Module 1

Power Semiconductor Devices

Lesson 8

Hard and Soft Switching of Power Semiconductors

This lesson provides the reader the following

(i) To highlight the issues related to device stresses under Hard switching;

(ii) To suggest means of reducing such stresses with external circuitry;

(iii) To propose alternative switching methods for stress reduction;

(iv) Enable the choice of the appropriate switching strategy

Soft and Hard Switching

Semiconductors utilised in Static Power Converters operate in the switching mode to maximise efficiency. Switching frequencies vary from 50 Hz in a SCR based AC-DC Phase Angle Controller to over 1.0 MHz in a MOSFET based power supply. The switching or dynamic behaviour of Power Semiconductor devices thus attracts attention specially for the faster ones for a number of reasons: optimum drive, power dissipation, EMI/RFI issues and switching-aid-networks.

With SCRs' 'forced commutation' and 'natural (line) commutation' usually described the type of switching. Both refer to the turn-off mechanism of the SCR, the turn-on dynamics being inconsequential for most purposes. A protective inductive snubber to limit the turn-on $^{di}/_{dt}$ is usually utilised. For the SCRs' the turn-off data helps to dimension the 'commutation components' or to set the 'margin angle'. Conduction losses account for the most significant part of total losses.

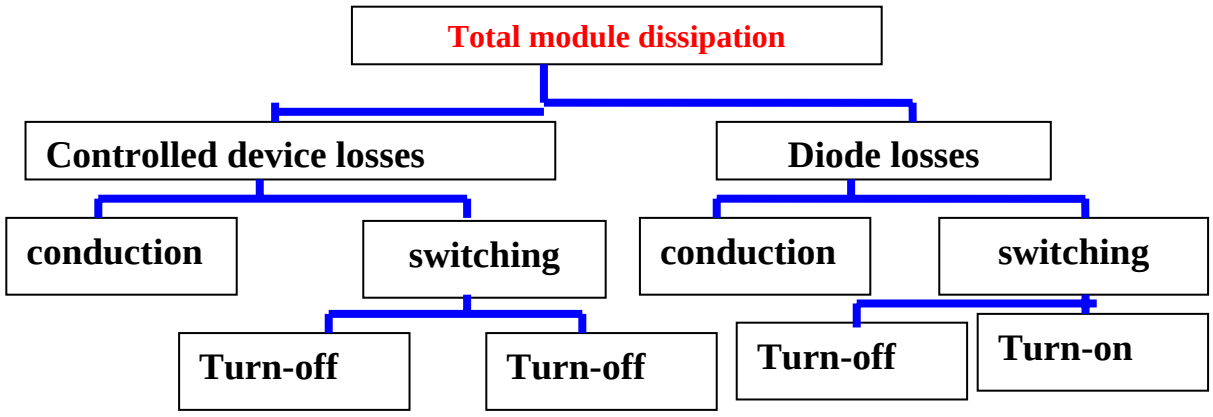
Present day fast converters operate at much higher switching frequencies chiefly to reduce weight and size of the filter components. As a consequence, switching losses now tend to predominate, causing the junction temperatures to rise. Special techniques are employed to obtain clean turn-on and turn-off of the devices. This, along with optimal control strategies and improved evacuation of the heat generated, permit utilisation of the devices with a minimum of deration.

This chapter first examines the switching process, estimates the device dissipation and indicates design procedures for the cooling system.

Losses in Power Semiconductors

A converter consists of a few controlled and a few uncontrolled devices (diodes). While the first device is driven to turn-on or off, the uncontrolled device operates mainly as a slave to the former. Power loss in the converter is the aggregate of these losses. Occasionally the diode and the controlled device are housed in the same module. The losses corresponding to each contribute to the temperature rise of the integrated module.

