6

Load, Switch,
and Commutation Considerations

Power switching devices are employed for controlling inductive, resistive or capacitive loads. Inductive loads include electrical machines, transformers, solenoids, and relays. High-current in-rush occurs with loads such as incandescent lamps, pulse-forming networks, snubbers, and motors. Incandescent lamps are essentially resistive, but the cold resistive in-rush current during turn-on is 12 to 18 times the steady-state current. This turn-on surge presents special switch-on problems. Capacitive loads, such as fluorescent lighting, also present high-current in-rush at turn-on. Electromechanical loads, such as shakers, present loads that vary between capacitive and inductive over their operating frequency range.

The interaction of the load circuit on the switch arrangement and its commutation depends on three inter-related factors.
- The type of load, usually inductive, and rarely purely resistive.
- Switching mechanism classification, how the load effects switching commutation, namely hard switching, resonant, etc.
- The switch characteristics required to fulfil the supply and load I-V requirements, such as a bidirectional current switch, an asymmetrical sustaining voltage switch, etc.

Each of the three factors and their interdependence with the switching mechanisms are considered separately.

6.1   Load types

The two principal load types of general interest in power electronics are
- the resistive load and
- the inductive load.

Turn-on and turn-off voltage and current switching waveforms, hence losses in a switch, depend on the type of load. Purely capacitive loads are rare, and tend to be inductive at the switch transient frequency.

6.1.1   The resistive load
A purely resistive load is rarely encountered in power switching applications (other than at load resonance). Figure 6.1 shows a simple resistive load being switched by a common emitter-connected IGBT transistor, which could equally be another appropriate semiconductor switch, for example, a MOSFET. When the gate is driven by the voltage waveform shown in figure 6.2a, the resultant collector voltage and current waveforms are as shown in figures 6.2b and 6.2c. These figures show that at turn-on, as the collector current increases, the voltage across the resistive load increases proportionally, as the collector voltage v_{ce} decreases at the same rate. That is, at turn-on, $v_{ce}(t) = V_s - i_c(t)R_L$, while at turn-off the inverse process occurs. Figure 6.2d shows transistor instantaneous power loss during turn-on and turn-off, which in each case has a peak value of $\frac{1}{4}V_sI_m$ when the collector voltage and current reach half their respective maximum values. The energy loss W during switching is given by

$$W = \int v_{ce}(t)\, i_c(t)\, dt \qquad \text{(J)} \qquad\qquad (6.1)$$

where the integration is performed over the switching transition period.
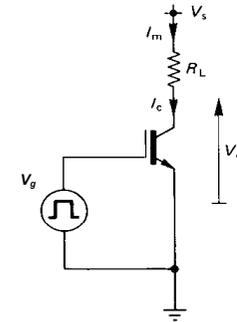
Figure 6.1. *A typical IGBT transistor switching circuit incorporating a resistive load.*

Figure 6.3 shows the safe operating area (SOA) characteristics for an IGBT, on logarithmic axes. Illustrated are the collector switch-on and switch-off trajectories, which are virtually coincident. In the off-state, point A on figure 6.2b, the transistor supports the supply rail voltage V_s while in the fully on-state, point C on figure 6.2b, the collector current I_m is V_s/R_L, neglecting the low on-state voltage of the transistor. During switching the collector voltage and current traverse the I-V switching trajectory between the steady-state operating conditions on $\rightarrow V_s/R_L$ and off $\rightarrow V_s$, as shown in figure 6.3.
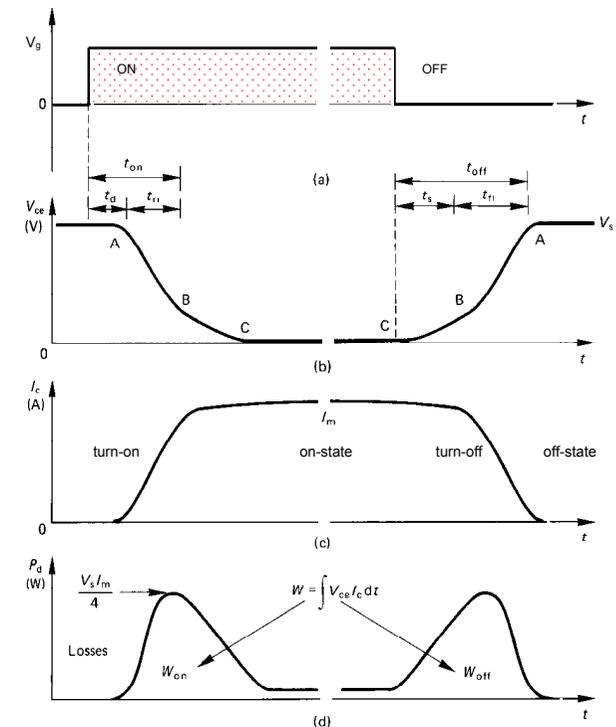

Figure 6.2. *Transistor switching waveforms for a resistive load: (a) on-off gate drive voltage; (b) collector-to-emitter voltage; (c) collector and load current waveform; and (d) instantaneous collector-emitter losses.*

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It is important that this trajectory does not exceed the shown SOA bounds set by the device voltage and current limits, and that the SOA region be traversed rapidly. For slow transitions, greater than a few microseconds, power dissipation considerations become the limiting design factor, which is a thermal limitation.

In order to perform the required thermal design calculations (for heatsink determination) it is necessary to be able to specify device-switching losses. To simplify analysis, the switching waveforms shown in figure 6.2 are linearised as shown in figure 6.4. As indicated on these waveforms, the collector voltage fall at turn-on is given by $v_{ce}(t) = V_s(1 - t/t_{on})$ while the collector current rise is $i_c(t) = I_m t/t_{on}$, where $I_m = V_s/R_L$. Combining $v_{ce}(t)$ and $i_c(t)$ by eliminating time t, gives

$$i_c = V_s(1 - v_{ce}/V_s)/R_L \qquad (6.2)$$

As shown in figure 6.3, this describes the linear turn-on transition of slope $-1/R_L$ from the on-state voltage with V_s/R_L collector current, shown as C, to the off-state at A where no current flows and the collector supports the supply V_s. Note figure 6.3 uses logarithmic axes, so the transition trajectory does not appear as a straight line (the inset figure is for linear axes).

Using equation (6.1), the switch-on loss for a resistive load is given by

$$W_{on}^r = \int_0^{t_{on}} V_s(1 - \frac{t}{t_{on}})I_m \frac{t}{t_{on}} dt$$

$$= \tfrac{1}{6}I_m V_s t_{on} \quad \text{or} \quad \tfrac{1}{6}\frac{V_s^2}{R_L}t_{on} \qquad \text{(J)} \qquad (6.3)$$

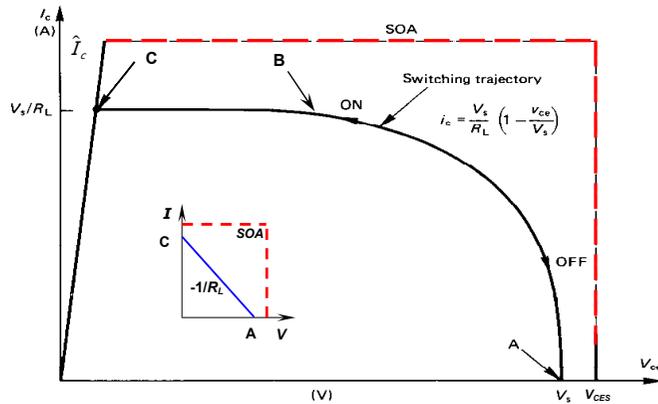where $I_m = V_s/R_L$ and t_{on} is the period of the switch-on interval, as shown in figure 6.4.

