}t)$ and the overvoltage (Figure 2.16). Control of the collector voltage is achieved within an analog fast-feedback loop. Stability requirements imply design of a controller with poles at a frequency higher than gate circuitry, with a pole in the range of 1–10 MHz. Control bandwidth of 50–90 MHz is achieved with high-performance operational amplifiers. A significant loss reduction can be achieved by controlling the IGBT voltage in closed-loop operations only near the peak rating. Open-loop operation can be considered for the rest of the operation range. This obviously complicates the control circuit.

The downside of this solution can be seen at inductive loads. The IGBT voltage cannot respond to the gate voltage turn-on control until the free-wheeling diode has turned-off. The closed-loop approach charges the gate quickly, producing a very high $(\mathrm{d}i/\mathrm{d}t)$.

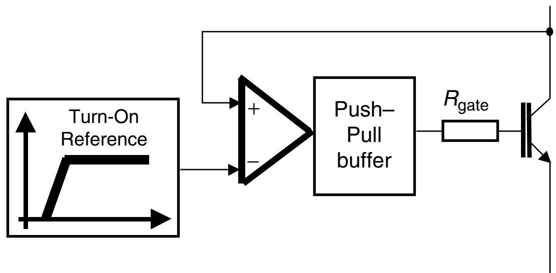

FIGURE 2.16 Active voltage balancing circuit used within series-connected IGBTs.

Historically, the second step in active gate control was in short-circuit protection. If the collectors circuit experiences a short-circuit, the protection circuit tries to shut the gate down and cut the current. This produces a large (di/dt) and a large overshoot. The equivalent gate resistance increases when the protection acts to turn the IGBT off with a soft shutdown. This avoids the large (di/dt) and the large voltage overshoot.

Voltage-overshoot protection can be achieved by including an additional transistor stage in the gate driver (Figure 2.17) [17]. At turn-off, Q_{prot} is turned on and the current is discharged through it. When the collector voltage reaches the breakdown voltage of the Zener diode, a current will flow through the gate of Q_{prot} and will

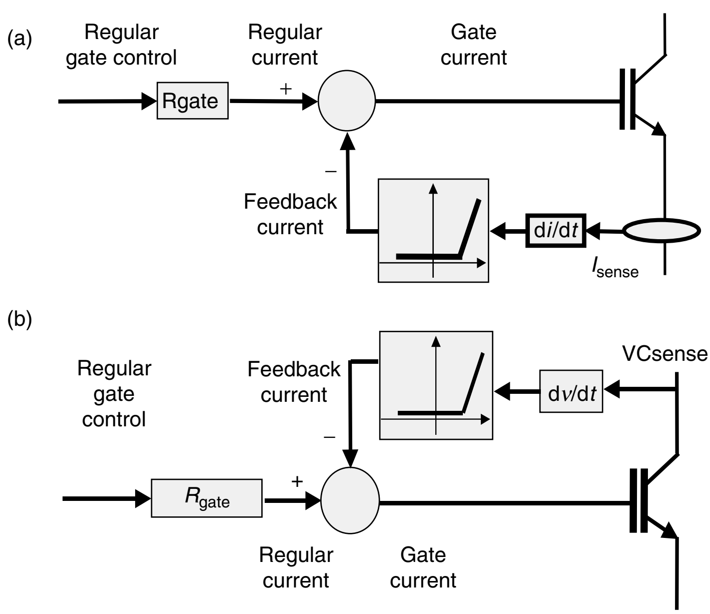

FIGURE 2.17 Principle circuits for active gate control: (a) based on (di/dt) optimization; (b) based on (dv/dt) optimization.

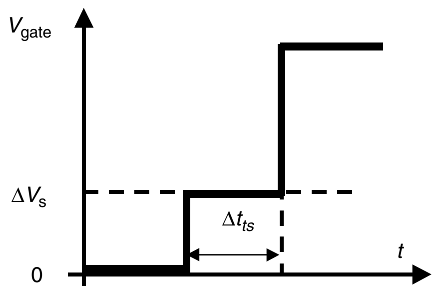

FIGURE 2.18 Principle of a simple experiment.

turn it off. The remaining current would flow through $R_{gate(off)}$ slowing down the (dv/dt) rate.

Modern gate drivers adjust (di/dt) and/or (dv/dt) independently according to criteria, such as electron magnetic interference (EMI) emission control or efficiency improvement through snubberless operation. For instance, the gate resistor can be optimized to reduce EMI emission through controlled (di/dt) and/or (dv/dt). Usually this value increases loss. These two constraints require different values of the gate resistors during operation.

There are different methods for active control of (di/dt) and (dv/dt) (Figure 2.17). The simplest control is the feedback control of the gate current based on the device current or voltage slopes. Sensing the current or voltage slopes is carried out with a shunt resistor, on Kelvin emitter, on the information resulting from the Miller effect sensing.

The active gate control is not easy to implement. The circuit designer faces several constraints related to a fast event time scale that does not allow delays within the analog circuit and to a feedback dependence on IGBT parameters.