The losses can be segregated as follows:



Conduction Losses

Conduction losses are caused by the forward voltage drop when the power semiconductor is on and can be described by (with reference to an IGBT)

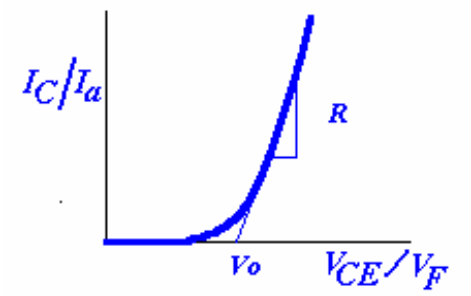


Fig. 3.1 Approximate forward voltage of IGBT and diode

$$W_C = V_{ce\ (sat)}(I_c).I_c$$

where I_c is the current carried by the device and $V_{ce(sat)}(I_c)$ is the current dependant forward voltage drop. This drop may be expressed as

$$V_{ce\ (sat)}\ (I_c) = V_0 + R\ .\ I_c$$

This relation defines the forward drop of an IGBT in a similar manner to a diode. A part of the drop is constant while another part is collector current dependent.
The given data should be used as follows: Using the numerical value is the most simple way to determine conduction losses. The numerical value can be applied if the current in the device is equal or close to the specified current - data sheet numerical values are specified for *typical* application currents.

The graph most accurately determines conduction losses. The conditions in which the data are used should correspond to the application. To estimate if a power semiconductor rating is appropriate, usually the values valid for elevated temperature, close to the maximum junction temperature T_{Jmax} , should be used to calculate power losses because this is commonly the operating point at nominal load.

Blocking Losses

Blocking losses are generated by a low leakage current through the device with a high blocking voltage.

$$W_B = V_b(I).I_L$$

Where I_L is the leakage current and $V_b(I)$ is the current dependemt blocking voltage. Data sheets indicates leakage current at certain blocking voltage and temperature. The dependence between leakage current and applied voltage typically is exponential; this means that using a data sheet value given for a blocking voltage higher than applied overestimates blocking losses. However in general, blocking losses are small and can often, but not always, be neglected.

Switching Losses

IGBTs are designed for use in switching converters and not for linear operation. This means switching time intervals are short compared to the pulse duration at typical switching frequencies, as can be seen from their switching times, such as rise time t_r and fall time t_f in the data sheets. Switching losses occur during these switching intervals.

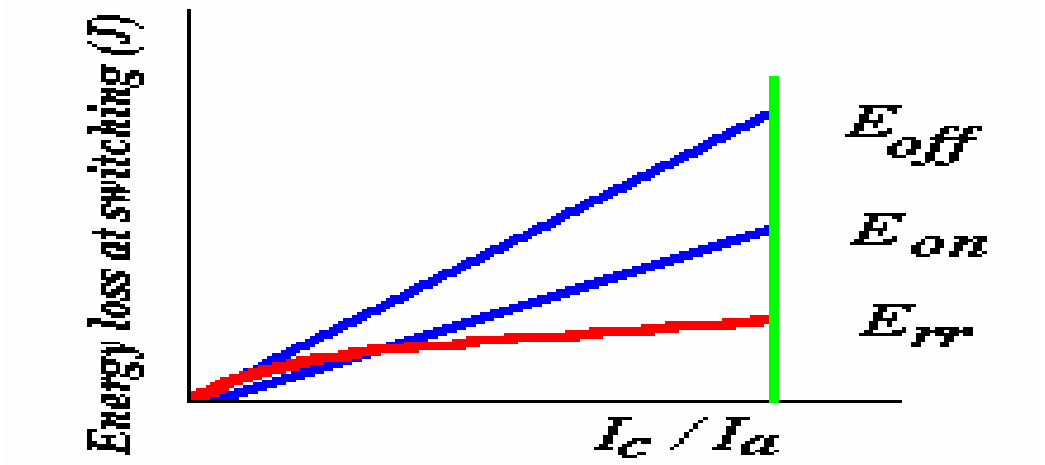


Fig. 3.2 Switching losses (appx)

For IGBTs they are specified as an amount of energy, $E_{on/off}$ for a certain switching operation. $E_{on/off}$ are the energy dissipated at turn-on/turn-off respectively. Using the numerical value is again the most simple way to determine switching losses. The numerical value can be applied if the switching operations are carried out at the same or similar conditions as indicated in the data sheet. Graphs for $E_{on}(I_C)/(R_G)$, $E_{off}(I_C)/(R_G)$ with collector current I_C and gate resistance R_G are provided.