Figure 6.3. *Transistor I-V characteristics showing safe operating area and the switching trajectory with a resistive load, on logarithmic axes, and inset, on linear axes.*

Similarly, using the time dependant collector voltage and current equations shown on figure 6.4a, the turn-off switching loss is given by

$$W_{off}^r = \int_0^{t_{off}} V_s \frac{t}{t_{off}} I_m(1 - \frac{t}{t_{off}}) dt$$

$$= \tfrac{1}{6}I_m V_s t_{off} \quad \text{or} \quad \tfrac{1}{6}\frac{V_s^2}{R_L}t_{off} \qquad \text{(J)} \qquad (6.4)$$

where t_{off} is the turn-off period as shown in figure 6.4.

The average power loss due to switching, which is required for the thermal design outlined in chapter 5, is obtained by multiplying energy loss W by the switching frequency f_s. That is, the turn-on switching loss is given by

$$P_{on} = \tfrac{1}{6}I_m V_s t_{on} f_s \qquad \text{(W)} \qquad (6.5)$$

while the turn-off loss is given by

$$P_{off} = \tfrac{1}{6}I_m V_s t_{off} f_s \qquad \text{(W)} \qquad (6.6)$$

Because of IGBT current tailing and voltage overshoot at turn-off, the practical switching losses will be larger than those given by the linear approximating methods outlined.
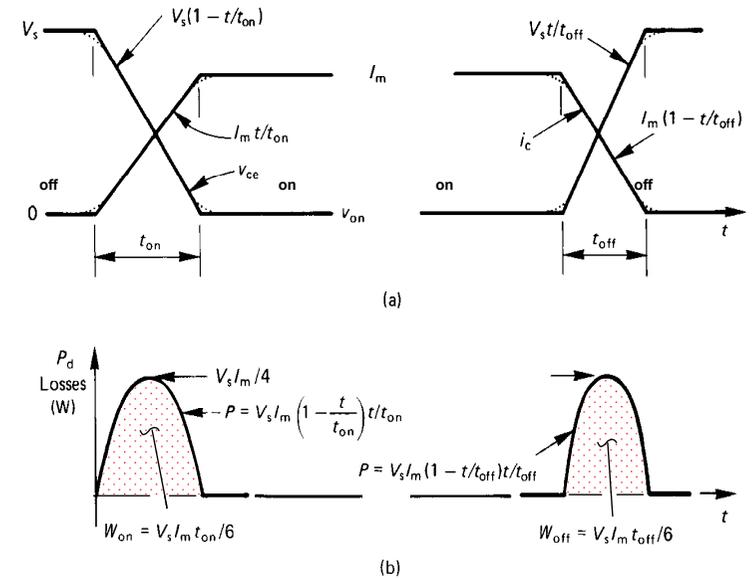
Figure 6.4. *Linear approximations of switching intervals for a purely resistive load:
(a) collector voltage and current linear waveforms and (b) corresponding energy and power losses.*

Example 6.1:   *Resistive load switching losses*

An IGBT switches a 10 ohms resistive load across a 100V dc supply. If the switch on-state duty cycle is 25%, ($\delta = \frac{1}{4}$), calculate the average load voltage and current. Calculate the switch losses if the switch-on time is $t_{on} = 1\mu s$, switch-off time is $t_{off} = 2\mu s$, and the on-state voltage is 2V.

Solution

When the switch is on, the current in the resistor is $I_L = V_s/R = 100V/10\Omega = 10A$.

The average load voltage is
$$V_o = \delta V_s$$
$$= 0.25 \times 100V = 25V$$
The average load current is
$$\bar{I}_o = V_o/R = 25V/10\Omega = 2.5A$$

The total switch losses P_T are made up of three components.

P_T = on-state loss + loss at switch-on + loss at switch-off

$$P_T = \delta \times v_{ce} \times I_L \ + \ \tfrac{1}{6}V_s I_L t_{on} f_s \ + \ \tfrac{1}{6}V_s I_L t_{off} f_s$$

$$= \tfrac{1}{4}\times 2V \times 10A \ + \ \tfrac{1}{6}\times 100V \times 10A \times 1\mu s \times 10kHz \ + \ \tfrac{1}{6}\times 100V \times 10A \times 2\mu s \times 10kHz$$

$$= \quad 5W \quad + \quad \tfrac{5}{3}W \quad + \quad \tfrac{10}{3}W$$

$$= \ 10W$$

Since the off-state leakage current and gate power losses are not specified, it is assumed these are insignificant. Technically the load current should be calculated based on 98V across the load since the switch supports 2V. Also the switching loss calculations should use a voltage of 98V, rather than 100V and a load current of 9.8A rather that 10A. The percentage error is small, and becomes increasingly insignificant at higher voltages.

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Example 6.2: *Transistor switching loss for non-linear electrical transitions*

Assume the transistor collector current at turn-off falls according to

$$i_c = \tfrac{1}{2}I_m(1 + \cos \pi t / T_0) \quad \text{for} \quad 0 \le t \le T_0 \tag{6.7}$$

For a resistive load, R_L

     i.   Calculate transistor loss at turn-off.
     ii.   Show that the switching trajectory across the SOA is as for the linear
           current fall case, as given by equation (6.2) and shown in figure 6.3.
     iii.   Calculate the peak power dissipation and the time when it occurs.

Solution

     i.   The collector voltage for a resistive load, on a dc supply V_s, is given by

$$v_{ce}(t) = V_s - i_c(t)R_L$$
$$= V_s - \tfrac{1}{2}I_m(1 + \cos \pi t / T_0)R_L$$

and since $V_s = I_m R_L$

$$v_{ce}(t) = \tfrac{1}{2}V_s(1 - \cos \pi t / T_0)$$

The turn-off energy loss is given by

$$W_{off} = \int_0^{T_0} p(t)\,dt = \int_0^{T_0} i_c(t)v_{ce}(t)\,dt$$
$$= \int_0^{T_0} \tfrac{1}{2}I_m(1 + \cos \pi t / T_0) \times \tfrac{1}{2}V_s(1 - \cos \pi t / T_0)\,dt$$
$$= \tfrac{1}{8}V_s I_m T_0$$

     ii.   Combining $v_{ce}(t)$ and $i_c(t)$ so as to eliminate the time variable, yields

$$i_c = \frac{V_s}{R_L}(1 - \frac{v_{ce}}{V_s})$$

which is the same straight line expression as in equation (6.2) and shown in figure 6.3, for the linear switching transition case.

     iii.   Instantaneous power dissipation is given by

$$P = v_{ce}i_c = v_{ce}\frac{V_s}{R_L}(1 - \frac{v_{ce}}{V_s})$$

Peak power $\hat{P}$ occurs when $dP/dv_{ce} = 0$, that is, when $v_{ce} = \tfrac{1}{2}V_s$, whence on substitution into the power expression P, yields

$$\hat{P} = \tfrac{1}{4}V_s^2 / R_L = \tfrac{1}{4}V_s I_m \quad \text{at} \quad t = \tfrac{1}{2}T_0$$
♣

Turn-on loss can be similarly analysed to yield virtually identical expressions, as required in problem 6.4.

6.1.2   The inductive load

The voltage spikes generated by inductive loads at turn-off may have high energy content, and the power generated may cause excessive device temperature, voltage stressing, and device failure.

At turn-off, the switch decreases the inductive load current from I_m to zero at a high *di/dt* and the resultant inductive voltage spike is given by

$$v(t) = L\frac{di}{dt} \quad\quad\quad \text{(V)}$$

where L is the load inductance. The spike energy to be absorbed by the switch is given by

$$W = \tfrac{1}{2}LI_m^2 \quad\quad\quad \text{(J)}$$

Both the voltage spike and its associated energy may be well outside the capabilities of the switching device. The peak voltage induced must be limited to a value below the breakdown rating of the device.