Consider the experiment shown in Figure 2.18 and the results in Figure 2.19 [6]. The gate voltage has an intermediate voltage level that decreases the gate current level on the first slope of turn-on. The voltage level ΔV_s and the length of the

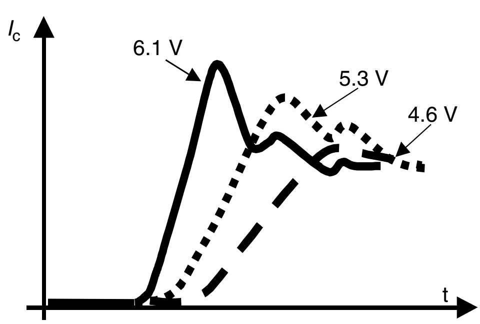

FIGURE 2.19 Results of the simple experiment.

time interval Δt_{ts} within this voltage level is adjusted. The IGBT/MOSFET behavior is a result of variable inductance. Despite the clear demonstration of the principle of active gate control, this experiment is not easy to implement.

Different solutions are reported in the literature for active gate control, but no one has yet seen its implementation in semiconductor devices [6,10–16,19,20]. Generally, the strong nonlinear character and the detection of the Miller plateau are a problem. Despite the hesitation to implement this method in semiconductor devices, the advantages that could be achieved are very important. It is therefore highly possible that we will see IC or discrete gate drivers use this method to control certain IGBT/MOSFET power modules.

2.4 GATE TURN-OFF THYRISTORS

As shown in the introduction to this chapter, gate turn-off thyristor (GTO) devices are used at high levels of current and voltage. They are a derivative of SCR devices, with a p–n–p–n structure, and can control both turn-on and turn-off processes. The operation is based on conventional recombination processes and the physics of junctions. The major drawback of these devices is the large current required in the gate circuit for turn-off. Moreover, the GTO has a very low gain which means it requires a sophisticated and expensive gate drive. It is therefore impractical to use a charge-extracting drive circuit, and so the GTO has a "tail effect," whereby the device still conducts while the minority carriers combine naturally.

2.5 ADVANCED POWER DEVICES

Despite the technology saturation in what concerns conversion circuits and converters, the power semiconductor sector is still dynamic. There is continuous development along existing devices like MOSFET and IGBT. New generations of IGBTs and MOSFETs are introduced each year in the market and their performance is continuously improving, especially through the design rule improvement (the pitch resolution in defining the shape of each semiconductor region). However, the most exciting news about totally new devices is their ability to change performance patterns through disruptive innovation.

A generic trend in emerging power semiconductor devices is the use of new substrate semiconductors made of silicon-carbide. These devices are not yet a market success, but show promise for the future.

More spectacular are the results in the development of power semiconductor switches for medium- and high-voltage applications. A good example is the integrated gate commutated thyristor (IGCT). Its architecture is combining the best features of an IGBT and a GTO. The new solid-state switch is for medium-voltage applications from 2 to 6.9 kV, with maximum ratings of 4000 A, which builds upon the drawbacks of IGBTs that have high conduction losses and GTOs that are slow and require additional circuitry.

The new IGCT makes possible designs that have not been feasible in the past. With these devices, engineers need not design around IGBT and GTO trade-offs, which often impose limits on starting torque and regeneration ability of motor drives.

Another power semiconductor device that is picking up in the market is the CoolMOS, a MOSFET with a special structure, rated up to 600 V, and able to switch up to 50 A. These devices change the entire way we think about power converters. Design of the power stage is limited by the parasitics of the implementation (printed circuit or busbar). The idea of switching, say, a 400 V bus at 250 kHz, pushes the designer to be very careful while designing circuit details.

2.6 PROBLEMS

P.2.1 Try to explain the variation of the gate-drain and gate-source capacitances.

P.2.2 The ON-resistance of a power MOSFET equals 120 mΩ at a junction temperature of 25°C and increases linearly with temperature up to 200 mΩ at 100°C. Calculate the conduction loss in function of the operation temperature if the load resistance is 10 Ω and the supply voltage is 150 V for a chopper operation.

P.2.3 Imagine a hybrid power switch made up of a bipolar transistor and a power MOSFET connected in parallel. What would be the benefits of such a device?

P.2.4 Qualitatively sketch the collector current versus time during turn-off for a short lifetime IGBT and for a long-lifetime IGBT and explain the differences.

P.2.5 Qualitatively sketch the collector current change during the turn-on of an IGBT device controlled through different gate resistors.

P.2.6 Qualitatively sketch the collector current change during the turn-off of an IGBT device controlled through different gate resistors.

P.2.7 Write a computer program for power-loss estimation based on the equations shown in this chapter and run this program for a simple case of a single IGBT switching a load resistor of 20 Ω, at 20 kHz, from a DC bus of 400 Vdc. Consider a real IGBT device along with the manufacturer datasheet and compare the results with those given in the datasheet.

P.2.8 Consider a MOSFET and an IGBT with the same breakdown voltage and the same current rating. How would you compare the gate-drain and gate-source capacitances of these two devices?

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