The graphs permit the most accurate determination of switching losses, given the parameters of the converter: R_G and converter current I_C.

Diode

A surge voltage occurs when the free-wheel diode recovers. Consider a converter leg. The lower device is off and that the load current is circulating through the free-wheeling diode of the upper device. Now if the lower device turns on, the current in the free-wheel diode of the upper device decreases during the overlap period and the load current begins to commutate to the lower device. It becomes negative during reverse recovery of the upper free-wheel diode. When the free-wheel diode recovers, the current in the circuit associated to the diode jumps to zero. The parasitic line inductance L_p develops a surge voltage equal to Lp di/dt in opposition to the decreasing current. This di/dt is dictated by the recovery characteristic of the free-wheel diode. Fast recovery "snappy" diodes can develop very high recovery di/dt when they are hard recovered by the rapid turn-on of a device in series with it in the same converter leg. These diodes take a smaller time to quench the reverse recovery current compared to a soft recovery diode.

The off-state losses of the main device and the turn-on dissipation may be neglected for most cases. With an IGBT driven DC-DC chopper as an example, the dissipation can be estimated as:

IGBT dissipation = Conduction losses + Switching losses

$= [\square . V_{ce(sat)} I_c] + [f_c(E_{on} + E_{off})]$ Watts

Diode dissipation = Conduction losses + Reverse recovery losses

$= [(1 - \square\square_F V_F] + [f_c E_{rr}]$

where, $\square$ is the conduction duty ratio, f_c the switching frequency and E_{on} , E_{off} , E_{rr} are the respective energy losses, Fig 3.2, data for which is provided by the device manufacturer.

The values of E_{on} , E_{off} , E_{rr} are at the rated values only and have to be adjusted to the working values of voltage (DC bus), V_{CE} (working) and load current, I_c.

$$E_{on} / E_{off} / E_{rr(working)} = E_{on} / E_{off} / E_{rr(working)} \bullet \left[V_{CE(working)} / V_{CE(rated)} \right]$$

$$E_{on} / E_{off} / E_{rr(working)} = E_{on} / E_{off} / E_{rr} \left[I_C / I_{C(rated)} \right]^{a/b/c}$$

Where, a, b and c are constants.

The power device in a converter mostly sees an inductive load. A simple circuit illustrating such a situation is shown in Fig. 3.3. Corresponding ideal waveforms are also indicated. The free-wheeling diode FWD, across the load is essential for clamping the induced voltages across the inductance when the device switches off. However, its presence causes the supply voltage, V_s to appear across the transistor whenever it carries part of the inductor current in overlap mode with the FWD during both turn-on and turn-off modes. This causes the transistor switching dissipation to increase.

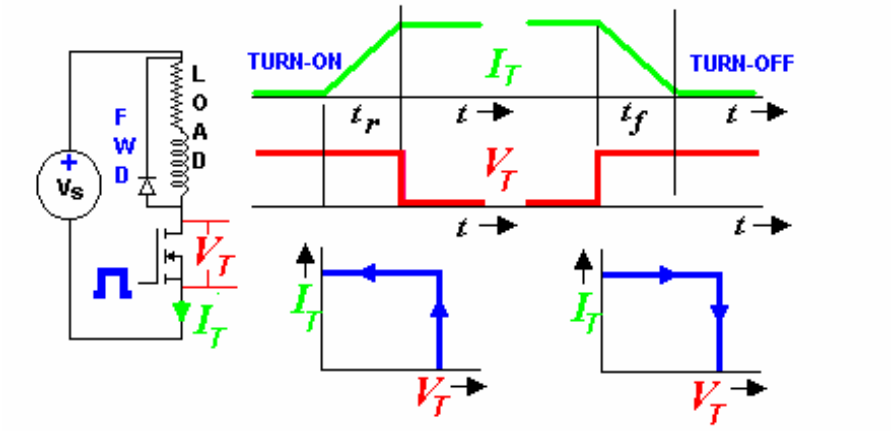


Fig. 3.3 Typical current and voltage transients during turn-on and turn-off of a clamped-inductive load and transitions in the V-I plane.