Four commonly employed voltage limiting techniques are shown in figure 6.5.
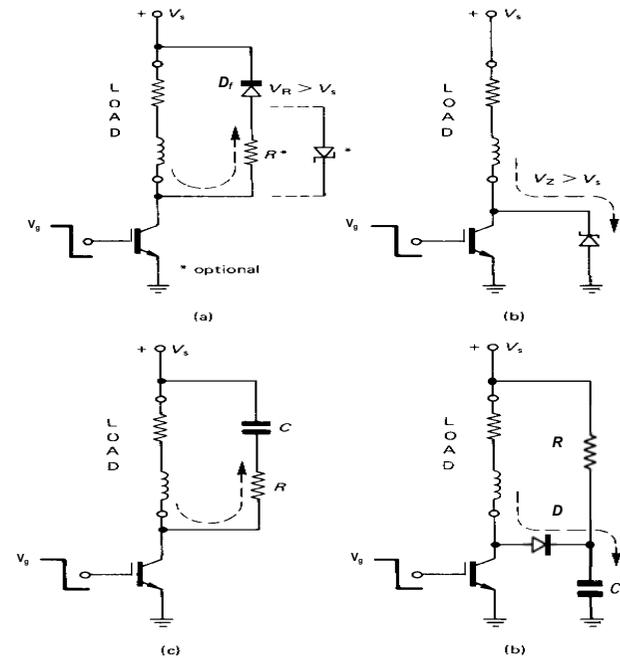

Figure 6.5. *Four methods of limiting inductive load turn-off voltage spike and of absorbing the associated energy: (a) freewheel clamping diode; (b) Zener diode clamp; (c) R-C snubber circuit; and (d) capacitor soft voltage clamp.*

The freewheel diode D_f in figure 6.5a is used to clamp the maximum device voltage to the supply rail voltage. The stored load energy is dissipated after turn-off as a result of the current that flows in the diode and load. The low impedance of the diode causes the current to decay slowly, since the inductor stored energy can only dissipate slowly in the freewheeling loop parasitic resistive components. A shorter current decay time can be achieved if series loop resistance R is added, as shown in figure 6.5a. Now the peak off-state voltage experienced by the switch is increased from V_s in the case of only the diode, to $V_s + I_m R$ because of the initial voltage drop across the optionally added resistor. This extra voltage drop, $I_m R$, decreases exponentially to zero. The resistor in figure 6.5a can be replaced by a Zener diode, thereby clamping the switch voltage at turn-off to $V_s + V_Z$. The load now freewheels at a fixed voltage V_Z thereby improving the rate of current decay, which is now constant. The inductive load current will fall linearly from I_m to zero in a time given by

$$t = LI_m / V_Z \quad\quad\quad \text{(s)}$$

An alternative Zener diode clamping circuit, as shown in figure 6.5b, can be employed in low power applications. The Zener breakdown voltage V_z is selected between the rail voltage V_s, and the switch breakdown voltage $(V_s < V_z < V_{BR})$. At turn-off, the Zener diode clamps the switch voltage to a safe level V_Z and absorbs the stored inductive load energy. The higher the clamping voltage level, the faster the energy is dissipated. The inductive load current decays linearly to zero in a time given by

$$t = LI_m / (V_z - V_s) \quad\quad\quad \text{(s)} \tag{6.8}$$

The two different Zener diode approaches perform the same switch clamping function in the same current decay time, if the voltage experienced by the switch is the same, but with different Zener diode losses. The desirable feature in the case of the Zener diode in parallel to the switch as in figure 6.5b, is that the protection component is directly across the element to be voltage protected. When placed in parallel with the load as in figure 6.5a, the switch is indirectly voltage protected, relying on the supply decoupling being a low inductance path. A reverse blocking diode D_f in figure 6.5a is mandatory.

The parallel-switch Zener diode approach in figure 6.5b has a number of disadvantages

- The Zener diode voltage rating must be in excess of the supply rail, V_s, while any Zener value can be used when the Zener diode is in parallel with the load.
- At higher voltages, >280V, Zener diodes will have to be series connected, thus the low inductance advantage of clamping with just one component is diminished.
- Assuming no resistance in the load, the energy dissipated with the two Zener diode approaches differs. When in parallel with the load, the load energy $\frac{1}{2}LI_m^2$ is dissipated while in the second case, load and supply energy are dissipated in the clamping Zener diode. The extra supply energy, in addition to $\frac{1}{2}LI_m^2$, dissipated in the Zener diode, is $\frac{1}{2}LI_m^2 V_s / (V_z - V_s)$. This is derived by recognising that, assuming a purely inductive load, the dc supply V_s delivers a current I_m which linearly falls to zero over the period given by equation (6.8).

The R-C snubbing circuit shown in figure 6.5c is commonly used in power conversion circuits to limit spikes caused by transformer leakage inductance, diode recovery, and interconnection wire inductance. The stored load energy is resonated to the snubber capacitor at switch turn-off. The reset resistor R (non-inductive) must overdamp the L-C-R oscillation by absorbing the transferred energy. The resistor also limits the snubber capacitor discharging current to a maximum of V_s /R at switch turn-on. For a purely inductive load, the snubber resistor power losses are given by the sum of the turn-off and turn-on losses, that is

$$P = \left(\frac{1}{2}LI_m^2 + \frac{1}{2}CV_s^2\right)f_s \qquad \text{(W)}$$

Figure 6.5d shows a capacitive voltage clamp used to soft clamp the switch voltage overshoot caused by the inductive energy stored in the load. The capacitor retains a charge of at least V_s. At switch turn-off, when the switch collector voltage reaches the capacitor (supply V_s) voltage level, the inductive stored load energy is transferred to the capacitor and concurrently, the capacitor discharges the energy in excess of V_s into the supply. When the capacitor is over charging, energy is taken from both the load inductance and the supply. When the capacitor discharges through the resistor back into the supply, the earlier energy taken from the supply is returned. The net effect is that only the energy $\frac{1}{2}LI_m^2$ is dissipated in the resistor. A reset resistor of low inductance is not necessary - a wirewound resistor can be used. This capacitive soft voltage clamp is analysed in detail in chapter 8.2.

Example 6.3:   *Zener diode, switch voltage clamping*

A reed relay coil of 1 mH inductance is switched at 20 kHz with a 20 per cent on-time duty cycle, across a 100 V dc rail. The energy stored in the coil at turn-off is dissipated in a 25 V Zener diode connected as shown in figure 6.5a.

- i.   Sketch the coil current and voltage, and the switch voltage waveforms.
- ii.   What is the average coil voltage?
- iii.   What Zener diode voltage is required for the circuit in figure 6.5b so as to produce the same coil current waveform as in figure 6.5a when using a 25 V Zener diode?
- iv.   For each circuit, calculate the power requirement of the Zener diode and the average power delivered from the 100 V supply.
- v.   Calculate the minimum resistance that replaces the Zener diode in figure 6.5a if the coil is to be switched on with almost zero current. Draw the coil current and switch voltage waveform, showing the switch peak voltage at turn-off.
- vi.   Discuss the relative features of each voltage clamping approach.

Solution

The three voltage clamping circuits being considered are shown in figure 6.6a.

- i.   With a 20kHz switching frequency, the coil current rises and falls every 50µs, with an on-state duty cycle representing 10µs for the current to increase in the coil and 40µs for the current reset decay to reach zero.
  From $V=Ldi/dt$, in steady-state, with zero coil resistance and zero initial current, the peak coil current is $I = V_s t /L$ = 100V×10µs/1mH = 1A. Thus the coil current rises linearly from zero to 1A in 10 µs. During reset, the coil current waveform depends on the reset circuit. For Zener diode (constant voltage) reset, the current falls linearly, while with a resistor the reset current decays with an L / R exponential time constant, as shown in figure 6.6b, for each case.

  The various circuit voltage and current waveforms are shown in figure 6.6b, where data derived from the rest of this example has been incorporated.
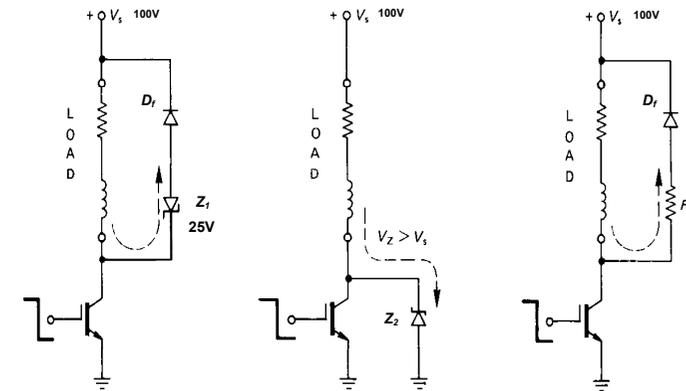
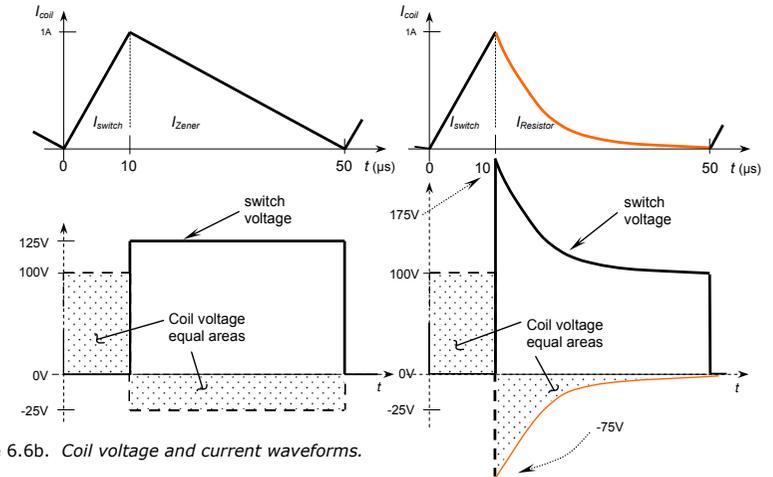Figure 6.6a. *Three inductive load clamping circuits.*

Figure 6.6b. *Coil voltage and current waveforms.*

- ii.   From $V=Ldi/dt$, for a steady-state continuous waveform, $\int V_L(t)dt = 0$, i.e. $1/T\int v(t)dt = V_{ave} = 0$, as shown on the coil voltage waveform (the coil voltage areas cancel).

- iii.   The parallel Zener diode requirement is $V_{Z2}= V_s+V_{Z1}$ = 100V+25V = 125V.

- iv.   **Zener diode V_{Z1}** in the parallel-load reset circuit:
  The energy $\frac{1}{2}LI^2$ is transferred from the coil to the Zener diode when the switch is turned off. The power dissipated in the Zener diode at 20kHz is therefore $\frac{1}{2}LI^2 f_s = \frac{1}{2}×1mH×1A^2×20kHz$ =10W. The total power drawn from the supply is the power stored by the coil at the end of the 10µs on-time, namely 10W.

  Zener diode V_{Z2} in the parallel-switch reset circuit:
  When the coil releases its stored energy (10W) into the Zener, current is also drawn from the supply. The total average power delivered by the supply over the 50µs period is given by $V_s I_{ave} = \frac{1}{2}×100V×1A = 50W$. This comprises $\frac{1}{2}LI^2$ (10W) from the supply into the coil when the switch is on for 10 µs, and the remainder (40W) into the Zener diode (plus the coil energy, 10W), when the switch is off for 40 µs. The Zener diode losses are 50W during the switch off period.

- v.   When a resistor is used in the reset circuit, the current decays exponentially from 1A to 0A. The resistance determines the peak switch voltage. The resistance does not affect the amount of energy dissipated, only the period over which the coil energy is released, dissipated as heat. Assume the coil current to be near zero after three L/R time constants, that is $3L/R$ = 40µs = t_{off}.

For L = 1mH, this gives R = 75Ω, with a power dissipation rating of 10W from part iv. At switch turn-off the collector voltage rises to (100V+1A×75Ω) 175V and then decays to 100V. Use an 82Ω (preferred value, exceeding 75Ω which reduces the time constant), 15W metal oxide resistor for low inductance.

vi. A Zener diode approach gives a fixed over-voltage on the switch, independent of current or stored energy. When clamping is in parallel with the switch, only one clamping element is needed, but its power requirement is significantly higher than when the clamp (Zener plus diode) is in parallel to the load. Any resistive element must have low inductance. This is restrictive given the power levels involved, and may result in only less effective wire wound elements being viable.

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By far the most common technique used to limit inductive switch-off voltage spikes in power circuits involves the use of a freewheel diode without R_{opt}, as shown in figure 6.5a and 6.7a. Typical switching waveforms for an inductive load clamped by a freewheel diode are shown in figure 6.7.

- **At turn-off**, the switching device conducts the full load current as the collector voltage rises to the supply rail. When the collector voltage reaches the supply rail level the freewheel diode becomes forward-biased and begins to conduct. Only then can the switch current fall to zero. The freewheel diode conducts the load current.

- **At switch turn-on**, assuming the diode is still freewheeling load current, the switch current increases, displacing freewheeling diode current, while the load is clamped to the rail voltage by the conducting freewheel diode. Only when the switch conducts the full load current can the freewheel diode recovered (and block), so that the switch voltage can fall to the low on-state level.
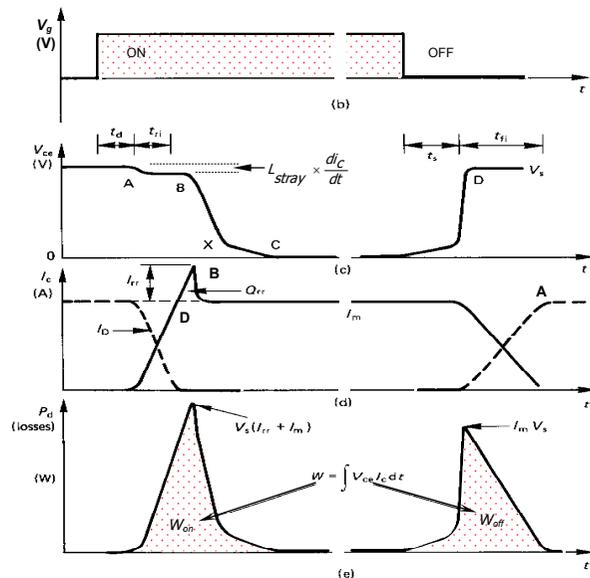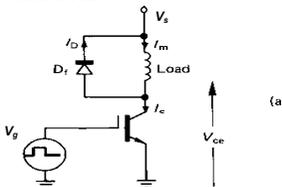

Figure 6.7. *Inductive load switching waveforms:*
(a) the circuit including the freewheel diode D_f; (b) on-off gate drive voltage; (c) collector-to-emitter voltage; (d) collector and freewheel diode current; and (e) switch instantaneous power losses.

It will be seen in figure 6.7 that during both turn-on and turn-off the switch must support instantaneously a maximum voltage, V_s, and full load current, I_m, condition. These severe electrical conditions are shown on the SOA characteristics in figure 6.8. In switching on from the operating point A to C, a maximum voltage and current condition (V_s, I_m) occurs at point D. Because of freewheel diode current reverse recovery effects at turn-on, an SOA trajectory point B is reached. At turn-off, due to stray inductance, voltage over shoot occurs and the point E is reached. By comparison with figure 6.2, it is seen that power losses during the switching intervals are higher for an inductive load than a resistive load.
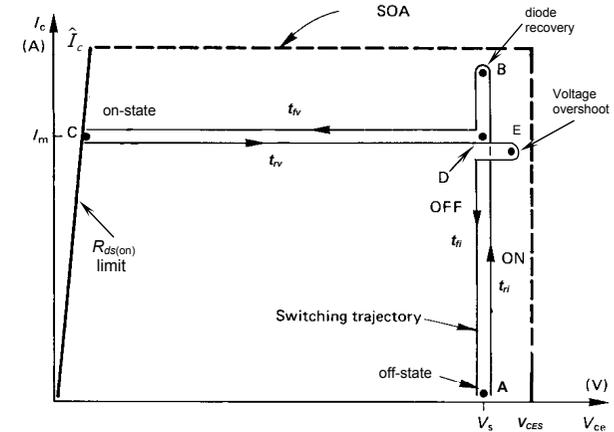

Figure 6.8. *I-V characteristics for an IGBT*
showing its safe operating area and switching trajectory for an inductive load (linear axes).