An RCD Switching-aid-network connected across the device reduces turn-off dissipation, Fig. 3.2. The controlled rise of the collector voltage of the transistor aids this process. However, turn-off energy is accumulated in the SAN, which is ultimately dissipated in the resistor. The RCD does not also help reduce turn-on dissipation when the reverse recovery current of the diode and the SAN current add up with the load current with V_s again appearing across the device.

Example 3.1

Derive the expression for the power dissipation during turn-on and turn-off of a transistor unassisted by a SAN. The supply voltage is V_m, peak load current I_m, and t_r, t_{off} being the turn on and turn-off times. Assume idealised waveforms.

Solution

The transition of the swichings in the V_C - I_C plane is rectangular. The energy dissipated in each turn-off switching cycle is

$$W_T = \int_0^{t_{off}} V_T . I_T dt = \frac{1}{2} . V_M . I_M . t_f$$

If actual waveforms are considered the dissipation is close to about double the above figure.

The dissipation at turn-on is, similarly 1/2. $V_M.I_M.t_{on.}$

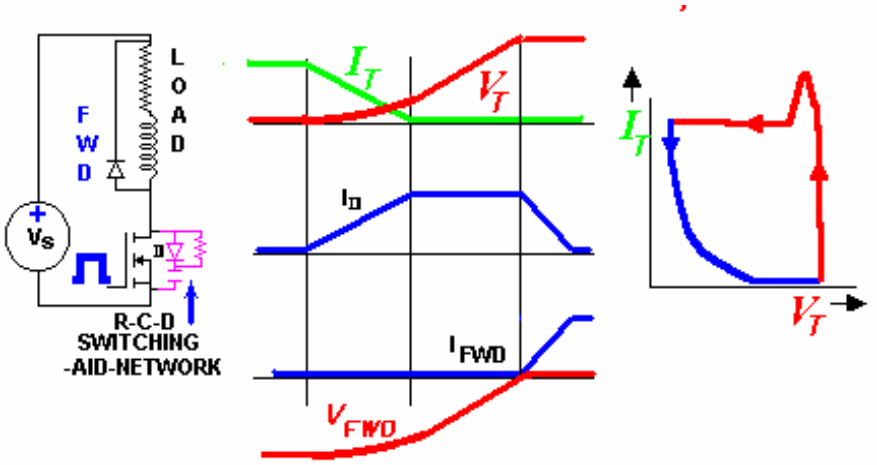


Fig. 3.4 Current and voltage waveforms at the Main Terminals of the switch with an R-C-D SAN, and in the associated FWD and SAN diode

Example 3.2

For a transistor carrying a collector current I_M and having a turn--off time t_f, find the details of a RCD SAN to restrict the voltage rise at the end of t_f to half the supply voltage. Calculate the corresponding losses in the transistor and in the SAN.

Solution

The action of the SAN in restricting the rise of transistor voltage till the current in it is extinguished is illustrated in Fig. 3.4.

Since the current is assumed to fall linearly during the period t_f, the collector voltage rises as:

$$V = V_0\left(\frac{t}{t_f}\right)^2 = \frac{I_M . t_f}{2C}\left(\frac{t}{t_f}\right)^2$$

Where V_0 is the voltage at the capacitor at the end of turn-off time t_f.
Thus,

$$V_0 = \frac{I_M\, t_f}{2C}$$

$$i = I_M\left(1 - \frac{t}{t_f}\right)$$

The Transistor current can be written as:
The dissipation in the transistor is

$$W_T = \int_0^{t_f} v.idt = \int_0^{t_f} \frac{I_M^2 . t_f}{2C}\left(1 - \frac{t}{t_f}\right)\left(\frac{t}{t_f}\right)^2 dt = \frac{I_M^2 . t_f^2}{2C} . \frac{1}{12} Watts$$

When the transistor switches off, the nearly constant load current linearly charges up the capacitor till it reaches the supply voltage. Subsequently, The FWD is positively biased and there

is a short period of over-lap between the FWD and the SAN diode. During this period, the capacitor over-charges to some extent.

If V_0 is the capacitor voltage when the transistor current is extinguished,

$$CV_0 = \int_{t_1}^{t_2} i.dt = \frac{1}{2} I_M t_f$$

If this V_0 is about 1/2 V_s,

$$C \geq \frac{I_M t_f}{V_s}$$

The energy dissipated in the SAN resistor which is also the energy shifted to the SAN from the transistor during turn-off is

$$P_R = \frac{1}{2} C V_M^2 F$$

Where F is the switching frequency. The resistance should be able to limit the transistor current to its peak rating. Thus,

$$R \geq \frac{V_s}{I_{CM} - I_M - I_{rr}}$$

I_{rr} is the reverse recovery current of the FWD.
If the capacitor has to discharge completely during the ON time,

$$C \geq \frac{I_M .t_f}{V_s - R.I_M}$$

In a Sine-PWM controlled converter with a peak value of the fundamental current equal to I_{cp}, the conduction losses in the IGBT would be

$$Wc = \delta.T \int_0^{\pi} I_c V_{ce(sat)} d\theta \qquad = \tfrac{1}{2} \delta.T \left[\frac{2\sqrt{2}}{\pi} I_{cp} V_o + I_{cp}^2 .R \right]$$

Where V_o and R are as shown in Fig 3.1. For the diode the dissipation is

$$W_F = \frac{1}{2} [1 - \delta] . \frac{2\sqrt{2}}{\pi} I_{cp} V_{od} + I_{cp}^2 R_d$$

Soft switching

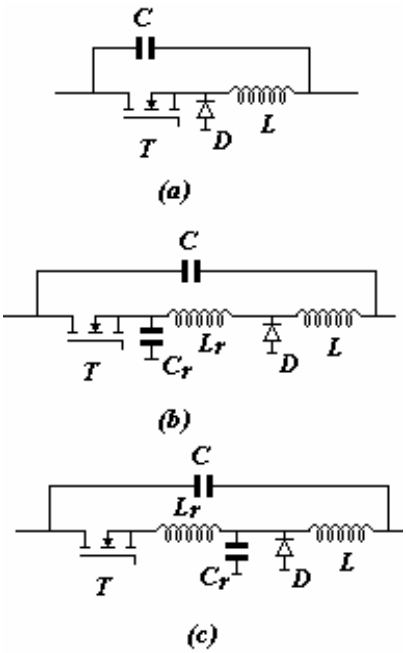


Fig. 3.5 Basic topologies for a) Hard switch, b) Zero-voltage switch and c) for a Zero current switch

Hard switching and its consequences have been discussed above. Reduction of size and weight of converter systems require higher operating frequencies, which would reduce sizes of inductors and capacitors. However, stresses on devices are heavily influenced by the switching frequencies accompanied by their switching losses. It is obvious that switching-aid-networks do not mitigate the dissipation issues to a great extent. Turn-on snubbers though not discussed, are rarely used. Even if used, it would not be able to prevent the energy stored in the junction capacitance to discharge into the transistor at each turn-on. Soft switching techniques use resonant techniques to switch ON at zero voltage and to switch OFF at zero current. There are negligible switching losses in the devices, though there is a significant rise in conduction losses. There is no transfer of dissipation to the resonant network which is non-dissipative. The two basic configurations are as shown in Fig. 3.5.