Switching losses can be calculated by using linear approximations to the switching transitions. It can be assumed that a silicon carbide Schottky freewheel diode is employed so as to allow reverse recovery effects to be neglected. Figure 6.9 shows the linearised switching waveforms for an inductive load, where maximum voltage V_s and current I_m occur simultaneously during both turn-on and turn-off. The equations for the collector voltage and current at turn-on and turn-off are also shown in figure 6.9.

The turn-on switching interval loss is given by the time integral over the current rise period plus the voltage fall period,

$$W_{on} = \int_0^{t_{ri}} V_s \times I_m \frac{t}{t_{ri}}\, dt \; + \; \int_0^{t_{fv}} V_s \left(1 - \frac{t}{t_{fv}}\right) \times I_m\, dt$$

$$= \tfrac{1}{2} V_s I_m t_{on} \qquad (J) \tag{6.9}$$

where $t_{on} = t_{ri} + t_{fv}$, as shown in figure 6.9. The current rise time at turn-on is termed t_{ri}, while the switch voltage fall time at turn-on is termed t_{fv}.

Similarly, from figure 6.9c, the turn-off loss is given by

$$W_{off} = \int_0^{t_{rv}} V_s \frac{t}{t_{ri}} \times I_m\, dt \; + \; \int_0^{t_{fi}} V_s \times I_m\left(1 - \frac{t}{t_{fv}}\right) dt$$

$$= \tfrac{1}{2} V_s I_m t_{off} \qquad (J) \tag{6.10}$$

where $t_{off} = t_{rv} + t_{fi}$, as shown in figure 6.9c. The switch voltage rise time at turn-off is termed t_{rv}, while the switch current fall time is termed t_{fi}.

Comparison of switching losses for a resistive load, equations (6.3) and (6.4), and an inductive load, equations (6.9) and (6.10), shows that inductive switching losses are three times those for the resistive load case. The peak power experienced by the switch during switching of an inductive load, $V_s I_m$, is four times greater than that experienced with a resistive load, $\tfrac{1}{4} V_s I_m$. As for the resistive load switching circuit, actual switch losses with an inductive load are also higher than those predicted by equations (6.9) and (6.10). The effects of current tailing, voltage over-shoot, and freewheel diode reverse recovery can together produce losses of the same order as those predicted for theoretical switching by equations (6.3), (6.4), (6.9), and (6.10).
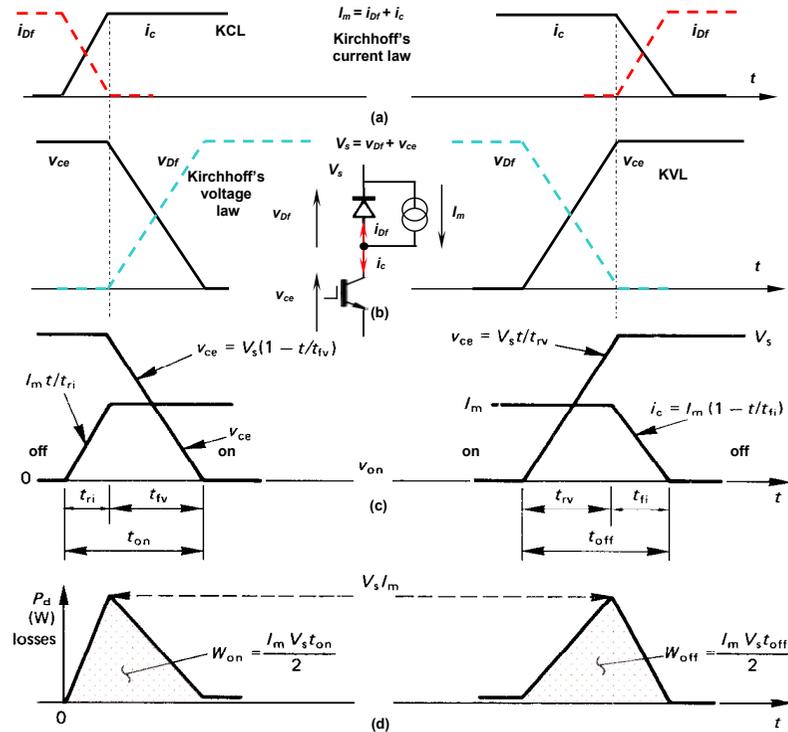
Figure 6.9. *Linear approximations of transistor switching intervals for an inductive load:*
(a) Kirchhoff's current law $I_m = i_{Df} + i_c$; (b) Kirchhoff's voltage law $V_s = v_{Df} + v_{ce}$; (c) collector voltage
and current waveforms with switching parameters defined; and (d) corresponding switching losses.

Example 6.4:   *Inductive load switching losses*

A power n-channel MOSFET switches a 10A, 100V dc, highly inductive load at 10kHz.
Calculate the worst case switch losses if the switch turn-on time is t_{on} = 1μs, switch turn-off time is t_{off} = 2μs, and the MOSFET channel on-state resistance is 0.2Ω at 10A.
Calculate the maximum instantaneous power dissipation in the switch, and determine when it occurs.

Solution
Maximum switch losses occur when the duty cycle approaches one ($\delta \rightarrow 1$) such the both turn-on and turn-off still occur (that is conduction is not continuous).
The total switch losses P_T are made up of three components

P_T = on-state loss   +   loss at switch-on   +   loss at switch-off

$P_T = \delta \times I_L^2 \times R_{ds(on)} +$ $\frac{1}{2}V_s I_L t_{on} f_s$ + $\frac{1}{2}V_s I_L t_{off} f_s$

= $1 \times 10^2 \times 0.2\Omega$ + ½×100V×10A×1μs × 10kHz + ½×100V×10A×2μs × 10kHz

=   20W   +   5W   +   10W

=   25W

Since the off-state leakage current and gate power losses are not specified, it is assumed these are insignificant. The switching loss calculations should use a voltage of 98V, rather than 100V, since (10A×0.2Ω) 2V is dropped across the channel resistance of the MOSFET. The percentage error is small, and becomes insignificant at higher voltages.

Maximum switch loss occurs when during the switching transitions, the drain current is 10A and the drain voltage is 100V. The maximum instantaneous loss is 10A×100V=1000W, ($I_L \times V_s$).

♣

6.1.3   Diode reverse recovery with an inductive load

When a silicon bipolar diode conducts the pn junction scl region accumulates charges. When the diode turns off and the current falls to zero, the junction retains charge that must recovery before diode reverse voltage can be supported. Negative diode current flows. This phenomenon was considered in chapter 4.2.2 and is shown in figure 6.10a. The maximum collector current at turn-on is increased above the load current level I_m by the reverse recovery current I_{rr}, to $I_m + I_{rr}$. The diode begins to support reverse voltage once the peak reverse recovery current is reached. As a consequence the switch turn-on losses are increased as shown in figure 6.10c.

The circuit current at peak recovery has a discontinuous derivative, and as a consequence, high circuit voltages are induced across circuit stray inductance due to $v = L di/dt$. High-frequency voltage ringing occurs as the stored energy in the stray inductance is dissipated and reverse voltages far in excess of V_s are experienced by the recovering diode.
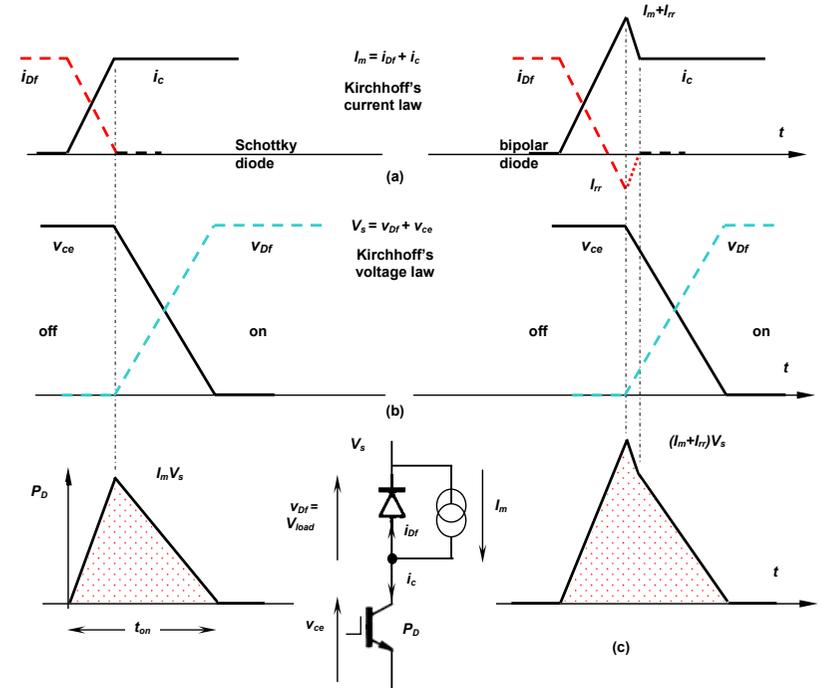

Figure 6.10. *Linear approximations of transistor switching turn-on interval for an inductive load showing freewheel diode reverse recovery effects on the right: (a) Kirchhoff's current law $I_m = i_{Df} + i_c$; (b) Kirchhoff's voltage law $V_s = v_{Df} + v_{ce}$; and (c) corresponding switching losses.*

Example 6.5:   *Inductive load switching losses with device models*

A MOSFET 340V dc chopper feeds an inductive dc motor load at 50kHz. In steady state the load current rises from 10A to 25A when the switch is on with a 75% on-state duty cycle ($\delta = \frac{3}{4}$).
The MOSFET switch turn-on time is t_{on} = 100ns, switch turn-off time is t_{off} = 200ns, and the channel on-state resistance is $R_{ds\,on}$ = 0.025Ω.
The freewheel diode is modelled by a 1V on-state voltage and on-state resistance of 0.05 Ω.
Neglecting diode recovery and diode turn-on loss, for this continuous load current case, calculate

i.   the MOSFET total losses
ii.   diode losses
iii.   power delivered to the motor load, if the armature resistance is 1 Ω and back emf is 170V
iv.   electromagnetic energy conversion efficiency and total circuit efficiency

Solution

i. The MOSFET losses comprise turn-on, turn-off, and conduction losses.
The rms current in the MOSFET is given by

$$I_{M-rms} = \sqrt{\frac{\delta}{3}\left(\hat{I}^2 + \hat{I} \times \check{I} + \check{I}^2\right)}$$

$$= \sqrt{\frac{0.75}{3} \times \left(25A^2 + 25A \times 10A + 10A^2\right)} = 15.6A$$

The MOSFET conduction losses are therefore

$$P_c = I_{rms}^2 R_{ds\ on} = 324.75 \times 0.025\Omega = 8.1W$$

The switching losses are

at turn-on $P_{t-on} = \frac{1}{2}V_s \check{I} t_{on} f_s = \frac{1}{2} \times 340V \times 10A \times 100ns \times 50kHz = 8.5W$

at turn-off $P_{t-off} = \frac{1}{2}V_s \hat{I} t_{off} f_s = \frac{1}{2} \times 340V \times 25A \times 200ns \times 50kHz = 42.5W$

Total MOSFET losses are $P_{MOSFET} = P_c + P_{t-on} + P_{t-off} = 8.1W + 8.5W + 42.5W = 59.1W$

ii. The diode RMS current is

$$I_{D-rms} = \sqrt{\frac{1-\delta}{3}\left(\hat{I}^2 + \hat{I} \times \check{I} + \check{I}^2\right)}$$

$$= \sqrt{\frac{0.25}{3} \times \left(25A^2 + 25A \times 10A + 10A^2\right)} = 9A$$

The average diode current is

$$\overline{I}_d = \frac{1}{2}(1-\delta)\left(\hat{I} + \check{I}\right) = \frac{1}{2} \times (1 - \frac{3}{4}) \times (25A + 10A) = 4.375A$$

The total diode losses are

$$P_{diode} = I_{D-rms}^2 R_{D-on} + \overline{I} \times V_{D-on}$$
$$= 81.25 \times 0.05\Omega + 4.375A \times 1V = 8.4W$$

iii. The power delivered to the load comprises losses in the 1Ω armature resistance and the power delivered into the 170V dc back emf.
The rms load current is given by

$$I_{M-rms} = \sqrt{\frac{1}{3}\left(\hat{I}^2 + \hat{I} \times \check{I} + \check{I}^2\right)}$$

$$= \sqrt{\frac{1}{3} \times \left(25A^2 + 25A \times 10A + 10A^2\right)} = 18A$$

The load resistor loss is $P_{Ra} = I_{a-rms}^2 R_a = 325 \times 1\Omega = 325W$

The average load current is

$$\overline{I}_a = \frac{1}{2}\left(\hat{I} + \check{I}\right) = \frac{1}{2} \times (25A + 10A) = 17.5A$$

The power delivered to the back emf is $P_{E-a} = \overline{I}_a E_a = 17.5A \times 170V = 2975W$

The total power delivered to the dc motor is $P_{motor} = P_{Ra} + P_{E-a} = 325W + 2975W = 3300W$

iv. The dc motor efficiency is

$$\eta_{dc} = \frac{\text{power output}}{\text{power input}} = \frac{2975W}{3300W} \times 100 = 90.2\%$$

Including switch and diode losses yields total circuit efficiency, that is

$$\eta_{circuit} = \frac{\text{power output}}{\text{dc supply power input}} = \frac{\text{power output}}{\text{chopper circuit losses + dc motor power input}}$$

$$= \frac{2975W}{(59.1W + 8.4W) + 3300W} \times 100 = 88.3\%$$

♣

6.2 Switch characteristics

Having considered the switching of inductive and resistive loads, the following are the electrical and thermal characteristics desirable of commutable switching devices (as well as low cost):

***off-state** (ideally open circuit):*
- Low, temperature independent leakage current in the off-state, to minimise off-state power loss, and to simplify resistive networks for device series connection.
- High forward and reverse voltage blocking ratings to reduce the need for series device connection, which would otherwise complication control and protection circuitry requirements. Series connection increases the on-state voltage, hence on-state loss. When a diode is used in antiparallel across the switch to allow reverse principal current flow, the switch does not require a significant reverse voltage blocking rating.
- High static off-state avalanche capability to absorb transient overvoltage stresses.
- High static and re-applied *dv/dt* capability to withstand high applied off-state voltages without avalanche or false turn-on, with minimal displacement current.

***on-state** (ideally short circuit):*
- Low on-state conducting voltage or low on-state resistance, in order to minimise on-state conduction power loss: with a slight positive temperature co-efficient at high current densities, to allow reliable parallel device connection.
- High on-state current density capability so as to avoid need for and problems associated with parallel device current sharing and differential thermal coefficients.
- Safe controlled switch off from a short circuit current condition.