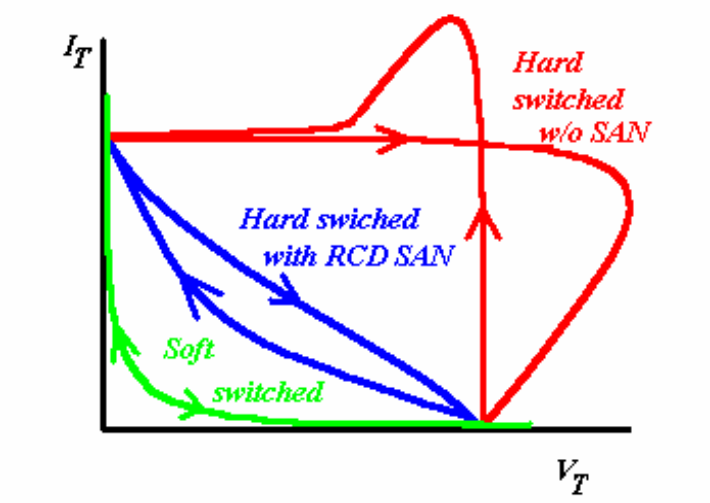


Fig. 3.6 Switching loci for a Hard-Switched converter without Switching-Aid-Networks, with the SAN and for a Soft-Switched converter operation

The switching trajectory in the voltage-current plane of a device is illustrated in Fig. 3.6 comparing the paths for that of a Hard-switched operation without any SAN, a Hard-switched with a R-C-D Switching-Aid-Network and a resonant converter. It is indicatve of the stresses and losses. A designer would prefer the path to be as close as possible to the origin.

A Zero Current Switch based converter is provided as illustration to the soft switching mechanism. It is equivalent to the topology shown above. The input capacitor and the one across the diode may be combined to arrive at this topology.

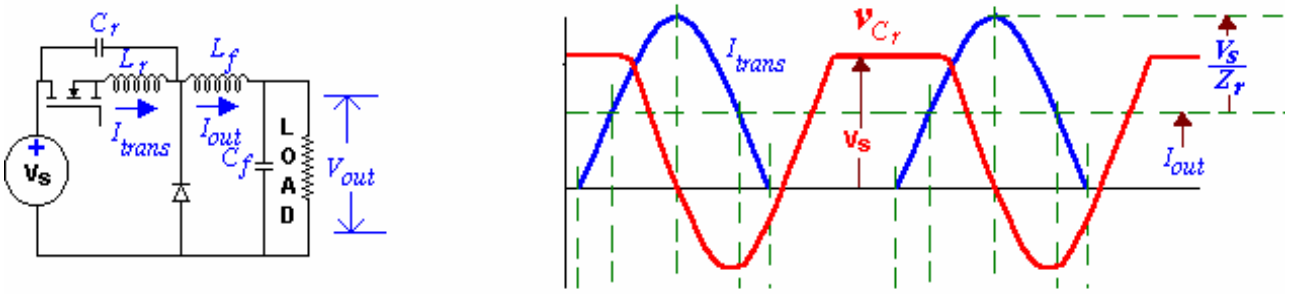


Fig. 3.7 A ZCS resonant buck converter

The ZCS converter is considered to be in stable operation with Load current I_{trans} flowing through the diode and the inductor L_f. The Capacitor C_r is charged to V_s. On switching the transistor ON the current in it ramps up from zero but the diode continues conduction till this current reaches the load current I_{out} level. Subsequently, the load current and the resonating current flows through the transistor. This current reaches a natural zero when the negative magnitude of the resonating current equals the load current. The transistor thus switches in the Zero Current mode for both turn on and turn off. The diode, on the other hand switches in the Zero Voltage mode under both situations. It must be noted that the peak current stress on the transistor is high . The peak voltage stress on the diode is also about twice the supply voltage. Both these stresses are significantly higher than that in a comparable Hard switched buck converter. Consequently,

while switching losses are practically eliminated in this resonant converter, conduction losses increase along with the device stresses. There is no scope of a SANs in resonant switching.

Objective type questions

Qs#1 Which component of power dissipation in a Power Semiconductor device is reduced by an RCD Switching –Aid –Network?

a) Off state losses
b) Turn-on losses
c) Turn-off losses
d) On-state losses

Ans: c) turn-off losses

Qs#2 Does an RCD SAN reduce total switching losses?

Ans: No. It transfers the losses from the device to itself.

Qs#3 Are resonant converters superior to the hard switched converter on all counts?

Ans: No. The resonant converter reduces switching losses at the cost of higher voltage/current stresses on the devices.