***switching** (ideally instantaneous):*
- Low control power to produce switching between states, with no 'Miller' interaction.
- Short, temperature independent, turn-on and turn-off times to result in low switching losses which will allow high frequency switching.
- High initial *di/dt* capability at turn-on to allow rapid low loss build-up of the turn-on principal current.
- High surge current capability to withstand transient over current fault conditions, resulting in better fault tolerance and nuisance tripping ride through.
- Large switching safe operating area, being able to simultaneously, but instantaneously, support rated voltage and rated current, without the need for switch snubber circuits.

thermal/mechanical:
- Easy to electrically connect and mechanically mount, with low thermal resistance and impedance for efficient heat removal.
- Mechanically, electrically, and thermally robust, with the ability to operate at high (and low) junction temperatures in high (and low) ambient, pressure, humidity conditions.
- Matching substrate structure and thermal properties to minimise stressing due to thermal, mechanical, and power stressing.
- Infinite storage and operational lifetime.

6.3 Switching classification

There are four principal *I-V* switching conditions during the commutation (turn-on or turn-off) of a switch, viz.:

- hard switching;
- soft switching;
- resonant switching; and
- naturally-commutated switching.

These four possibilities are classified in terms of the *switching time* t_s and the *commutation time* t_q, where $t_q \leq t_s$. Figure 6.11 shows the four electrical cases and specifies the switching and commutation times for each.

- *Switching time* t_s is the time for a switch to change from fully on ($v=0$, $i=I_L$) to fully off ($v=V_s$, $i=0$), such that no further change occurs in the switch voltage and current due to the change of state.
- *Commutation time* t_q is associated with the external circuitry and is defined as the time the switch takes to reach zero current at turn-off or to reach zero volts at turn-on. Alternatively, commutation time is the period of switch power loss at turn-on or turn-off, due to the switch changing states.
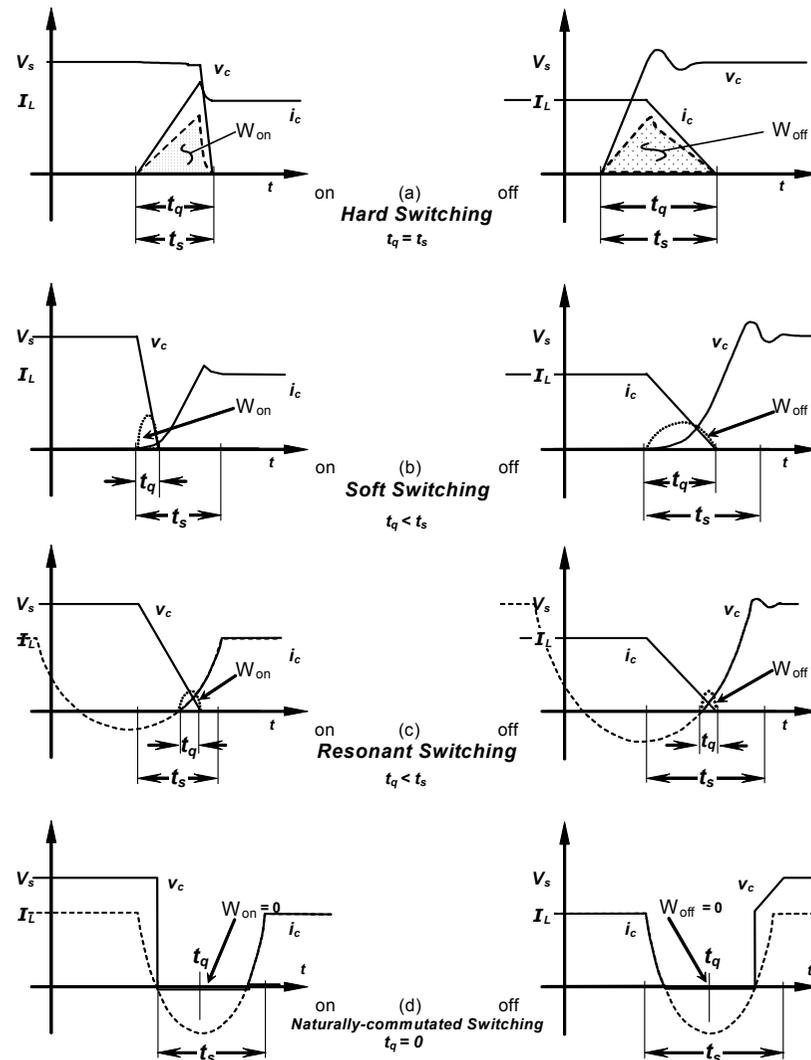
Figure 6.11. *Switch voltage (v_c), current (i_c), and power loss (W_{on} and W_{off}) of four switching classifications: (a) hard switching; (b) soft switching; (c) resonant switching; and (d) naturally-commutated switching.*

Generally, the switch loss magnitude (stress) for a given set of electrical and thermal operating conditions, decreases when progressing from severe hard switching through to virtually lossless naturally-commutated switching.

6.3.1 Hard switching: $t_q = t_s$

The turn-on and turn-off switching waveforms in figure 6.11a for an inductive load show that hard switching is characterised by $t_q = t_s$. The resistive and inductive switching considered in sections 6.1.1 and 6.1.2 are examples of hard switching. In figure 6.4 for a resistive load, the switching periods t_{on} and t_{off} (t_s) correspond to the period of switch losses (t_q) during each state transition. In figure 6.9 for the inductive load, the t_q periods correspond to the power loss periods at switching ($t_{rv} + t_{fi}$ and $t_{fv} + t_{ri}$).

6.3.2 Soft switching: $t_q < t_s$

Figure 6.11b shows typical soft-switching waveforms for turn-on and turn-off. The switching losses are complete before the switch has reached its final steady-state condition. That is, $t_s > t_q$ such that the periods t_s and t_q both commence at the same time.

At turn-on, the switch voltage reaches zero before the switch current reaches the steady-state full-load value I_L. Once the switch voltage reaches zero, the rising current no longer results in a power loss. This *I-V* characteristic at turn-on (usually involving inductance in series with the switch) is a form of quasi *zero current switching*, ZCS.

The inverse occurs at turn-off. The switch current reaches zero before the switch voltage has settled at the supply voltage level V_s. (Usually involving capacitance in parallel with the switch.) This is a form of quasi *zero voltage switching*, ZVS.

Soft-switching results when auxiliary stress diverting circuits, called snubber circuits, are used, as will be considered in chapters eight and nine.

6.3.3 Resonant switching: $t_q \ll t_s$

Resonant-switching waveforms at turn-on and turn-off are shown in figure 6.11c, with switching periods t_s shown. Resonant switching occurs if the switching period is associated with either the switch voltage or current being zero, due to external load circuit conditions which have diverted the load current or voltage. That is $t_s > t_q$.

Switching of the voltage when the switch current is zero, usually at turn-on, is called *zero current resonant switching*, ZCRS, while commutating the current while the switch voltage is zero, usually at turn-off, is called *zero voltage resonant switching,* ZVRS. Because the exact instant of zero may vary, being load circuit dependant, some control restriction is inevitable. Zero voltage or current switching can be readily attained with ac mains converter circuits since switching can be synchronised with supply zero voltage crossing, or zero current when the load current reverses due to the supply voltage reversal.

### 6.3.4 Naturally-commutated switching: $W = 0$, $t_q = 0$

Figure 6.11d shows switching when the voltage and current are both zero, called *naturally-commutated switching*. This was a commonly used technique for force turn-off of thyristors before the emergence of the GTO thyristor. Current from an auxiliary commutation circuit displaces (exceeds) the device principal current and reverse biases the device, at turn-off. The method was not used at turn-on. Commutated turn-on and turn-off occurs in inverter circuits where the switch has an anti-parallel connected diode. When the diode conducts and the switch is on but not conducting, if the load power factor causes the current to reverse, then the main switch automatically starts conducting with the switch voltage at zero because the diode was previously conducting, clamping the switch voltage slightly negative.

Naturally-commutated switching occurs for ac mains zero crossing switching, with a purely resistive load such that the load V and I are in phase. Switching losses are virtually zero.

6.4 Switch configurations

Most semiconductor switches are unipolar, that is, allow current and/or voltage to be supported in one direction. The MOSFET allows uncontrolled reverse current flow; hence can not support reverse voltage because of its parasitic body diode. Some structures, like the RCT considered in chapter 3.3.3, integrate an anti-parallel diode with a thyristor. Generally, such integrated approaches sacrifice some electrical characteristics. Many applications require a bi-directional current and/or bi-directional supporting voltage switches, so the basic switches can be configured as shown in figure 6.12, to give the necessary *I-V* characteristics. The net effect of the bi-directional voltage arrangements is good dynamic electrical characteristics but poor static characteristics. Specifically, the switching performance is as for the principal switch but the on-state loss is that of two series connected devices. In the case of the bidirectional blocking thyristor, the on-state voltage is increased slightly because an n-buffer can not be used in its fabrication. But the symmetrical GCT offers lower on-state voltage and significantly higher current ratings than an equivalent a high voltage IGBT with a series connected diode. The bi-directional conducting thyristor discussed in chapter 3.3.4 attempts to minimise the sacrificed on-state voltage limitation. A reverse blocking IGBT can also be realised. Die edge passivation of the diode region by a through the die p+ diffusion, plus guard rings, increase processing complexity, and hamper voltage ratings. A punch through IGBT version with reverse voltage blocking properties, is therefore problematic. On-state voltages are increased for a given switching speed and, as with the MOSFET body diode, the non-optimal diode recovery characteristics are a compromise because of the overriding n-substrate low resistivity requirements. See chapter 3.2.4.

- Controllable switching devices with reverse blocking capability are usually required for ac to ac converters, half-wave resonant converters, and current fed inverters.
- Voltage source inverters, full-wave resonant converters, and dc to dc converters usually do not require switching devices with reverse blocking properties, but may use an antiparallel connected diode.
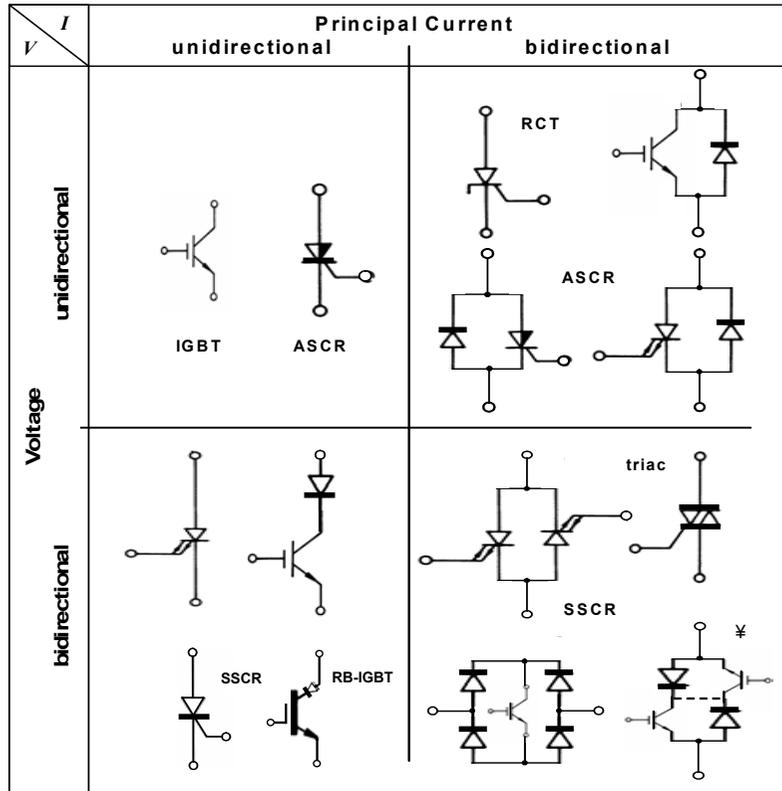
¥ Can be arranged so that emitters are at the same potential. Switches may be reverse blocking IGBTs.

Figure 6.12. *Switch configurations for uni-directional and bi-directional I-V characteristics.*

6.5   **Power converter configuration classification**

A power electronic converter processes the available energy form to another form having a different frequency and/or voltage or current magnitude. There are four basic types of electrical energy converters depending upon the conversion function performed:

Table 6.1   **Four basic conversion processes**

Conversion from-to	Topology	Function	Symbol
dc to dc	chopper smps	constant to variable dc or variable to constant dc	
dc to ac	inverter	dc source to ac of desired voltage and frequency	
ac to dc	rectifier converter	ac to unipolar dc current	
ac to ac	cycloconverter matrix converter	ac of desired frequency and/or magnitude from an ac source	

Reading list

Peter, J. M., *The Power Transistor in its Environment,*
Thomson-CSF, Sescosem, 1978.

Problems

6.1.   During turn-on and turn-off of a power transistor the current-voltage relationships are as shown in figure 6.13. Calculate the energy loss during both turn-on and turn-off periods and the mean power loss if the transistor is being switched at a frequency of 10 kHz. What is the maximum instantaneous power dissipated?
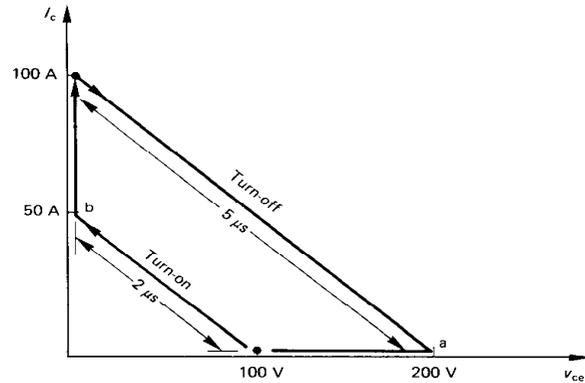   [1.66 mJ, 16.6 mJ, 183 W, 5kW]

Figure 6.13.

6.2.   The equivalent circuit in figure 2.4a involving parameters E_o and R_o can be extended to model a thyristor by replacing the ideal diode by an ideal thyristor. Derive general expressions for the thyristor mean power loss P_d and rms current i_o with a constant load current I_o and switch on-time duty cycle δ.
          If E_o = 1 V and R_o = 0.01 Ohms, for I_o = 50 A and a 25 per cent on-time duty cycle, calculate the thyristor:
    i.   On-state voltage, V_F
    ii.   Mean power, P_d
    iii.   rms current, i_o.
          [See example 2.1: 1.5 V, 18.75 W, 25 A]

6.3.   If the collector voltage at turn-on falls according to
    $v_c = \tfrac{1}{2}V_s(1 + \cos \pi t / T_o)$ for $0 \leq t \leq T_o$
    i.   For a resistive load, R_L, calculate transistor loss at turn-off.
    ii.   Show that the switching trajectory across the SOA is as for the linear current fall case.
    iii.   Calculate the peak power dissipation and when it occurs.

6.4.   A transistor is used to switch an inductive load with a current of I_m.
          At transistor turn-off, the collector voltage rises to the supply rail V_s according to
          $v_{ce} = \tfrac{1}{2}V_s (1 - \cos \pi t / T_{ov})$ for $t \leq T_{ov}$,
          then the collector current falls according to
          $i_c = \tfrac{1}{2}I_m(1 + \cos \pi t / T_{oi})$ for $t \leq T_{oi}$.
          Using the same integration form as in equation (6.10), show that the turn-off loss is
          $P = \tfrac{1}{2}V_sI_m T_o$ where $T_o = T_{ov} + T_{oi}$.

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